Radar Technology Encyclopedia

(Electronic Edition)

David K. Barton

Sergey A. Leonov

Editors

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INTRODUCTION

The *Radar Technology Encyclopedia* is a joint product of leading United States and Russian radar experts with decades of experience on design, development, and test of state-of-the-art radar systems and technology. The *Encyclopedia* covers the entire field of radar fundamentals, design, engineering, systems, subsystems, and major components. It contains about 5000 entries, each giving the depicted term definition, and, if applicable, the standard notation, brief description, evaluation formulas, relevant block diagrams, performance summary, and a reference to the literature in which the more detailed information is available. The purpose is to provide, in a single volume, the reference material for researchers and engineers in radar and related disciplines, representing the most modern information available in both the former Soviet Union and in the West. It includes an extensive bibliography of sources from both regions. This bibliography covers practically all monographs and textbooks in radar and related subjects published after World War II in English (in the U.S.A. and England) and Russian (in the former Soviet Union) languages that covers the overwhelming majority of the worldwide library of radar books.

The *Encyclopedia* format is alphabetical by subject. It consists of top-level articles, which are identified with bold capital letters (e.g., **MAGNETRON**), and, if applicable, are followed by subarticles, which are identified in lowercase bold (e.g., **rising-sun magnetron**). The top-level articles are arranged in the way so the key word (typically, a noun) determines its alphabetical position (e.g., *microwave antenna* is cited as **ANTENNA, microwave**, *radar targets* as **TARGET, radar**, *data smoothing* as **SMOOTHING, data**). Subarticles within a top-level article are given in a conventional word order typically used in literature and alphabetically arranged, for example:

AMPLIFIER, microwave
amplifier-attenuator
amplifier chain
aperiodic amplifier
backward-wave tube amplifier
balanced amplifier
bandpass amplifier

and so forth.

The subarticles are alphabetized without regard to whether the qualifying adjective precedes or follows the main word: **broadband antenna** precedes **antenna control**.

Within each article and subarticle, if applicable, the cross-reference to another subarticle is indicated in lowercase bold, e.g.:

> "The RCS of this type of clutter is calculated using the volume of the clutter cell V_c and the volume reflectivity η_v (see **volume clutter**). "

That subarticle is found alphabetically within the same top-level article, e.g., **CLUTTER**. If the cross-reference refers to another top-level article, then the name of this article is given in capital letters. For example, a reader is referred to an article **NOISE**, and will find that article alphabetically under **N**.

Parentheses in the name of an article or subarticle mean that the word is optional. For example, **phased array (antenna)** means that the term is used both as **phased array** or **phased array antenna**. Square brackets mean that the word in the brackets can be used instead of the previous one. For example, **bed of spikes [nails] ambiguity function** means that the term is used as **bed of spikes ambiguity function** or **bed of nails ambiguity function**.

For definitions of terms, extensive use has been made of *IEEE Standard Dictionary of Electrical and Electronics Terms* and *IEEE Standard Radar Definitions*. The standard definitions reproduced from these dictionaries and other acknowledged sources are put into quotes. The *Encyclopedia* does not contain separate articles with the description and performance of concrete radar stations and facilities, because even brief description of the major radars developed throughout the world requires to provide additional volume as thick as this one. This information is systematized in *Jane's Radar and Electronic Warfare Systems*, updated and issued annually, and the *Encyclopedia* does not duplicate this material. However, where applicable, extensive examples of modern radars are provided.

Each article and subarticle contains references, primarily to textbooks, which are listed alphabetically by author in the *Alphabetical Bibliography* at the end of *Encyclopedia*. The combination of the surname of the first author and a year of edition identifies the cited book:

Ref.: Skolnik (1980)

refers to the book listed in the bibliography as:

Skolnik, M. I., *Introduction to Radar Systems***,** McGraw-Hill, 1980;

and the brief reference:

Ref.: Barton (1969)

identifies the book listed with both authors and two editions or publishers:

Barton, D. K., and Ward, H. R., *Handbook of Radar Measurement*, Prentice-Hall, 1969; Artech House, 1984.

In rare cases where there is no applicable textbook, reference is made to a professional journal article. Typically, each article is followed by references to the major current books, as listed in the *Alphabetical Bibliography,* and for the readers interested in a full bibliography on a corresponding subject the *Bibliography by Subject* is provided. It contains a full bibliography list of the identifiable radar and radar-related books published during the last 50 years and is arranged in 35 sections by subject. Within each section the books are given in chronological order, and alphabetically by author within one year. At the end of *Encyclopedia* is a list of the most common radar abbreviations and acronyms.

The author of each article and subarticle is identified by the corresponding initials following the entry, when that entry exceeds a few lines of definition (see *About the Authors*). The original generation of the list of entries, compiling of the *Bibliography,* and final editing of *Encyclopedia* material was done by David K. Barton and Sergey A. Leonov.

David K. Barton and Sergey A. Leonov,
Editors

Use of Hypertext Links

In this electronic edition of the *Radar Technology Encyclopedia,* hypertext links have been added to transfer rapidly from one article to a related or referenced subject. The words or phrases from which links can be exercised appear in blue text. Clicking on any blue entry initiates an immediate transfer to the related entry. The program keeps track of the history of these transfers, and the reader can retrace steps by clicking in either the right or left page margins.

A

ABSORBER, radar. The term *absorber* refers to a radar-absorbing structure or material (RAS or RAM), the purpose of which is to soak up incident energy and reduce the energy reflected back to the radar. Its main objective is to achieve reduction in the radar cross section (RCS) of radar targets. Other applications are to suppress wall reflections in anechoic chambers and reflections from nearby structures at fixed radar sites.

Absorbers can be classified from the point of view of scattering phenomena as specular and nonspecular types, and from the point of view of their bandwidth as narrowband RAS and wideband RAS. The major representatives of narrowband RAS are the Salisbury screen and the Dahlenbach absorber. Wideband RAS are represented by $\mu_r = \varepsilon_r$ type absorbers, circuit analog absorbers, frequency-selective surfaces, geometric transition absorbers, Jaumann absorbers, and graded absorbers. Some of these types can be combined to form hybrid absorbers with improved performance. All these types are specular absorbers designed to reduce specular reflections from metallic surfaces. Nonspecular absorbers are intended primarily for suppression of surface traveling-wave echoes. *SAL*

Absorbers for anechoic chambers are applied to the internal surfaces of an anechoic chamber to absorb the incident radio waves. The basic requirements are wideband performance and low reflection coefficient.

Usually the absorber is a plastic foam frame with filler that readily absorbs radio waves (microspheres of polystyrene, teflon, etc.), the density of the material and the concentration growing with depth. Radar-absorbing material is most convenient in the form of pyramids with an angle of 30° to 60° at the apex, which assures multiple re-reflections that increase absorption. To reduce the reflection coefficient to –20 dB, the height of the pyramids must be 0.5λ to 0.6λ, but to reduce it to –50 dB, a height of 7λ to 10λ is required. In this case thinner structures are used, made, for example, from ferrite absorbing materials. *IAM*

Ref.: Finkel'shteyn (1983), p. 145; Knott, 1993, pp. 528–532.

Chirosorb absorbing material is a novel RAM typically fabricated by embedding randomly oriented identical chiral microstructures (e.g., microhelices), in an isotropic host medium. In comparison with conventional RAMs, it possesses an excellent low-reflectivity property and may be practically invisible to radar. *SAL*

Ref.: Bhattacharyya (1991), p. 233.

Circuit analog (CA) absorbers are sheets of low-loss material on which specific conducting patterns have been deposited. The patterns constitute resistance, inductance, and capacitance. The deposited film can be represented by an equivalent RLC circuit, parameters of which can be controlled by the geometric configuration, film thickness, and conductivity of the deposition on the film. An example of a pattern deposited on a CA sheet in shown in Fig. A1.

CA absorbers can be tuned, as with an RLC circuit, enabling the designer to improve the bandwidth of the multi-sheet configuration. In general, CA absorber is a lossy version of a class of printed patterns known as frequency-selective surfaces (FSS). *SAL*

Ref.: Knott (1993) p. 326; Bhattacharyya (1991), pp. 215–217.

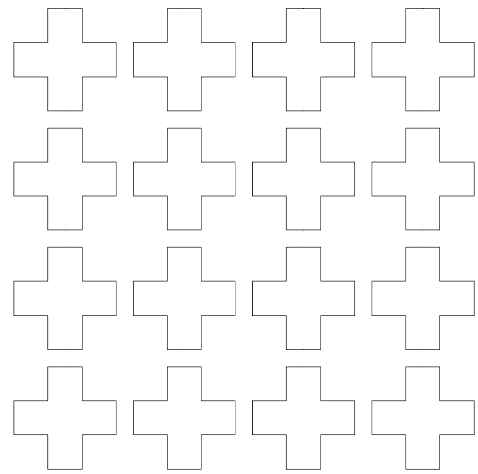


Figure A1 Circuit analog absorbers (after Knott, 1993, Fig. 8.18, p. 326).

A **Dahlenbach absorber** (Fig. A2) consists of a thick homogeneous lossy layer backed by a metallic plate. It is a simple narrowband absorber that is flexible and can be applied to different kinds of curved surfaces. It is characteristic of single-

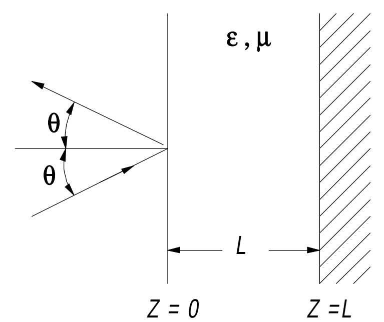


Figure A2 Dahlenbach absorber (after Bhattacharyya, 1991, Fig. 4.65, p. 211).

layer absorbers backed by metal plates that it is impossible to achieve zero reflection because the layer material must be such that low reflection occurs on its front face, and using physically realizable materials it is impossible to force reflection from both the front face and the metal backing to zero. The main objective in this case is to choose electrical properties of the layer to make two reflections to cancel each other. Reflectivity curves for dominantly electrical and magnetic layer materials are shown in Figs. A3 and A4, respectively. The optimum layer thickness in the first case is near a quarter wavelength, in the second case it is near a half wavelength. *SAL*

Ref.: Knott (1993), pp. 314–320; Bhattacharyya (1991), pp. 208–212.

A **dielectric absorber** uses dielectric absorbing materials for its construction. An example of a simple, single-layer dielectric absorber is the Salisbury screen. In practical applications, multilayer dielectric absorbers are used, such as Jaumann absorbers and graded dielectric absorbers. Practical graded dielectric absorbers are made of discrete layers with properties changing from layer to layer. *SAL*

Ref.: Knott (1993), pp. 313–327.

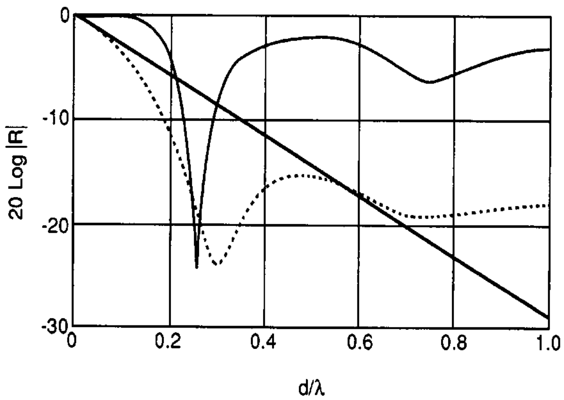


Figure A3 Reflectivity of dominantly electric materials. Solid trace: $|\varepsilon_r| = 16$, $|\mu_r| = 1$, $\delta_\varepsilon = 20°$, $\delta_\mu = 0°$; dashed trace: $|\varepsilon_r| = 25$, $|\mu_r| = 16$, $\delta_\varepsilon = 30°$, $\delta_\mu = 20°$; diagonal trace: $|\varepsilon_r| = |\mu_r| = 4$, $\delta_\varepsilon = \delta_\mu = 15°$. $\varepsilon_r = |\varepsilon_r|\exp(i\delta_\varepsilon)$ and $\mu_r = |\mu_r|\exp(i\delta_\mu)$ are the complex permittivity and permeability of the material relative to those of free space (from Knott, 1993, Fig. 8.12, p. 319).

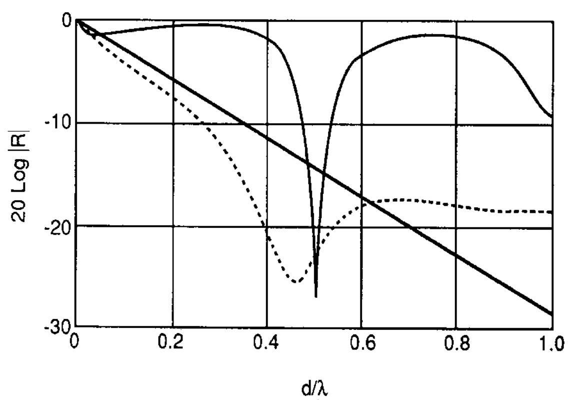


Figure A4 Reflectivity of dominantly magnetic materials. Solid trace: $|\mu_r| = 16$, $|\varepsilon_r| = 1$, $\delta_\mu = 10°$, $\delta_\varepsilon = 0°$; dashed trace: $|\mu_r| = 25$, $|\varepsilon_r| = 16$, $\delta_\mu = 20°$, $\delta_\varepsilon = 30°$; diagonal trace: $|\varepsilon_r| = |\mu_r| = 4$, $\delta_\varepsilon = \delta_\mu = 15°$ (from Knott, 1993, Fig. 8.13, p. 319).

Ferrite absorbing material provides attenuation of a radio wave passing through it. Ferrite absorbing coatings are marked by their low weight and thickness. Usually they are used for masking the warheads of ballistic missiles and various reflective parts of short-range missiles. They provide an attenuation of 15 to 30 dB. With a thickness of 5 mm, a square meter of coating has a weight of up to 5 kg. Ferrite absorbing materials are used for camouflage in a wide waveband, from the meter to the centimeter range.

Ferrite material is used for coatings of anechoic chambers, taking the form of a layer of tightly placed tiles or an absorbing wall consisting of individual magnetic rods arranged vertically and horizontally. *IAM*

Ref.: Stepanov (1968), p. 62; Bhattacharyya (1991), pp. 177, 217–218.

Frequency-selective surface (FSS) types of absorbers usually take the form of a thin metallic patterns etched into or deposited onto lossless substrates or films. The desired effect is to pass waves of a given range of frequencies, or all waves except those in a required band (bandpass or bandstop filtering). Other uses are high-pass or low-pass filtering. Some configurations used in FSS are shown in Fig. A5. Frequency selective surfaces find many practical applications: in antenna reflectors, wave polarizers, RCS control, and so forth. The Jaumann and circuit analog absorbers are versions of FSS. *SAL*

Ref.: Bhattacharyya (1991), pp. 224, 228.

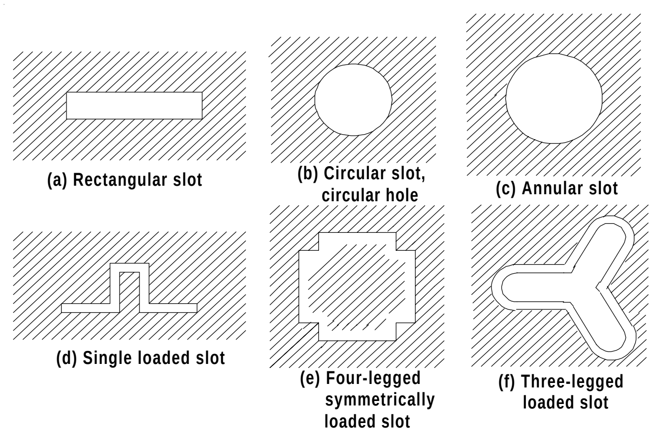


Figure A5 Frequency-selective surfaces (after Knott, 1993, Fig. 8.22, p. 330).

A **geometric transition absorber** is based on geometric transition from free space to the highly lossy medium that provides an effective dielectric gradient and minimizes reflections. The major shapes available are convoluted, wedge-shaped, twisted-wedge-shaped, rectangular, triangular, conical, and pyramidal. The pyramidal profile is most often used, usually having the structure of a planar array of pyramidal absorbers (Fig. A6). Geometric transition absorbers are used in anechoic chambers to reduce reflection from the

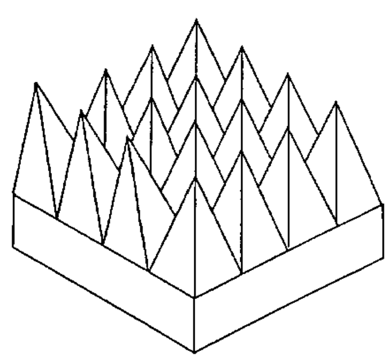


Figure A6 Geometric transition absorber (from Knott, 1993, Fig. 8.18, p. 326).

walls. This type of absorber can provide reflectivity reduction in excess of 50 dB and bandwidth from 100 MHz to 100 GHz. *SAL*

Ref.: Knott (1993), pp. 326, 528–532; Bhattacharyya (1991), p. 219.

A **graded absorber** is constructed from discrete layers with properties changing from layer to layer. The most common use layers of dielectric materials. One commercial example is a three-layer graded dielectric absorber about 1 cm thick with properties shown in Fig. A7. In the commercial productions of graded dielectric absorbers, five or more layers have been used. Commercial graded magnetic absorbers appear to have been limited to three layers. *SAL*

Ref.: Knott, (1993), p. 324.

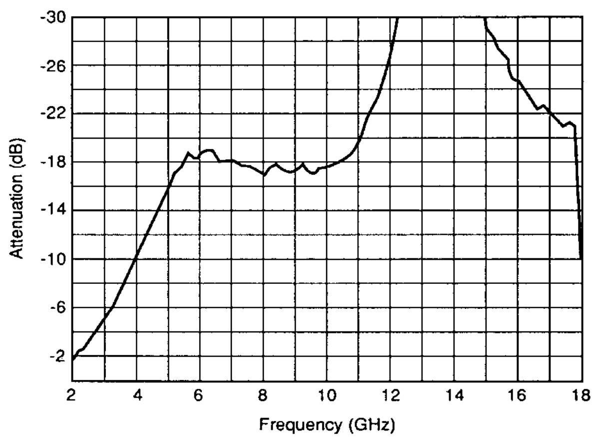


Figure A7 Measured reflectivity of a three-layer graded dielectric absorber (from Knott, 1993, Fig. 8.17, p. 325).

A **hybrid absorber** combines different types of absorbers to provide broader bandwidth or improved performance within the same band. For example, magnetic and circuit analog absorbers, or Jaumann and graded dielectric absorbers can be combined. Reflection coefficients as a function of frequency for a three-layer Jaumann, a graded dielectric, and a hybrid absorber are shown in Fig. A8. *SAL*

Ref.: Knott (1993), pp. 339–343.

Interference absorbing materials, when used as coatings, constitute resonant absorbers, consisting of one layer of dielectric applied to the metal surface that is to be protected. The thickness d and the constants ε (the permittivity) and μ (the permeability) of the material are selected for a given wavelength, λ, to meet the condition $d = \lambda/4(\varepsilon\mu)^{1/2}$.

The coating is usually made of plastic or rubber, filled with graphite powder or carbonyl iron. Such materials are narrowband absorbers and operate well only at angles of incidence close to normal. Materials of the interference type can also be used for effective absorption over a broad frequency band, with several layers having thickness and structure optimized for different wave lengths. This is achieved through a specific combination of dielectric and magnetic constants of the absorber. The material can also contain dipoles made from metal fiber, filamentary crystals, or fibers made from plastic with a metal coating.

Interference materials are made either with metal or nonmetal substrate having a high relative dielectric constant (100 to 200), the latter simplifying the attachment of the coating to the masked structure.

Multilayer interference materials provide signal attenuation from 20 to 40 dB at X-band, and from about 7 to 12 dB at C-band (Fig. A8). *IAM*

Ref.: Stepanov (1968), p. 55; U.S. Patent no. 3,568,195, cl. 343-18, 3-2-71.

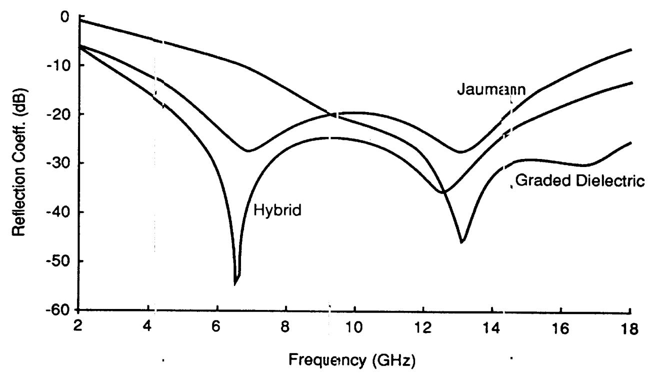


Figure A8 Reflection coefficient as a function of frequency for Jaumann, graded dielectric, and hybrid absorbers (from Knott, 1993, Fig. 8.28, p. 342).

Jaumann absorber is a wideband multilayer structure. It is made from alternating layers of lossy film and relatively thick layers of low-loss materials. The cascade process used in multilayer absorbers considerably improves the bandwidth of the absorption. Figure A9 shows the calculated reflected

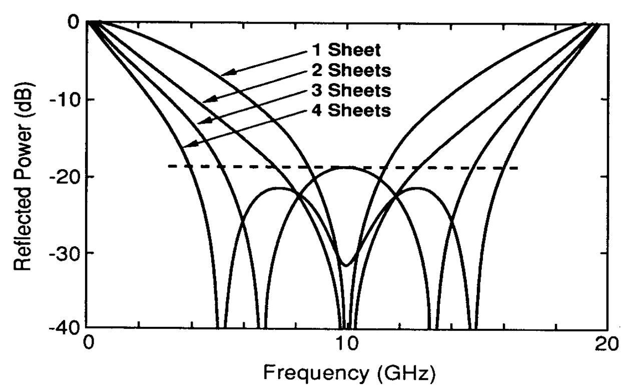


Figure A9 Jaumann absorber (from Knott, 1993, Fig. 8.15, p. 322).

power versus frequency for Jaumann absorbers containing variable numbers of resistive sheets. *SAL*

Ref.: Skolnik (1990), p. 11.48; Bhattacharyya (1991), p. 215; Knott (1993), pp. 320–323.

A **magnetic absorber** uses magnetic radar absorbing material such as ferrite slabs. It has an advantage over dielectric absorbers, because usually it requires only 1/10 of the thickness of dielectric absorbers to cause the same RCS reduction. As an example, absorption characteristics of a two-layer magnetic absorber, constructed from a ferrite-resin mixture impregnated with short metal fibers, are shown in Fig. A10. *SAL*

Ref.: Bhattacharyya (1991), pp. 217, 218; K. Hatakeyama and T. Inui, "Electromagnetic Wave Absorber Using Ferrite Absorbing Materials Dispersed with Short Metal Fibers," *IEEE Trans.* ***MAG-20,*** no. 5, Sept. 1984, pp. 126–1263.

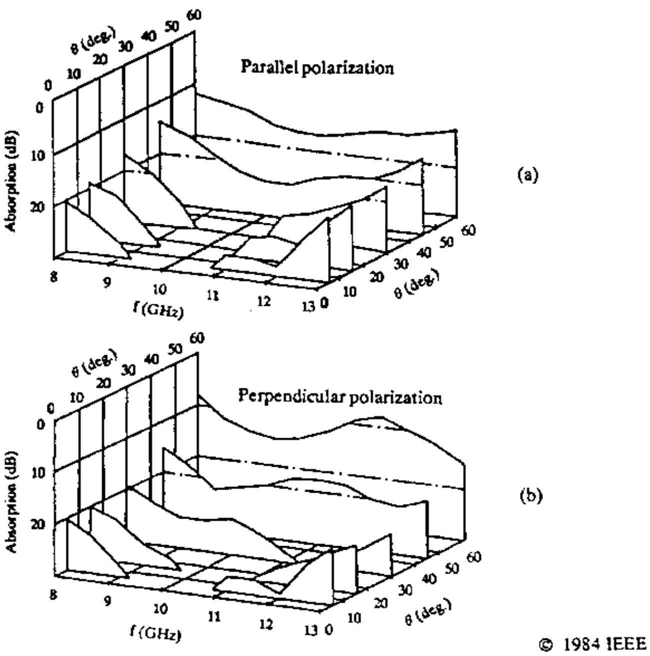


Figure A10 Absorption characteristics of a two-layer magnetic absorber (from Hatakeyama and Inui).

The $\mu_r = \varepsilon_r$ **type absorber** has performance based on the following theorem: If a target has equal values for relative permittivity and permeability, the far-zone backscattered fields are zero if shape and material of the body remain unchanged for a 90° rotation of the body around the direction of incidence. In the case of the $\mu_r = \varepsilon_r$ absorber, the intrinsic impedance of the medium is equal to that of free space, and so theoretically there will be no reflection from the interface with free space for illumination by normally incident plane waves. In practice the material always has some loss and the desired matching cannot be achieved, so we have some residual reflection. But using a layer of ferrite material with $\varepsilon_r = \mu_r$ makes it possible to reduce RCS over a considerable bandwidth. The magnitude of reflection coefficients at the plane interface between free space and a $\mu_r = \varepsilon_r$ absorber depends on the angle of incidence with $|\mu\varepsilon|$ as the parameter (Fig. A11). *SAL*

Ref.: Bhattacharyya (1991), p. 216.

Narrowband absorbing material usually is a single-layer interference material. The small thickness of the coating is an advantage of such material. *IAM*

Ref.: Finkel'shteyn (1983), p. 145; Bhattacharyya (1991), p. 204.

Nonspecular absorbing materials are RAMs designed to suppress returns that arise primarily from surface traveling waves, edge waves, or creeping waves. The main design approaches use magnetic and dielectric surface coatings to

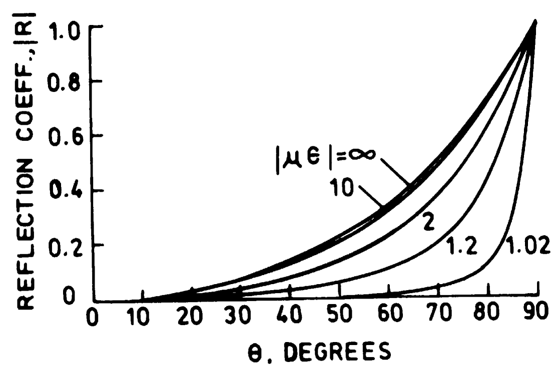


Figure A11 The angular performance of $\mu_r = \varepsilon_r$ absorber for different values of $|\mu\varepsilon|$ (from Bhattacharyya, 1991, Fig. 4.68, p. 215).

reduce surface currents (and so to suppress traveling and creeping waves echoes) and use tapered resistive strips to suppress edge diffraction returns. *SAL*

Ref.: Knott (1993), pp. 343–355.

Pyramidal absorber is the term sometimes used for a geometric transition absorber with pyramidal profile. *SAL*

Ref.: Bhattacharyya (1991), p. 219.

The **Salisbury screen absorber** is a classical resonator absorber that is the simplest specular narrowband radar-absorbing structure (Fig. A12). A Salisbury screen can be electric or magnetic and usually consists of a resistive sheet or screen in front of a conducting plane, separated by a dielectric or magnetic slab called a *spacer.* In practice, the resistive layer is glued to a light plastic foam or honeycomb spacer backed by metal foil.

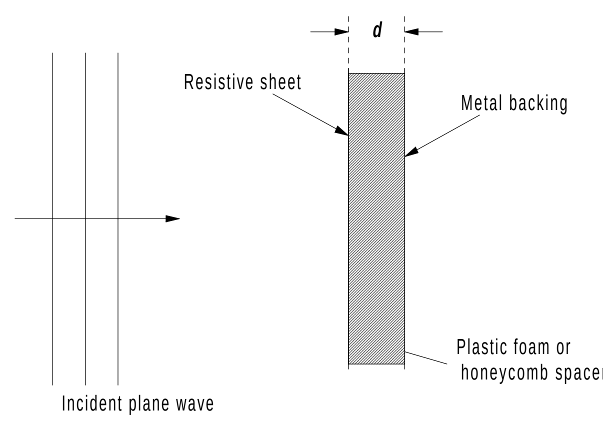


Figure A12 Salisbury screen absorber (after Knott, 1993, Fig. 8.8, p. 314).

Reflection coefficients of Salisbury screens depend on the angle of incidence. Recently, multiple electric and magnetic Salisbury screens were designed. This implementation provides a relatively large reduction of RCS in the specular direction, and the RCS reduction does not deteriorate too much in directions away from the normal, or if the surface is curved or contains fabrication errors. *SAL*

Ref.: Skolnik (1990), p. 11.46; Bhattacharyya (1980), pp. 204–208; Knott (1993), pp. 314–318.

Screening absorbing material is intended to attenuate undesirable radiation to protect the operators of the radar station and other groups of servicing personnel operating in the zone of the high-intensity microwave radiation. Screening radio-absorbing materials are intended to absorb high-intensity radiation (3 to 4.5 W/cm^2) and to operate jointly with cooling systems (air and water). A multilayer structure having conical recesses on its back surfaces is used for screening. Each layer consists of hollow ceramic microspheres bound by a glue-cement. Diameters of the microspheres diminish from layer to layer in the direction of the rear surface. The absorber withstands a temperature of up to 1,315°C. For screening of electronic apparatus and antenna cowlings, fabric material is used with a pyramidal structure and resistive surface intended for frequencies above 2.4 GHz and having a weight of 500 g/m^2. *IAM*

Ref.: Electronics, 1970, vol. 43, no. 1, p. 81; Patent CAN 3,441,933 cl. 343-18 of 4-29-69.

Structural absorption material is used to achieve a combination of physical strength and absorption. This technique is based on replacing the original structure with composites of absorbing material and nonmetallic structure, or combinations of filaments of wave absorbing materials and metallic or nonmetallic structure. *SAL*

Ref.: Morchin (1993), p. 123.

Surface-wave absorbing material is a thin layer of absorber, typically ferrite and synthetic rubber paint. To improve the performance multiple layers can be used (see **Jaumann absorber**). *SAL*

Ref.: Morchin (1993), p. 122.

A tunable absorbing material is one whose absorption band can be adjusted within certain limits. One example is a radio-absorbing grid of synthetic fiber or metal wire with a diameter of less than 0.1λ with absorbers attached to it. Depending on the wavelength, narrowband absorbers with the necessary dimensions are selected, and the distance between them on the grid is adjusted.

These absorbers are multilayer structures consisting of reflective and absorbing layers. When such a coating is used as a masking means for ground equipment, absorbers are combined that are effective at various wavebands from 0.1λ to 10λ, where λ is the longest wavelength of the radiation. *IAM*

Ref.: Paliy (1974), p. 197; U.S. Patent no. 3,427,619, cl. 348-18, dated 2-11-69.

Wideband absorbing material is effective over a wide frequency band. Depending on the application, wideband coatings of various types can be used. For example, for masking aerospace craft a material in the form of an elastic silicon-organic foam capable of operating for a long time at high temperatures (up to 260°C) in a waveband shorter than 4 cm is used. For masking of stationary or slowly moving objects, multilayer materials may be used, made from porous rubber mixed with coal dust, or coatings of pressed grains of polystyrene foam surrounded by a strong coal film. The front part of such coatings is usually corrugated. Ferrite radio-absorbing materials have wideband properties. *IAM*

Ref.: Stepanov (1968), p. 59; Bhattacharyya (1991), pp. 212–220.

ABSORPTION is "the irreversible conversion of the energy of an electromagnetic wave into another form of energy as a result of its interaction with matter." In radar, the major interest is in absorption along the path of propagation in the atmosphere, through precipitation, foliage, and so forth (see **ATTENUATION**). *SAL*

Ref.: IEEE (1993), p. 3; Skolnik (1990), pp.11.46–11.51; Knott (1993), pp. 297–359; Bhattacharyya (1991), pp. 176–220.

accumulator (**see INTEGRATOR**).

ACCURACY is the quality of freedom from mistake or error. In radar applications accuracy is usually considered in regard to the process of radar measurement and is characterized by measurement errors (see also **ERROR**). *SAL*

The **fundamental accuracy of radar measurement** is that corresponding to the minimum measurement error that can be attained for the measurements in a noise background. In other words, it is the minimum error due to the fundamental limitation: the presence of random noise. It cannot be reduced but can only be increased in a real system because of nonideal characteristics of radar subsystems and introduced losses. For estimation of parameters of coherent signals in a white noise background, the fundamental accuracy for separate measurement of delay time (range), doppler frequency (velocity) is

$$\sigma_t = \frac{1}{q\sqrt{|\ddot{x}_t(0,0)|}} = \frac{1}{qB_{ef}}$$

$$\sigma_f = \frac{1}{q\sqrt{|\ddot{x}_f(0,0)|}} = \frac{1}{q\tau_{ef}}$$

where σ_t and σ_f are the rms errors of delay time and frequency measurement, q^2 is the signal-to-noise ratio, $\ddot{x}(0,0)$ is the second derivative of the ambiguity function, $\chi(t_d, f_d)$ for $t_d = f_d = 0$, and B_{ef} and τ_{ef} are the effective bandwidth and duration of the signal. The analogous approach can be applied to angular coordinates measurement using the four- coordinate response concept. Finally, fundamental accuracy of a basic radar parameter, α, can be written as:

$$\sigma_{\alpha_{min}} = K_\alpha / q$$

where for range, $\alpha = R$, $K_R = c/2B_{ef}$; for doppler velocity, $\alpha = v_r$, $K_v = \lambda/2\tau_{ef}$, λ = wavelength; and for either angular coordinate, $\alpha = \theta$, $K_\theta = \lambda/L_{ef}$, where L_{ef} is the effective aperture width for the specified angular coordinate. Hence, the radar measurement accuracy of all radar parameters is higher when the signal-to-noise ratio increases. For a fixed signal-to-noise ratio, the wider the spectrum of radar waveform, the higher the range measurement accuracy; the greater the duration of radar waveform, the higher the velocity measurement accuracy; and the wider the antenna aperture in the plane of the estimated angle, the higher the angular measurement accuracy. *SAL*

Ref.: Skolnik (1980), pp. 400–411; Shirman (1981), pp. 200–205; Leonov (1988), p. 25.

The **fundamental accuracy of monopulse** is achieved by comparison of signal amplitudes in beams formed simultaneously (monopulse estimation or simultaneous lobing) permits the radar to approach the fundamental accuracy limit given by

$$\sigma_\theta = \frac{\theta_3}{k_m\sqrt{2(E/N_0)}}$$

where $k_m \cong 1.6$ is the monopulse slope constant and $E/N_0 = 2n(S/N)_m$ is the total energy ratio for a target on the sum beam axis. When the target is displaced from the axis, the error will increase for two reasons: (a) the energy ratio in the sum channel will decrease, and (b) a second component of error, caused by noise in the normalization process (by which Δ/Σ is formed), will appear. Figure A13 shows the ratio by which the off-axis target error will increase from the on-axis error. *DKB*

Ref.: Barton (1969), pp. 24, 43; Skolnik (1990).

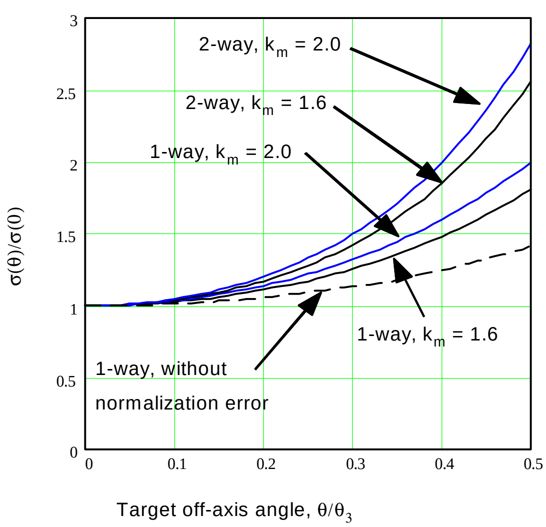


Figure A13 Ratio of monopulse off-axis error to on-axis error (from Barton, 1969, Fig. 2.12, p. 43).

The **fundamental accuracy of sequential lobing** can be achieved by any of three distinct methods of estimating target angle from observations of target amplitudes in beam positions sequenced in time:

(a) The beam may be scanned continuously, as with a mechanically scanned antenna, exchanging multiple pulses with varying amplitude during passage across the target position;

(b) The beam may be step-scanned across the target position, as with an electronically scanned antenna, with one or more pulses per step; or

(c) The beam may be scanned in a circle around the target position (see **RADAR, conical-scan**). The fundamental accuracy in each case, when measuring a nonfluctuating target, is limited by thermal noise, and is measured by the random noise error component σ_θ:

$$\sigma_\theta = \frac{\theta_3}{k\sqrt{2(E/N_0)}}$$

where θ_3 is the one-way half-power beamwidth, k is a pattern slope constant, and E/N_0 is the applicable signal-to-noise energy ratio. The different scanning options lead to different values of k and E/N_0.

For the case of continuous (linear or sector) scanning, the slope constant becomes $k_p = 1.66$ and the energy ratio is that of n pulses received with the on-axis ratio $(S/N)_m$ divided by a beamshape loss, $L_p = 1.33$.

For a step-scanned beam, the slope constant is given by

$$k = \frac{df}{d(\theta/\theta_3)} = \frac{d(G_2/G_1)}{d(\theta/\theta_3)}$$

where f is the voltage pattern of the beam, and G_1 and G_2 are the one-way power gains of the two beams nearest the target. This slope constant depends on the illumination function of the antenna and the spacing between the two beams. Figure A14 shows the normalized slope $K' = k(\lambda/L\theta_3) = k/0.886$ for a uniformly illuminated aperture of width L, as a function of target position in the beam.

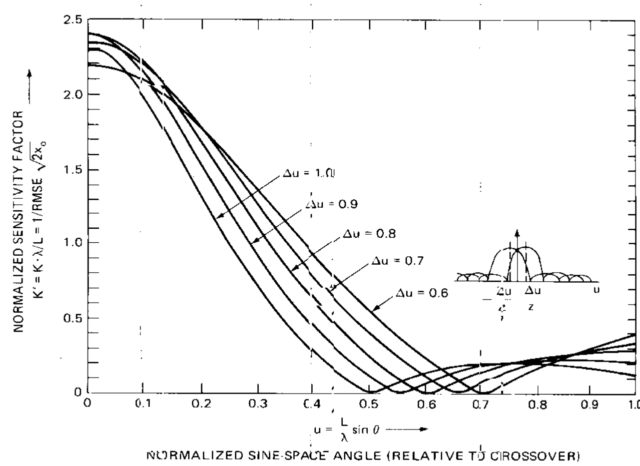


Figure A14 Normalized slope constant K' vs. target position for different beam spacings (from Skolnik, 1990, Fig. 20.4, p. 20.22, reprinted by permission of McGraw-Hill).

The applicable energy ratio is given by

$$\frac{E}{N_0} = \frac{(E/N_0)_1}{1+f^2}$$

where $(E/N_0)_1$ is the energy ratio in beam 1 and f is as defined in the slope equation. In this formulation, using only the energy ratio for beam 1, and that reduced for off-axis targets, the slope constant takes on a higher value than would appear if the total received energy were used.

For a conical scanning tracker, the slope constant $k = k_s$ is a function of the beam squint angle, as shown in Fig. A15. The energy ratio is that for n on-axis pulses divided by $2L_k$, or $E/N_0 = n(S/N)_m/2L_k$, where the crossover loss L_k is shown for

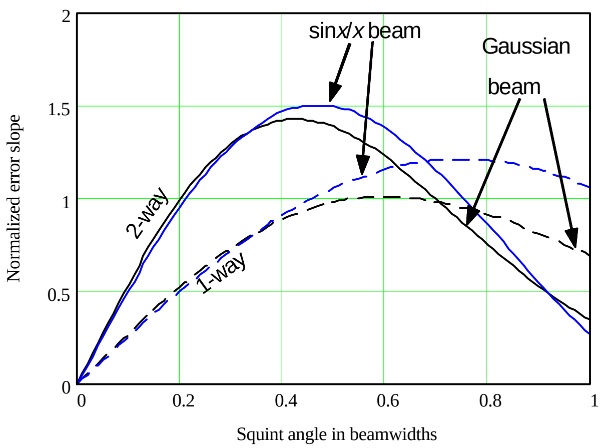


Figure A15 Conical-scan error slope (after Barton, 1969, Fig. 2.9, p. 36).

the two-way case in Fig. A16 (the one-way loss is one-half that shown).

These fundamental limits to accuracy are seldom the only significant errors in angle estimation, since the usual target will fluctuate at a rate such as to cause a scintillation error component equal to at least several hundredths of the beamwidth. *DKB*

Ref.: Barton (1969), pp. 33–37; Skolnik (1990).

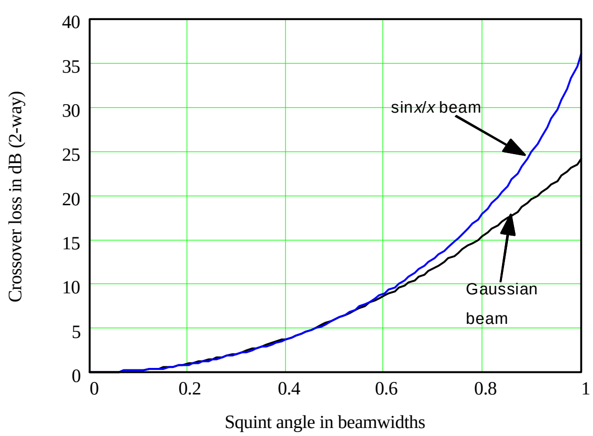


Figure A16 Conical-scan crossover loss (after Barton, 1969, Fig. 2.10, p. 37).

ACQUISITION is the process of establishing a stable track on a target that is designated in one or more coordinates. A search of a given limited volume of coordinate space is usually required because of errors or incompleteness of the designation. Acquisition usually involves target detection based on considerations of *S/N* threshold and integration procedure to accomplish a given probability of detection with a given false-alarm rate, after which radar automatically implements its tracking loops. A typical circuit for automatic acquisition is shown in Fig. A17. The main acquisition parameters are acquisition probability, acquisition range, and acquisition time. *SAL*

Ref.: IEEE (1993), p. 12; Barton (1964), pp. 437–466; Barton (1988), pp. 451–458; Skolnik (1980), p. 177, (1990), p. 18.26; Neri (1991), pp. 147–150.

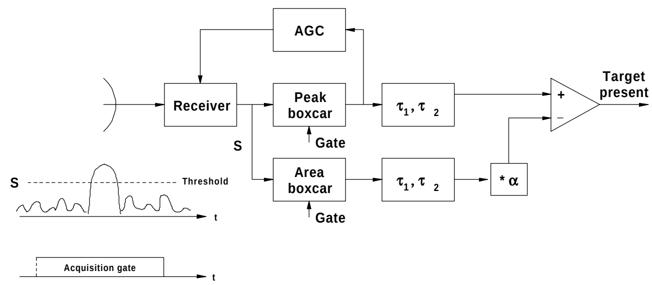


Figure A17 Automatic acquisition circuit (after Neri, 1991, Fig. 2.72, p. 148).

Acquisition probability is “the probability of establishing a stable track on a designated target”. The usual notation is P_a. This probability is defined as the summation over all resolution elements n_v, covered by the radar, of the product of two probabilities: P_v, the probability that the target lies within the scan volume, and P_d, the probability of detection of a target if it lies within this volume:

$$P_a = \sum_{i=1}^{n_v} P_{vi} P_{di}$$

Usually single-scan acquisition probability and cumulative acquisition probability are distinguished. *SAL*

Ref.: IEEE (1993), p. 12; Barton (1988), pp. 441, 453.

Cumulative acquisition probability is the overall probability of acquisition of the target on at least one of the k scans. The usual notation is P_c. If both the probability of target detection P_d and probability that the target lies within the scan volume P_v are independent from scan to scan, the cumulative probability of acquisition is:

$$P_c = 1 - (1 - P_a)^k$$

where P_a is the single-scan acquisition probability. *SAL*

Ref.: Barton (1964), p. 445, (1988), p. 455.

Acquisition range is the target range at the moment of time when the acquisition procedure can be considered complete. The usual notation is R_a. Acquisition range describes the acquisition capability of the radar and can be found from the basic search radar equation:

$$R_a^4 = \frac{P_{av} A_r t_s \sigma}{4\pi \psi_s k T_s D_0(1) L_s}$$

where P_{av} = transmitter average power, A_r = effective receiving aperture, t_s = search frame time, ψ_s = search solid angle, σ = RCS of target, k = Boltzmann’s constant, T_s = effective system noise temperature, $D_0(1)$ = detectability factor of single sample from steady target, and L_s = total search loss. *SAL*

Ref.: Barton (1964), pp. 451, 456–458.

Reacquisition is the process of acquiring a target that has been under track but has been lost. Even if lock-on has been successful, the target may be lost as a result of fading or increase in interference level. The tracking radar and its designation system should have means to provide rapid reacquiring based not only on designation sources but also on information obtained during the previous acquisition process and subsequent tracking. This process of reacquisition is the simplest for targets with highly predictable trajectory parameters, such as satellites. In the case of a satellite, a relatively short track with a precision radar can be used to generate orbital elements for reacquisition of a satellite at its next revolution. *SAL*

Ref.: Barton (1964), p. 458.

Single-scan acquisition probability is the probability of acquisition of the target on a single scan. The usual notion is P_a. If probability P_a is independent from scan to scan it is defined and related to cumulative acquisition probability by:

$$P_a = P_v P_d = (1 - P_c)^{1/k}$$

where P_v is the probability that the target lies within the scan volume, P_d is the probability of target detection, P_c is the required cumulative probability of acquisition, and k is the number of scans. *SAL*

Ref.: Barton (1964), p.445, (1988), p. 455.

Acquisition time is the time a radar needs to acquire a target with the required acquisition probability. The usual notation is t_a. During t_a the radar may perform one or more scans so that the cumulative acquisition probability reaches the required value. *SAL*

Ref.: Barton (1988), p. 451.

ADAPTER, microwave. A microwave adapter is a device providing the connection between two transmission lines. One distinguishes between narrowband and wideband adapters, and adapters between transmission lines of one type or different types. When adapters are designed, attention is paid to achieving high quality of transmission line matching over the frequency band while ensuring the required power handling capability. Based on the type of transmission line, there are adapters between waveguides of various shapes (e.g., rectangular or round), coaxial waveguides, coaxial- and waveguide-strip adapters, and others. These adapters, as a rule, are narrowband devices.

To match active loads, stepped and smooth adapters are used between transmission lines of a single type, ensuring wideband matching for a given reflection coefficient. *IAM*

Ref.: IEEE (1993), p. 15; Montgomery (1947), Ch. 10; Sazonov (1988), pp. 57, 140; Rakov (1970), vol. 2, p. 246.

A **coaxial-waveguide adapter** is an adapter between a coaxial and waveguide transmission line. Coaxial-waveguide adapters provide excitation of a rectangular waveguide with a H_{10}-wave, and of a round waveguide with a E_{01}-wave from a coaxial waveguide with a T-wave (see **WAVE, electromagnetic**). The basic component of a coaxial-waveguide adapter is the pin, around which a current flows, which is located in a waveguide short-circuited on one side, parallel to the lines of force of the electrical field.

The maximum bandwidth of such adapters reaches 20% with a traveling-wave ratio of 0.95. *IAM*

Ref.: Montgomery (1947), p. 336; Sazonov (1988), p. 57; Lavrov (1974), p. 331.

A **smooth adapter** is one whose cross section changes smoothly. Essentially, smooth adapters are the extreme case of stepped adapters with an unlimited increase in the number of steps and a tendency toward zero in the length of each of them. In type of frequency characteristic, a smooth adapter is equivalent to a high-pass filter. Good matching is achieved in all frequencies above some boundary frequency.

A smooth adapter is preferable to a stepped adapter when the power to be transmitted is high. The minimum length of a smooth adapter must be three to four times longer than the wave length in the line. *IAM*

Ref.: Montgomery (1947), p. 339; Sazonov (1988), p. 144; Rakov (1970), vol. 2, p. 246.

A **stepped adapter** is one whose cross section changes in a step-wise manner. The simple stepped adapter is a single-stage adapter with step length equal to one-fourth of the wave length, the quarter-wave transformer, but it has limited bandwidth. Multistage adapters are used to extend the band. The selected step length is the same, and the necessary shape of the frequency matching characteristic is assured by selection of the wave resistances of the steps. A stepped adapter is equivalent to a bandpass filter. In comparison with a smooth adapter, a stepped adapter has a shorter length with the identical wave resistances and mismatch tolerances, but is inferior to the smooth adapter in power handling capability. *IAM*

Ref.: Sazonov (1988), pp. 45, 140; Rakov (1970), vol 2., p. 246.

ALGORITHM

The **Burg algorithm** is an implementation of the maximum entropy method technique to overcome disadvantage of poor spectral resolution in spectral estimation tasks. The Burg algorithm uses the output data of the doppler filter (Fig. A18) to estimate the feedback coefficients of an all-pole network, whose output is given by difference equation:

$$Y(n) = \varepsilon(n) + \sum_{k=1}^{p} a_{pk} y(n-k)$$

where $Y(n)$ are the complex voltages at the N taps of the adaptive doppler processor; $\varepsilon(n)$ is the white noise sequence exciting the network, and a_{pk} are the filter coefficients.

The Burg algorithm employs recursive relations to determine the coefficients a_{pk}, which determine the locations of poles in the transfer function. After these coefficients are esti-

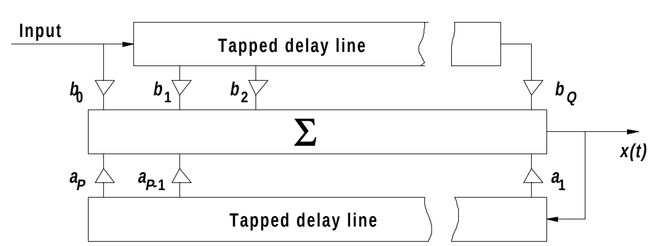


Figure A18 General filter model (after Nitzberg, 1992, Fig. 11.7, p. 297).

mated, the spectrum is computed in a regular way, by inserting them into the filter transfer function. *SAL*

Ref.: Nitzberg (1992), pp. 296–298; Galati (1993), p. 391.

The **Cooley-Tukey algorithm** is a time-saving computational technique in spectral analysis. The exponential Fourier transform of periodic function $F(t)$,

$$G(\omega) = \frac{1}{2\pi}\int_{-\infty}^{\infty} F(t)\exp(j\omega t)\,dt$$

can be written in discrete matrix form as

$$[G_n] = [W^{nk}][F_k] \qquad (1)$$

$$n = 0, 1, 2, \ldots (N-1)$$
$$k = 0, 1, 2, \ldots (K-1)$$

where

$$F_k = \frac{T}{2\pi K}F(t_k),$$

$W = e^{-2\pi j/N}$, $\omega_n = n\Delta\omega$, $t_k = k\Delta t$, and $\Delta\omega$, Δt are increments of sampling in frequency and time domain correspondingly.

The Cooley-Tukey method of computing the $[G_n]$ matrix is based on expressing the $[W^{nk}]$ matrix in terms of products of Y square matrices, where Y is an integer in the increment equation $K = 2^Y$. It offers the possibility of reducing the number of multiplications to YN (where N is the order of these square matrices) as compared to N^2 for the direct evaluation of the matrix of (1). The Cooley-Tukey algorithm is extensively applied in spectral analysis and digital signal processing. *SAL*

Ref.: Cooley (1965), pp. 297–301; Hovanessian (1984), pp. 251–264.

The **dynamic programming algorithm (DPA)** is used to define the optimum paths over which a system can make transitions from one state to another. The possible paths the system can take are assigned numerical values through a merit function. The most common of these functions are efficiency and cost. The DPA was originally developed by Bellman for control problems. Later it found uses in target detection, signal processing, and radar system analysis and tradeoff studies. For these uses it is more commonly known as the Viterbi algorithm, which is a forward DPA, recursively updating the merit function through consecutive stages of the system. *SAL*

Ref.: Bellman (1957); Bar-Shalom, (1990), pp. 85–153.

Gram-Schmidt algorithm (see **CANCELER, Gram-Schmidt**).

Howells-Applebaum algorithm (see **CANCELER, Howells-Applebaum**).

Sidelobe cancelation (SLC) algorithms are the algorithms used in the sidelobe cancellation technique to place a null in the required direction. For an adaptive processor with N degrees of freedom, the N input channels are each multiplied by a complex weight and summed to form the output:

$$\overrightarrow{RW} = \mu\vec{S}$$

where $\vec{R}$ is the matrix of the cross-covariances of the signals in the N channels; $\vec{W}$ is a vector of the weights; μ is an arbitrary nonzero constant, and $\vec{S}$ is vector representation of the desired signal in each channel.

The main algorithms to determine the optimum weights $\vec{W}$ are Kalman methods, the Howells-Applebaum control loop, the Widrow algorithm, the Gram-Schmidt algorithm, direct matrix inverse, and inverse matrix updating. The possible implementations of SLC algorithms are all-analog, all-digital, and hybrid. The principal advantage of analog implementation is simplicity of RF channel matching. The major disadvantages are lack of flexibility and slow convergence. On the contrary, the all-digital SLC is flexible and possesses high-speed convergence but meets difficulties in matching RF-to-digital receiver chains. The most common hybrid SLC is the cascaded A/D canceler, which offers the advantages of both digital and analog systems. *SAL*

Ref.: Cantafio (1989), pp. 465–467; Monzingo (1980).

The **step transform algorithm** is a special algorithm in matched filter processing using subaperture processing to reduce the size of fast Fourier transform (FFT). A block diagram of the step transform matched filter is shown in Fig. A19. A short duration signal is mixed with the incoming signal and the output is fed to the FFT processor. The time aperture of the processor is equal to the duration of the reference waveform but is less than the total waveform length. The use of a series of short-aperture FFTs gives the possibility of reducing hardware requirements by up 50% from conventional FFT approaches. The step transform algorithm is especially efficient for synthetic aperture radar (see also **TRANSFORM, Fourier**). *SAL*

Ref.: Brookner (1977), pp. 163–169.

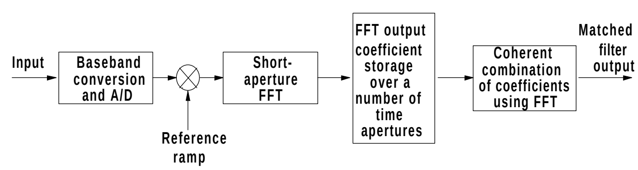


Figure A19 Step transform matched filter (after Brookner, 1977, Fig. 4, p. 165).

Viterbi algorithm (see **dynamic programming algorithm**).

The **Volder algorithm** is a recursive procedure for determining the coordinates of a vector when it is rotated through a given angle. It was developed by J. E. Volder in 1956 for the calculation of trigonometric and hyperbolic functions. The algorithm uses a sequence of rotation angles $\alpha_i = \arctan(2^{-i})$, which makes it possible to reduce the rotation of the vector to a series of additions, subtractions, and shifts (division by 2), which are readily implemented in digital hardware. The algorithm is also known as the coordinate rotation digital computer (CORDIC) algorithm.

The Volder algorithm is the basis of algorithms for calculating elementary functions, various trigonometric and transcendental functions, and solutions to equations containing these functions. The Volder algorithm is used to construct processors for transforming coordinates in radar and navigation systems, and processors for computing the discrete or fast Fourier transform. *IAM*

Ref.: Volder, J. E., *IRE Trans.* ***EC-8***, no. 9, 1959.

The **Widrow algorithm** is a single-step algorithm for unconditional minimization. It is used in adaptive sidelobe cancellation for phased array antennas. With this algorithm, the weighting vector under stationary conditions at the *n*th iteration is calculated using the formula:

$$\vec{w}_n = \vec{w}_{n-1} - \gamma\varepsilon_{n-1}\vec{x}_{n-1}$$

where $\varepsilon_{n-1} = (\vec{w}^* - \vec{w}_{n-1})^T\vec{x}_{n-1}$ is the error between the model and the $(n-1)$th measurement, and $\vec{w}^*$ is the optimum stationary value of the weighting vector when the signal is received from the given direction. The goal of the adaptation in the algorithm is the minimization of the mean square error ε^2 and, as a result, the suppression of interference in the sidelobes. The rate of convergence and validity of the algorithm depend upon the parameter γ, which is estimated through a combination of measurements.

The Widrow algorithm is self-training, based on the fastest descent method. The applicability of the algorithm is limited by the stationarity of the signal conditions and interference. The algorithm is appropriate for use in adaptive arrays, when a distinction between the modulation characteristics of the signal and interference may be expected. Implementation of the algorithm requires one correlator for each antenna element. *IAM*

Ref.: Widrow, B., et al., Adaptive antenna systems, *Proc. IEEE* ***55***, no. 12, Dec. 1967, pp. 2143–2159; Monzingo (1980), p. 11; *Radiotekhnika*, no. 11, 1986, p. 8; Galati (1993), p. 394.

ALIASING is the process of distortion in a sampled data system induced by sampling at a rate that is less than that required for ideal sampling. Examples of aliasing in the time and frequency domains are shown at Figs. A20 and A21.

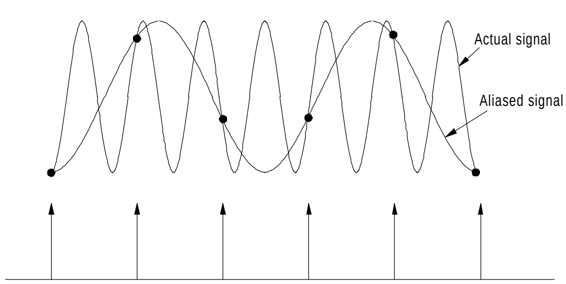


Figure A20 Aliasing in the time domain (after Currie, 1989, Fig. 6.6, p. 183).

Aliasing can be removed by increasing the sample rate to the value required to pass the frequency components of the signal, or with an antialiasing filter added before sampling

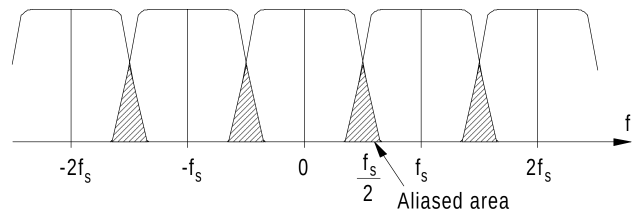


Figure A21 Aliasing in the frequency domain (after Currie, 1989, Fig. 6.7, p. 183).

(Fig. A22). To achieve a high-enough rate is not always feasible due to hardware constraints, and additional filters introduce time delay and are expensive. Thus, the solution of aliasing problems is usually a combination of sample rate selection and presample filtering. *SAL*

Ref.: Barton (1969), p. 185; Currie (1989), pp. 182–184.

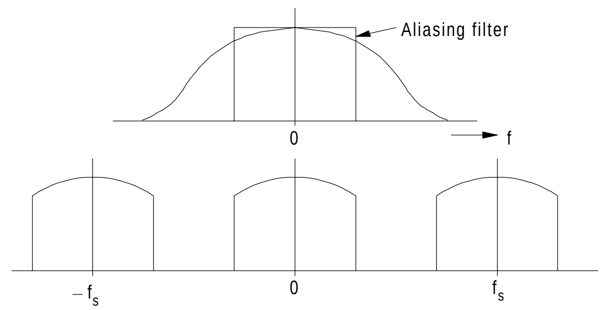


Figure A22 Antialiasing filter (after Currie, 1989, Fig. 6.9, p. 185).

ALTIMETER, RADAR. An altimeter is defined as "an instrument which determines the height of an object with respect to a fixed level, such as sea level." A radar altimeter is one using radar principles for height measurement of a flying vehicle. Types of radar altimeters are divided into frequency-modulated continuous wave (FMCW) altimeters and pulse altimeters, depending on the waveforms used. FMCW altimeters are further classified into broad-beamwidth types and narrow-beamwidth types from the point of view of antenna beamwidth. Pulse altimeters can be referred to short-pulse altimeters or pulse-compression altimeters depending on whether intrapulse modulation is used. Altimeters operating in optical bands are termed *laser altimeters*.

The main function of a radar altimeter is to measure and indicate the height of a flying vehicle like an aircraft or a spacecraft. Usually, low-altitude altimeters, operating up to 3,000m, and high-altitude altimeters (more than 3000m) are distinguished. The Federal Communications Commission has assigned to radar altimetry two frequency bands, centered approximately at 1,600 and 4,300 MHz. Radar altimeters are also termed *absolute altimeters*, as distinct from barometric altimeters that determine height by sensing local air pressure. (See also **RANGE EQUATION, altimeter**). *SAL*

Ref.: IEEE (1993), p. 30; Hovanessian, (1984) pp.326–333; Skolnik (1980) pp. 84–86, 14.34–14.36; Cantafio (1989) pp. 229–279.

An **FMCW altimeter** is one using frequency-modulated continuous-wave waveforms. The most common modulation waveforms are triangular and sawtooth linear frequency modulation. A basic mechanization of an FMCW altimeter is shown in Fig. A23. The frequency modulator generates a modulated waveform that is applied to the voltage-controlled oscillator (VCO). The VCO and a power amplifier, if need be, constitute the transmitter, from which energy is directed toward the surface by the transmitting antenna. The linearizer adjusts the scale factor by setting the slope of the frequency modulation.

Backscattered energy arriving through the receiving antenna is applied to the receiver. A sample of transmitted signal is mixed with the received signal, and the difference frequency, containing information about the measured altitude, is amplified and applied to the signal processor. With respect to antenna beamwidth, broad- and narrow-beamwidth FMCW altimeters are distinguished. The first is one using a beamwidth of typically tens of degrees. This type of altimeter is used when the altitude has to be measured to the closest point below the vehicle, or when measurements are needed while the vehicle is subject to large pitch and roll. The second uses a beamwidth of a few degrees. This type of altimeter is used when it is necessary to measure range to the surface along a given axis. An example is the radar altimeter for the Surveyor spacecraft, with beamwidth equal to 4°. *SAL*

Ref.: Skolnik (1980), pp. 84–86; Hovanessian (1984), pp. 329–331; Cantafio (1989), pp. 245–247.

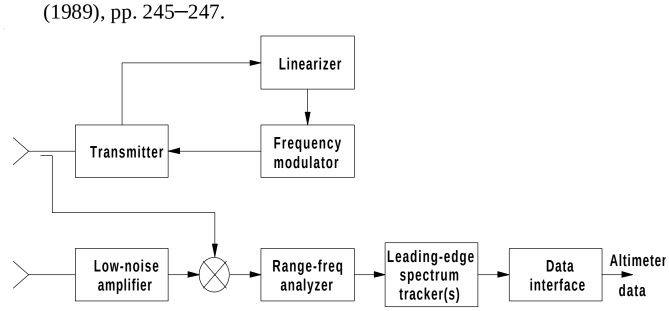


Figure A23 Block diagram of basic FMCW altimeter (from Cantafio, 1989, Fig. 7.11, p. 247).

A **laser altimeter** is one operating in the optical band. In practice it is a pulsed laser rangefinder providing accurate ranging information to a few meters maximum error. *SAL*

Ref.: Brookner (1977), p. 354.

A **pulse altimeter** is one using pulsed waveforms. In principle it is a pulsed radar making range measurement in the direction toward the earth. Usually it has low-altitude and high-altitude modes in which sensitivity time control (STC) voltage varies to control receiver gain. The major types of pulse altimeters are the short-pulse altimeter and the pulse-compression altimeter. A pulse-compression altimeter is one using complex waveforms with intrapulse modulation. It is a useful system for high-resolution radar altimetry; for example, for spacecraft vehicles to ensure reasonable values of peak transmitted power producing measurement from orbital altitudes. A pulse-compression technique was employed on the radar altimeters used on the Skylab and GEOS-C earth-orbiting spacecraft.

The major types of pulse compression altimeters are linear frequency modulated (or chirp) altimeters and phase-shift keyed (PSK) altimeters. A variant of the chirp pulse-compression system is the stretch technique, which uses a chirped local oscillator in conjunction with series of bandpass filters to process the received signal and has the potential for resolution of a few centimeters. The PSK altimeter has some advantages relative to short-pulse and FMCW altimeters, as the waveform usually is digitally generated and processed so that both modulation and processing errors are minimized.

The short-pulse altimeter is one using a pulsed waveform with short pulsewidth without intrapulse modulation. The pulsewidth may vary from tens of nanoseconds at low altitudes to several microseconds at high altitudes. The principle of operation is common to pulsed radar without pulse compression. Any of several implementations of range trackers, differing basically in the range-gating arrangement, can be used. *SAL*

Ref.: Cantafio (1989), pp. 230, 237–242.

AMBIGUITY, in radar measurement, is the effect of erroneous measurement of target location when more than one value of a target coordinate corresponds to the single value of radar return parameters. Usually angle ambiguity, doppler ambiguity, range ambiguity, and range-doppler ambiguity are distinguished. *SAL*

Ref.: Barton (1969), p. 12; Hovanessian (1984), p. 332; Cantafio (1989), pp. 237–257.

Angle ambiguity leads to an erroneous angle measurement due to the periodic character of an antenna pattern resulting from its multilobe structure. The problem of angle ambiguity is especially common in measurement by phased arrays, interferometers, and antenna measurement with multipath propagation effects. The elimination of angle ambiguity is achieved by using special design solutions for the specific antenna system and conditions of its application. *SAL*

Ref.: Barton (1964), pp. 54–56.

A **distal ambiguity** is a system response far from the desired target location, that is undesirable if additional targets or clutter may appear at such a location.

Ref.: Nathanson (1990), p. 285.

Doppler ambiguity expresses the possibility of assigning different values of radial velocity v_r to a given doppler frequency f_d. It is the result of the periodic character of the doppler spectrum in pulsed-doppler radars operating with discrete numbers of pulses (pulse trains). The maximum unambiguous doppler frequency is

$$f_{du} = \pm f_r / 2$$

where f_r is the pulse repetition frequency. Correspondingly, the maximum unambiguous radial velocity is

$$v_{ru} = \pm \lambda f_r / 4$$

where λ is wavelength.

Doppler ambiguity is absent, by definition, when a high pulse repetition frequency (PRF) waveform is used, although this solution leads to ambiguous range measurement (range ambiguity). *SAL*

Ref.: Long (1992), p. 98.

Phase ambiguity is the result of the periodic nature of a sinusoidal wave, and sometimes of the inability to determine the quadrant in which a given signal lies. Quadrant ambiguities in phase measuring can be resolved, for example, by using coherent in-phase/quadrature detection of the received signal. *SAL*

Ref.: Currie (1987), p. 499.

Range ambiguity is the result of the periodic structure of a transmitted pulse train. If the pulse repetition interval is t_r, and measured time delay from transmission to the received target echo pulse is t_d, the target range is normally taken to be $R_0 = t_d\, c/2$, where c is the velocity of light. However, the target may actually be at a range $R = R_0 + iR_u = (t_d + it_r)c/2$, where R_0 is the apparent range at which target is detected, and i is any positive integer. Target echoes received from an earlier pulse transmission are called multiple-time-around echoes. They can be distinguished from unambiguous echoes if the radar operates with varying PRF or completely eliminated if it is operated at low PRF where, by definition, the maximum range at which targets are expected is less than the unambiguous range $R_u = c/2f_r$, where f_r is the pulse repetition frequency. The upper bound on the PRF to provide unambiguous range measurement is called the *range-ambiguity limit*. *SAL*

Ref.: Skolnik (1980), p. 53; Cantafio (1989), p. 127.

Range-doppler ambiguity refers to the joint range and doppler ambiguity in pulsed-doppler radar due to effects of range and doppler ambiguities. This case applies by definition to the medium PRF mode of operation. The choice of proper PRF and waveform (e.g., noiselike waveforms or bursts at different PRFs) can be helpful when unambiguous target location in both range and velocity is required. *SAL*

Ref.: Nathanson (1990), pp. 306–310.

AMBIGUITY FUNCTION. The ambiguity function is "the squared magnitude $|c(t_d, f_d)|^2$ of the function that describes the response of a radar receiver to targets displaced in range (time delay, t_d) and doppler frequency, f_d, from a reference position, where the function $|\chi(0, 0)|^2$ is normalized to unity. Mathematically,

$$\chi(t_d, f_d) = \int_{-\infty}^{\infty} u(t)u^*(t + t_d)\exp(j2\pi f_d t)dt$$

where *u(t)* is the transmitted envelope waveform, suitably normalized, positive *t* indicates a target beyond the reference delay, and positive f_d indicates an incoming target. Used to examine the suitability of radar waveforms for achieving accuracy, resolution, freedom from ambiguities, and reduction of unwanted clutter." The function was first introduced by Woodward. *SAL*

Ref.: IEEE (1993), p. 31; Woodward (1953); Skolnik (1980), pp. 411–420; Nathanson (1990), pp. 283–301; Sloka (1970), pp. 45–61.

Ambiguity functions for typical waveforms. There are three general classes of ambiguity function: the knife-edge (ridge), the bed of spikes (nails), and the thumbtack ambiguity function. Ambiguity functions for several common radar waveforms are given in Table A1, and two-dimensional diagrams of these ambiguity functions are given in Figs. A24 through A30.

The following notations are introduced: f_0 is the waveform carrier frequency; τ is the transmitted pulse width; τ_e is the effective duration of the waveform (equal to τ for a rectangular transmitted pulse); B is the frequency excursion for a chirp waveform during time τ; K_w is the compression ratio:

$$B_e = \frac{\tau_e^2}{\int_{-\infty}^{\infty} G^2(f)df} \quad \text{and} \quad t_e = \frac{\tau_e^2}{\int_{-\infty}^{\infty} A_m^4(t)dt}$$

are Woodward functions; $A_m(t)$ is the amplitude modulation function; and $G(f)$ is the energy spectrum. *SAL*

Ref.: Skolnik (1970), Ch. 3; Sloka (1970), pp. 40–43.

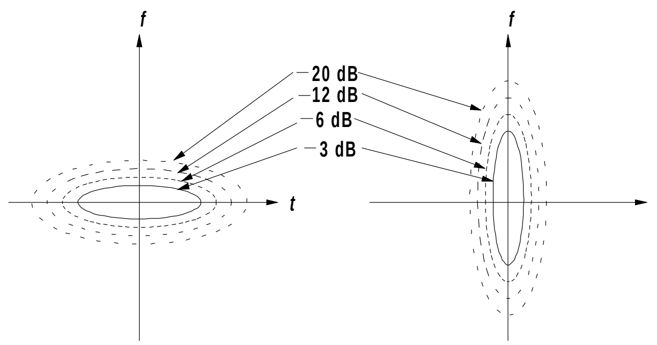


Figure A24 Two-dimensional ambiguity diagram of a single Gaussian pulse. (a) long pulse; (b) short pulse. The contours indicate levels relative to the central response.

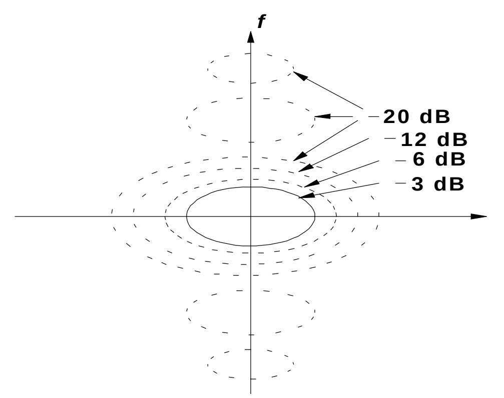


Figure A25 Ambiguity function of a rectangular pulse.

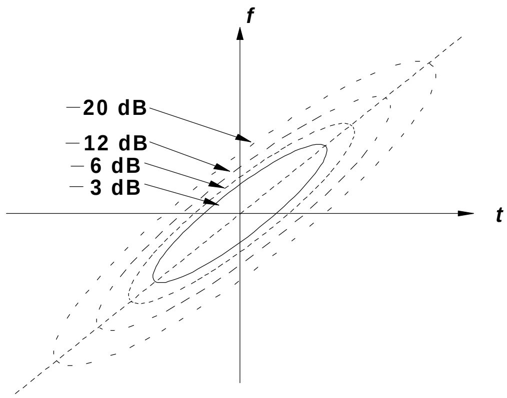


Figure A26 Ambiguity function of an LFM Gaussian pulse.

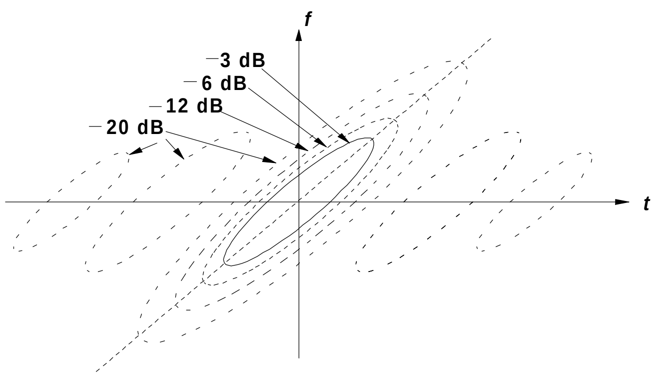


Figure A27 Ambiguity function of an LFM rectangular pulse.

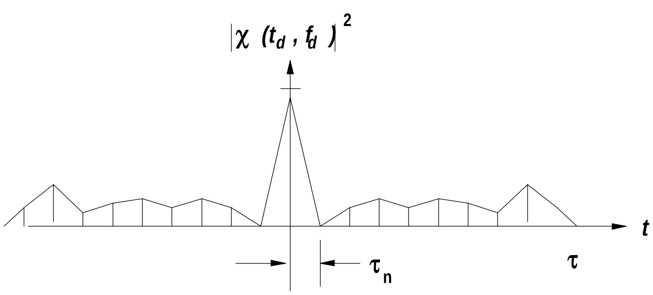


Figure A28 Ambiguity function of a PSK pulse.

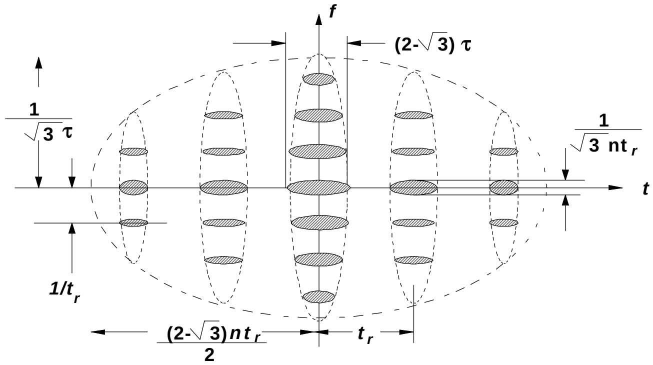


Figure A29 Ambiguity function of a train of Gaussian pulses (after Sloka, 1970).

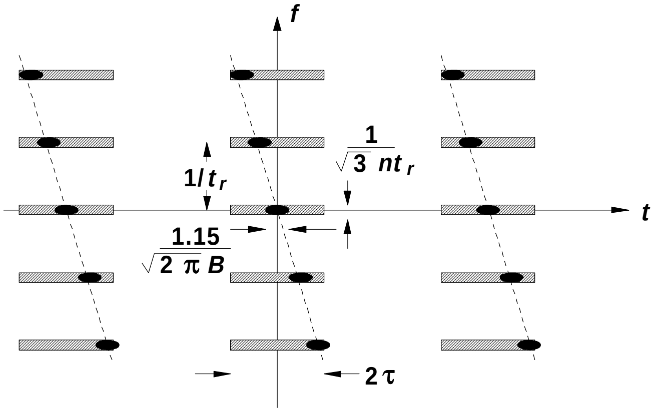


Figure A30 Ambiguity function of a train of rectangular

The **bed of spikes [nails] ambiguity function** applies to periodic pulse trains. The internal structure of its components depends on the waveform of individual pulses in the train. The response regions for this waveform appear equally spaced in the delay-doppler plane (Fig. A31). *SAL*

Ref.: Skolnik (1980), p. 418.

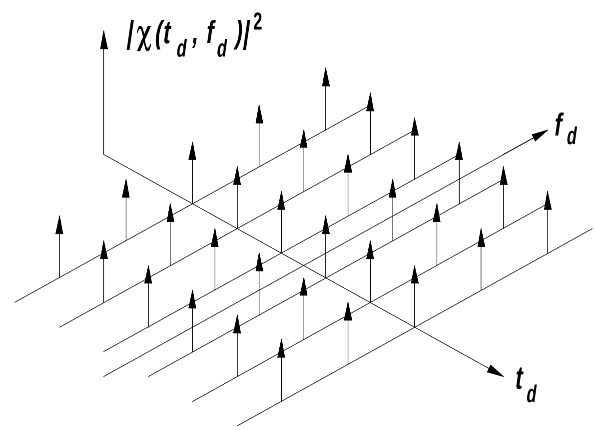


Figure A31 Bed of nails ambiguity function (after Skolnik, 1980, Fig. 11.13, p. 419).

The **cross-ambiguity function** is a description of radar receiver response when the receiver filter is mismatched to the transmit waveform:

$$|\chi(t_d, f_d)|^2 = \left|\int_{-\infty}^{\infty} u(t)v^*(t+t_d)\exp(j2\pi f_d t)\mathrm{d}t\right|^2$$

where the variable u refers to the waveform properties and v to the mismatched filter. *SAL*

Ref.: Rihaczek (1969), pp. 153–157; Nathanson (1990) p. 286.

 ambiguity functions for typical waveforms

Table A1
Ambiguity Functions of Common Waveforms

Waveform	Ambiguity Function	RMS Widths in Time and Frequency		Correlation Widths (at 0.75 level)		Resolution Factors	
$u(t)$	$\lvert\chi(\tau_d, f_d)\rvert$	α	β	τ_c	f_{dc}	$1/B_e$	$1/\tau_e$
1. Gaussian pulse $\exp\left(-\frac{\pi t^2}{2\tau_e^2}\right)\exp(j2\pi f_d t_d)$	$\exp\left(-2\pi\left(\frac{t_d^2}{4\tau_e^2}+\tau_e^2 f_d^2\right)\right)$	$\sqrt{2\pi}\tau_e$	$\frac{\sqrt{\pi/2}}{\tau_e}$	$\sqrt{\frac{2\times 1.15}{\pi}}\tau_e$	$\frac{\sqrt{1.15}}{\sqrt{2\pi}\tau_e}$	$\sqrt{2}\tau_e$	$\frac{1}{\sqrt{2}\tau_e}$
2. Rectangular pulse $\exp(j2\pi f_0 t), \quad \lvert t\rvert \le \tau/2$ $0, \quad \lvert\tau\rvert > t_0/2$	$\left(\frac{\sin(\pi f_d(\tau_d - \lvert t_d\rvert))}{\pi f_d \tau}\right)^2, \lvert t_d\rvert \le \tau$ $0, \quad \lvert\tau\rvert > t_0/2$	$\frac{3\pi}{2\tau}$	$\frac{\pi\tau}{\sqrt{3}}$	$(2-\sqrt{3})\tau)$	$\frac{1}{\sqrt{3}\tau}$	$\frac{2\tau}{3}$	$\frac{1}{\tau}$
3. LFM Gaussian pulse $\exp\left(-\frac{\pi t_d^2}{2\tau_e^2}\right)\exp\left(j2\pi\left(f_0+\frac{Bt}{2\tau}\right)t\right)$	$\exp\left(-2\pi\left(\frac{(1+4K_w)^2 t_d^2}{4\tau_e^2} + \tau_e^2 f_d^2 + 2K_w f_d t_d\right)\right)$	$\sqrt{2\pi}B$	$\sqrt{2\pi}\tau_e$	$\frac{\sqrt{1.15}}{\sqrt{2\pi}B}$	$\frac{\sqrt{1.15}}{\sqrt{2\pi}\tau_e}$	$\frac{1}{\sqrt{2}B}$	$\frac{1}{\sqrt{2}\tau_e}$
4. LFM rectangular pulse $\exp\left(j2\pi\left(f_0+\frac{Bt}{2\tau}\right)\right), \lvert t\rvert \le \tau/2$ $0, \lvert t\rvert > \tau/2$	$\left(\frac{\sin\left(\pi\left(\frac{K_w t_d}{\tau}+f_d\tau\right)\left(1+\frac{\lvert t_d\rvert}{\tau}\right)\right)}{\pi\left(\frac{K_w t_d}{\tau}+f_d\tau\right)}\right)^2$	$\sqrt{2\pi}B$	$\frac{\pi\tau}{\sqrt{3}}$	$\frac{1}{\sqrt{\pi}B}$	$\frac{1}{\sqrt{3}\tau}$	$\frac{1}{B}$	$\frac{1}{\tau}$
5. PSK pulse $\sum_{i=0}^{m-1} u(t-it_r, t_r) g_i \exp(j2\pi f_0 t)$ $\lvert t\rvert \le \tau/2$ $0, \quad \lvert t\rvert > \tau/2$	$\left(\frac{\sin(\pi f_d\tau)}{\pi f_d\tau}\right)^2\left(1-\frac{\lvert t_d\rvert}{\tau}\right), \lvert t_d\rvert \le \tau_n$ $\frac{1}{N}\left(\frac{\sin(\pi f_d\tau)}{\pi f_d\tau}\right)^2, \lvert t_d\rvert > \tau_n$	$\frac{3\pi}{2\tau_n}$	$\frac{\pi\tau}{\sqrt{3}}$	$(2-\sqrt{3})\tau)$	$\frac{1}{\sqrt{3}\tau}$	$\frac{7\tau}{6}$	$\frac{1}{\tau}$
6. Pulse train $\sum_{i=0}^{n} A_{pt}(t) a_{pt}(t-it_r, t_r)$ $\times \exp(j2\pi f_0 t), \quad \lvert t\rvert \le t_0/2$ $0, \quad \lvert\tau\rvert > t_0/2$	$\sum_{i=-n}^{n} \lvert\chi_p(t_d - it_r, f_d)\rvert^2$ $\times \sum_{k=-\infty}^{\infty} \lvert\chi_a(t_d, f_d - k/t_r)\rvert^2$	As for single-pulse a_{pt}	As for envelope A_{pt}	As for single-pulse a_{pt}	As for envelope A_{pt}	$\frac{(1/B_e)_p}{t_r} \times \left(\frac{1}{B_e}\right)_a$	$t_r\left(\frac{1}{t_e}\right)_p \times \left(\frac{1}{t_e}\right)_a$

Note: g_i is a series of unities, e.g., (1, –1, –1, ...) defining the binary code; τ_n is the subpulse width; A_{pt} is the pulse-train envelope; a_{pt} is the single-pulse modulation function; t_r is the pulse repetition interval; τ is the transmitted pulse width; $\lvert\chi_p(t_d, f_d)\rvert^2$ is the single-pulse ambiguity function; and $\lvert\chi_a(t_d, f_d)\rvert^2$ is the pulse-train envelope ambiguity function.

The **ideal ambiguity function** is that of the waveform with ideal resolution of targets no matter how close they are located. It consists of a single peak of infinitesimal thickness at the origin and is equal to zero everywhere else (Fig. A32). The ideal ambiguity function is an idealized notion, and can never be achieved in practice because of the fundamental properties of the ambiguity function. *SAL*

Ref.: Skolnik (1980), p. 412.

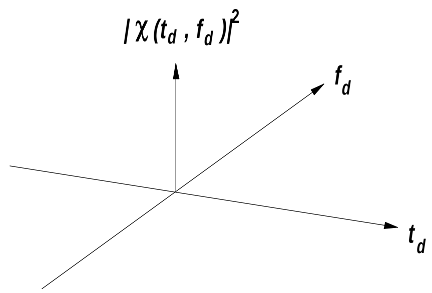


Figure A32 Ideal ambiguity function (after Skolnik, 1980, Fig. 11.7, p. 413).

The **knife-edge [ridge] ambiguity function** is that of a single pulse waveform. Its orientation is along the time-delay axis for a long pulse, along the frequency axis for a short pulse, or can be rotated by the application of linear frequency modulation (Fig. A33). *SAL*

Ref.: Skolnik (1980) p. 418.

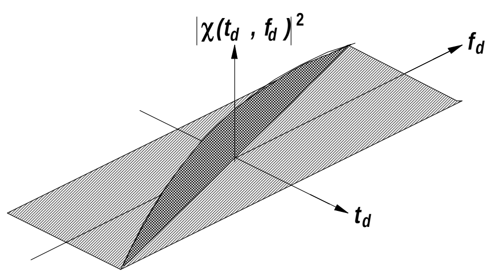


Figure A33 Knife-edge ambiguity function (after Skolnik, 1980, Fig. 11.13, p. 419).

The **thumbtack ambiguity function** is common to noiselike or pseudonoise waveforms. By increasing the bandwidth or pulse duration one can make the width of the spike narrow along the time or the frequency axis, respectively (Fig. A34). *SAL*

Ref.: Skolnik (1980), p. 418.

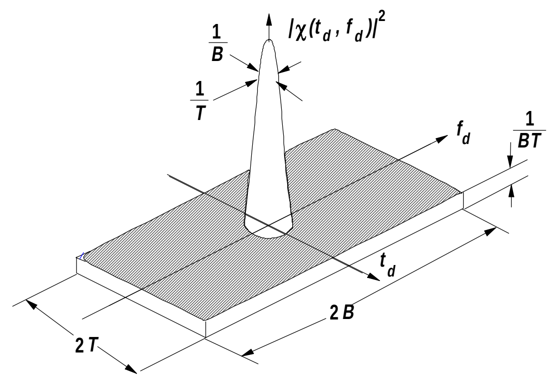


Figure A34 Thumbtack ambiguity function (after Skolnik, 1980, Fig. 11.13, p. 419).

AMPLIFIER, microwave. An amplifier is "a device that enables an input signal to control a source of power, and thus is capable of delivering at its output a reproduction or analytic modification of the essential characteristics of the signal." A microwave amplifier amplifies a microwave input signal using the energy of an external source. In radar applications there are two primary methods of classifying amplifiers: the frequency band of the signal to be amplified, and the type of basic component employed as the principal amplifying device. From the point of view of frequency, amplifiers are categorized as radio-frequency (RF) amplifiers, intermediate-frequency (IF) amplifiers, and video(-frequency) amplifiers. From the point of view of the principal amplifying device, (vacuum-)tube amplifiers and solid-state amplifiers are distinguished. The first are based on microwave tubes: backward-wave tubes, gyrotrons, klystrons, magnetrons, traveling-wave tubes, and twystrons. The second are based on solid-state components, primarily diodes and transistors. The main types of such amplifiers are Gunn diode, IMPATT diode, TRAPATT diode, tunnel diode, field-effect transistor, and transferred electron effect amplifiers. Some special kinds of amplifiers, such as difference and gain-controlled amplifiers, can be used to achieve specified characteristics of the output signals. To achieve the required power level, single amplifiers can be cascaded to form an amplifier chain.

The main characteristics of microwave amplifiers are frequency band, bandwidth, output peak and average power, gain (amplification factor), and efficiency. *SAL*

Ref.: IEEE (1993), p. 32; Fink (1982), pp. 13.60–13.70, 13.100–13.117.

An **amplifier-attenuator** is one whose gain can be controlled within given limits. This is achieved by means of a control voltage. The basic characteristics of a amplifier-attenuator are gain control range, range of the control voltage, and maximum output power delivered to the load.

In radar receivers amplifier-attenuators using tunnel diodes have gains from –30 to +20 dB, and maximum power outputs of the order of 10^{-5}W. Amplifiers with field-effect tetrodes are used to obtain higher output powers. *IAM*

Ref.: Rudenko (1971), p. 92.; Musiyachenko, V. A., Microwave control amplifier-attenuators using tunnel diodes, *Radiotekhnika*, no. 6, 1986, p. 31.

An **amplifier chain** is a system of cascaded amplifiers designed to achieve the required power level. Such chains are usually found in radar transmitters (see **TRANSMITTER, amplifier chain**).

Ref.: Skolnik (1990), p. 4.9.

An **aperiodic amplifier** is one without resonant circuits or frequency-selective elements, resulting in a wide passband. It is sometimes termed an *untuned amplifier.*

Ref.: Popov (1980), p. 40.

A **backward-wave(-tube) amplifier** uses a backward-wave tube as its basic element. There are linear-beam and crossed-field backward wave amplifiers. The linear beam backward wave amplifier operates as a synchronized generator. Stable synchronization is achieved by matching the intrinsic frequency of the tube with the frequency of excitation. The amplitron is also a backward-wave amplifier, which combines the operating principles of the crossed-field backward wave amplifier with that of a magnetron.

In linear-beam backward-wave types amplification occurs at lower levels of beam current, which is the reason why these amplifiers do not achieve high power levels. The amplifier gain and passband width vary widely over the frequency tuning range. These amplifiers are used as voltage-tunable preselectors. The amplifier gain is 3 to 20 dB.

Backward wave amplifiers other than amplitrons are seldom used in radar applications. *IAM*

Ref.: Andrushko (1981), p. 66; Skolnik (1990), pp. 4.12–4.14.

A **balanced amplifier** is "an amplifier in which there are two identical signal branches connected so as to operate in phase opposition and with input and output connections each balanced to ground." It typically consists of a parallel combination of two single-stage signal branches and two bridge circuits for input and output. The balanced circuit is used to increase the output power and reliability. Balanced amplifiers have more stable amplitude and phase characteristics and exhibit higher stability than unbalanced amplifiers. Shortcomings of balanced amplifiers are, as a rule, lower sensitivity and the necessity of using active elements with identical gain and noise figures.

In radar applications, balanced amplifiers are typically used in low-noise input circuits. The balanced amplifier is sometimes termed a *push-pull amplifier. IAM*

Ref.: IEEE (1993), p. 33; Gassanov, (1988), p. 171.; Rudenko (1971), p. 85.

A **bandpass amplifier** passes signals in fixed-frequency bands and possesses a constant gain. It is used to separate of signals that have spectral components in the frequency range of the amplifier. To achieve the desired frequency characteristics, bandpass amplifiers use coupled resonator filters or coupled, tuned tank circuits. *IAM*

Ref.: Buda (1986), p. 86; Fink (1982), p. 3.38.

A **cascode amplifier** is a two-stage tube IF amplifier connected in a common-cathode or common-grid circuit. As compared with pentodes, the use of triodes results in lower noise figures (1.3 to 1.4 dB). High amplifier stability is assured by using the triodes in a common-grid connection. A high amplifier input resistance is achieved by connecting the first stage in a common-cathode circuit.

In older radar applications cascode amplifiers were used in the first stage of intermediate frequency amplifiers. *IAM*

Ref.: Valley (1948), p. 440; Fradkin (1969), p. 55; Benson (1986), p. 14.92.

A **class-A (vacuum-tube) amplifier** is one in which "the grid bias and alternating grid voltages are such that anode current in a specific tube flows all the time."

A **class-AB (vacuum-tube) amplifier** is one in which "the grid bias and alternating grid voltages are such that anode current in a specific tube flows for appreciably more than half but less than the entire electrical cycle."

A **class-B (vacuum-tube) amplifier** is one in which "the grid bias is approximately equal to the cutoff value so that the anode current in a specific tube flows for approximately one half of each cycle when alternating grid voltage is applied."

A **class-C (vacuum-tube) amplifier** is one "in which the grid bias is appreciably greater than the cutoff value so that the anode current in a specific tube is zero when no alternating current is applied, and so that anode current flows for appreciably less than one half of each cycle when alternating grid voltage is applied." *SAL*

Ref.: IEEE (1993), p. 33.

corporate structure [-combined] amplifier (see **power amplifier**).

crossed-field amplifier (see **CROSSED-FIELD AMPLIFIER**).

A **difference [differential] amplifier** produces an output signal proportional to the difference between input signals. It can also be realized as a nonlinear integrated circuit whose function is to output logic "1" or logic "0," depending on which of two inputs is more positive than the other. One of the applications is in EW crystal detector receivers. *SAL*

Ref.: IEEE (1993), p. 344; Wiegand (1991), p. 140.

A **diode amplifier** uses a diode as its active element. A circulator is customarily required to separate the input and output signals. Based on the type of active element, diode amplifiers are classified into tunnel diode, IMPATT-TRAPATT diode, or Gunn diode amplifiers. The fundamental parameters of diode amplifiers are gain, output power, noise figure, bandwidth, and efficiency. Another important characteristic of the diode amplifier is its stability. Diode amplifiers cover practically the entire microwave band from 1 to 300 GHz. They are used in various radar subsystems, from input circuits (tunnel diode amplifiers) to power amplifiers (IMPATT-TRAPATT diode amplifiers). *IAM*

Ref.: Howes (1976).

A **doppler and range frequencies amplifier** is used in continuous-wave radars to amplify the (converted) signal at the frequency $f_d + f_r$ to a level necessary for the reliable operation of the unit converting the frequency shift to a current or voltage level shift (where f_d is the doppler shift and f_r is the range frequency increment). The amplifier parameters depend on the type of radar. In radars with frequency modulation with $f_r > f_d$, the amplifier has a linear frequency characteristic with a gain slope of 12 dB/octave (6 dB/octave for operation against extended targets) over its range of operating frequencies. In doppler radars and in radars having frequency modulation with $f_r < f_d$, the amplifier has a flat frequency characteristic and a nonlinear (usually logarithmic) amplitude characteristic. To compensate for the effects of fluctuation of the target return and the effects related to wave propagation, automatic gain control is used. *IAM*

Ref.: Vinitskiy (1961), p. 267.

Amplifier efficiency is the ratio of the power P_o delivered to the output load to the power P_s accepted from the amplifier power source:

$$\eta = P_o / P_s$$

Ref.: IEEE (1993), p. 407.

The **extended interaction amplifier (EIA)** is one based on the klystron as the principal amplifying device (and is sometimes termed the extended interaction klystron amplifier, EIKA). Typical EIA characteristics when it is used at millimeter waves are shown in Table A2. *SAL*

Ref.: Currie (1987), p. 455.

Table A2
Typical EIA Tube Characteristics

Model	Frequency (GHz)	Bandwidth (MHz)	Power (W)	Gain (dB)	Duty (%)
VKE2406	50 - 80	-	100	-	10
VKB2400T	94 - 96	200	1000	30	10

A **field-effect tetrode amplifier** uses field-effect tetrodes (dual-gate FETs) as the active elements. These devices differ from field-effect transistors by the presence of a supplementary gate. The properties of the field-effect tetrode are equivalent to those of two series-connected, single-gate field-effect transistors. For low-noise operation, the tetrode utilizes a supplementary noise-suppression electrode. Field-effect tetrode amplifiers exploit the higher gain of the tetrode over the field-effect transistor, as well as the ability to control the gain by changing the level of dc voltage on one of the gates. The depth of modulation of the gain can exceed 40 dB at frequencies up to 10 GHz. Corresponding circuits for tetrode amplifiers do not differ significantly from those of transistor amplifiers. *IAM*

Ref.: *Foreign Radioelectronics*, no. 6, 1982, p. 80.; Liechti, C. A., *IEEE Trans MTT-23*, 1975, no. 6; Fink (1982), p. 9.72.

A **field-effect transistor (FET) amplifier** is one based on the FET as the principal amplifying device. The current technology typically employs the monolithic FET power amplifier operating at a frequency of 1 to 20 GHz with output power up to 1W. *SAL*

Ref.: Ostroff (1985), pp. 47–49; Schleher (1986), pp. 505–511.

An **amplifier figure of merit** is the product of the amplifier gain and its bandwidth. The figure of merit is sometimes called the *gain-bandwidth product*. Figure of merit is one of the basic parameters of microwave amplifiers. *IAM*

Ref.: Druzhinin (1967), p. 377; Fink (1982), p. 13.3.

Amplifier gain is the ratio of the current, I_o, voltage, V_o, or power, P_o, at the amplifier output to the corresponding parameter at the input (I_i, V_i, P_i). The gain, G, is unambiguous when expressed in decibels:

$$G_{\text{dB}} = 20\log\frac{I_o}{I_i} = 20\log\frac{V_o}{V_i} = 10\log\frac{P_o}{P_i}$$

When the amplifier and its load contain reactances, amplifier gain is frequency-dependent. Typically amplifier gain is stated for the middle frequencies in the specified band, where it is not sensitive to the frequency. Gain is often called available gain, and when specified as a voltage ratio it is called the amplification factor. *SAL*

Ref.: IEEE (1988), p. 35; Mamonkin (1977), p. 11; Fink (1982). p. 13.3.

A **gain-controlled amplifier** is one with variable gain to maintain the echo level within the available dynamic range of the receiver under different situations of radar operation (differences in target RCS, meteorological conditions, range, etc.). The primary techniques used to control echo level in radar receivers are sensitivity time control (STC) and automatic gain control (AGC). *SAL*

Ref.: Skolnik (1990), p. 3.17.

A **Gunn diode amplifier** uses a Gunn diode as its active element. The diode is operated in a stable mode, which is determined by the external circuit and also by the properties of fabrication and doping (impurity levels) of the semiconductor diode. A typical amplifier in the band from 9 to 11 GHz has a gain of 8 dB and a noise figure of 9 to 10 dB. The highest output power is about 1W, and the efficiency can reach several percent. The highest operating frequency of a Gunn diode amplifier can approach 80 GHz. The gain and output power can be increased by connecting several amplifiers in series by circulators and bridge circuits. *IAM*

Ref.: Howes (1976), p. 304; Fink (1982), pp. 13.105–110.

A **gyrotron amplifier** is intended for power amplification at microwave frequencies. The family of gyrotron amplifiers also includes devices (gyro-klystrons and gyro-TWTs) that are hybrids of gyrotrons, klystrons, and TWTs. Typical basic parameters of gyro-klystrons are gain ≈ 40 dB, bandwidth ≈ 1% at 28 GHz. Gyro-TWTs are available for frequencies up to 94 GHz. At that frequency the gain is ≈ 30 dB with an output power ≈ 20 KW over a 2% bandwidth. Gyrotron amplifiers are under active development. *IAM*

Ref.: Curry (1987), pp. 466–470.; Granatstein, V. L., and Park, S. Y., "Report at International Electron Devices Meeting," Dec. 1983, Washington, DC.

An **intermediate-frequency (IF) amplifier** is one following the mixer of a superheterodyne receiver, whose function is to amplify received signals at intermediate frequency. Typically, IF amplifiers have more stages in the amplifier chain than RF amplifiers, are more stable, and have greater gain per stage. Thus, IF amplifiers provide most of the amplification of the received signal, which ensures the proper operation of receiver circuits, primarily of the detector. In radar applications, transistors and integrated circuits are widely used as IF amplifiers.

The basic amplifier parameters are intermediate frequency, gain, passband width, passband shape, and selectivity. The passband shape defines the frequency response levels relative to the maximum gain in the passband. The selective properties of the amplifier are determined by the choice of the frequency-selective filters. At the customary frequencies of IF amplifiers, 10 to 120 GHz, these are *L-C* filters, quartz crystals, solid-state piezoelectric, and supersonic surface wave filters. *SAL, IAM*

Ref.: Sokolov (1984), p. 249; Popov (1980), p. 448; Leonov (1988), p. 63; Skolnik (1990), pp. 3.17–3.32.

A **klystron amplifier** is one intended for the amplification of energy at microwave frequencies. For amplifier applications, a klystron with multiple resonators and having one or several electron beams is generally used. The klystron amplifier gain ranges from 30 to 70 dB, bandwidth 1 to 8%, efficiency up to 30%. Output circuit resonators with extended distributed interaction regions provide bandwidths of 10 to 15% with efficiencies of up to 65%. Klystrons are used in radars to achieve high microwave output power levels. Power klystrons have peak power levels exceeding 10 MW (see also **KLYSTRON**). *IAM*

Ref.: Chaikov (1974), p. 5.; Skolnik (1990), p. 4.14.

A **lin(ear)-log(arithmic) amplifier** is an automatic gain control amplifier that operates in a linear manner for low-amplitude input signals and in a logarithmic manner for high-amplitude input signals. *SAL*

Ref.: Johnston (1979), p. 62.

A **log(arithmic) amplifier** is one whose output voltage is proportional to the logarithm of its input voltage. The logarithmic amplitude characteristic is typically obtained by shunting the loads of amplifier stages, by nonlinear circuits, or by successive summing of the output voltages of several stages. The basic properties of the logarithmic amplifier are the compression of the dynamic range of the amplified signal, and the inverse relationship of the fluctuation of the output signal to the intensity of the input signal. Logarithmic amplifiers are widely used in radar applications (e.g., as angle discriminators of monopulse radars, and in circuits for reducing interference from weather returns). (See CFAR, log-FTC**).**

Ref.: Leonov (1970), p. 88.; Finkel'shteyn (1983), p. 349; Hughes (1986); Skolnik (1990), pp. 3.25–3.30.

A **low-noise amplifier** is one having a low noise temperature. It usually has high gain, wide bandwidth, and a large dynamic range. The most frequently used low-noise amplifier for connecting signal sources and loads are reflection and balanced amplifiers. Typical values of noise temperature for various amplifier types in the frequency range from 1 to 10 GHz are shown in Table A3.

In radar applications, low-noise amplifiers are used as the first amplifier stage in receivers and are often mounted near the antenna feed. Low-noise amplifiers are important elements of transmit-receive array modules. *IAM*

Ref.: Skolnik (1980), pp. 351–353; Gassanov (1988), pp. 17, 156.

Table A3
Noise Performance of Amplifiers

Low-Noise Amplifier Component	Noise Temperature (K)
Tunnel diode	200 – 800
Bipolar transistor	100 – 700
Field-effect transistor	70 – 100
Parametric (uncooled)	30 – 50
Field-effect transistor (cooled)	13 – 30
Parametric (cooled)	10 – 20
Quantum parametric (cooled)	4 – 5

A magnetron amplifier is one based on a crossed-field device. It represents a microwave device wherein the amplification of an electromagnetic wave, propagating in a slow-wave structure, is realized by an extended interaction with the electronic beam moving within crossed electric and magnetic fields. Magnetron amplifier types are built using either linear or circular (annular) electrodes (Figs. A35, A36). Most magnetron amplifiers use the circular geometry, which allows reduction of the overall dimensions of the device and simplifies the construction of the magnetic circuit. Magnetron amplifiers operate both in CW and pulse modes. The range of operating frequencies generally extends from 0.4 to 17 GHz, with gains to 20 dB, and peak power outputs up to 10 MW (see also **CROSSED-FIELD AMPLIFIER**). *IAM*

Ref.: Fink (1982), p. 13.117; Leonov (1988), p. 49.

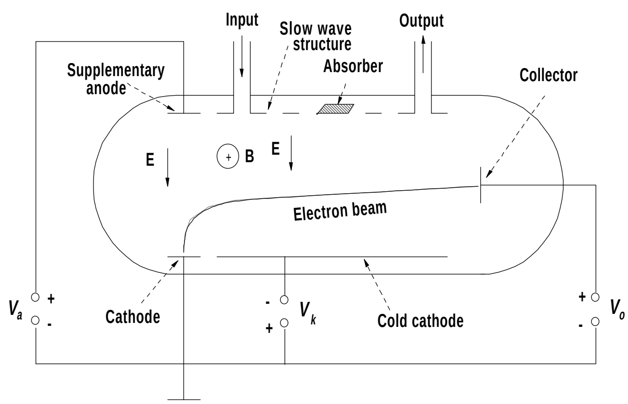


Figure A35 Linear magnetron amplifier.

A **master-oscillator power amplifier (MOPA)** is a transmitter amplifier chain that consists of a stable low-power oscillator followed by a power amplifier. *SAL*

Ref.: Skolnik (1980), p. 106; Skolnik (1990), pp. 14.8–14.11.

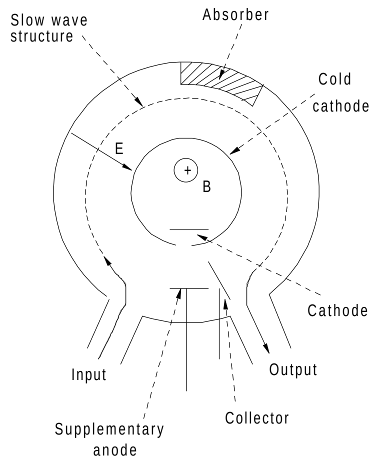


Figure A36 Annular magnetron amplifier.

A **negative resistance amplifier** uses the effects of negative resistance to achieve very low noise and good amplification. Typically used in ultrasensitive radar receivers, such as those based on parametric amplifiers or masers (see also **RESISTANCE, negative**). *SAL*

Ref.: Sauvageot (1992), p. 17.

An **operational amplifier (op amp)** is a dc amplifier with high gain and negative feedback that can perform operations such as summing, differentiating, and integrating, with input analog signals. It can be used in video circuits of radar receivers before the received signal is converted to digital form. *SAL*

Ref.: IEEE (1993), p. 889; Wiegand (1991), p. 151; Leonov (1988), p. 49.

A **parametric amplifier** is a device wherein the amplification of the input signal is done by a microwave power source (pump generator) that periodically changes the reactive parameter of the circuit. The latter is represented by the barrier capacity of a parametric diode, or by the variable inductance of a ferrite, placed in a high-frequency magnetic field created by the pump generator. Uncooled parametric amplifiers usually have noise temperatures of 50 to 80K. Amplifiers have high gains (up to 20 dB per stage) but relatively narrow bandwidths (up to 10%). A significant reduction in the noise temperature can be achieved by cooling the amplifier (10 to 20K at a temperature of 20K, and 4 to 10K at 4.2K) at frequencies from 1 to 10 GHz. Some electron-beam parametric amplifiers are based on the periodic change of the reactive resistance of the extended resonator as the bunched electrons in the beam pass through it.

In radar receivers, parametric amplifiers are used as RF amplifiers. *IAM*

Ref.: Gassanov (1988), p. 162; Druzhinin (1967), p. 357; Zhelerkovskiy (1971), p. 10; Fink (1982), pp. 13.64–13.66.

A **power amplifier** is intended for the delivery of high output power. The power amplifier is characterized by its frequency, the level of power output, efficiency, amplification factor, level of regulation, and other parameters common to different amplifiers. The active elements in power amplifiers can be either vacuum tubes or semiconductor devices. Table A4 shows typical values of output power levels, in watts, for different types of power amplifiers.

Table A4
Range of Amplifier Output Powers (W)

Type of power amplifier	Operating frequency (GHz)		
	1.0	10	100
Vacuum tubes:			
Klystron	$5{\times}10^5$	10^4	
TWT	$3{\times}10^4$	$3{\times}10^3$	10^2
Semiconductors:			
Avalanche transit time diode		55	30
Gunn diode			15
Bipolar transistor	50		
Schottky field-effect transistor		45	10

Power amplifiers are used in intermediate and output stages of radar transmitters and also in modules of antenna arrays.

There are two main configurations of power amplifiers: the corporate-combined amplifier and the space-combined amplifier (Fig. A37). The first is typically used in solid-state transmitters feeding conventional antennas, and the second in modular phased arrays. *IAM*

Ref.: Gassanov (1988), p. 174; Gilmour (1986), Chaps. 8-14; Ostroff (1985); Skolnik (1990), pp. 4.9–4.25, Ch. 5.

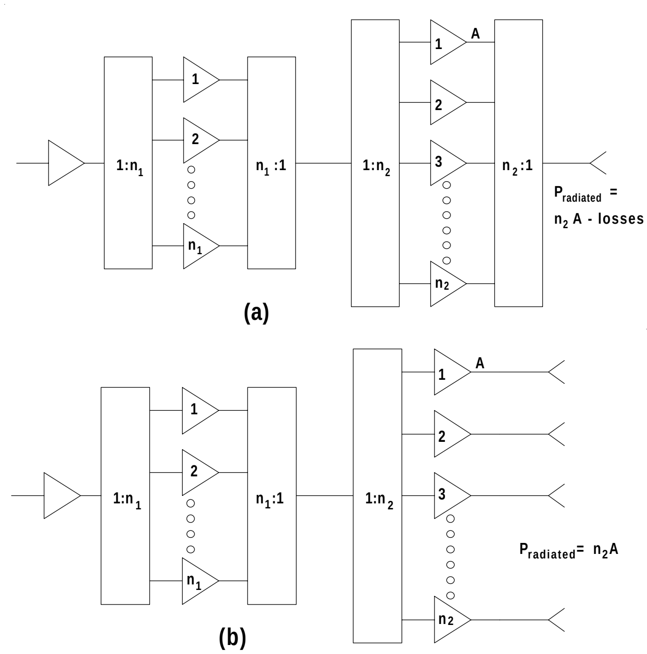


Figure A37 Block diagram of (a) corporate-combined power amplifier and (b) space-combined power amplifier (after Skolnik, 1990, Fig. 5.6, p. 5.12).

A **pulse amplifier** is intended for the amplification of pulse waveforms. One distinguishes between linear and nonlinear amplifiers. Linear amplifiers reproduce the form of the pulse with a given accuracy. Nonlinear amplifiers intentionally produce a distortion of the pulse by changing the amplitude and/or duration of the pulse. To compensate for undesired pulse distortion, a flat frequency response of the gain over the low, middle, and high frequencies is achieved by feedback and compensation circuits. The most important parameter of the linear pulse amplifier, the level of amplifier distortion, is determined by the amplifier transfer function. *IAM*

Ref.: Agakhanyan (1970), pp.13–15.

push-pull amplifier (see **balanced amplifier**).

A **(quantum) paramagnetic amplifier** is a microwave amplifier that uses the energy transfer from excited ions of the active element (paramagnetic substance) to electromagnetic waves by transitions of electrons from higher to lower energy levels. The energy levels of the paramagnetic substance depend on the external magnetic field, and a pump generator creates the population inversion of the levels. Quantum paramagnetic amplifiers are classified as resonant and traveling-wave types. In amplifiers of the traveling-wave type, the active element is distributed in a waveguide, within which propagates the wave to be amplified. Because a cavity resonator is absent, the traveling-wave amplifier is more broadband. The highest frequencies attained by the quantum paramagnetic amplifier reach 100 GHz. Typical amplifier parameters for the frequency range from 1 to 10 GHz are shown in Table A5.

Table A5
Typical Quantum Paramagnetic Amplifier Parameters

Parameter	Amplifier type	
	Resonant	Traveling-wave
Gain (dB)	15–20	20–30
Bandwidth	< 1%	3.5%
Noise temperature (K)	20–100	5–10

Traveling-wave type amplifiers can have tuning ranges of up to 20%. Quantum paramagnetic amplifiers are used in high-sensitivity input circuits of sensitive radar receivers. *IAM*

Ref.: Andrushko (1981), p. 155.

A **radio-frequency (RF) amplifier** is intended for the amplification of RF input signals from the antenna, maintaining the lowest noise figure. It is also used for the suppression of interference in the auxiliary receive channels. The basic amplifier parameters are noise figure, power gain, and efficiency. The type of amplifier to be used depends on the operating bandwidth. At microwave frequencies quantum-paramagnetic, parametric, transistor, TWT, and tunnel diode amplifiers are typically used. *IAM*

Ref.: Rudenko (1971), p. 8; Fradkin (1969), p. 22; Fink (1982), pp. 13.34–13.43.

A **resistive amplifier** is a transistor or vacuum tube amplifier where the load circuit, to obtain the output signal, uses an active resistance. It is characterized mainly by a flat amplification across a wide bandwidth, reaching several megahertz. Resistive amplifiers are simple in construction and are used as the components of pulse amplifiers. *IAM*

Ref.: Popov (1980), p. 369.

A **resonant amplifier** is one using resonant circuits. An example of the resonant amplifier is the bandpass amplifier. Among the basic characteristics of the resonant amplifier are the resonant amplification factor (gain at the resonant frequency), selectivity, noise figure, signal distortion levels, and stability. It is employed as an amplifier at radio frequencies (RF) and intermediate frequencies (IF). Sometimes this type of amplifier is termed the *tuned amplifier*. *IAM*

Ref.: Fink (1982), pp. 3.35, 13.3; Buda (1986), p. 59.

A **single-port amplifier** is a reflection-type amplifier based on a single-port structure, such as shown in Fig. A38. It requires a circulator for coupling in and out, and its performance sometimes is limited by this additional component. Examples of single-port amplifiers are Gunn diode amplifiers and IMPATT diode amplifiers. *SAL*

Ref.: Currie (1987), p. 415.

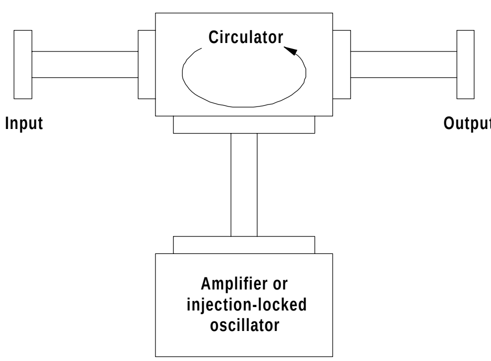


Figure A38 Single-port amplifier (after Currie, 1987, Fig. 8.8, p. 406).

A **solid-state amplifier** is one entirely based on solid-state components (semiconductor devices and integrated circuits). The primary types of solid-state amplifiers used in radar applications are transistor amplifiers, Gunn diode amplifiers, diode amplifiers, and transferred electron effect device (TED) amplifiers. Typically, they are arranged in corporate-combined or space-combined structures (see **power amplifiers**). *SAL*

Ref.: Fink (1982), Ch. 13; Leonov (1988), p. 51.

space-combined amplifier (see power amplifier).

A **summing amplifier** is one whose output signal, V_{out}, is equal to the sum of its input signals, V_i, $i = 1, 2, \ldots, n$:

$$V_{out} = \sum_{i=1}^{n} a_i V_i$$

The weighting factors a_i are the amplification factors for the individual input signals and are selected in accordance with the required rule of summation. Summing amplifiers are usually employed as operational amplifiers. The weighting factors are determined by passive elements in the input and feedback circuits. *IAM*

Ref.: Korn (1952), pp. 11–14, 121–124; Jordan (1985), p. 20.46; Mamonkin (1977), p. 278.

A **transferred electron effect device (TED) amplifier** is one based on a TED as the principal amplifying device. The TED amplifier is a form of single-port amplifier, so it uses circulator or hybrid techniques similar to IMPATT circuits. Typical values of bandwidth, gain, and efficiency are 35%, 20 dB, and 10%, respectively. *SAL*

Ref.: Fink (1982), p. 13.105.

A **transistor amplifier** uses transistors as the basic active elements. The most widely used types in the microwave band are bipolar transistors and field-effect transistors.

Transistor amplifiers are used in receiver front ends, preamplifiers, and power amplifiers, solid-state transmitter chains, and other radar subsystems and devices. Low-noise amplifiers at a receiver front end are primarily based on Schottky-barrier field-effect transistors and have the following characteristics: operating band of 1 to 60 GHz, gain of 15 to 5 dB/stage, and the noise figure of 0.5 to 8 dB. The passband is from several percent up to a few octaves. Power transistor amplifiers have the following characteristics per stage: bipolar-transistor-based amplifiers: operating frequency of up to 7 or 8 GHz, output power is about 5 W, gain is 6-8 dB, and efficiency at higher frequencies is about 30%; field-effect-transistor-based amplifiers: operating frequency of up to 80 or 90 GHz, output power of up to 10 W at 10 GHz frequency, and efficiency of about 20%. *IAM*

Ref.: Gassanov (1988), pp. 168, 196; Fink (1982), pp. 13.106–13.113.

A **traveling-wave tube (TWT) amplifier** is one based on a traveling-wave tube as the basic active component. Depending on the TWT, linear-beam TWT amplifiers and crossed-field TWT amplifiers are distinguished (the latter is termed the *magnetron amplifier*). Linear-beam TWT amplifiers are classified as low-power amplifiers (up to 1W), power amplifiers (more than 100W), and superpower amplifiers (more than 100 kW). According to the mode of operation, they are classified as continuous-wave and pulsed amplifiers.

Low-power amplifiers operate in the band of 0.25 to 110 GHz, amplification band is 30 to 70%, gain is 25 to 35 dB, and the noise figure is 6 to 16 dB, for frequencies of 1 to 20 GHz and 16 to 20 dB for frequencies of 20 to 100 Ghz. They are typically used in the broadband receivers front ends, where the requirements to noise levels and size are not very stringent.

Power and superpower TWT amplifiers are used in radar transmitters. Their performance is: efficiency of up to 40%, pulse power of up to 10 MW, gain of 30 to 70 dB, and passband of 10 to 15%. These devices require a high-voltage supply (up to 100 kW). *IAM*

Ref.: Skolnik (1990), pp. 4.15–4.19; Andrushko (1980), p. 58; Gilmour (1994).

tuned amplifier (see **resonant amplifier**).

A **tunnel-diode amplifier** is one using a tunnel diode as the basic component, and frequency-selective circuits or directional filters. Typically, circulators are used at the input of such amplifiers. The operating frequency band is about 0.25 to 110 GHz, gain is 10 to 20 dB, the relative passband is 3.5 to 60%, and the typical noise figure is 3 to 7 dB. Along with the common structures of amplifiers based on coaxial, waveguide, and strip lines, the integrated structures based on thin-film technology are also used. The use of these amplifiers is expedient in devices where size and weight are the basic constraints, and the requirements to noise levels, dynamic range, temperature range, and input power are not high. In radars, they are typically used as the amplifiers in phased arrays modules, and in the second stage of amplification in the receiver after parametric amplifiers. *IAM*

Ref.: Rudenko (1971), pp. 14–16, 92–122; Fink (1982), pp. 13.60–13.64.

The **twystron amplifier** is one based on the twystron tube as the principal amplifying device. Twystron amplifier bandwidth is from 6 to 15%, peak power is in the megawatt range and efficiency is about 30%. Typical twystron amplifier characteristics are given in Table A6. *SAL*

Ref.: Brookner (1977), pp. 314, 315; Skolnik (1990), p. 4.17.

Table A6
Typical Twystron Tube Characteristics

Tube type	VA-45	VA-146	VA-915
Frequency band (GHz)	2.7–2.9, 2.9–3.1, 3.0–3.2	5.4–5.9	3.4–3.6
Peak power (MW)	3.5	4.0	7.0
Average power (kW)	7.0	10.0	28.0
Pulse width (ms)	10.0	20.0	40.0
Efficiency (%)	35.0	30.0	30.0

untuned amplifier (see **aperiodic amplifier**).

A **vacuum-tube amplifier** is one in which a vacuum tube is the principal amplifying element. The main amplifying tubes used in modern radar applications are radio-frequency devices (see **backward-wave tube amplifier; crossed-field amplifier; klystron amplifier; magnetron amplifier; traveling-wave tube amplifier; twystron amplifier; power amplifier).** *SAL*

A **video(-frequency) amplifier** is one that has an objective to amplify the video pulse up to a level sufficient to ensure the proper operation of displays. Typically, a video amplifier consists of several stages: video pulses amplifier, video pulse amplitude limiter, and cathode follower for load matching. Logarithmic amplifiers are widely used in such chains to limit the amplitude of the pulses. The main features of a video amplifier are broadbandness, large dynamic range, and efficiency. *IAM*

Ref.: VanVoorhis (1948), pp. 515–544; Fradkin (1969), p. 62.

The **AMPLITRON** is a reentrant beam, continuous cathode, crossed-field amplifier in which the electron current moves through crossed electric and magnetic fields and interacts with a back harmonic wave. It is distinguished by the combination of an open slow-wave circuit with an electron current closed in a ring (see **CROSSED-FIELD AMPLIFIER**).

An amplitron is similar in construction to a multicavity magnetron, and differs only in the open-circuit multicavity anode block and the existence of an input port (Fig. A39). Typical gain is 6 to 15 dB, bandwidth is 8 to 15%, and efficiency is 40 to 80%. Physically, it looks like a magnetron oscillator, but it has both input and output ports. This type of tube was invented in 1953 by W. C. Brown. *SAL*

Ref.: Brookner (1988), p. 263; Zherebtsov (1989), p. 342.

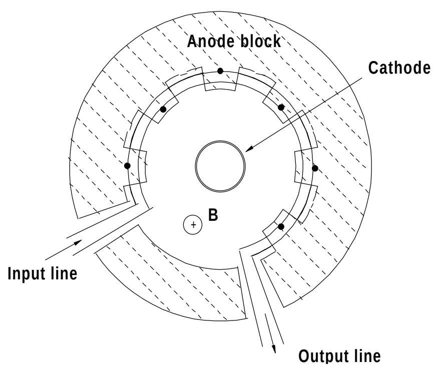


Figure A39 Amplitron amplifier.

AMPLITUDE is the maximum value of a quantity that varies periodically or as any other function. "Amplitude of a sinusoidal electromagnetic wave is the maximum value of a field quantity in space or time." For the simple sine wave:

$$A\sin(\omega t + \theta)$$

the amplitude is positive real A, where A, ω, and θ are constants. In this case, amplitude is synonymous with maximum or peak value. *SAL*

Ref.: IEEE (1993), p. 34

ANALYZER

The **Fourier analyzer** is a processor performing Fourier transforms in analog form. Typically, chirp surface-acoustic-wave (SAW) transmission lines and charge-coupled devices (CCDs) can be used. In many cases a digital fast Fourier transform (FFT) technique is more expedient in radar applications, as it is more flexible and provides greater accuracy. *SAL*

Ref.: Brookner (1977), pp. 186, 315.

A **frequency analyzer** is (1) a device for determining the target range in a frequency-modulated, continuous-wave radar, (essentially a spectrum analyzer), or (2) a device for determining characteristics of an intercepted radar waveform, including its carrier frequency and modulations imposed thereon. *IAM*

Ref.: Vasin (1977), p. 26; Wiley (1993), pp. 199–236.

A **spectrum analyzer** is a device that determines the amplitude of the frequency components in a spectrum. A basic analyzer is a harmonic resonator, which may be an electrical resonator or a bandpass filter with constant parameters, a piezoelectric or other electromechanical resonator, or other similar device. Depending on its principle of operation, a spectrum analyzer may be categorized as a simultaneous (full-band) analyzer, a sequential analyzer with a tuned resonator, or a sequential analyzer with a tuned oscillator. When analyzing wideband signals, it is common to use a sequential analyzer consisting of a mixer, oscillator, resonator, and an indicator. Various portions of the spectrum being analyzed are successively presented within the passband of the resonator.

If an optical device is used to determine the amplitude (and sometimes phase) of the frequency components of a spectrum, the analyzer is termed an *optical spectrum analyzer*. Its operation is based on the property of an optical lens by which it transforms the amplitude and phase distribution of an optical signal into a spatial spectrum in the focal plane. One advantage of an optical spectrum analyzer is that the incident signal is processed in real time. The fundamental parameter of the analyzer – its resolution – depends upon the frequency of the signal and the characteristics of the optical system. Optical spectrum analyzers are employed in radars using optical signal processing. In such a system the radio frequency signal is first transformed into an optical signal with the help of an optical modulator, which is usually an ultrasonic delay line.

The main parameters of a spectrum analyzer are its resolution and analysis period. Spectrum analyzers are used in radars with continuous frequency modulation to obtain range coverage of an observation region. *IAM*

Ref.: Supryaga (1974), p. 51; Vinitskiy (1961), p. 318; Scheer (1993), pp. 346–367; Finkel'shteyn (1983), p. 459; Zmuda (1994), pp. 435–442.

ANECHOIC CHAMBER. An anechoic chamber is a closed, screened premise inside which the conditions of free-space propagation are simulated by using special methods and equipment. Literally, it is a room "without echoes." For radar testing, an anechoic chamber is a room lined with a material designed to absorb RF energy without reflection so that transmitted and received signal measurements will accurately represent the true characteristics of the unit under test without interference from spurious sources. Most anechoic chambers designed for radar testing are lined with broadband geometric transition absorbers, which are wedges or pyramidal shapes of carbon-loaded synthetic sponge rubber.

Typically, anechoic chambers have rectangular or horn-type shapes with the dimensions from few meters to tens of meters. Usually there are two sections: a room for equipment and a measurement (test) section.

The main characteristic of the chamber is an anechoic coefficient, which is defined as the ratio of spurious scattered power density at the specified point to the transmitted power density. The value of the anechoic coefficient is determined by the quality of the radar-absorbing material, the shape of the chamber, and the place of measurement. The admissible value of the anechoic coefficient is defined by the antenna sidelobe levels and the required measurement accuracy. Typically, this value is about –40 to –60 dB.

The basic underlying data to design the measurement section of anechoic chamber is the frequency band, assumed characteristics of the objects to be tested and their dimensions, absorbing materials performance, and cost constraints. Typically, anechoic chambers are used to measure antenna performance and for RCS measurements. Recently, radar holography methods came into use to measure fine structure of radar target backscattering. *PCH, IAM*

Ref.: Mayzel's (1972), p. 109; Strakhov (1985), p. 102; Tuchkov (1985), p. 149; Van Nostrand (1983), p.154; Fink (1982), p. 6.28.

ANGEL (ECHO). An angel echo is "a radar echo caused by meteorological conditions, such as clouds, atmospheric inhomogeneities, lightning, or by birds or insects." There are two general classes of angel echoes: dot angels arising from birds and insects that are point targets, and distributed angels arising from inhomogeneities of the refractive index of the atmosphere. The degrading effects of dot and distributed angels depend on the radar cross section of the source. Birds and insects, especially in large concentrations, can also appear as distributed targets and have a degrading effect on radar operation. Other primary angel sources are clear-air turbulence, boundary surfaces between differentially moistened surface air masses over adjacent cold and warm water, mineral and organic particles carried into the air by heavy winds and thunderstorms, smoke particles and debris caused by forest and dump fires, and so forth (see also **CLUTTER**). *SAL*

Ref.: IEEE (1993), p. 38; Skolnik (1980), pp. 508–512.

ANGLE

Aspect angle is the angle between the radar line of sight and the longitudinal axis of a target (Fig. A40).

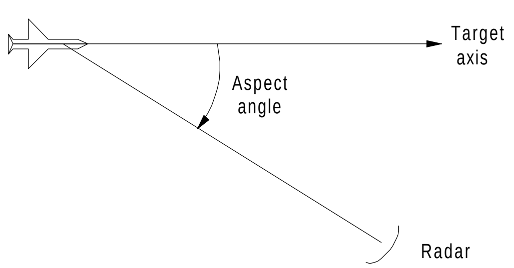


Figure A40 Aspect angle defined.

Azimuth angle is "the angle between a horizontal reference direction (usually north) and the horizontal projection of the direction of interest, usually measured clockwise" (see Fig. A41).

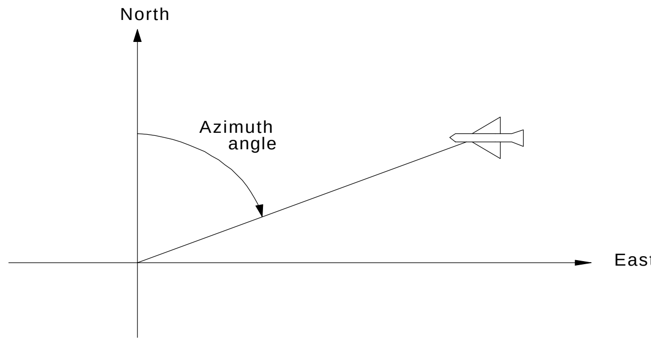


Figure A41 Azimuth angle, measured from north.

Bearing angle is "the angle in the horizontal plane between a reference line and the horizontal projection of the line joining two points." It is usually expressed in degrees measured clockwise form the reference. Relative bearing is to some arbitrary reference: absolute bearing is to North. *AIL*

Ref.: Popov (1980), p. 280; IEEE (1993), p. 102; ITT (1975), p. 32.8.

Bistatic angle is the angle between the lines of sight from the transmitter and to the receiver of a bistatic radar system (Fig. A42). When the angle approaches 180°, the system operates in the forward-scattering mode. *DKB*

Ref.: Willis (1991), p. 2.

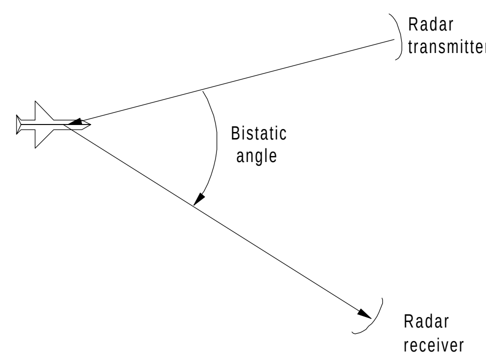


Figure A42 Bistatic angle defined.

The **Bragg angle** is the angle between the normal and diffracted laser beam in the acousto-optic Bragg-cell receiver, or between the mean sea surface and the direction of enhanced backscatter caused by a regular pattern of waves or ripples. The usual notation is α_B. *SAL*

Ref.: Long (1983), p. 81; Neri (1991), p. 297; Zmuda (1994), p. 417

The **Brewster angle** is the grazing angle from the interface between two media at which the reflection coefficient of light is zero. For radar waves, the reflection coefficient may not go to zero. The *pseudo-Brewster angle* is then defined as that corresponding to the minimum reflection coefficient, Γ, of the surface:

$$\Gamma = \rho \exp(j\phi)$$

where ρ, the magnitude of Γ, describes the change in amplitude and the argument ϕ describes the phase shift on reflection. The angle for which $\phi = \pi/2$ and ρ is at a minimum is the pseudo-Brewster angle, given by

$$\sin\psi_B = \frac{1}{\sqrt{1+\varepsilon_r}}$$

SAL

Ref.: Kerr (1951), p. 399; Skolnik (1980), p. 444; Meeks (1982), p. 15.

angle deception (see **ELECTRONIC COUNTERMEASURES**).

Depression angle, measured in the elevation plane, is the amount by which the antenna main beam is depressed below the radar's local horizon.

Elevation angle is the angle between horizontal reference plane and line of sight in the direction of interest, measured upward (see Fig. A43).

Ref.: IEEE (1993), p. 431.

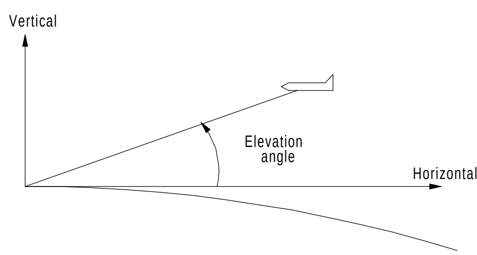


Figure A43 Elevation angle to radar line of sight.

The **Faraday rotation angle** is the angle of the polarization plane rotation of a wave with linearly polarization taking place due to the Faraday effect. The usual notation is θ_R. For the ionosphere

$$\theta_R = 81\pi \times 10^{-18} N_t f_m \cos(\theta / c f_0^{\,2})$$

where N_t is the number of electrons in a vertical column through the ionosphere having 1 m^2 cross section, f_m is the earth's gyromagnetic frequency in hertz, θ is the angle between the earth's magnetic field and the direction of propagation, c is the velocity of propagation of light, and f_0 is the frequency of the signal.

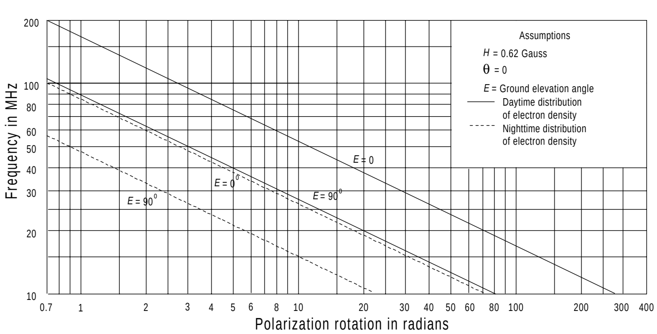


Figure A44 Polarization rotation as a function of frequency for a target at 1000-km altitude, two-way path, and longitudinal propagation (after Berkowitz, 1965, Fig. 1.23, p. 363).

The total rotation for two-way paths to 1000-km altitude is shown in Fig. A44 as a function of frequency for different elevation angles. *SAL, DKB*

Ref.: Berkowitz (1965), pp. 360–364; Morchin (1993), p. 328.

Grazing angle is the angle measured in the vertical plane between a ray and a reflecting surface (Fig. A45) It is the complement of the incidence angle, and, for short ranges at which the flat-earth approximation is valid, is equal to the depression angle from the antenna. *DKB, SAL*

Ref.: Barton (1964), p. 95; (1991), p. A-8.

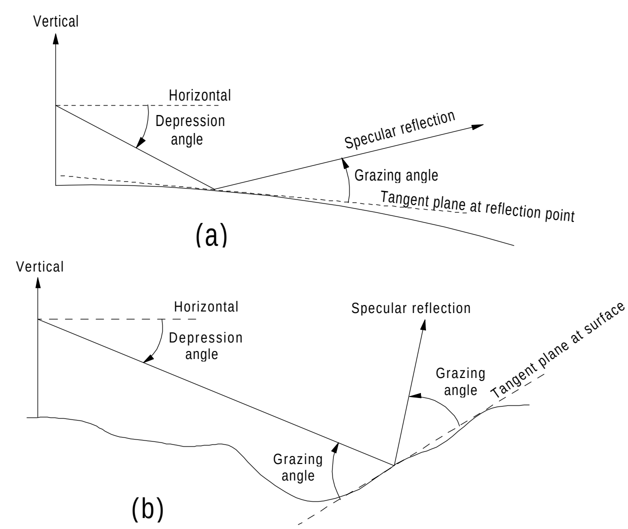


Figure A45 Grazing angle for (a) smooth surface and (b) undulating surface.

The **incidence angle** is the complement of the grazing angle.

pseudo-Brewster angle (see **Brewster angle**).

The **Rayleigh critical angle** is the grazing angle, ψ_c, in the equation of Rayleigh roughness criterion below which a surface having given height deviations appears to be "smooth." At this angle, the specular scattering coefficient is $1/e$, indicating transition from specular to diffuse scattering. For an rms surface height deviation σ_h, the angle is

$$\psi_c = \operatorname{asin}\left(\frac{\lambda}{4\pi\sigma_h}\right)$$

where λ is the wavelength. *DKB, SAL*

Ref.: Beckmann (1963), pp. 9–10; Currie (1987), p. 213.

The **search angle** is the solid angle over which a search radar scans:

$$\psi_s = A_m(\sin\theta_m - \sin\theta_0)$$

where A_m is the azimuth sector, θ_m is the maximum elevation, and θ_0 the minimum elevation. *DKB*

Ref.: Barton (1964), pp. 134–135.

Solid angle is the three-dimensional angle defining a volume in space. In the search radar equation, the search volume ψ_s is expressed in terms of the solid angle. The measure of solid angle is the steradian. One steradian is the solid angle that encloses a surface area of a sphere equivalent to the square of the radius. By definition, there are 4π steradians of solid angle in a unit sphere. Figure A46 demonstrates that, except for small angles (such as the pencil beam of a tracking radar), the solid angle is not simply the product of the azimuth and elevation sectors in radians. *PCH*

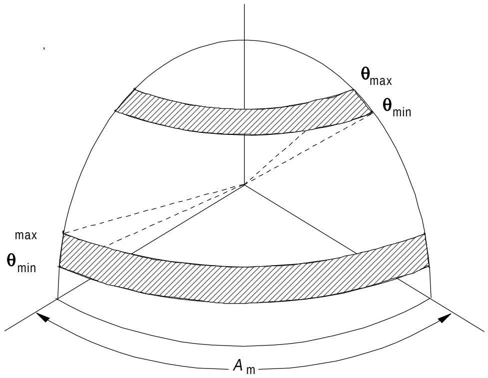


Figure A46 The solid angle of search, in steradians, for a ground- based radar.

Squint angle is the angle between the beam axis and (1) the tracking axis of a conical scan antenna or (2) the face of an array. In the first case (Fig. A47), the squint is intentional, to provide sensing of target position relative to the tracking axis. In the second case (Fig. A48), it is the result of the choice of frequency and the spacing of the radiating elements relative to the wavelength within the waveguide. This angle is given by

$$\sin\theta = \lambda\left(\frac{1}{\lambda_g} - \frac{1}{\lambda_{go}}\right)$$

where λ is the wavelength, λ_g is the wavelength in the waveguide, and λ_{go} is the wavelength that would provide a broadside beam ($\theta = 0$). *SAL, DKB*

Ref.: IEEE (1993), p. 1,268; Skolnik (1970), pp. 13.2–13.5; (1980), pp. 155, 158.

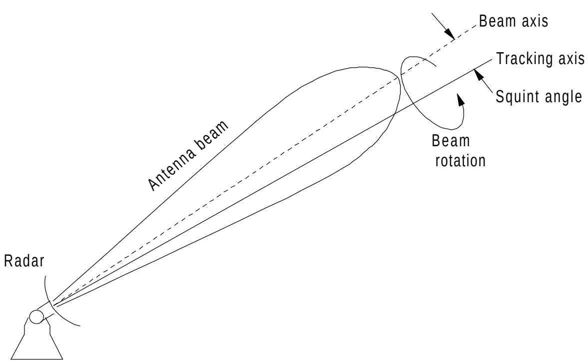


Figure A47 Squint angle in conical scanning.

Tilt angle is "the vertical angle between the axis of measurement and a reference axis; the reference is normally horizontal." It is used in the radar context to describe (1) the angle between the face of a planar array and the vertical, (2) the angle between the electrical axis of an antenna and the ground plane, or (3) the angle between the major axis of a polarization ellipse and a reference direction. In each case, the angle is measured in the vertical plane. *DKB*

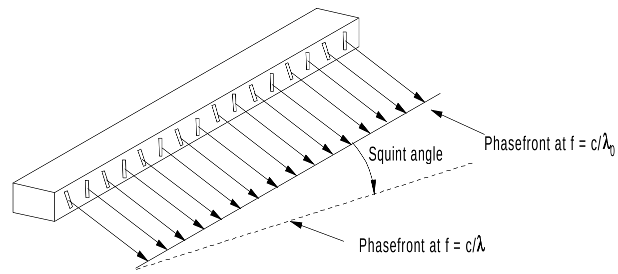


Figure A48 Squint angle in radiation from slotted waveguide.

Ref.: IEEE (1993), p. 1374.

ANTENNA. An antenna is defined as "a structure associated with the region of transition between guided wave and free-space wave, or vice versa." An important property of antennas, as stated by the reciprocity theorem, is that the antenna pattern is identical for transmitting and receiving modes of antenna operation provided that nonlinear circuits (or unilateral devices) are not employed. In radar applications the main function of antenna is to concentrate a radiated energy into the beam of required shape, referred to as the *antenna pattern*, to transmit it in the desired direction, and to receive the energy returning from targets. Radar antennas typically are directional antennas providing angular resolution of observed targets and angular coordinate measurement. The main parameters of radar antennas are operating frequency band, antenna pattern shape, directive gain (or directivity), power gain (often referred to as simply *antenna gain*), beamwidth, sidelobe level, polarization type, voltage standing-wave ratio, and (for transmitting antennas) power handling capability.

Radar antennas vary widely in design. They can be classified in groups based on specific features: they are first classified as aperture-type antennas or antenna arrays. Aperture-type antennas can be omnidirectional (used mainly in electronic warfare applications) or directional antennas; the latter are represented mainly by horn antennas, lens antennas, and reflector antennas. Antenna arrays also represent a large class of discrete antennas and are described in a separate entry (see **ARRAY, antenna**). Subsequent classifications of microwave antennas can be based on specific configuration features (e.g., conformal antennas, deployable antennas); technology features (e.g. microstrip antenna); specific signal processing involved (e.g. synthetic aperture antenna); and so forth.

The radar antenna is perhaps the most important subsystem, defining to a great extent the radar operational capabilities and cost. In radar applications there are two main classes of antennas used: array antennas and reflector antennas. The first provides inertialess electronic scanning, independent tracking of many targets in the conditions of complicated interference environment, and flexibility in synthesizing of different types of radiation patterns. They are more complicated and more expensive than reflector antennas, being preferred in military radars where high performance under conditions of ECM is a predominant factor.

Reflector antennas feature relative simplicity, high gain, and lower cost, but they are more vulnerable to jamming and their scanning capabilities are restricted only to mechanical and electromechanical scanning. Nevertheless, in commercial radar applications where cost factors are very important, reflector antennas are still the competitive choice. *SAL*

Ref.: Johnson (1984); Skolnik (1980), pp. 223–277, (1990), pp. 6.1–6.59, 7.1–7.69; Rudge (1982).

active antenna (see **ARRAY, active**).

An **adaptive antenna** is an antenna "having circuit elements associated with its radiating elements such that some of the antenna properties are controlled by the received signal." Typically, this type of antenna is used to receive signals from desired sources while suppressing those from undesired (interference) sources. Usually the main component of an adaptive antenna system is a phased array (see **array, adaptive)**, a multiple-beam antenna, or a combination of the two. The basic configuration of an adaptive antenna (Fig. A49) incorporates N ports, N complex weights, a signal-summing network, and a weight-determining algorithm. *SAL*

Ref.: IEEE (1993), p. 16; Johnson (1993), Ch. 22.

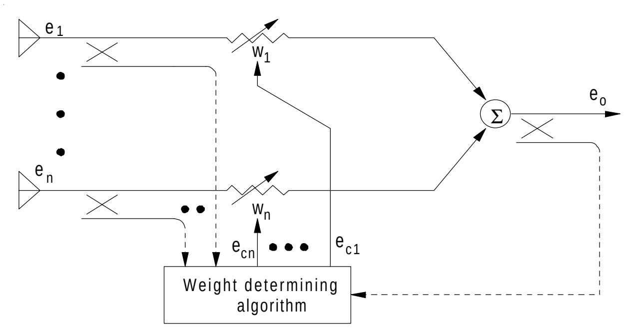


Figure A49 Fundamental adaptive nulling circuit (after Johnson, 1993, Fig. 22.1, p. 22.3).

antenna aperture (see **APERTURE, antenna**).

An **aperture(-type) antenna** is one with a flat aperture forming a directional pattern. The aperture is usually a portion of a plane surface near the antenna, which is perpendicular to the direction of maximum radiation and through which most of the radiation flows. The most widely used types of aperture antennas are reflector, lens, and horn antennas.

Approximate formulas to predict the beamwidth, θ, and gain, G, of an aperture antenna are

$$\theta_3 = K_\theta \frac{\lambda}{D} \text{ deg}$$

$$G = \frac{K_g}{\theta_1 \theta_2} = \frac{4\pi}{\theta_1 \theta_2 L_n}$$

where λ is the wavelength, D is the aperture dimension in the plane of the pattern, $K_\theta \approx 60°$ is a dimensionless constant, θ_1 and θ_2 are the 3-dB beamwidths in two orthogonal principal planes, and $K_g \approx 10$, $L_n \approx 1.3$ are constants for θ in radians ($K_g \approx$ 30,000 for θ in degrees). More accurate estimates should take into account the aperture illumination and antenna efficiency. For cosecant-squared elevation patterns, about 1.5 dB should be subtracted from the gain as estimated from the equation above. The obsolete term for aperture antennas is wavefront antennas. *AIL*

Ref.: Sazonov (1988), p. 360; Johnson (1984), pp. 1.6, 1.15; Mailloux (1994), pp. 267–275.

antenna array (see **ARRAY**).

antenna axis (see **AXIS, antenna**).

A **backfire antenna** is one "consisting of a radiating feed, a reflector element, and a reflecting surface such that the antenna functions as an open resonator, with radiation from the open end of the resonator."

Ref.: IEEE (1993), p. 83; Johnson (1993), p. 12.14.

Antenna bandwidth is an operating band of frequencies within the limits of which other parameters do not exceed the limits of the tolerances called for by the technical requirements. Usually, the parameter depending the most on frequency defines the limits of the operating frequency band. For example, a change in the position of the radiation pattern maximum, expansion of the beam and a drop in directive gain, and so on can stipulate a frequency band limitation. The following breakdown of antennas is conditionally accepted depending on bandwidth: narrowband (with a ratio of upper and lower frequency limits) up to 1.2:1; broadband up to 5:1; ultrawideband up to 10:1 and higher. *AIL*

Ref.: Sazonov (1988), p. 204; Barton (1988), pp. 185–187; Mailloux (1994), pp. 34–40.

antenna beam (see **BEAM, antenna**).

A **broadband antenna** is an antenna used for reception and transmission of signals in a broad frequency band or on various frequencies. This type of antenna is capable of operating with an upper and lower frequency ratio of up to 5:1 and more, with a slight change in radiation pattern shape. Log periodic and helical antennas fall in this category.

Broadband antennas are finding use mainly in COMINT and ECM systems. These antennas sometimes are called *frequency-independent antennas*. *AIL*

Ref.: Sazonov (1988), pp. 262–272; Johnson (1993), Chaps. 14, 43.

An **antenna with built-in functional circuits** is one in which the functions of the antenna and several radar units are combined from a design standpoint. This type includes antennaverters, antennafiers, and antennamitters. Based on functions accomplished, an antenna correspondingly is combined with a frequency converter, radio-frequency amplifier, and transmitter. Advantages of such devices include better performance figures, increased reliability, and compact design. *AIL*

Ref.: Popov (1980), p. 35.

A **Cassegorian antenna** is a multiple-reflector antenna with a subreflector that has a hyperbolic arc in the vertical plane (as in the Cassegrainian antenna) and an elliptical arc in the other plane (as in the Gregorian antenna). This type of subreflector is used as an aberration-correcting subreflector in conical-torus scanning. *SAL*

Ref.: Johnson (1993), p. 17.47.

A **Cassegrainian antenna** is a reflector antenna comprising a main parabolic reflector and a hyperbolic subreflector placed between the focus and center of the main reflector (Fig. A50). If one considers the subreflector as a hyperbolic reflector creating a mirror image of the feed at point F_r which is at the focus of the parabola, then the antenna with a subreflector can be considered as a conventional single-reflector parabolic antenna, but with increased focal length. This is an important special feature of such antennas since the increase in focal length is equivalent to amplification equal to

$$K_h = \frac{\tan(\alpha_0/2)}{\tan(\alpha_r/2)} = \frac{l_r + 1}{l_r - 1}$$

where l_r is the subreflector eccentricity.

On the other hand, use of a subreflector leads to aperture shadowing, decreasing gain, and increasing sidelobe level. It is possible to decrease subreflector dimensions while simultaneously increasing the feed directivity or bringing it closer to the subreflector to decrease shadowing. For minimum shadowing, the subreflector diameter must be equal to the feed dimensions.

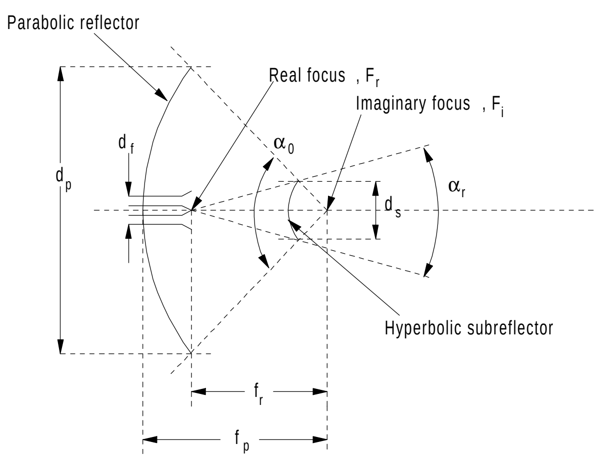


Figure A50 Cassegrainian reflector antenna. (after Leonov, 1986, Fig. 2.3, p. 15).

Cassegrainian antennas are widely used in monopulse radars. The main advantage is the capability to locate the feed behind the mirror, reducing the length of the feed line and decreasing the angle error caused by the phase difference between feed line segments. Another advantage is the ability to provide electromechanical scanning by tilting the subreflector.

If a movable subreflector is used for scanning the beam, while the feed and paraboloid remain fixed, the antenna is called an inverse Cassegrainian antenna. The beam, collimated by a radome-supported, wire-grid paraboloid, is reflected by a flat polarization-rotating reflector and the reflected wave, with polarization rotated 90°, passes without

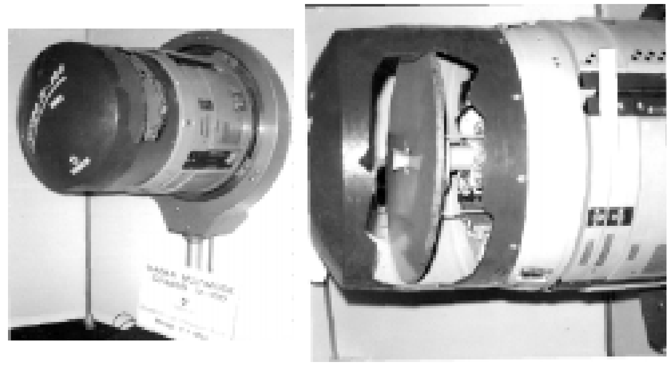


Figure A51 Inverse Cassegrainian antenna: the Cyrano airborne radar.

loss through the grid paraboloid (Fig. A51). The main advantage of such a technique is the compact structure for a given coverage requirement and relatively rapid (for reflector antennas) beam scan with low servo drive power. These features make it attractive for airborne radar applications. *AIL, SAL*

Ref.: Leonov (1986), p.15; Johnson (1993), pp. 17.33–17.41; Skolnik (1990), pp.18.32, 19.35.

Cheese-type antenna (see **pillbox antenna**).

A **circular antenna** is an antenna with a circular aperture. The usual coordinate system for a circular antenna is given in Fig. A52. The circular antenna is typically used to generate a pencil beam.

Ref.: Bogush (1989), p. 158.

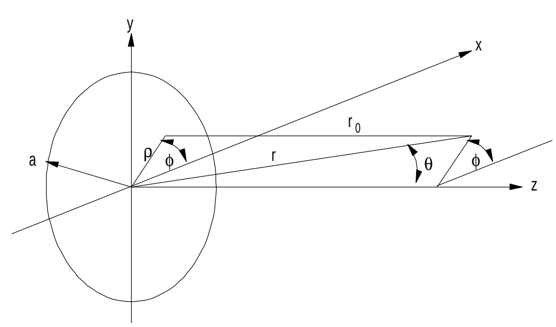


Figure A52 Planar circular antenna and coordinate system (from Bogush, 1989, Fig. 3.31, p. 146).

A **conformal antenna** is one conforming to a nonplanar surface. Typically, these antennas are classified as those with aperture dimensions much less than the local radius of curvature and those with dimensions comparable to that radius, and it is the latter type that is more commonly regraded as conformal. These types of antennas are especially appropriate for construction of air- and missile-borne scanning antennas, where they offer reduction in aerodynamic drag for flush-mounted conformal geometry. This geometry also offers advantages in coverage (e.g., for hemispherical coverage from a hemispherical surface or 360° azimuth coverage from a cylindrical body). Conformal antennas have found practical applications as conformal arrays, typically in cylindrical, conical, ring, and spherical shapes. *AIL, SAL*

Ref.: Johnson (1993), Ch. 21; Voskresenskiy (1981), p. 150.

A **conical-scan antenna** is a tracking radar antenna that uses the conical scan technique. One of the simplest conical-scan antennas is a parabola with an offset feed rotated about the axis of the reflector. If the feed maintains the plane of polarization fixed as it rotates, it is called a *nutating feed*, and if the feed causes polarization to rotate, it is called a *rotating feed*. The tracking bandwidth of a radar employing conical-scan antennas is low, and this technique was used primarily in early radar development. *DKB*

Ref.: Skolnik (1980), p. 155.

A **cosecant-squared antenna** is one forming a cosecant-squared beam in elevation (see **BEAM, cosecant-squared**). This beam can be generated with an array or shaped reflector antenna. In the latter case, either the true parabola with a properly designed set of multiple feed horns, or distorted section of parabola may be used. Doubly-curved reflectors are typically used to generate cosecant-squared beams in search radars (see **REFLECTOR, doubly-curved**). An example of a cosecant-squared antenna is a doubly-curved reflector antenna used in the Raytheon ASR10SS radar (Fig. A53). *SAL*

Ref: Johnson (1984), p. 17–19; Skolnik (1980), p. 55.

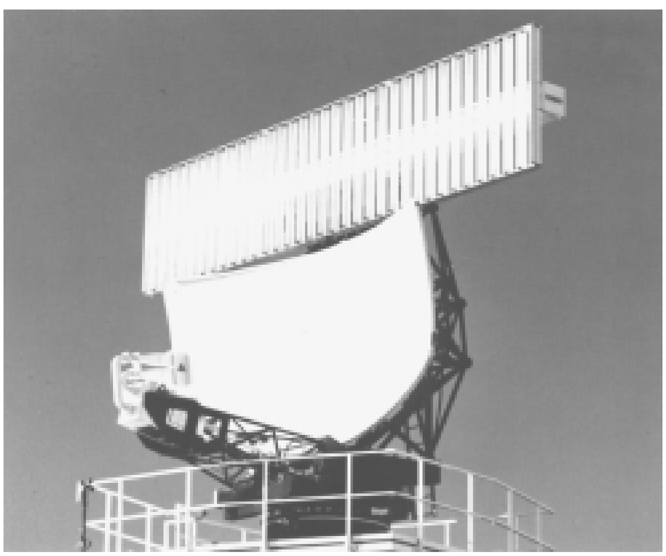


Figure A53 Cosecant-squared antenna for ASR10SS radar.

A **deployable antenna** is one that changes its dimensions and shape during transition from travel to operating configuration. These antennas are typically categorized as self-deploying or manually deploying, and as inflatable or mechanical, based on how the reflector is packed in travel configuration. Inflatable antennas may be solid or tubular. Mechanical antennas are categorized as umbrella, truss, circular, prefabricated, or sectoral. Deployable antennas are used in mobile ground-based radars and in space-based radars.

One widely used configuration is the deployable umbrella antenna, in which rigid, folding, or flexible ribs are used. The reflectors in such antennas are formed by a flexible material that is stretched over a frame made of radial ribs, attached to a central bushing, that fold like an umbrella. Antennas with rigid ribs are stable against wind loads and are highly reliable. Umbrella antennas with folding ribs are used both in ground-based and in space radars. Umbrella antennas with flexible ribs are used only in space due to their lack of rigidity. *AIL*

Ref.: Gryannik (1987), pp. 4, 7; Cantafio (1989), pp. 499–524.

A **dielectric [polyrod] antenna** (Fig. A54) is one consisting of a dielectric rod having circular or rectangular cross section, with a length of a few wavelengths. The rod is fed by a section of circular or rectangular waveguide. In the dielectric rod, the lowest hybrid slowed electromagnetic wave, HE_{11}, is used. The slow-wave factor of phase velocity along the dielectric rod is defined by the diameter and taper of the rod.

The half-power beamwidth of antenna, θ_3, and its directivity, D, may be evaluated by using expressions

$$\theta_3 = (107° - 61°)\sqrt{\lambda/L}$$

$$D = (4.0 - 7.2)L/\lambda$$

where L is the length of the rod.

Dielectric rod antennas are used as independent antennas and as the array elements with broad patterns. *AIL*

Ref.: Johnson (1993), pp. 12.21–12.22; Sazonov (1988), p. 303; Wolff (1988), pp. 394–405.

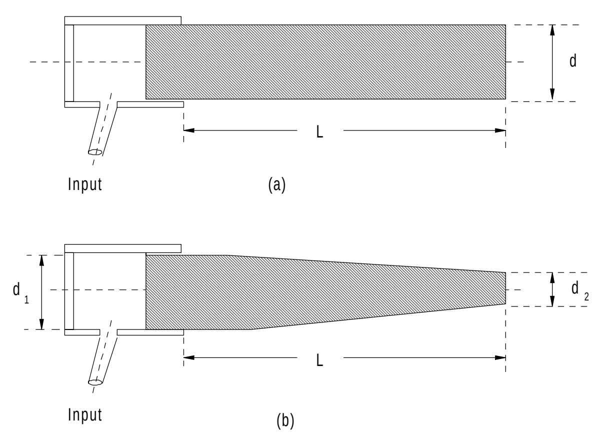


Figure A54 Dielectric antennas: (a) cylindrical rod; (b) conical rod.

A **diffraction antenna** is one in which, besides a feed and a radiator, there is a diffraction element shaping the radiation pattern. A horn, lens, mirror, slot, or dielectric rod usually is employed as the diffraction element. *AIL*

Ref.: Popov (1980), p. 115.

A **dipole antenna** is one in which one or several dipoles are used. Individual dipoles are rarely used as independent antenna systems. Multidipole antennas may comprise dipoles assembled in several tiers with the corresponding distances between tiers. A metallic reflector is located at a specific distance behind the dipoles to obtain unidirectional radiation. Dipoles are used in metric and decimetric wave bands. *AIL*

Ref.: Johnson (1984), p. 4.3; Sazonov (1988), p. 222; Mailloux (1994), pp. 240–267.

A **directional antenna** is one radiating and receiving electromagnetic energy along one or several main directions. Basic directivity characteristics include directive gain, power gain, and beamwidth. The most widely encountered pattern shapes are pencil beams, fan beams, and cosecant-squared beams. These antennas are also termed *high-gain antennas*. *AIL*

Ref.: Fradin (1977), p. 77; Johnson (1993), p. 1.13.

Antenna directivity [directive gain] is "a measure of the ability of an antenna to concentrate radiated power in a particular direction." It is defined as the ratio of achieved radiation intensity, $\Phi(\theta,\phi)$, at angles θ,ϕ, to that of an isotropic radiator radiating the same total power:

$$D(\theta,\phi) = \frac{\Phi(\theta,\phi)}{\Phi_{avg}} = \frac{\Phi(\theta,\phi)}{P_r/(4\pi)}$$

where

$$P_r = \int_0^{2\pi}\int_0^{\pi} \Phi(\theta,\phi)\sin\theta(d\theta)d\phi$$

is the total power radiated by the antenna.

When the antenna aperture is illuminated with uniform, equiphase distribution, the directivity achieved is maximized, and is known as the *standard directivity.* For other illuminations, the ratio of the directivity on the axis of maximum radiation to the standard directivity is called the *normalized directivity* or *antenna (aperture) illumination efficiency.*

The directivity and power gain of the antenna are related through the radiation efficiency, η:

$$G(\theta,\phi) = \eta D(\theta,\phi)$$

SAL

Ref.: IEEE (1993), pp. 362, 1057; Johnson (1984), p. 1.5.

dish antenna (see **reflector antenna**).

A **displaced phase center antenna (DPCA)** is one in which doppler spread due to platform motion is compensated by physically or electronically displacing the phase center of the antenna between pulses. It is used in radars based on movable platforms (e.g., airborne or spaceborne).

Ref.: Cantafio (1989), pp. 430–435; Skolnik (1990), pp. 16.8–16.14.

A **Dolph-Chebyshev antenna** is one with a pattern optimized to have the narrowest beamwidth for a specified size and sidelobe level, or the minimum sidelobe level for the required size and beamwidth. Typically, a Dolph-Chebyshev antenna is a linear or two-dimensional array of equally spaced elements with a specially adjusted current in each of them. Dolph derived the aperture illumination with this property by forcing a correspondence between the Chebyshev polynomial and the polynomial describing the antenna pattern. In practice, the high edge currents required in Dolph-Chebyshev antennas limit their applicability in radar applications where high gain and low sidelobes are desired. What is more, there is a practical upper limit to the size of an antenna having a Dolph-Chebyshev pattern, and therefore a lower limit to the beamwidth. *SAL*

Ref.: Skolnik (1980), p. 257.

antenna drive (see **DRIVE, antenna**).

A **dual-beam antenna** is one that permits reduction in the receiving gain close to the horizon for short-range targets, without suppressing targets lying at the higher elevations above the strong clutter. A typical dual-beam antenna coverage pattern is shown in Fig. A55.

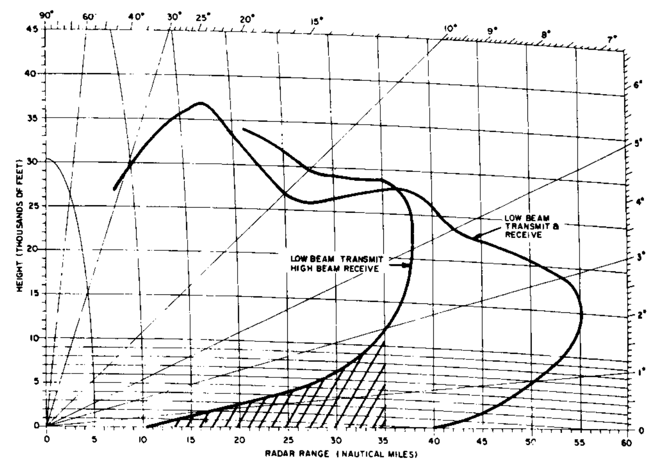


Figure A55 Typical coverage pattern obtained with dual-beam antenna (from Barton, 1988, Fig. 7.2.8, p. 344).

Typically, a second feed horn is added to the conventional reflector system and the receiver is switched to this horn for the first portion of the interpulse period. The transmitter remains connected to the low beam horn. The transition between the two receiving beams is made in steps, and only in the specific regions of strong clutter. This preserves detection of the lowest targets as long as they do not lie directly over the strong clutter region. This techniques is common to two-dimensional search radars. *DKB, SAL*

Ref.: Johnson (1987), pp. 9.7–9.10; Barton (1988), p. 343.

A **dual-[double]-reflector antenna** consists of a main reflector and a subreflector. The feed illuminates the surface of the subreflector, from which the reflected wave travels to the surface of the main reflector (Fig. A56). In this case the length of feed transmission lines can be reduced considerably since the feed can be located near the surface of the main reflector. The most common configurations of dual-reflector antennas are the Cassegrainian (convex subreflector) and Gregorian (concave subreflector). *SAL*

Ref.: Johnson (1993), pp. 17.33–17.44; Leonov (1988), p. 34.

A **dummy antenna** is a device that has the necessary impedance characteristics of an antenna and the necessary power-handling capabilities but does not radiate or receive radio waves. In receiver practice, the portion of impedance not included in signal generator is often called the dummy antenna. *SAL*

Ref.: Johnston (1979), p. 58.

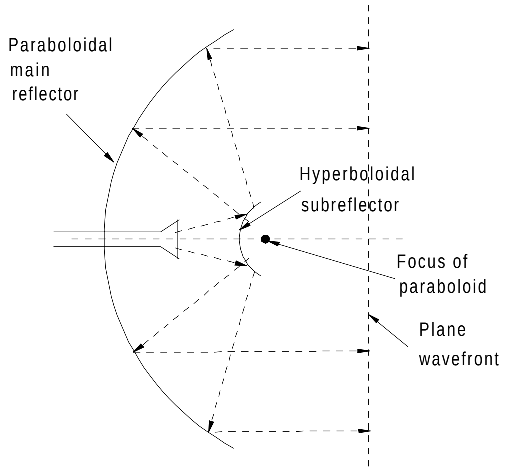


Figure A56 Geometry of the Cassegrainian dual-reflector antenna.

A **dynamic antenna** is one having parameters periodically changing over time. The changing parameters may be amplitude or phase distribution of the field in the antenna aperture; antenna dimensions; time of activation of an individual antenna element; and so forth. Dynamic antennas are typically arrays. Problems of synthesis of a radiation pattern with a low sidelobe level may be solved in dynamic antennas by switching array elements. They sometimes are called *antennas with time-modulated parameters*. *AIL*

Ref.: Bakhrakh (1989), p. 17; Johnson (1993), Ch. 22.

The **antenna effective area** (in a given direction) is "the ratio of the power available at the terminals of an antenna to the incident power density of a plane wave from that direction polarized coincident with the polarization that the antenna would radiate." (See also **APERTURE, effective**).

Ref.: IEEE (1993), p. 43.

antenna efficiency (see **APERTURE efficiency**).

An **electrically large antenna** is one with aperture greater than 60 wavelengths. They are typically used in radar astronomy, antiballistic missile radar, space-surveillance radar, and other applications where high gain and narrow beamwidth are required. *SAL*

Ref.: Johnson (1993), p. 17.31.

A **fan-beam antenna** is one generating a fan beam, typically used in search radars or surveillance radars with the wide dimension in the vertical plane and scanning in azimuth. It is also used in height-finding radar, where the wide dimension is in a horizontal plane and scanning is done in elevation. The fan beam can be produced by array antenna or by a reflector antenna with the mirror truncated in the plane of the wide antenna pattern. *SAL*

Ref.: Johnson (1993), p. 1-13; Skolnik (1980), p. 54.

antenna feed (see **FEED, antenna**).

An **antennafier** is an antenna with an embedded amplifier.

A **fixed-beam antenna** is one without scanning or beam switching. Fixed-beam antennas are primarily used in electronic warfare systems, and the main types of such antennas are broadband dipoles, broadband monopoles, spirals (planar or conical), horns and slots, log periodics, helices, and reflector types. *SAL*

Ref.: Schleher (1986), pp. 474–482.

frequency-independent antenna (see **broadband antenna**).

A **fuselage-mounted antenna** is an airborne radar transceiving antenna installed along an aircraft fuselage on the inside or in a special pod. It may or may not be used to generate a synthetic aperture. One without synthetic aperture capability is a long antenna designed for observation of the terrain or surface objects in accordance with the radiation pattern it shapes. The antenna is attached in such a way that its beam is perpendicular to the aircraft flight trajectory and surveillance occurs by displacement of the beam during flight. Two antennas, one on the right side and the other on the left, often are used to increase the area of the observed sector. Each antenna has a narrow (1 to 1.5°) beam in the horizontal (azimuth) plane and a relatively wide (50 to 60°) beam in the vertical (elevation) plane. Each antenna has a waveguide feed and usually is a slotted waveguide antenna. *AIL*

Ref.: Kondratenkov (1983), p. 64.

Antenna gain, $G(\theta,\phi)$, is "a measure of the ability to concentrate in a particular direction the power accepted by the antenna." It is related to the antenna directivity, $D(\theta,\phi)$ in the direction specified by angles θ and ϕ through the radiation efficiency, η:

$$G(\theta,\phi) = \eta D(\theta,\phi)$$

and to the effective (aperture) area, $A_r(\theta,\phi)$ by

$$G(\theta,\phi) = \frac{4\pi A_r(\theta,\phi)}{\lambda^2}$$

where λ is the wavelength.

For aperture-type antennas, $A_r = \eta_a A$, where η_a is the aperture efficiency and A is the physical area. In this case,

$$G = \frac{4\pi A}{\lambda^2}\eta_a$$

(see also **aperture-type antenna**).

For array antennas, the gain in direction θ_0 from normal to the array surface can be approximately estimated as

$$G(\theta_0) \approx \pi\eta N\cos\theta_0$$

where N is the number of array elements and η is the radiation efficiency for the array antenna (the cosine function may be raised to the 3/2 power for many practical arrays, depending on the matching of the elements to space).

Antenna gain is often expressed in decibels:

$$G_{\mathrm{dB}} = 10\log G$$

and for a voltage gain pattern with maximum value F

$$G_{dB} = 20 \log F$$

SAL

Ref.: Johnson (1984), p. 1.5; Skolnik (1990), p. 7.2; Barton (1988), p. 166.

A **Gregorian (reflector) antenna** is a reflector antenna with two concave mirrors (Fig. A57). The main mirror is of parabolic shape, and subreflector has elliptic shape. One of the foci, F_1, of the subreflector is located at the focus of the main mirror. The phase center of the feed is placed in the second focus, F_2, of the ellipse. The waves going from the feed illuminate the subreflector, are focused in the point F_1 and travel to the main reflector, creating an in-phase front in its aperture. Typically, this type of antenna is used in electrically large reflector antennas, as used in tracking radars and radar astronomy. *AIL*

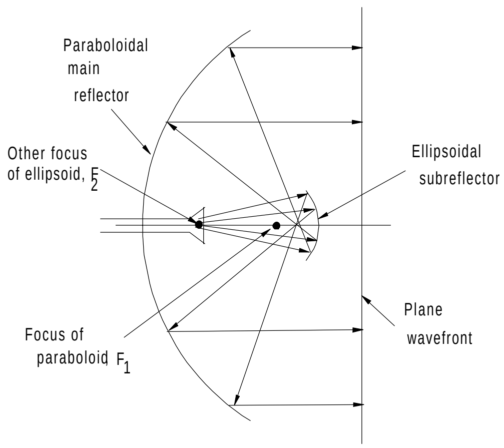


Figure A57 Dual-reflector Gregorian antenna.

Ref.: Johnson (1993), pp. 17.33–17.46; Sazonov (1988), p. 387.

A **helical antenna** is a broad-beam antenna typically consisting of a helix radiator that resembles an uncoiled watch spring (Fig. A58). The polarization of radiated waves is circular, and the direction of radiation is along the axis of the helix when the circumference of the helix is nominally one wavelength. For a typical helix with constant diameter, the bandwidth ratio is about 1.5 to 1, the gain is the function of the number of turns and is about 10 dB for a 6-turn helix. The bandwidth can be increased by 10 to 1 if the helix is wound on the surface of a cone with spacing between the turns increased logarithmically with the turn diameter. Beamwidths up to 90° are typically obtained. Helices have higher gain than spiral antennas, relatively lower dissipation losses, but their structure is larger than spiral antennas. *SAL*

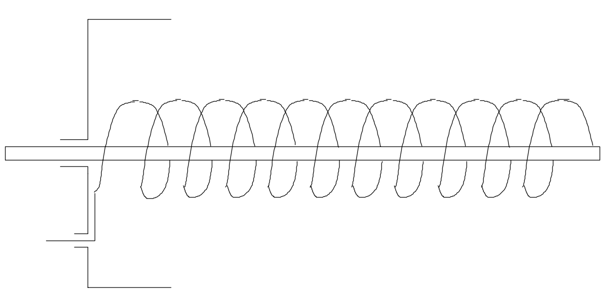


Figure A58 Helix antenna (after Macnamara, 1995, Fig. 5.18, p. 161).

Ref.: Saad (1971), vol, 2, p. 48; Johnson (1984), p. 13.1; Schleher (1986), p. 480.

A **high-gain antenna** is one with relatively high gain (tens of decibels) in the main lobe. These are necessarily directional antennas, which can be of the array or aperture type.

Ref.: Johnson (1993), p. 1.15.

A **"honey" antenna** is a broad-beam biconical horn antenna that provides an omnidirectional pattern in the azimuth plane. It is formed by a biconical horn. Outputs with amplitudes proportional to the sine or cosine of the signal's angle-of-arrival over the 360° azimuth coverage are formed by specially combining four probes at the bottom of the feed. The main range of application is direction-finding and electronic warfare systems. *SAL*

Ref.: Johnson (1993), p. 40.2; Schleher (1986), p. 479.

A **horn antenna** is a directional antenna made as an extension of the waveguide feeding the horn, in a manner similar to the wire antenna. Antenna directivity increases as horn aperture increases and as the angle of the horn decreases (e.g., the length of the horn increases). The most popular types of horns are pyramidal, diagonal, conical, and corrugated conical (Fig. A59).

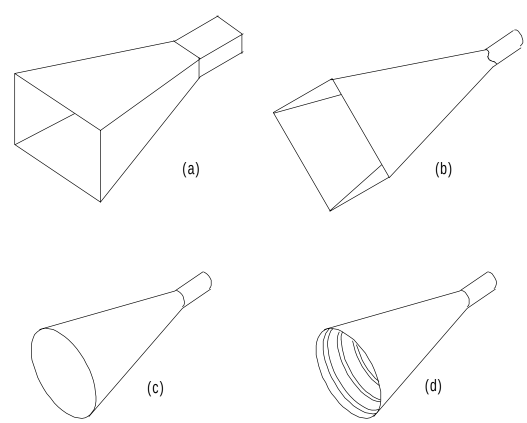


Figure A59 Common types of horn antennas: (a) pyramidal horn; (b) diagonal horn; (c) conical horn; (d) corrugated horn (after Currie, 1987, Fig. 12.12, p. 539).

The pyramidal horn is one of the most common structures, as it can be made cheaply from the flat pieces of metal that are cut along the straight lines. Its gain G_p is equal to

$$G_P = 10[1.008 + \log(a_\lambda \cdot b_\lambda)] - (L_E + L_H)$$

where a_λ and b_λ are the aperture dimensions in wavelengths λ, and L_E and L_H are the losses in decibels due to phase error in the E and H planes as given in Fig. A60.

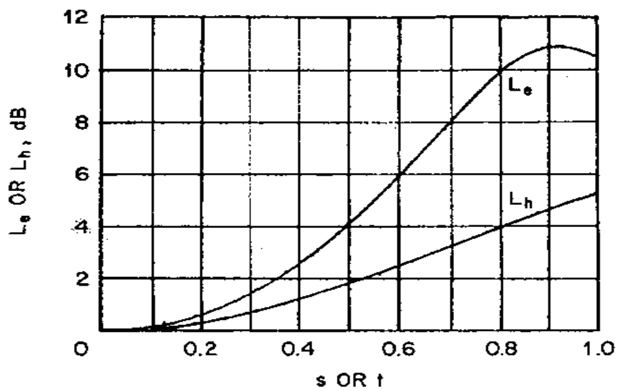


Figure A60 Loss corrections for phase error in sectoral and pyramidal horns. The parameters $s = b^2/(8\lambda l_e)$, and $t = a^2/(8\lambda l_h)$, where l_e and l_h are the lengths of the horn flares (from Johnson, 1993, Fig. 15.8, p. 15.10, reprinted by permission of McGraw-Hill).

For a conical horn, the gain G_c has been determined to be

$$G_c = 20\log C_\lambda - L$$

where C_λ is the circumference of the horn aperture and L is the gain loss due to phase error given in Fig. A61. To reduce the sidelobes in both E and H planes, grooves inside the wall of the conical horn can be made to produce a corrugated conical horn. There are also other types of horn antennas such as sectoral, biconical (sometimes called "honey antenna"), and so forth. Because of relatively good overall performance, structural simplicity, and predictable design performance, horn antennas are widely used in the entire radar band from decimeter to millimeter waves, primarily as feeds for reflector and lens antennas. As independent antennas they are used mostly in electronic warfare and direction-finding applications, short-range field tests, and laboratory experiments. *SAL*

Ref.: Fink (1982), pp. 18.2–18.32; Johnson (1993), Ch. 15; Schleher (1986), pp. 478–480; Currie (1987), pp. 537–538.

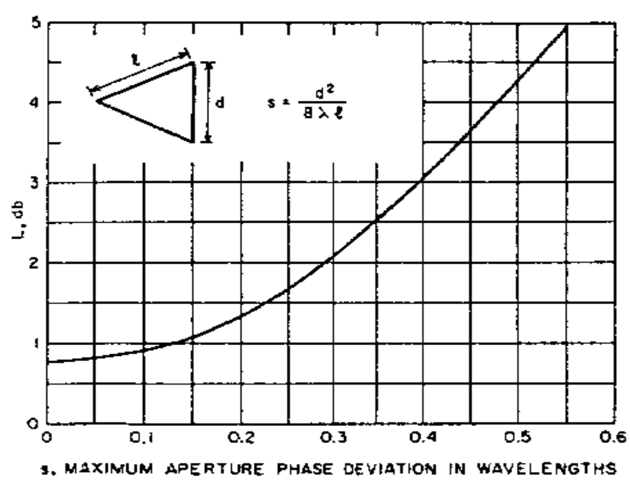


Figure A61 Loss correction for phase error in conical horns (from Johnson, 1993, Fig. 15.11, p. 15.17, reprinted by permission of McGraw-Hill).

An **ideal antenna** is a lossless antenna for which the radiation efficiency is $\eta = 1$. In this case the antenna directivity and gain are identical.

Ref.: Johnson (1993), p. 1.5.

An **image antenna** is an imaginary antenna located below the reflecting surface and pointed downward at an angle equal to the upward tilt angle of the real antenna (Fig. A62). Such a formulation is used to describe the effects of surface reflection on search radar coverage and tracking radar accuracy. (See **PROPAGATION over the earth**).*SAL*

Ref.: Barton (1964), p. 167.

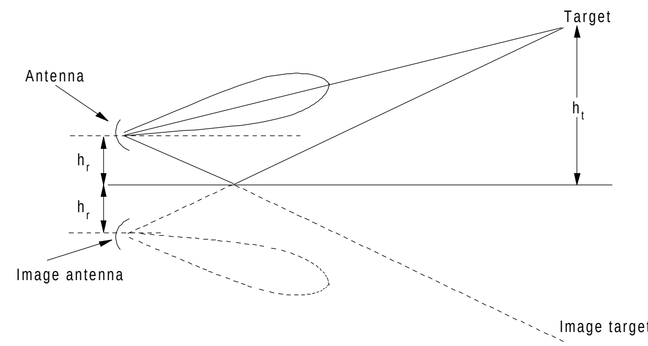


Figure A62 Geometry of the "image" antenna.

An **impedance antenna** is one comprising a driver and a director. The driver is designed to transmit as great a portion of the energy as possible from an oscillator to the director. A wave being propagated along the director is a surface wave. The director is the radiating element of the antenna. The director is an impedance structure and is either a dielectric (see **dielectric antenna**) or a metal rib. Examples of impedance antennas are the dielectric rod antennas shown above in Fig. A54, and the design shown in Fig. A63. Impedance antennas found wide use on airborne vehicles. A shortcoming is the relatively high sidelobe level. They sometimes are referred to as *surface-wave antennas. AIL*

Ref.: Sazonov (1988), p. 308.

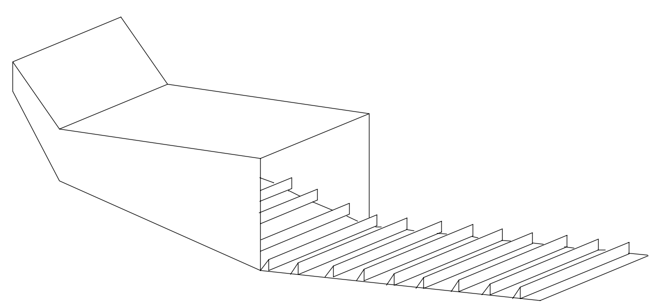


Figure A63 Impedance antenna.

An **inflatable antenna** is a deployable antenna in the shape of a cylinder or sphere and inflated just before use. It may be solid or tubular and is made of strong radio-transparent film and aluminum foil. The antenna is kept in a container when folded up. Inflatable antennas are used in space-based radars. *AIL*

Ref.: Gryannik (1987), p. 7; Cantafio (1989), p. 651.

Antenna input impedance defines the ratio of voltage to current at transmitting antenna input and characterizes the antenna as the load for the transmitter. This parameter mainly is used for linear antennas. *AIL*

Ref.: Lavrov (1974), p. 28; Blake (1984), p. 159; Johnson (1984), p. 2.11.

inverse Cassegrainian antenna (see **Cassegrainian antenna**).

An **isotropic antenna** is one providing equal gain in all directions (isotropic pattern).

A **large vertical aperture (LVA) antenna** is a secondary radar antenna using an array of significant height to control the elevation beamwidth and the illumination of the surrounding ground surface. Its advantages over the more conventional antenna with small vertical aperture (basically a flared horn radiator) are higher gain and less multipath illumination of transponders, which causes gaps in coverage and false triggering of transponders in the conventional secondary radar. (See also **RADAR, airport surveillance**).*DKB*

Ref.: Stevens (1988), p. 46.

A **leaky-wave antenna** is a waveguide antenna with a longitudinal slot along the entire length or holes of varying shape in the narrow or wide walls. In these antennas, a wave that partially leaks to the outside through the longitudinal slot or through the array of holes is propagated along the waveguide. The waveguides may be round or rectangular. A plane flange is situated at the slot to insure unidirectional radiation. The holes in the arrays are nonresonant and are situated at short distances from one another (much less than the length of the wave in the waveguide). The radiation field is a spherical wave with the radiation pattern maximum located at some squint angle to the axis of the waveguide. Since this angle depends on frequency, these antennas may be used for frequency scanning. *AIL*

Ref.: Fradin (1977), p. 375; Johnson (1993), Ch. 10.

A **lens antenna** is one shaping a radiation pattern through use of the optical properties of electromagnetic waves on the edge of the separation of two media. In the general case, a lens antenna comprises a feed and a lens. The feed is a point or a linear source from which radiation in the form of a spherical or cylindrical wave strikes the illuminated side of the lens. Figure A64 depicts formation of the field in a lens antenna.

A spherical wave from source F illuminates the lens. Being refracted on the convex illuminated surface of the lens, the spherical wavefront is converted into a plane wavefront. The lens output surface is plane and, upon exit from it, the wavefront remains a plane. Since the dimensions of the lens output surface are great compared with wavelength, its radiation is highly directional.

Lens antennas are classified based on the shape of the lens, materials used, and radiation pattern formed. There are

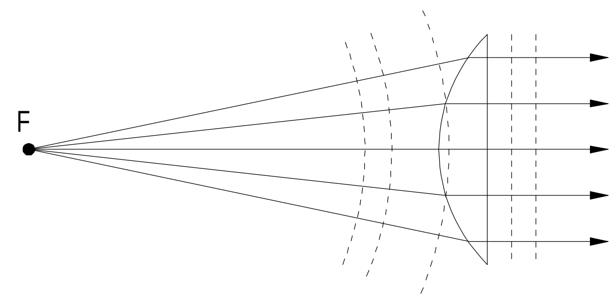


Figure A64 Formation of a field in a lens antenna.

lenses in the shape of a rotation figure, cylindrical, conical-parabolic, and spherical. Depending upon materials used, lens antennas are categorized as dielectric or metal-plate (see **LENS**). Based upon the radiation pattern shaped, they are classified as highly directional or broad-beam, using scanning or multibeam configuration. Lens antennas have several advantages: they are broadband, and it is relatively easy to obtain patterns with a low sidelobe levels, shape the radiation pattern, and scan over large sectors. Moreover, aperture blockage is absent in lens antennas. Shortcomings include low efficiency due to loss in materials, large volume occupied, and manufacturing complexity. *AIL*

Ref.: Zelkin (1974), pp. 3–23; Johnson (1993), Ch. 16; Bey (1987), pp. 6–10; Sletten (1988, 1991).

A **lens-horn antenna** is one obtained by combining a dielectric lens with a small horn.

If the lens is too thick or heavy, the grooved lens can be applied by cutting out steps of thickness Δ. The phase center of the horn is placed at the focal point of the lens (Fig. A65). The lens converts the spherical wavefront into a planar one. The lens-horn structure is more compact than a horn antenna, the length of which can become excessive when a high-gain antenna is required. Sometimes the lens-horn antenna is simply termed a *lens antenna*, omitting mention of the feed. *SAL*

Ref.: Currie (1987), p. 538.

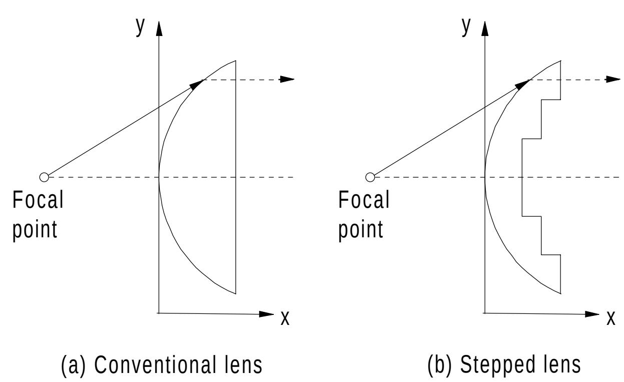


Figure A65 Lens-horn geometry (after Currie, 1987, p. 541).

A **logical synthesis antenna** is one in which the radiation pattern is shaped during comparison of the amplitude of signals from separate antennas, using logic devices of the "YES-NO," "OR," "AND," and "MORE-LESS" type. Use of these operations and of an auxiliary antenna makes it possible to suppress sidelobes, locking out the main receiving channel for all signals below a given level. Logical synthesis antennas are used to obtain a radiation pattern with a given mainlobe width and with a low level of sidelobes, which cannot be achieved using conventional methods. *AIL*

Ref.: Bakhrakh (1989), p. 15.

A **log-periodic antenna** is a broadband antenna with parameters which have periodic logarithmic dependence on the operating frequency. Radiating structures of such antennas may have different shapes. The spatial log-periodic antenna with an angle $\psi = 90°$ ensures the radiation in the direction of the apex of the structure (Fig. A66). The direction of the maximum radiation pattern coincides with the bisector of the angle ψ. In the case when $\psi = 0$, both annexes of antenna coincide and the planar log-periodic antenna is formed. Such a system can be considered as linear array of symmetrical dipoles with the length changing monotonically. Such types of antennas are mainly used in electronic countermeasures systems. *AIL*

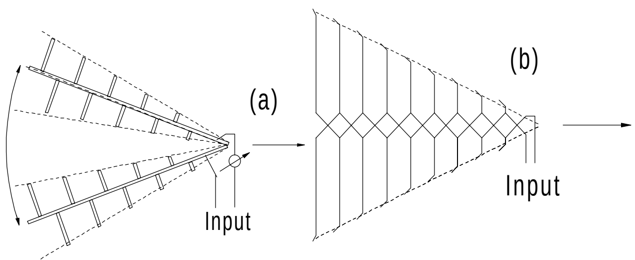


Figure A66 Log-periodic antennas: (a) spatial; (b) planar.

Ref.: Johnson (1993), pp. 14.32–14.53; Sazonov (1988), p.271.

A **log spiral antenna** is a helical antenna with logarithmic parameter periodicity depending upon frequency. A conical helical antenna with variable distance between coils of the spiral and a multiple plane helical antenna are log spiral antennas. The spiral in a conical antenna is wound such that distance l of the point of a coil from the apex of the cone is proportional to radius r at this point (Fig. A67a). The parameters of such an antenna are as follows: l_{min} = minimum distance from cone apex; r_{min} = minimum coil radius; N = number of coils; α = step angle; α_c = angle at cone apex.

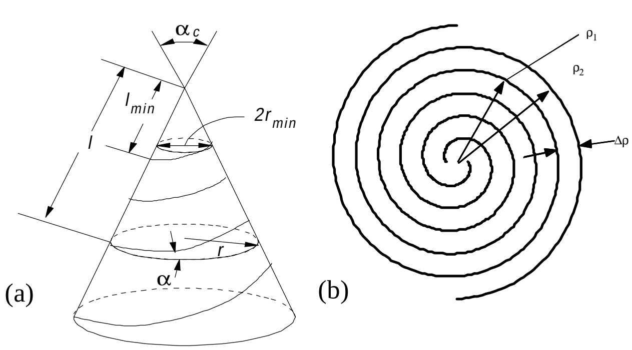


Figure A67 Log-spiral antennas: (a) conical spiral; (b) planar spiral.

Figure A67b depicts an double logarithmic Archimedes spiral. The antenna is made of two conductors in a printed fashion on a thin sheet of high-frequency dielectric. Distance $\Delta\rho$ between conductors and conductor width ρ are constant relative to angle. The distance selected between conductors usually equals the conductor width. A screen is positioned underneath the spiral for unilateral radiation. However, this leads to a reduced bandwidth. *AIL*

Ref.: Johnson (1993), pp. 14.4–14.32; Sazonov (1988), p. 265; Fradin (1977), p. 212.

lossless antenna (see **ideal antenna**).

A **low-noise antenna** is one in which special measures are taken to obtain a low noise temperature. For reflector antennas such measures include: reduction in the illumination at its edges; use of supporting feed structures with minimum reflecting properties; and use of parabolic reflectors having small f/D, where f = focal distance; D = parabolic reflector diameter. However, all these methods reduce noise temperature through a decrease in the reflector aperture efficiency. It is possible to reduce noise temperature without a noticeable decrease in the aperture efficiency by employing special feeds such as a circular dipole array. This ensures almost uniform radiation of the area of the parabolic mirror and a sharp drop in radiation towards its edge. Other methods also are possible. *AIL*

Ref.: Skolnik (1970), p. 10.18; Johnson (1984), Ch. 16.

A **low-sidelobe antenna** is one having sidelobe levels less than some specified threshold. Low-sidelobe, very-low-sidelobe (VLSA), and ultra-low-sidelobe (ULSA) antennas may be distinguished (see Table A7).

The basic principles of achieving low and ultra-low sidelobes are to provide the proper amplitude-phase distributions (including use of ad hoc weighting functions such as Dolph-Chebyshev, Kaiser-Bessel, etc.) and to compensate for numerous error sources in the design, fabrication, assembly, and the siting of the antenna. These steps obviously increase the antenna cost and introduce some difficulties in its manufacture.

Table A7
Definitions of Sidelobe Levels

Antenna Description	Sidelobe Levels		
	dB below mainlobe		dB below isotropic
	Peak	Average	Average
Normal	> −25	> −30	> −3
Low-sidelobe	−25 to −35	−35 to −45	−3 to −10
Very-low-sidelobe	−35 to −45	−45 to −55	−10 to −20
Ultra-low-sidelobe	< −45	< −55	< −20

Among reflector antennas the parabolic cylinder with line feed (Fig. A68) and the doubly-curved reflector with complex feed give reasonable compromises between efficiency and low sidelobe level. As to phased arrays, the most promising ULSA technique is the electrically fixed array of slotted waveguides. In electronically scanned arrays, antennas with digital beam forming (DBF) offer low-sidelobe performance for receiving. *SAL*

Ref.: Brookner (1988), Ch. 6; Barton (1988), pp. 188–196; Skolnik (1990), pp. 7.37–7.49; Farina (1992), pp. 13–58.

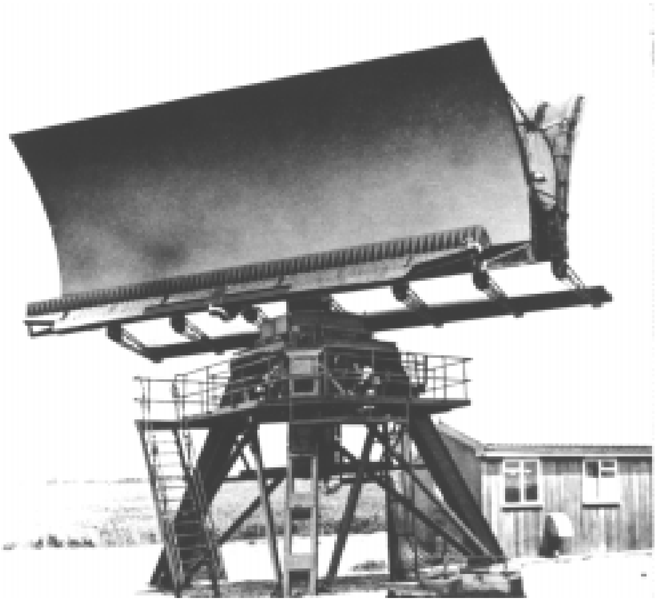


Figure A68 Low-sidelobe antenna using line-fed parabolic cylinder.

antenna measurement (see **TEST, antenna**).

A **microstrip antenna** is one manufactured with printed-circuit technology. The main constituent parts of a microstrip antenna (Fig. A69) are a metal radiator, a ground plane, and the dielectric substrate. Excitation is by the coaxial line through the hole on the ground plane and substrate, as shown in the figure, or by a strip line in the plane of the radiator. The main advantages of such antenna are the simplicity, relatively small volume and weight, and small aerodynamic resistance when the antenna is mounted on an aircraft. The main disadvantage is narrowbandness associated with the resonant mode of antenna operation. Sometimes these antennas are termed *printed-circuit* or *patch antennas*. *AIL*

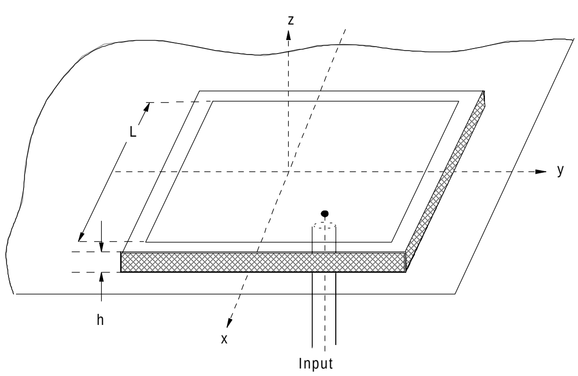


Figure A69 Microstrip antenna.

Ref.: Gupta (1988); Bhartia (1991); Johnson (1993), Ch. 7; Sazonov (1988), p. 258; Zurcher (1995), Ch. 2; Sainati (1996).

Mill's cross antenna (see **ARRAY, Mill's cross**).

A **minimum scattering antenna** is a lossless antenna, fed from an N-port matched and uncoupled waveguide system, that is rendered invisible when all the ports are terminated in open circuits. This concept is used in antenna radar cross section reduction techniques. *SAL*

Ref.: Knott (1985), p. 416.

mirror antenna (see **reflector antenna**).

An **antennamitter** is an antenna with an embedded transmitter.

An **antennamixer** is an antenna with an embedded mixer.

A **monopulse antenna** is one forming the patterns appropriate for monopulse angle-sensing and -tracking. The antenna system for conventional monopulse radar must form three simultaneous patterns: a sum beam, a difference beam for azimuth and a difference beam for elevation. Typically, reflector antennas, phased arrays, lens antennas, and spiral antennas are used. Monopulse patterns are formed using four-horn or multiple-horn feeds for reflector antennas or space-fed arrays, and dual-ladder feeds for constrained feed arrays. *SAL*

Ref.: Leonov (1986), p. 13; Johnston (1984), p. 19.10; Johnson (1993), Ch. 34; Barton (1988), pp. 198–204.

A **multielement antenna** is one comprising several elements. The main types are antenna arrays and traveling-wave antennas. Various types of multielement antennas are widely used in radar since they can provide a high directive gain and a narrow beamwidth in conjunction with excellent capabilities for adaptation to interference environments. *AIL*

Ref.: Bakhrakh (1989), p. 11.

A **multiple-beam antenna** is one forming a series of beams simultaneously. The main types of multibeam antennas are stacked-beam antennas and monopulse antennas based on reflector antenna or space-fed array techniques, illuminated by multiple-horn feed clusters or array feeds. *SAL*

Ref.: Johnson (1993), pp. 17.41–17.52; Barton (1988), pp. 196–207.

A **multiple-reflector antenna** is one using additional secondary reflectors (subreflectors) to form the antenna pattern. The secondary reflector makes it possible to relocate the feed close to the source or receiver, to produce low spillover, to produce a low-sidelobe distribution, and to provide other advantages. If only one additional subreflector is used, the term *dual-reflector antenna* is used. The most common multiple-reflector antenna in radar applications is the Cassegrainian type. *SAL*

Ref.: Skolnik (1990), pp. 6.23–6.25; Johnson (1993), pp. 17.33–17.44.

antenna noise temperature (see **TEMPERATURE**).

An **omnidirectional antenna** is one capable of providing 360° coverage in the azimuth plane. Such antennas are typically used in electronic warfare systems. *SAL*

Ref.: Neri (1991), p. 285.

An **optical antenna** is one in which operation is based on principles of geometrical optics. Optical antennas typically include two subgroups: reflector antennas and lens antennas. *SAL*

Ref.: Skolnik (1990), p. 6.2.

An **"out-phasing" antenna** is an auxiliary antenna or group of antennas used with the basic antenna system to pick up jamming signals, as input to a sidelobe cancelation system in the receiver. *SAL*

Ref.: Johnston (1979), p. 64.

antenna (radiation) pattern (see **PATTERN, antenna**).

patch antenna (see **microstrip antenna**).

A **pencil-beam antenna** is one forming a pencil beam of circular cross section. These antennas typically are phased arrays or circular reflector antennas used in precision-tracking radars or multifunction phased array radars where a pencil beam is used both for search and track functions. *SAL*

Ref.: Johnson (1993), p. 17.31; Skolnik (1980), p. 54.

A **pillbox antenna** is a reflector antenna formed by a section of a paraboloid sandwiched between two parallel plates. Collimation of the beam in one plane is produced by the pillbox and collimation in the other plane is produced by a parabolic cylinder reflector (Fig. A70). Such an antenna is illuminated by a waveguide or horn. Typically, this antenna is used when a narrow pattern in only one of the planes is desired. It can also be used as a linear feed for a highly elliptical reflector. Sometimes these antennas are termed *cheese-type* antennas. *AIL*

Ref.: Skolnik (1970), p. 10.13.

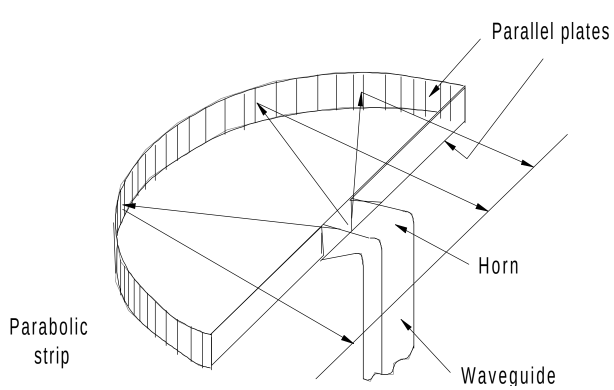
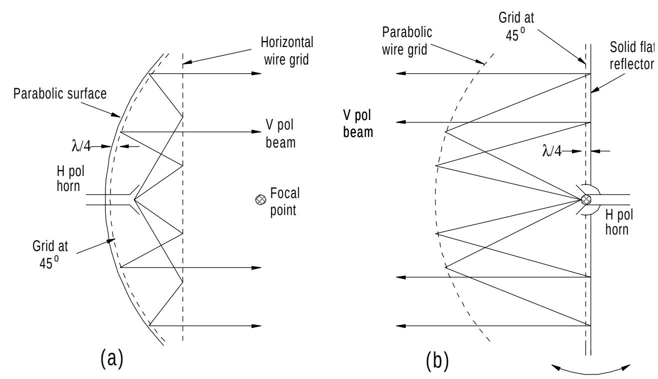


Figure A70 Pillbox ["cheese"] antenna.

A **planar antenna** is one whose elements are located in one plane. It is also termed a *flat antenna*.

Ref.: Fradin (1977), p. 184.

A **polarization-twist antenna** is a dual-reflector antenna in which the feed illuminates the first reflector, which consists of wires matched to the feed polarization Fig. A71). Rays from this wire grid are reflected to the second reflector, consisting of a solid surface overlaid with a grid (or ribs) at 45°. Upon reflection from the second reflector, the polarization is rotated 90° to pass through the first reflector without blockage.



(a) Folded parabolic geometry, (b) mirror-scan geometry.

In the configuration (a), the second reflector is a conventional parabolic surface, and the first reflector (grid) is a planar surface which folds the focal point back to a region near the center of the parabola, reducing the length of the feed line and the mechanical inertia of the antenna, The inverse Cassegrainian antenna shown previously in Fig. A51 uses the polarization-twist configuration shown in (b). The feed-horn illuminates a wire-grid parabolic reflector, embedded in a plastic cover in front of the antenna system. This reflects power back to the tilting plate, on which are a series of ribs oriented at 45° to the incident polarization and λ/4 deep, that rotate the linear polarization through 90° and reflect it as a beam which passes through the parabolic reflector and into space.

Polarization-twist antennas can also use either conventional Cassegrainian or Gregorian configurations. *DKB*

polyrod antenna (see **dielectric antenna**).

printed circuit antenna (see **microstrip antenna**).

antenna radiating element [radiator] (see **RADIATING ELEMENT, antenna**).

Antenna radiation regions in front of an antenna are divided into (a) a near-field reactive region, (b) a near-field radiating region, and (c) a far-field radiating region (Fig. A72). For measurement of antenna patterns, the normal criterion for far-field conditions is $R_f = 2D^2/\lambda$, where D is the diameter of the antenna and λ is the wavelength. Accurate measurement of low sidelobe levels may require even greater distances from the antenna. The actual power density along the axis of circular aperture with taper to create −25-dB sidelobes is shown in Fig. A73. (See also **FIELD, antenna**). *DKB*

Ref.: Hansen (1964), pp. 24–46; Saad (1971), pp. 32–38; Johnson (1993), p. 1-11.

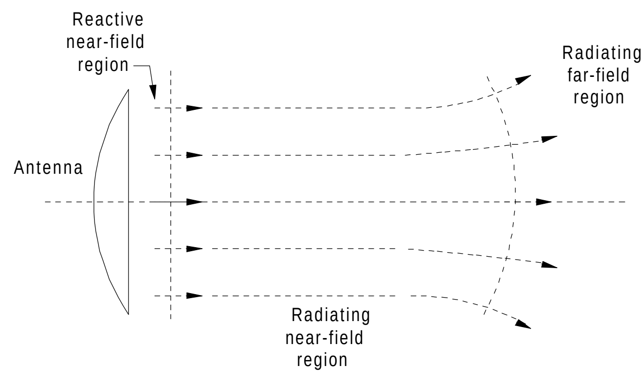


Figure A72 Electromagnetic field regions in front of antenna (after Johnson, 1993, Fig. 1-7, p. 1.11).

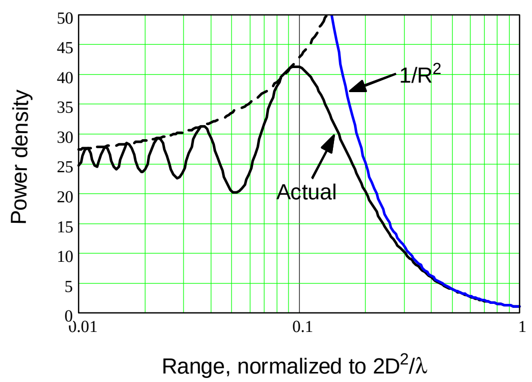


Figure A73 Power density along axis of circular aperture (after Saad, 1971, Vol. 2, p. 34).

antenna radome (see **RADOME**).

A **receiving antenna** is one designed for reception of radio waves and conversion into radio-frequency currents. The main characteristics of such an antenna are gain, directive gain, effective area, antenna aperture efficiency, and frequency response. *AIL*

Ref.: Popov (1980), p. 318.

A **rectangular antenna** is one with a rectangular aperture. The coordinate system for a rectangular antenna is given in Fig. A74.

Ref.: Bogush (1989), p. 139.

A **reflector antenna** is an aperture-type antenna with a feed radiating toward a reflector that shapes the radiation to obtain the desired antenna pattern. The main types of reflector antenna feeds are the dipole feed, slot feed, waveguide feed, horn feed, and linear feeds of different types. The main types of reflectors are the parabolic reflector, spherical reflector, conical reflector, and special-profile reflectors. Reflector antennas can employ single or multiple reflectors. In the latter case, the antenna has a main reflector and one or more secondary reflectors (subreflectors).

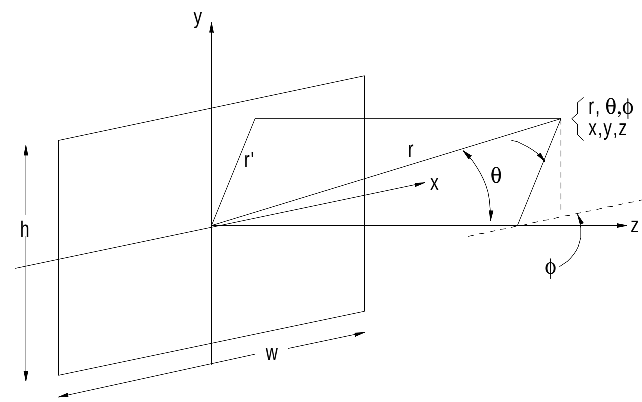


Figure A74 Planar rectangular aperture and coordinate system (after Bogush, 1989, Fig. 3.29, p. 139).

The main type of parabolic reflector is the paraboloid of rotation, formed by rotating the arc of parabola about the line joining the vertex and the focal point. Such a reflector is termed a *mirror* or *dish*, the antenna is termed a *paraboloidal reflector*, *mirror*, or *dish antenna*. This configuration is the most common for high-gain antennas. The most common types of double-reflector dish antennas are Cassegrainian and Gregorian antennas, named after the inventors of optical telescopes of corresponding configurations. An antenna combining the geometry of both antennas sometimes is termed a *Cassegorian antenna*. Other parabolic antennas are represented by parabolic cylinder and parabolic torus reflector shapes. In the parabolic-cylinder antenna, the elements of the cylinder are all perpendicular to the plane of the parabolic arc, and because the surface is singly curved, construction and tolerance problems are less than in the paraboloidal reflector antenna. The feed can be a point-source (e.g., horn feed), but more often a linear feed is employed. This antenna can be employed for generating either pencil or fan beams, and it is a sensible choice in applications where a beam is to be shaped or steered in only one coordinate. An example of parabolic-cylinder antenna was shown in the low-sidelobe design of Fig. A68.

In the parabolic-torus antenna the reflector is formed by rotating the parabolic arc (symmetrical or offset) about a vertical axis positioned on the concave side of the arc (Fig. A75). The section of the torus in one plane has the form of a parab-

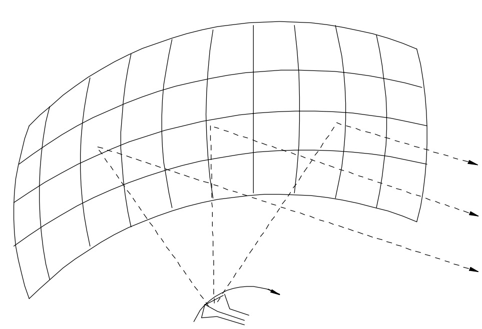


Figure A75 Parabolic-torus antenna geometry (after Johnson, 1993, Fig. 17.46, p. 17.47).

ola, and in the orthogonal plane it is a circle. The combination of focusing capabilities of the parabolic mirror and wide-angle capabilities of the spherical mirror provides scanning angles up to 120° when the displacement of the feed along the focal arc is employed. Such reflectors can be used to provide wide-angle coverage in very-large-aperture radar antennas, such as antenna systems of early-warning radars (e.g., the AN/FPS-50).

A shaped-reflector antenna uses a reflector shaped to produce the desired pattern and scan characteristics. Among the shapes used for radar applications, the spherical reflector has the shape of a sector of a sphere. It can provide wide-angle electromechanical scanning by feed displacement, but the construction is rather bulky and this antenna is used primarily when very large apertures are required (e.g., in radar astronomy). The Gregorian antenna configuration is the most typical for such antennas. A conical reflector antenna uses the reflector antenna of conical shape. It requires a special structure of feed to produce a conical wave illuminating the reflector that blocks the center of cone and gives large sidelobes, and typically are not used in radar applications. The patterns formed are typically fan or cosecant-squared beams, the latter using symmetrical or asymmetrical cut-outs from the paraboloidal reflector (truncated paraboloids) to form the required pattern. Doubly-curved reflectors are the most common configurations for cosecant-squared beams.
SAL, AIL

Ref.: Rusch (1970); Johnson (1984), pp. 17.1–17.54; Skolnik (1990), pp. 6.1–6.63; Sazonov (1988), pp. 371–394; Sletten (1988, 1991).

antenna scanner (see **SCANNER**).

A **signal-processing antenna** forms a radiation pattern in the receive mode by nonlinear signal processing or in both receive and transmit modes by means of multifrequency excitation or modulation of parameters. Signal-processing antennas typically include monopulse phased-array antennas with phase scanning and adaptation, logical synthesis antennas, and space-time antennas. *AIL*

Ref.: Bakhrakh (1989), p. 24; Fradin (1977), p. 345; Steinberg (1976), Chaps. 11, 12.

A **slot antenna** comprises one or several parallel narrow slots cut in current-carrying surfaces of differing shape. Many varieties of slot antennas have been designed. They are distinguished by their shape, size, positioning, and the shape of the surface in which they are cut. A slot system's radiation maximum is directed perpendicular to the plane in which the slot is cut. Several slots may be used to intensify the directional characteristics of the antenna. The ability to fabricate slot antennas flush to a metallic surface makes them convenient to use on airborne vehicles. They also are used as the elements of antenna arrays and as feeds for other complex antennas. *AIL*

Ref.: Johnson (1993), Ch. 9; Sazonov (1988), p. 253.

A **slotted-waveguide antenna** has the form of slots in the walls of a rectangular waveguide. The slots have a length close to $\lambda/2$, where λ is the wavelength. The voltage in the slot varies approximately quadratically, having a zero value at the edges and the maximum in the center. Multislot antennas, which have a series of slots cut in the wall of the waveguide at equal spacings, are used to create highly directional radiation. It is best to cut longitudinal slots in the narrow wall of a waveguide and transverse slots in the wide wall. Such slot placement leads to significant radiation pattern sidelobes. For this reason, it is most advisable to use a multislot antenna with longitudinal slots on the wide wall of the waveguide. The distance between slots and their tilt angle are selected so the radiation of all slots is cophasal. *AIL*

Ref.: Fradin (1977), p. 118; Johnson (1993), Chaps. 8, 9.

A **space-time antenna** is one having one or several parameters that are a function of time. This category includes, for instance, the synthetic aperture antenna. *SAL*

Ref.: Bakhrakh (1989), p. 17.

spiral antenna (see **log-spiral antenna**).

A **stacked-beam antenna** forms a multiple-beam pattern that covers the entire elevation search sector. Up to 1975 it was usually a doubly-curved reflector with an array of feed horns of front of it (Fig. A76). Modern stacked-beam antennas use

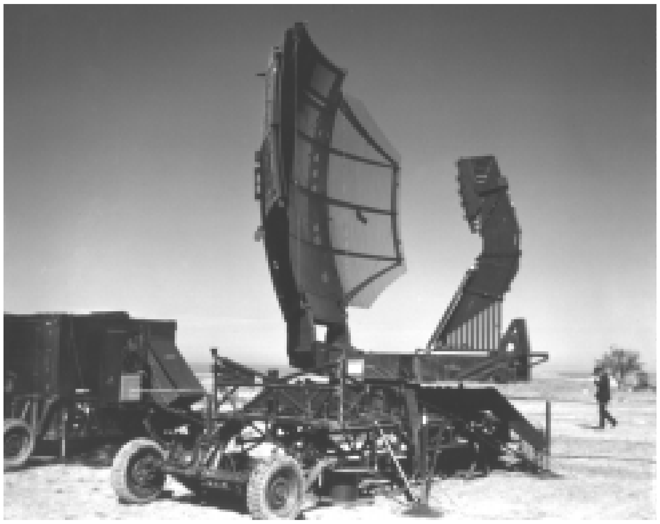


Figure A76 AN/TPS-43 stacked-beam reflector antenna.

planar arrays in which each row of elements is fed from an elevation beam-forming matrix. Mechanical rotation in azimuth is still used, especially when 360° coverage is required. The transmitter is normally connected through circulators to all beam ports, with coupling coefficients such as to produce a single cosecant-squared beam pattern.

The elevation beam-forming matrix can operate at RF, using a Rotman lens, a Butler matrix, or equivalent network; at intermediate frequency; or it can be a digital beam-former. *SAL*

Ref.: Barton (1988), pp. 196–198.

strip(line) antenna (see **microstrip antenna**).

superbroadband antenna (see **ultrawideband antenna**).

A **superdirective antenna** is a linear antenna array of finite length with a large number of radiators making it possible to realize an array factor as an arbitrary function. The capability of unlimited increase in directive gain of an antenna of finite length is referred to as antenna superdirectivity. Analysis of the solution demonstrates that, given number of elements, $N \to \infty$, element spacing, $d \to 0$, and constant array length $Nd = \text{const}$, the directive gain will increase without limit. However, computations and research illustrate that the transition to the superdirectivity mode is impractical because the increase in directive gain compared with normal directional antennas is extremely small. *AIL*

Ref.: Johnson (1993), p. 2.40; Sazonov (1988), p. 358; Mailloux (1994), pp. 19, 80.

surface-wave antenna (see **impedance antenna**).

A **synthetic aperture antenna** is one in which aperture synthesis is achieved based on the effect of the displacement of a physical antenna along the flight trajectory of a moving platform, using special signal processing. The physical antenna, which may take the form of a parabolic cylinder reflector, slotted waveguide, or a phased array, is small and has a sufficiently wide radiation pattern to illuminate the observed surface over a significant period of platform motion. The essence of the synthesis method is based on *a priori* information of the trajectory of motion of the antenna platform and involves the reception of signals during motion, remembering them, and subsequent coherent addition to form the high-resolution image. The sector of trajectory of the platform in which the signal is formed and processed is the synthetic aperture. This makes it possible to obtain high spatial resolution in the angular coordinate, corresponding to a conventional antenna with an aperture of hundreds or thousands of wavelengths.

Digital signal processing and an antenna with a digital synthetic aperture (see **RADAR, synthetic aperture**) are used for signal processing and for obtaining a radar image of the earth's surface directly on board the aircraft. Digital synthetic aperture antennas have several advantages, the main ones being: responsiveness; ability to synthesize the antenna aperture during random maneuvers of the airborne platform; automation of the detection and measurement of the coordinates of objects; and multiple reproduction of the recorded information. *AIL*

Ref.: Goryankov (1988), p. 22; Kondratenkov (1983), p. 40; Skolnik (1990), Ch. 21.

antenna temperature (see **TEMPERATURE**).

antenna testing (see **TESTING, antenna**).

A **test antenna** is one designed for measurement of antenna performance. Horns, dipoles, and paraboloid antennas are used as measurement antennas. Selection of antenna type depends on antenna measurement methods, parameters measured, and frequency band. *AIL*

Ref.: Strakhov (1985), p. 10.

A **transmitting antenna** is one designed for radiation of electromagnetic waves. For transmitting antennas, along with directivity characteristics, minimum losses of electromagnetic power to heating the antenna conductors and dielectrics, and high power-handling capability are of important significance. *AIL*

Ref.: Sazonov (1988), p. 201.

A **traveling-wave antenna** is one in which an axial radiation mode is realized. They are made on the basis of slow-wave structures capable of supporting surface waves. A traveling-wave antenna is excited through selection of slow-wave structure parameters. They usually have a frequency band from several to tens of percent.

The advantage of such antennas is the slight dimensions of the transverse cross section of the radiating system. This makes it possible to house them on the flat surface of the bodies of airborne vehicles. Dielectric rod, helical, impedance, and director antennas are categorized as traveling-wave antennas. These antennas are also called *end-fire antennas*, and previously were referred to as *surface-wave antennas*. *AIL*

Ref.: Johnson (1993), Ch. 12; Sazonov (1988), p. 302.

ultra-low-sidelobe antenna (ULSA) (see **low-sidelobe antenna**).

An **ultrawideband antenna** is one used for transmission and reception of ultrawideband (UWB) signals and operating with a signal bandwidth exceeding 50% of the center frequency. The conventional log periodic and helical broadband antennas cannot be used for transmission and reception of UWB signals. The reason is the strong dispersion of the phase-frequency response inherent to the aforementioned antennas and leading to significant distortion of the shape of the ultrawideband signals.

A plane log periodic dipole after appropriate modification can be used to transmit UWB signals (see **log-periodic antenna**). This modification involves changing antenna geometry (selection of dipole length) so the dipole resonant frequencies are subordinate to a law of arithmetic progression rather than the geometric progression found in the conventional log periodic antenna, that is,

$$\omega_n = \omega_1 + (n-1)\delta\omega, n = 1, 2, \ldots N$$

where ω_1 is the resonant frequency of the first dipole, $\delta\omega$ is the step of the progression, and N is the number of dipoles. Here, dipole lengths must be the following

$$l_n = \frac{l_1}{1 + \frac{(n-1)\delta\omega}{\omega_1}}$$

Such a change in the law of progression makes possible the transition from a logarithmic dependence of the phase-frequency response to a linear dependence. The aforementioned departure from logarithmic periodicity leads to an insignificant deterioration in the band properties of the antenna, but makes it possible to transmit a UWB signal without distortions. *AIL*

Ref.: Yatskevich, V. A., and Fedosenko, L. L., *Antennas for Radiation of Ultrawideband Signals*, Moscow, *Electronics*, vol. 29, 1986, p. 69 (in Russian); Taylor (1995), Ch. 5.

An **antennaverter** is an antenna with an embedded frequency converter.

wave-channel antenna (see **ARRAY, Yagi-Uda**).

wave-front antenna (see **aperture-type antenna**).

A **waveguide antenna** consists of a waveguide fed in the dominant mode and opening onto a conducting ground plane. In radar applications it is mostly used as a feed for reflector antennas or frequency-scanning arrays. *SAL*

Ref.: Fink (1982), pp. 18.23–18.26; Johnson (1993), Ch. 9; Barton (1988), p. 167.

Yagi antenna (see **ARRAY, Yagi-Uda**).

APERTURE, antenna. The antenna aperture is “a surface, near or on an antenna, on which it is convenient to make assumptions regarding field values for the purpose of computing the field at external points.” More generally, the aperture of an antenna is its physical area projected on a plane perpendicular to the mainbeam direction. Radar antennas are sometimes classified by the geometrical shape of their apertures (e.g., circular, elliptical, and rectangular). *SAL*

Ref.: IEEE (1993), p. 46.

An **active aperture** is one that incorporates active transmitter elements within the antenna structure itself, as opposed to a passive aperture, which focuses energy received from a source (transmitter or target) outside the antenna (see **ARRAY, active**). *PCH*

Aperture blockage occurs when a source of physical interference, usually the antenna feed structure of a reflector antenna, blocks a portion of the aperture from taking part in the exchange of radar energy. The blocked portion of the aperture is said to be shadowed and therefore will not contribute constructively to the field. The far-field antenna pattern will then consist of the sum of the intended pattern, plus three unintended components: blockage, feed scattering, and spillover. *PCH*

Ref.: Johnson (1993), pp. 17.32, 30.34–30.40.

A **circular aperture** can be a parabolic reflector, a lens, or an array having an active area that is circular in shape. The beamwidth of a circular aperture with uniform illumination over diameter D is $\theta_3 = 1.02\lambda/D$, and the first sidelobe level is –17.6 dB. Tapered illuminations are more commonly used, with beamwidth constants near 1.2 and sidelobes –20 to –25 dB. *DKB*

Ref.: Skolnik (1980), p. 233; Johnson (1993), p. 2.19.

A **continuous aperture** is one over which the illumination function is smooth, producing no grating sidelobes. The mesh of a reflector must be fine enough to prevent significant leakage of the wave through its surface. In the case of an array antenna, the spacing d between discrete elements must be such as to meet the criterion for avoidance of grating lobes at the maximum scan angle θ:

$$d \le \frac{\lambda}{1 + |\sin\theta|}$$

DKB

Ref.: Johnson (1993), p. 2.31.

aperture distribution (see **aperture illumination**).

The **effective aperture** is an important concept in radar antenna theory; it may be regarded as a measure of the effective area (see **antenna gain**) presented by the antenna to the incident wave. If G_r is the radar receiving antenna gain, and λ the wavelength of the radiation, then the effective aperture (or effective area) is $A_r = G_r\lambda^2/4\pi$. For a uniformly illuminated aperture, an ideal (lossless) antenna will have maximum directivity or gain; that is, the antenna takes full advantage of the physical area A and the antenna gain is at its maximum value, G_0. Then $A = G_0\lambda^2/4\pi$. *PCH*

Ref.: Johnson (1993), p. 1.6.

Aperture efficiency, η_a, relates the effective aperture area, A_r, to the physical antenna area A; that is, $A_r = \eta_a A$. Aperture efficiency is also equal to the ratio of the actual directivity of an antenna to the maximum possible directivity of that antenna: $\eta_a = G_r/G_0$. In a rectangular aperture with separable illuminations, the aperture efficiency will have two components, $\eta_a = \eta_x\eta_y$. *PCH*

Ref.: IEEE (1993), p. 47; Johnson (1993), p. 1.6.

An **elliptical aperture** can be a parabolic reflector, a lens, or an array having an active area that is elliptical in shape. The analysis is performed as with a circular aperture, but the beamwidth and sidelobe level in each principal plane will depend on the dimension D and the illumination function in that plane. *DKB*

Ref.: Barton (1988), p. 155.

aperture excitation (see **aperture illumination**).

Aperture illumination is the electric field distribution across an antenna aperture. It is also called the *aperture distribution*, *excitation*, *taper*, or *weighting function*. If this function is known, the radiation pattern, or electric field intensity, can be found as a function of x, y, and z coordinates relative to the antenna centroid. The far-field (Fraunhofer) electric field intensity for a one-dimensional aperture of width a in the z dimension (perpendicular to the direction of the main beam), where $a \gg \lambda$, can be expressed by the Fourier transform

$$E(\phi) = \int_{-a/2}^{a/2} A(z)\exp\left(j\frac{2\pi z}{\lambda}\sin\phi\right)dz$$

where $A(z)$ is the aperture illumination, or current at distance z, flowing in the x-direction, and ϕ is the angle off beam center in the *y-z* (azimuth) plane. $A(z)$ is a complex function, having both amplitude and phase:

$$A(z) = |A(z)| \exp j\Psi(z)$$

where $|A(z)|$ = illumination amplitude and $\Psi(z)$ = illumination phase. At an angle ϕ, the contributions from a particular point on the aperture will be advanced or retarded in phase by $2\pi(z/\lambda)\sin\phi$ radians. Each of these contributions is weighted by the factor $A(z)$. The field intensity is the integral of these individual contributions across the face of the aperture.

For a given field intensity pattern, $E(\phi)$, the aperture illumination can be found from

$$A(z) = \frac{1}{\lambda}\int_{-\infty}^{\infty} E(\phi)\exp\left(-j\frac{2\pi z}{\lambda}\sin\phi\right)d\sin\phi$$

which may be used as the basis for synthesizing an antenna pattern, or finding the aperture illumination that yields the desired antenna pattern $E(\phi)$.

Maximum antenna gain and minimum beamwidth are realized with a uniform aperture illumination; *i.e.*, one that is constant over the aperture and zero everywhere else. Such illumination will lead to an antenna pattern of the $(\sin x)/x$ form, with the intensity of the first sidelobe 13.3 dB below peak gain. There exists a large variety of useful illumination functions, and low sidelobes can be obtained by using functions such as a cosine or a Taylor function. Antennas with the lowest sidelobes are those with illumination functions for which the amplitude tapers to a small value at the aperture edges. Lower sidelobes are produced with greater amplitude tapering, but at the expense of a wider mainbeam width and lower mainbeam gain. Controlling the antenna pattern in this way is referred to as *aperture tapering*. A table describing many of the functions commonly used, and listing the sidelobe levels, beamwidth factors, and efficiencies, is given in the article on **WEIGHTING**. Some typical illumination functions are given in Table W3. (see also **PATTERN, antenna**). *PCH, SAL*

Ref.: Skolnik (1990), p. 7.37.

aperture illumination efficiency (see antenna directivity).

aperture matching (see **ARRAY aperture matching**).

A **passive aperture** is one that focuses energy received from a true source outside the antenna (as opposed to an active aperture).

A **rectangular aperture** consists of rows and columns of radiating elements with an aperture illumination defined along those coordinates. The beamwidth of a uniformly illuminated rectangular aperture of width w is $\theta_3 = 0.886\lambda/w$, and the sidelobe level is –13.3 dB. In many cases, the illumination functions in the *x*-and *y*-coordinates are separable:

$$g(x, y) = g(x)g(y)$$

In that case, the antenna patterns in each coordinate may be calculated separately, each being the Fourier transform of the corresponding illumination function.

While the physical shape of a reflector or lens antenna may be rectangular, the aperture distribution (illumination) function produced by a feed horn generally has an elliptical shape, leading to beamwidths and sidelobe levels that are better predicted as resulting from an elliptical aperture. *DKB*

Ref.: Johnson (1993), p. 2.19.

A **synthetic aperture** is created by taking advantage of the motion of a radar platform to extend the effective antenna aperture beyond the limits of its actual physical dimensions. In a side-looking synthetic aperture radar (SAR), the effective beamwidth of the radar is no longer described by conventional (real) antenna relationship $\theta_3 = K_\theta\lambda/D$, where the beamwidth constant $K_\theta \approx 1$ depends on aperture weighting, but by $\theta_s = K_\theta'\lambda/L_e$, where L_e is the effective length of the aperture and $K_\theta' \approx 0.5$. Narrowing of the beamwidth in a synthetic aperture radar is subject to certain fundamental constraints: (1) the length L_e can be no longer than the width of the region illuminated by the real aperture: $L_e \le R\theta_3$, and (2) $L_e \le (R\lambda)^{1/2}$. The second limitation applies to an unfocused SAR, where the aperture size must be such that the phase front can be considered a plane wave. This limitation can be removed in a focused SAR by compensating for the curvature of the spherical wavefront (*i.e.*, applying a phase correction at each "element" of the synthetic array). See **RADAR, synthetic aperture**; **ANTENNA, synthetic aperture.** *PCH*

Ref.: Skolnik (1980), Ch. 7; Skolnik (1990), Ch. 6

aperture tapering (see **aperture illumination**).

APPROXIMATION

detection probability approximation (see **DETECTION probability**).

The **flat-earth approximation** is applicable to very-short-range targets in height finding, giving a sufficiently good estimate of target height as

$$h_t = h_r + R\sin\theta_t$$

where h_a is the radar antenna height, R is the measured target range, θ_t is the measured or estimated target elevation angle, and h_t is the estimated target height. *SAL*

Ref.: Skolnik (1990), p. 20.14.

The **four-thirds earth radius approximation** expresses the refractive effects of the troposphere in terms of straight ray paths over a spherical earth whose radius is ka, where $k = 4/3$ and $a = 6.5 \times 10^6$ is the true radius of the earth (see **ATMOSPHERE, refraction**). The refractive index implied by this model varies linearly in the lower atmosphere:

$$n(h) = n_0 - k_1 h$$

where $n_0 \approx 1.000313$ is the refractive index of the atmosphere at the surface and $k_1 \approx 4 \times 10^{-8}$ per meter is the assumed gradient of refractive index with respect to height, h. In general, the factor k can be expressed for arbitrary value of the gradient as

$$k = \frac{1}{1 - k_1 a / n_0}$$

that results in $k = 4/3$ for paths at low or medium altitude in the standard atmosphere *DKB, SAL*

Ref.: Blake (1980), p. 184.

North's approximation (see **DETECTION probability**).

The **spherical earth (parabolic) approximation**, used in height finding, accounts for the earth's curvature as parabolic in range and gives a target height for a radar located near the surface of the earth as

$$h_t = h_a + R \sin\theta_t + \frac{R^2}{2ka}$$

where a is the radius of the earth, $k \cong 4/3$ is the factor taking into account refraction (see **four-thirds earth radius approximation**) and the other parameters are defined in the flat-earth approximation. *SAL*

Ref.: Skolnik (1990), p. 20.14.

ARRAY (ANTENNA). An array antenna is "an antenna comprised of a number of identical radiating elements in a regular arrangement and excited to obtained a prescribed radiation pattern." The main types of array antennas used in radar applications are phased arrays and frequency-scanned arrays. *SAL*

Ref.: IEEE (1993), p. 55.

An **active array** is one in which an active element (oscillator, amplifier, or mixer) is connected to the path of each radiator. These elements, along with the radiator, form the array module. Active antenna arrays are categorized as receiving, transmitting, and transceiving. Active antenna array advantages include the capability to increase radiated power, decrease thermal losses, increase reliability, and reduce the length of the paths between radiators and transceiving circuits (see also **amplifier array**). *AIL*

Ref.: Sazonov (1988), p. 396; Mailloux (1994), p. 41.

An **adaptive array** consists of an N-element array (usually in the receiving mode), where the useful signal is maximized based on an analysis of the signal-to-interference ratio. An important aspect of adaptive arrays is the appropriate choice of weighting coefficients $W(t)$, which are placed between the antenna elements and a combining network. In the general case, the vector $W(t)$ must have the capability of changing the amplitude and phase of the received signal from each element. The rates of change must correspond to the rates of change of the signal-to-interference ratio, and the range of change must correspond to the dynamic range of the levels of signal and interference, and the range of phase relationships between the different array elements.

The criteria for array performance in suppression of interference may be (a) the maximum ratio of signal to interference at the output of the array; (b) the minimum mean square deviation of the received signal from the given reference level at the output of the array; (c) minimum interference power at the array output; or (d) the maximum probability of detection of the desired target signal. *AIL*

Ref.: Bakhrakh (1989), p. 167; Steinberg (1976), Chaps. 11, 12; Mailloux (1994), pp. 167–186.

An **amplifier array** is one with the final transmitter amplifier and the first receiving amplifier placed at the array element. In constrained feed systems, these amplifiers may be placed at any level in the dividing network: at the individual radiating element, at row or column level, or at subarrays. An advantage of placing amplifiers at the element is that the phase shifter may be placed on the feed side of the amplifier, reducing its power rating and the effect of phase shifter loss on system performance. In a typical solid-state modular array, a common phase shifter at each element amplifier is switched between the transmit and receive paths, while the radiating element is connected to the amplifier through a circulator (Fig. A77).

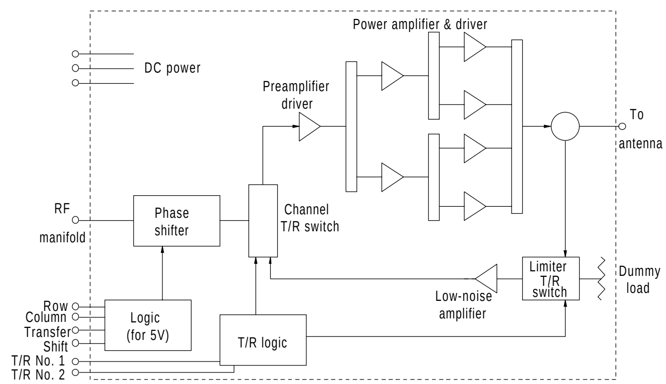


Figure A77 Typical phased-array transmit-receive module (after Brookner, 1977, Fig. 3, p. 266).

The total power of the amplifier array is limited only by the available prime power, the RF power that can be generated within the volume associated with each array element, the heat that can be dissipated from this volume, and cost considerations. *SAL*

Ref.: Brookner (1977), Chaps. 19, 20; Barton (1988), pp. 179–181; Mailloux (1994), p. 41.

annular array (see **ring array**).

An **array of arrays** is a term sometimes used to define a two-dimensional array consisting of a number of identical linear arrays.

Ref.: Johnson (1993), p. 3.29.

Array bandwidth is the range of frequencies within which the array performance meets specified requirements. The basic elements establishing array bandwidth are the radiating elements, phase shifters, and feed networks. Most radiating elements are well matched over a broad band of frequencies, and hence the main limitations to array bandwidth are the feed networks and phase shifters. *SAL*

Ref.: Johnson (1984), p. 20.60.

The **array blinding effect** is the inability of a phased antenna array to radiate in a given direction, caused by the mutual coupling of the radiators. Depending on the structure of the radiating elements, the blinding effect is explained either by the occurrence of concealed resonance owing to the propagation of a surface wave in the structure of the radiators or by the suppression of the field in the aperture as a result of the excitation in the array waveguides of higher frequency waves caused by the asymmetry of the external field. The blinding phenomenon is observed when orienting the beam of the array within the limits of the operating scanning sector and therefore, it leads to the reduction of the scanning sector of the phased antenna array. *IAM*

Ref.: Voskresenskiy (1981) p. 27; Mailloux (1994), pp. 339–355.

A **broadside array** is one having N elements, producing a beam with the mainlobe perpendicular (broadside) to the array.

Ref.: Sauvageot (1992), p. 29; Johnston (1984), p. 3.1.

conformal array (see **ANTENNA, conformal**).

A **conical array** is one whose radiators are positioned on a conical surface. In conical arrays it is possible to scan with a constant pattern and gain in the principal plane normal to the cone axis, and with the usual gain reduction and beam broadening over the sector in the plane of the generator of the cone. In hemispherical scans, some sections of the conical surface radiate at very large angles, and it is therefore expedient to use less than 0.5λ for the array element spacing. To use the radiators efficiently, their beam axes are pointed *not* at the normal to the generator of the cone but at the direction required to have the maximum gain. To obtain narrow beamwidths, the number of elements in a conical array must be of the order of 10^4. Such antennas are used in cases where it is required to position a hemispherically scanning array on the conical body of an aircraft (or other flying vehicle), and when the maximum gain must lie in the axial direction or in a direction near the axis of the cone. *AIL*

Ref.: Voskresenskiy (1981), p. 153; Johnson (1993), p. 21.20; Mailloux (1994), p. 234.

A **cylindrical array** is one whose radiators are positioned on a cylindrical surface. The radiators used are wire and slot dipoles, open-ended waveguides and horns, and spiral and dielectric rod antennas. The selection of the type of radiators depends on the wavelength and the required bandwidth, on the application and the operating conditions, and on the construction requirements of the array. To obtain a narrow beamwidth, the number of elements must be close to 10^4. Cylindrical arrays are used in cases where azimuthal scanning with a constant beam shape and gain is required. *AIL*

Ref.: Voskresenskiy (1981), p. 82; Johnson (1993), p. 21.9; Mailloux (1994), pp. 194–233.

A **digital array** is a phased array in which the signal received by each antenna element is converted into digital code and further processing (including forming of the antenna pattern and signal processing is performed in a digital computer. Practically it consists of transceiver modules (each employing an array radiator, transceiver, and analog-to-digital converter), a digital computer, and a data bus connecting the elements with the computer (Fig. A78). The computer performs digital beam-forming, including all operations to generate the antenna pattern and to control the beam shape and its direction in space. For these purposes, a general-purpose or special digital computer can be used executing adaptive beam-forming algorithms, discrete Fourier transforms, pattern correction, and other required calculations.

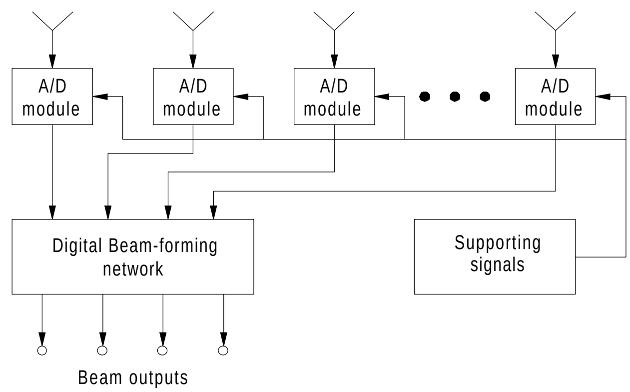


Figure A78 Block diagram of a digital array antenna.

The basic advantages of the digital array as compared with the conventional receiving array are the capabilities for:

(1) Instantaneous shaping of the array pattern in any direction;
(2) Generation of monopulse patterns with different width and crossover level;
(3) Generation of adaptive array patterns of arbitrary shape;
(4) Fast cancellation of array distribution errors resulting in a very stable shape of the antenna pattern and accurate positioning of the beam in space.

The difficulties in practical realization of digital phased array technology lie in the complexity of hardware and software implementation. Hundred or thousands of transceiver modules with complicated transmission links are used, requiring advanced construction technology (the use of compact fiber-optic transmission lines is considered promising). More efficient and less time-consuming computing algorithms for target detection and measurement have to be designed to reduce the complexity of the required software. *SAL*

Ref.: Leonov (1988), p. 164; Bakhrakh (1989), pp. 88–98; Skolnik (1990), p. 7.8.

Array directivity is "the ratio of power density per unit solid angle at the peak of the main beam to the average power radiated per unit solid angle over all space." For a linear array of N elements spaced a distance d apart, the directivity, D, in the broadside direction is

$$D \approx \frac{2\eta N d}{\lambda}$$

where η is the aperture efficiency and λ is the wavelength. For a two-dimensional array,

$$D \approx 4\pi\eta \frac{Nd_x d_y}{\lambda^2} \cos\theta_0$$

where d_x and d_y are the spacings in the two coordinates and θ_0 is the angle between the main beam and the normal to the array surface (see also **ANTENNA directivity**). *SAL*

Ref.: Johnson (1984), pp, 20.15, 20.22.

A **dome array** is a lens array that provides hemispherical coverage. The simplest configuration of a dome array consists of a dome (constrained-constant-thickness lens) and a planar array that feeds the lens (Fig. A79). An alternative to a constrained lens is the use of a homogenous dielectric material with graded thickness in the vertical plane. This type of antenna was developed as an alternative to the use of several planar arrays to scan a beam over a full hemisphere. *SAL*

Ref.: Johnson (1984), p. 16.23.

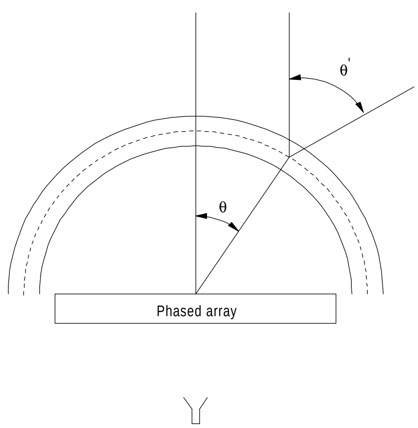


Figure A79 Dome antenna configuration (after Johnson, 1993, Fig. 16.19, p. 16.24).

A **dual-polarized array** represents a construction of linear arrays connected by their individual lines to a common microstrip feed line. Each linear array is terminated in a matched load. The array is fed from a single source. The creation of the corresponding distribution of fields across the array aperture is achieved by placing tabs of different sizes along the linear arrays and by varying the widths of the feed lines. The advantage of such arrays lies in their simplicity, small size and low weight. A shortcoming is their narrow bandwidth. *AIL*

Ref.: Leonov (1988), p. 168.

An **array element** is a radiating element (a small independent microwave antenna) the set of which constitutes the array aperture. The most widely used are dipoles, slots, small horns, and waveguides (see **RADIATING ELEMENT**); spirals, microstrip disks or patch elements are also employed. In array theory, radiators are typically considered to have broad isotropic patterns. In practice, radiation patterns of real radiating elements are nonisotropic, and a radiator in the array environment has a pattern that differs from the pattern of isolated element in amplitude, phase, and sometimes in polarization also (see **mutual coupling in array**). Mutual coupling results in the phenomenon wherein radiator impedance varies as a function of scanning. The selection of the radiator for particular array is based on the consideration of its physical dimensions and environmental requirements, polarization and power-handling capability, and appropriate aperture matching over the required scanning range. *SAL*

Ref.: IEEE (1993), p. 55; Johnson (1984), pp. 20.25–20.31.

An **end-fire array** is one in which the element phase settings are such as to radiate a beam along or near the plane containing the elements. The gain and beamwidth of optimized end-fire arrays of length L are

$$\theta_3 \cong 0.96\sqrt{\frac{\lambda}{L}}$$

$$G \cong \frac{7L}{\lambda}$$

This array is also called a *traveling-wave* or *surface-wave antenna*. *SAL*

Ref.: Johnson (1993), Ch. 12.

The **array factor** is the phased array pattern $f_a(\theta, \phi)$ when the element factor $f_e(\theta, \phi)$ is isotropic. The complete radiation pattern of phased array is the product of the array factor and the array element factor, $f(\theta, \phi) = f_e(\theta, \phi)\, f_a(\theta, \phi)$. The array factor characterizes the directivity capabilities of an array as a system of the radiators (see **PATTERN, array**). *SAL*

Ref.: IEEE (1993), p. 55; Johnson (1984), p. 20.5; Skolnik (1990), p. 7.10; Leonov (1988), p. 36.

array feed network (see **FEED**).

A **fiber-optic array** is one in which the link (analog or digital) among separate phased-array antenna elements is accomplished using fiber-optic transmission lines. The latter are used in both passive (Fig. A80) and in active phased-array

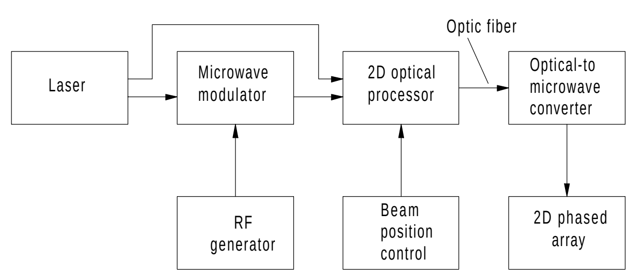


Figure A80 Passive phased-array antenna with optical circuits.

antennas. Meanwhile, in active arrays, all signals going to the transceiver module or in it may be transmitted via one optical fiber. Figure A81 provides a possible diagram of the distribution of signals in an active fiber-optic phased-array antenna.

Use of fiber-optic phased-array antennas is especially promising in millimeter-wave radars. *AIL*

Ref.: Zmuda (1994), Ch. 11.

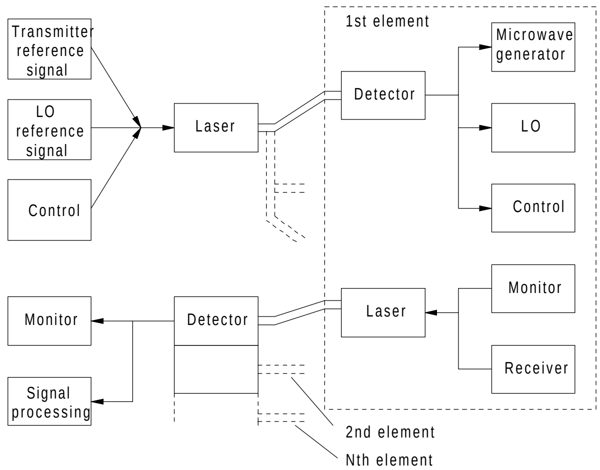


Figure A81 Signal distribution in an active fiber-optic phased-array antenna.

A **frequency-scan array** is one "in which the direction of the radiated beam is controlled by changing of the operating frequency." The frequency scan principle is based on the fact that the change of frequency produces a change in phase of the signal passing through the length of a transmission line (see **SCANNING, frequency**). Typically, the array is passive radiators are excited directly from the beam-forming network. Either serial or parallel feed configurations can be used. The most common types of feeds are the sinuous feed (and tandem-sinuous feed for monopulse technique) and the dummy-snake feed. A practical example of frequency-scan array with dual scan bands is the AN/SPS-48 (Fig. A82).

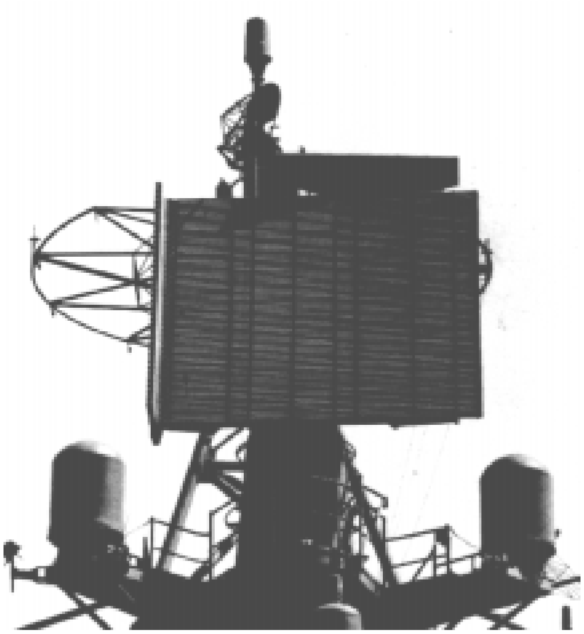


Figure A82 AN/SPS-48 frequency-scanning 3D radar antenna.

The disadvantage of the conventional frequency scan array lies in the fact that when the entire available bandwidth is used to steer the beam, then each direction in space is associated with a definite frequency. The antenna becomes vulnerable to jamming since a jammer can concentrate its energy over a narrow frequency band. The most practical solution lies in frequency agility when a multiple-beam-forming network is used with multiple interleaved arrays. See also **RADAR, frequency-scan; PATTERN, antenna**. *SAL*

Ref.: Johnson (1984), p. 19.1; Skolnik (1980), p. 298; Voskresenskiy (1981), p. 64.

A **linear array** is one consisting of a group of identical elements placed in one dimension along a given direction. Linear arrays may have equidistant or nonequidistant element spacings. They are used in the analysis of the directional properties of arrays in antenna theory (see **PATTERN, array**), and as building blocks for forming an array of arrays. *AIL*

Ref.: Venenson (1966), p. 72; Mailloux (1994), pp. 72–81; Johnson (1993), pp. 3.1–3.29, 20.3–20.15.

A **low-sidelobe array** is one for which the antenna sidelobe level is maintained below some specified level (see **ANTENNA, low-sidelobe**). Since for phased arrays the amplitude of each element can be controlled individually, good sidelobe control in comparison with reflector antennas can be achieved. The sidelobe reduction is achieved at the expense of gain reduction and increased beamwidth. It also increases the cost of the antenna and the complexity of tolerance control, and it imposes requirements for operation in an environment free from obstructions that would cause sidelobe increase. *SAL*

Ref.: Brookner (1988), Ch. 6; Skolnik (1990), p. 7.37.

Array (aperture) matching is the matching of the array with its feed network and with free space. Usually wide-angle matching is used, which allows the improvement of the characteristics at all scanning sectors of the array. Methods of wide-angle matching of a phased array can be divided into two groups: (1) methods related to a modification of the physical construction of the array excitation and (2) methods based on placing dummy elements in front of the array aperture, whose reflection reduces the change in the output impedance of the radiators during scanning.

Group 1 includes the following ways of wide-angle matching:

(1) Use of connecting circuits between elements.

(2) Filling of the waveguide horns with dielectric. The size of the waveguide horns and the parameters of the filling medium are selected so that a wave can propagate in only a single mode.

(3) Use of waveguide radiators with several types of propagating modes.

Group II uses the following methods:

(1) Placing a conducting partition parallel to the E-plane of the dipole radiators. These partitions reduce the change in the reflection coefficient during scanning in the E-plane but do not effect the match during scanning in the H-plane.

(2) Placing a thick dielectric plate over the array aperture.

(3) Using a thin dielectric sheet having a high dielectric permeability, positioned a small distance from the waveguide array.

(4) Using a close spacing of the radiating elements. The most wideband methods are those that reduce element spacings, utilize thin dielectric sheets, and use partitions between the elements. *AIL*

Ref.: Oliner (1972); Voskresenskiy (1981), p. 38; Skolnik (1990), p. 7.22; Mailloux (1994), pp. 367–387.

A **microstrip array** is one using integrated circuit technology. The aperture of such arrays is formed by using microstrip elements, positioned at small spacings on a conducting ground plane. Between the ground plane and the microstrip elements is a thin dielectric substrate. A problem in designing such arrays is the placement of all necessary components on the ground plane. As a solution, the dimensions of microstrip radiators, phase shifters, power dividers, and so forth, are reduced. Sometimes the array extends over several ground planes. For the dielectric substrate there is a wide variety of materials having good mechanical, electrical, and temperature characteristics. In the development of microstrip arrays, computer-aided design is widely used. A photo mask of a millimeter-wave microstrip array is shown in Fig. A83.

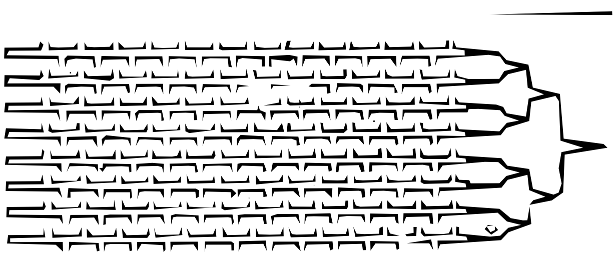


Figure A83 Photo mask for a microstrip array antenna (from Leonov, 1988, Fig. 5.4, p. 168).

Basic applications of microstrip arrays are found in radars in centimeter and millimeter wave bands. *AIL*

Ref.: Johnson (1993), Ch. 7; Sazonov (1988), p. 258; Zurcher (1995), Ch. 2.

A **Mill's cross array** consists of two linear arrays (Fig. A84), placed perpendicularly to each other. The central elements of the two arrays coincide. The pattern of this array is formed after processing the signals by multiplication. The beamwidth depends on the number of elements N_1 and N_2 in each linear array, and the spacing between the elements d_1 and d_2. The array antenna is named after its inventor. *AIL*

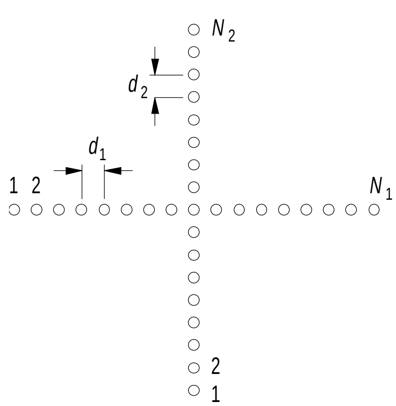


Figure A84 Mill's cross antenna array.

Ref.: Steinberg (1963), pp. 78–82; Fradin (1977), p. 348.

An **array module** is a device comprising an amplifier, active elements, and the elements controlling them. A module has small transverse dimensions (0.6λ to 0.7λ). Semiconductor devices and integrated circuits based on microstrip radiators and microstrip transmission lines are used in these modules. Modules are used in active arrays.

Figure A85 shows a possible structure of a transceiving module. The same radiator serves for reception and radiation of signals in transceiving active arrays. The module receiving channel comprises a limiter, a low-noise amplifier, and a control phase shifter.

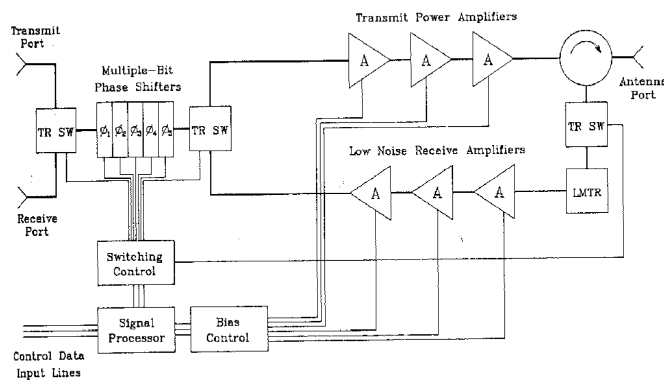


Figure A85 Transceiving module diagram (from Skolnik, 1990, Fig. 5.9, p. 5.17, reprinted by permission of McGraw-Hill).

The transmitting channel comprises a phase shifter and a power amplifier. Power switches that switch the heterodyne and antenna in "receive" and "transmit" modes are common to both module channels. Transmitting and receiving modules are built using the same principles examined above.

Antenna array modules usually have unequal amplitude and phase responses. Phase autotuning and amplification stabilization circuits are included in each module to eliminate this shortcoming. As a result, phase and amplitude errors are reduced to acceptable values. *AIL*

Ref.: Voskresenskiy (1981), pp. 247–258; Fradin (1977), pp. 343–345; Brookner (1977), Chaps. 19, 20.

A **monopulse array** is one supporting monopulse angle sensing in azimuth and elevation planes. It can be either a space-fed or constrained-feed phased array, although the generation of appropriate monopulse patterns in the latter case is a more difficult problem. A basic problem in generation of monopulse patterns is the compromise between efficiency of sum and difference patterns and reasonable sidelobe levels. The solution is typically a choice of appropriate feed configuration (e.g., multihorn or dual-ladder feeds). An example of a monopulse space-fed is the *Grill Pan* shown in Fig. F18, and a constrained-feed monopulse array is the AN/SPY-1 shown in Fig. A86. *DKB, SAL*

Ref.: Barton (1988), p.198; Leonov (1986), p. 23.

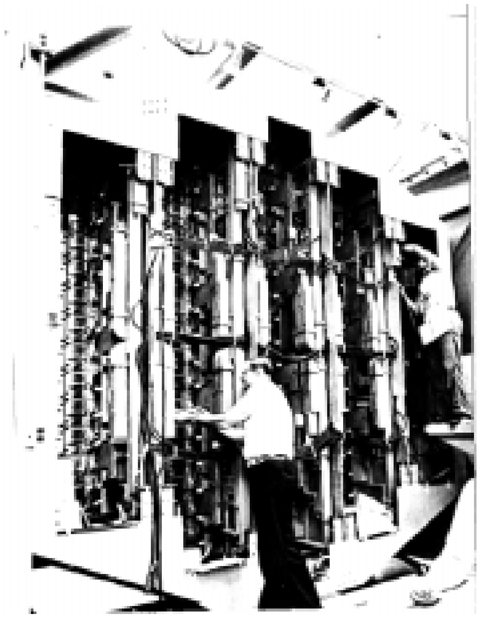


Figure A86 AN/SPY-1 Aegis monopulse phased-array antenna viewed from the rear, during assembly.

A **multibeam array** supports the generation of several beams that can be used simultaneously for surveillance of a given sector. Each beam has its corresponding separate input channel. The basic element supporting generation of several beams is the multiple beam-forming network.

The essential elements of the beam-forming networks are quadrature directional couplers and fixed phase shifters. The general shortcomings of multibeam arrays using beam-forming networks are the large numbers of connecting structures, phase shifters, and complicated branching feeds. *AIL*

Ref.: Bakhrakh (1989), p. 142; Sazonov (1988), p. 405; Mailloux (1994), pp. 167–182, 423–446.

A **multifaced array** is a system of planar arrays. It can be either a set of two to four independent planar arrays covering different azimuth sectors, or a set of subarrays arranged on the faces of a convex polyhedron and interconnected to produce a single scanning beam. If the number of subarrays is ten or fewer, they are usually positioned on the faces of a regular or truncated pyramid. For larger numbers of subarrays, they are located on the faces of regular polyhedra. The patterns of pyramidal phased arrays are similar to those of planar arrays. Multifaced phased arrays of this type are quasiconformal arrays. *AIL*

Ref.: Voskresenskiy (1981), p. 146; Mailloux (1994), pp. 231–233.

A **multifrequency array** is one that operates over several frequency bands. There are several approaches to achieving multifrequency operation:

(1) Multifrequency, overlapping arrays formed by the merging of one array into another in such a way that all the radiators are located in the same aperture, and the radiators for one frequency band are dispersed between the radiators of the other bands.

(2) Convex multifrequency arrays with distributed multifrequency radiators positioned on convex, curvilinear surfaces.

(3) Multifrequency arrays constructed using multifrequency or wideband radiators and frequency separation filters.

(4) Multifrequency combined arrays that operate as ordinary arrays in higher frequency bands, while at lower frequencies the array aperture forms a shaped impedance structure that is illuminated by a feed positioned at the periphery of the aperture. *AIL*

Ref.: Bakhrakh (1989), p. 113.

Array mutual coupling is the effect of coupling among adjacent radiators. The degree of coupling depends on the distance between elements, the pattern of the individual elements (element factor), and the structure in the vicinity of the element. The main effect of mutual coupling is the change of impedance and pattern of the element, which vary as functions of scan angle. The coupling is typically significant for elements spaced up to several wavelengths from each other. It can be described by a mutual-coupling coefficient which relates the voltage induced in the *mn*th element to the voltage at the *pq*th element. *SAL*

Ref.: Johnson (1984), pp. 20–25; Skolnik (1990), p. 7.23.

A **passive (antenna) array** is one in which all elements are excited from a common oscillator or connected to a common receiver. Therefore, an immutable part of a passive array is the feed network connecting the elements. Passive antenna arrays are categorized as receiving, transmitting, and transceiving. They are finding wide use in variable-purpose radars. *AIL*

Ref.: Sazonov (1988), p. 397; Mailloux (1994), p. 40.

array pattern (see **PATTERN, array**).

A **phased array** is an "array antenna whose beam direction or radiation pattern is controlled primarily by the relative phases of the excitation coefficients of the radiating elements." Physically it is composed of a group of individual elements that are arranged in a linear or two-dimensional (typically planar) spatial configuration. Usually the following basic features are used to classify phased arrays:

(1) Scanning methods,

(2) Radiator feed methods;,

(3) Positioning of radiators in the array.

The main scanning methods are phase scanning and frequency scanning. Typically, in radar literature, it is the phase-scanned arrays that are referred to as *phased arrays*. From the viewpoint of feed methods, arrays are divided into constrained-feed arrays and space-fed arrays, the latter taking the form of reflectarrays or transmission arrays (see **array feed networks**). With regard to element positioning, phased arrays are divided into uniformly spaced and unequally spaced arrays.

Phased arrays are the most advanced type of antenna used in modern radars. They provide the radar with flexibility and adaptation to the assigned task: ability to change beam position in space almost instantaneously (electronic scanning); generation of very high powers from many sources distributed across the aperture; high directivity and power gain; possibility of synthesizing any desired radiation pattern (including formation of pattern nulls in the directions of undesired interference sources); capability of combining search, track, and recognition functions when operating in multiple-target and severe interference environments (including jamming); enhanced target throughput capability; and compatibility with digital computers and digital signal processing algorithms. On the other hand, they are the most complicated and expensive types of modern antennas. (See also **PATTERN, array**). *SAL*

Ref.: IEEE (1993), p. 941; Johnson (1993), Chaps. 19, 20; Skolnik (1990), p. 7.1; Sazonov (1988), p. 396; Amitay (1972); Mailloux (1994).

A **phase-frequency array** is one using frequency scanning in one coordinate and phase scanning in the other.

Ref.: Skolnik (1980), p. 303.

A **phase-phase array** is one that uses phase shift to steer the beam in both coordinates. This array is typically referred to as a *phased array* in the literature.

Ref.: Skolnik (1980), p. 303.

A **piece-wise linear array** is one consisting of a combination of flat modular subarrays in the shape of an incomplete cube, octahedron, or other regular polyhedron. The basic difficulty in creating these arrays lies in the electrical combination of the individual subarray modules without incurring excessive energy losses. They have found applications in receiving antennas for the formation of spatial directivity characteristics as a resource for coherent optics and holography. *AIL*

Ref.: Voskresenskiy (1986), p. 70.

A **planar array** has all elements located in a single plane occupying a definite area. Planar arrays have different configurations of elements: rectangular, triangular, or hexagonal, in which the elements are positioned at the vertices of the rectangles, right triangles, regular hexagons and also at the center of the hexagon. *AIL*

Ref.: Fradin (1987), p. 184; Mailloux (1994), pp. 20–27, 34, 81–87, 112–162.

A **quasiconformal array** is one whose radiating elements are positioned on the surface of a polyhedron having a large number (up to 400) of nearly identical facets having identical flat subarrays. The subarrays can be realized in the form of a belt or as polygonal structures. The number of radiators in each subarray is determined by the required total quantity of radiators in the array, the minimum allowable number of subarrays N_{sa}, and the convenience of usage. Quasiconformal phased arrays are used to minimize the gain of the oscillators required in the scan region. Small oscillator gains are achieved by displacing the excited regions around the center of symmetry of the array with the help of electrical commutators. The larger the number of subarrays, the smaller the gain of the individual oscillator. For N_{sa} = 100, the parameters of a quasiconformal phased array can be approximated by those for a spherical array. *AIL*

Ref.: Voskresenskiy (1981), p. 159.

array radiator (see **RADIATING ELEMENT**).

A **radio-optical array** is an active array that uses an active antenna array in which the processing of the received radio signals occurs at optical frequencies. A top-level block diagram of a radio-optical array is given in Fig. A87. Shown are the array (1) and a spatial-temporal optical modulator (2). This modulator imparts a spatial-temporal modulation to the phase of the coherent light output from the laser (3) corresponding to the parameters of the signals received from objects imaged with the same from the output optical model of the received radiation. This is followed by a transformation in a coherent processor (4). As a result, an optical image of the received spatial-temporal signal is formed at the processor output. This information is registered on the photo detector (5) and with the help of converter (6) is input to the computer (7) or the analog device for final processing and decision-making.

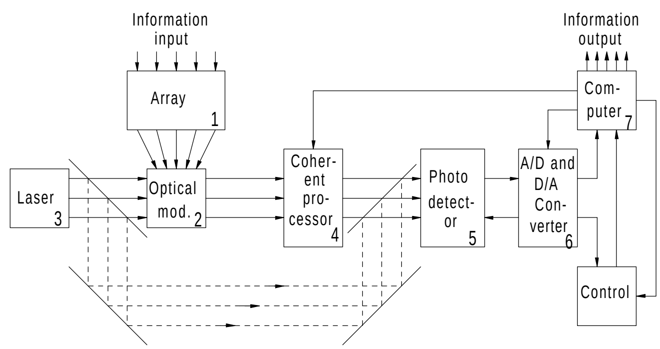


Figure A87 Block diagram of a radio-optical array.

Depending on the type of light modulator used, one distinguishes the following radio-optical arrays:

(1) Multichannel acousto-optical arrays.

(2) Electro-optical modulators with electron beam addressing.

(3) Multichannel modulators addressed by an electrical voltage.

Addressing is the mechanism by which the information is supplied to the light modulator.

Radio-optical arrays have a number of advantages compared with ordinary antenna arrays. Radio-optical arrays permit: the realization of parallel surveillance by placing receive elements on flat or curved surfaces; real-time signal processing and simultaneous panoramic surveillance in one spatial dimension; and efficient suppression of jamming signals in the receiving direction. These arrays do not require phase shifters, do not have bulky beam-forming networks, and are wideband.

The shortcoming of radio-optical arrays includes a widening of the beam that results in a slight decrease in directional gain. *AIL*

Ref.: Voskresenskiy (1986), pp. 14–28; Zmuda (1994), Ch. 11.

A **reflectarray** is a space-fed phased array in which the elements are illuminated from the front by a feed and reradiate a controlled phase front that produces a beam scanned relative to the broadside direction (see **FEED, space**).

re(tro)directive array (see **Van Atta array**).

A **ring array** is one whose radiators are placed along one or several rings. A disadvantage of single-ring arrays is the relatively high sidelobe level. To reduce the sidelobes multi-ring arrays are used. Ring arrays can be constructed in the form of a circle, ellipse, or sphere. The advantage of a ring array is its ability to radiate in any direction. This array is also termed an *annular array*. In practical radars such arrays are seldom used. *AIL*

Ref.: Benenson (1966), p. 238, Mailloux (1994), pp 197–204.

A **Rotman array** is a multibeam array of elements coupled to beam ports through a Rotman lens or a stack of such lenses. The resulting beams are fixed in space and may be coupled to multiple, parallel transmit or receive channels or to a single channel switched among the beam positions. Figures A88 and A89 show one- and two-dimensional Rotman lens-fed arrays. *DKB*

Ref.: Barton (1988), p. 178; Mailloux (1994), pp. 505–511.

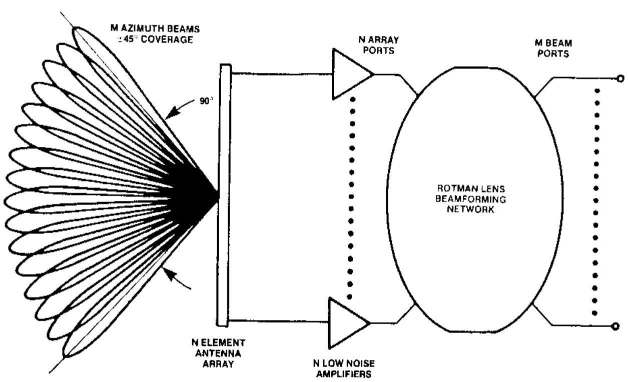


Figure A88 Rotman lens-fed array for one coordinate (from Barton, 1988, Fig. 4.3.7, p. 178).

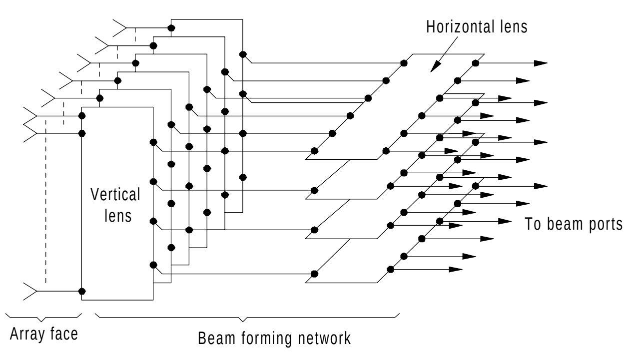


Figure A89 Rotman lens-fed array for two coordinates (from Barton, 1988, p. 179).

array scanning methods (see **SCANNING**).

A **self-phased [-focusing] array** is an antenna-reflector reradiating the energy in the direction of the incident electromagnetic wave. As opposed to the other antenna-reflectors (e.g., the Van-Atta array) where array elements are coupled in pairs, self-phasing arrays use other methods to introduce the desired phase shift. This may involve use of frequency conversion, where the local oscillator (LO) frequency is exactly equal to or twice the frequency of the incidence wave and a difference frequency is used; use of double frequency conversion where the LO frequency is close to the frequency of the incident wave; using two phase shifters, and so forth. Sometimes this array is termed a *self-focusing array. AIL*

Ref.: Steinberg (1976), p. 214; Fradin (1977), p. 343.

A **signal-processing array** is one in which special signal processing of the received signals is used to enhance the quality of extraction of information. Usually the following signal-processing techniques are used: temporal modulation of the antenna parameters (see **ANTENNA, space-time**), logical processing (see **ANTENNA, logical synthesis**), and methods applied for self-phasing antennas. *AIL*

Ref.: Fradin (1977), p. 345.

A **slotted waveguide array** is an array antenna consisting of many slots fed by a common waveguide. Typically, resonant half-wavelength slots are used, cut in either the wide or narrow walls of rectangular waveguide excited by a TE_{10} mode wave (see **WAVEGUIDE**). Longitudinal and transverse slots excite fields with linear polarization. To obtain circular polarization, cruciform slots are used, obtained by collocating the centers of longitudinal and transverse slots. These arrays are widely used in high-directivity on-board antennas. They are sometimes called *slot antennas*. *AIL*

Ref.: Voskresenskiy (1981), p. 107; Johnson (1993), Ch. 9.

space-tapered array (see **unequally spaced array**).

A **spherical array** is one whose radiators are placed on the surface of a sphere with almost constant density. Scanning is done by commutating the feed to the radiators while maintaining a constant pattern shape and gain. The center of the excited (illuminated) region is located in the direction of the main beam. By turning off some radiators and by controlling the shape of the illuminated region, it is possible to obtain patterns having different characteristics. The radiators of a spherical array must have circular or controllable polarization. The number of radiators in spherical arrays commonly ranges from 10^4 to 10^6. These arrays are used in cases where it is necessary to scan a hemispherical volume with a minimum change in the beam shape and gain. *AIL*

Ref.: Voskresenskiy (1981), p.155; Mailloux (1994), pp. 233–234.

A **subarray** of an array is a part of the antenna aperture. The subarrays are combined to form the required overall array pattern. *AIL*

Ref.: Skolnik (1970), pp. 11.20, 11.45.

A **surface array** has its elements or subreflectors positioned on a surface of arbitrary shape. The main types of surface arrays are conformal and quasiconformal arrays. *AIL*

Ref.: Voskresenskiy (1981), pp. 150–157.; Samoilenko (1983), p. 214; Johnson (1993), p. 21.18; Mailloux (1994), pp. 193–235.

A **thinned [sparse] array** is one in which the number of radiating elements is reduced as compared with the number required to fill the aperture completely. The thinning is done so as not to significantly affect the shape of the mainlobe of the array. However, the mean level of the sidelobes increases proportionally to the number of elements omitted. The density of elements decreases toward the edge of the array aperture so as to taper the amplitude distribution. The radiating elements must be positioned to preclude the formation of grating lobes by coherent interaction of signals. The elements of a thinned array can be positioned either randomly or according to some specific law. This type of array is also termed a sparse array. *AIL*

Ref.: Skolnik (1970), pp. 11.24, 11.58; Mailloux (1994), pp. 91–108.

An **unequally spaced array** is one with a unequal distance between elements. This element positioning eliminates periodicity, thanks to which higher order grating lobes are eliminated (or significantly diminished in magnitude). Therefore, it is possible to

(1) Reduce the number of radiators without significant increases in the width of the mainlobe and in side lobe level,

(2) Expand beam-scanning limits and operate in a broader waveband.

(3) Control the level of sidelobe radiation in different sectors.

(4) Simplify the system for aperture excitation.

Shortcomings include the complexity of array synthesis and analysis that requires use of methods of statistical antenna theory. This type of array is also called a space-tapered array. *AIL*

Ref.: Benenson (1966), p. 82.

A **uniformly spaced array** is one all of whose radiating elements are positioned at regular (equal) intervals. When uniform aperture illumination is used with such an array, the pattern has relatively high sidelobe levels. The sidelobes can be reduced by tapering the amplitude distribution, but this results in a decrease of the directivity of the antenna. An alternative approach to reducing sidelobe levels and optimizing the radiation pattern, while feeding each element with equal power, is to use unequally spaced arrays. *AIL*

Ref.:Skolnik (1970), p. 11.15; Fradin, (1977), p. 147.

A **Van Atta array** is an antenna-reflector that allows positioning the principal reflected lobe in the direction of the incoming wave. It is named for its inventor, L. C. Van Atta. The elements of the array are connected to each other with corresponding lines of equal length. Van Atta arrays may be active or passive (Fig. A90). In a passive array (Fig. A90a), the signals received by the elements located to the right of the array center are retransmitted by the elements located at the mirror image in the left half of the array (with equal delays created when the signals pass through sections of equal lengths). Signals received by the elements on the left are retransmitted by elements located at the right. As a result, the elements on the left illuminate with advanced phases, corresponding to the phases of the wave on the right half of the array, while the elements on the right illuminate with delayed phases, corresponding to the phases of the left elements. There occurs a change in the sign of the phases of the radiators. As a result, the principal maximum of the secondary illumination propagates in the direction of the incoming wave (the dashed line in Fig. A90a).

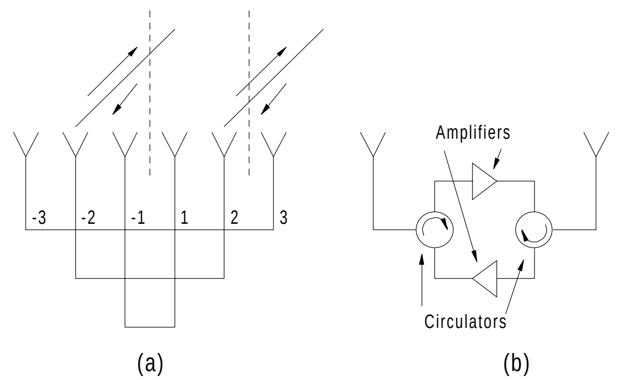


Figure A90 Van Atta arrays: (a) passive array; (b) a pair of radiators of an active array.

In an active Van Atta array, the active elements are connected to pairs of radiators (Fig. A90b). The active array increases significantly the level of the return radiation. The principle of operation of the active array is identical to that of the passive one. Van Atta arrays are applied as reflectors in ECM systems. *AIL*

Ref.: VanBrunt (1978), pp. 144, 371, 627.

A **wideband array** is one operating over a wide frequency band or at several different frequency bands. Design approaches to avoid the degradation in power or accuracy characteristics can be implemented either through specific beam-steering or signal-processing techniques. For scanning over a wide frequency band with the required slope of the phase of the wave front, it is necessary that the excitation of the individual radiators be either advanced or delayed in time. Phased arrays therefore use parallel-feed networks because the excitation of the elements then does not depend on frequency. The frequency band of operation is limited solely by the dependence of the beam characteristics on phase. One can also use phased arrays divided into subarrays. Each subarray uses its own increment of time delay. In this case the wideband-ness is provided by a rearrangement in the frequency-independent elements of the time delay. In the signal-processing approach one can use a set of filters that are matched to the signals at the different angular beam positions. To direct the beam to any particular angle, a command is given to the signal processor which inserts the optimum filter corresponding to the chosen angular beam. *AIL*

Ref.: Skolnik (1970), pp. 11.43–11.50; Bakhrakh (1989), p. 66.

A **Yagi-Uda array** is one formed from a series of dipoles located in parallel in a common plane and forming a "wave channel." One of the dipoles is the actively driven element (Fig. A91) (1) and the rest are passive. One of the passive elements located behind the actively fed antenna plays the role of reflector (2), while the others, placed in front of the actively fed antenna, play the role of directors (3). The reflector length is somewhat greater than $\lambda/2$, while driver length is somewhat less than $\lambda/2$.

This array is a type of end-fire array named after S. Uda and H. Yagi, who were the first to describe it correspondingly in Japanese and in English. Yagi-Uda arrays with a large number of dipoles can be treated as surface-wave antennas. The main advantages of such antennas are design simplicity, high directive gain, and low weight. The antenna's narrow bandwidth is a drawback. They are used in VHF radars and sometimes are called *wave-channel antennas*. *AIL*

Ref.: Fradin (1977), p. 194; Johnson (1993), pp. 3.13–3.15.

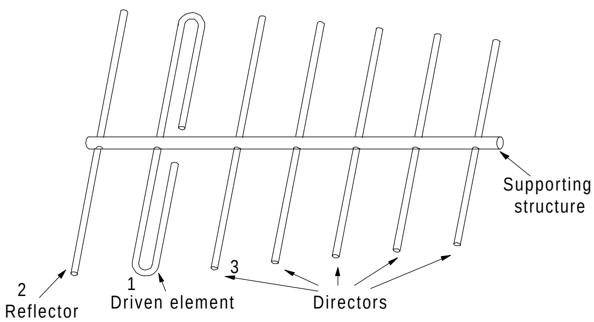


Figure A91 Yagi-Uda array.

ASTRONOMY, radar. Radar astronomy is the branch of astronomy investigating celestial objects with radar methods. The main problem in radar astronomy arises from the fact that tremendous distances are involved, so extreme receiver sensitivity and transmitter power are required for the detection of weak signals. The detectability of radar targets relative to the moon is shown in Table A8. It shown that very sophisticated equipment is required to detect the distant targets. The general block diagram of an astronomical radar is shown in Fig. A92.

The most common type of antenna is the large, steerable parabolic reflector. The Cassegrainian antenna is a good solution for astronomical radar antennas because the feed is closer to the main mirror in double-reflector antennas, so the transmission line losses are less, as lengthy transmission lines can be eliminated and the receiver can be mounted at the feed since it is easily accessible for maintenance. The transmitters typically must be coherent and capable of handling high average powers (the main difference between operation mode of transmitters for radar astronomy relative to conventional radars is that they require higher average power rather than high peak power). The receiver is usually a superheterodyne receiver with parametric amplifiers to reduce self-generated noise and increase sensitivity. The performance of some facilities used in radar astronomy are given in the Table A9. *SAL*

Ref.: Evans (1968); Hovanessian (1984), p. 349-357; Skolnik (1970), pp. 33.1–32.24.

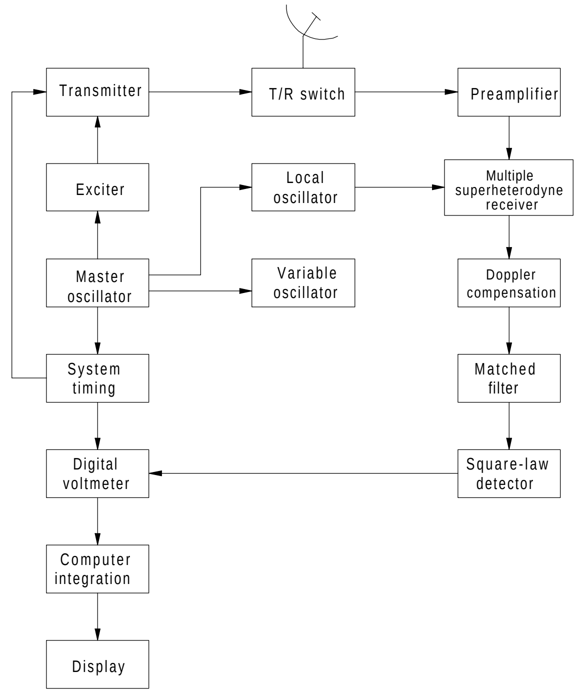


Figure A92 Block diagram of a radar astronomy system (after Hovanessian, 1984, Fig. 13-5, p. 353).

Table A8
Detectability of Radar Targets Relative to the Moon.

Target	σ/R^4 (relative to value for moon)	Level in dB
Large aircraft	1,000	30
Moon	1	0
Sun	1×10^{-5}	–50
Venus	2×10^{-7}	–67
Mars	1.3×10^{-8}	–78.9
Mercury	1.7×10^{-9}	–87.7
Jupiter	3.3×10^{-10}	–94.8
Saturn	1.7×10^{-11}	–107.7
Uranus	1.7×10^{-13}	–127.7
Neptune	2.3×10^{-14}	–136.4

ATMOSPHERE. Earth's atmosphere consists of several concentric shells containing gases, vapors, and other material in suspension and bound to the earth by gravitational force. The composition of the atmosphere by weight is approximately 76.8% nitrogen and 23.2% oxygen. The atmosphere

Table A9
Comparison of Some Radar Astronomy Installations (from Evans and Hagfors, 1968)

Institute	Location	Frequency (MHz)	Antenna diameter (ft)	Gain (dB)	Aperture (m^2)	Average power (kW)	Peak power (kW)	Pulse length (μs)	PRF (Hz)	T_s (K)
Cornell Univ.	Arecibo, PR	430	1,000	57.0	20,000	150	2,500	0.1–10	Variable	400
Crimean Deep Space Tracking Station	Crimea, USSR	700	8×50*	46.8	700	60	60	CW	CW	100
California Inst. of Tech, JPL	Goldstone Lake, CA	2,388	85	54.2	355	100	100	CW	CW	30
Manchester Univ.	Jodrell Bank, U.K.	408	250	47.3	2,300	1.5	1.5	30	1	1,200
Massachusetts Inst. of Tech. Lincoln Lab	Westford, MA (Millstone)	1,295	84	46.5	190	150	150	0.1–4	Variable	80
Massachusetts Inst. of Tech. Lincoln Lab	Massachusetts (Haystack)	8,000	120	66.8	525	100	100	CW	CW	100

*This antenna consists of an array of eight 16-m parabolas fixed to a frame.

extends from the earth's surface to 600 to 1,500 km in space, but more than 75% of the atmosphere lies below about 10.7 km (35 kft) and it is in this region where most of earth's weather effects occur. The atmosphere can affect radar operation in several important ways: (1) by absorbing energy from the radar wave (attenuation), (2) by bending the path of the radar energy (refraction), (3) by contributing interfering signals (clutter) due to the radar energy reflecting off rain (backscatter) and other forms of precipitation, and (4) by adding noise to the radar receiver. The magnitude of these effects is frequency (hence wavelength) dependent, and therefore the atmosphere is a major consideration at the design stage of a radar, in the selection of operational frequency. See **PROPAGATION**; **ATTENUATION**; **atmospheric refraction**; **NOISE, antenna**. *PCH*

Ref.: Van Nostrand (1983); Blake (1982), p. 177

Atmospheric absorption [attenuation] is the loss of radar energy due to absorption in the propagation medium (air, clouds, precipitation, and the ionosphere). See **ATTENUATION; LOSS, atmospheric**.

Atmospheric backscatter from clouds and precipitation in the form of rain, hail, or snow may be considered an unwanted source of interference, or clutter, to radars whose mission is to detect targets other than the weather itself. Rain clutter is especially important in that, on a global basis, it occurs most often, can be extended over large areas, and falls at high rates. The magnitude of the precipitation clutter signal as seen by a radar (i.e., its radar cross section), depends on several factors, including the volume clutter reflectivity (m^2/m^3) of the precipitation (a function of the rain rate); the dimensions of the radar resolution cell ($\Delta R \times \theta_{Az} \times \theta_{El}$); and the number of ambiguities in the radar waveform. Radars typically reject weather clutter through the use of coherent waveforms and processing, which discriminate moving targets from clutter on the basis of a measurement of their doppler frequency. The capability of a radar to reject atmospheric clutter is defined by the radar's clutter attenuation factor, or clutter improvement factor. (See **CLUTTER**). *PCH*

Atmospheric ducting is a mode of anomalous propagation in which specific atmospheric conditions create a confined conduit, or duct, that follows the earth's curvature. At certain radar frequencies, the duct acts like a waveguide, permitting propagation of the radar wave beyond that expected under normal atmospheric conditions and enabling the radar to detect targets beyond the radar horizon. Ducts occur as a consequence of an atmospheric inversion of either temperature, humidity, or both, in which the gradient of the index of refraction, dn/dh, increases at a rapid rate with altitude. The large decrease in the index of refraction with altitude causes the energy trapped within the duct formed at low altitude to propagate along the earth's curvature. Ducts near the ground or sea surface are more common than elevated ducts, supporting only certain modes of propagation and are usually not deep in the vertical dimension. For these reasons and others, extended propagation by atmospheric ducting is more likely to be experienced by horizon-oriented surface radars operating at the higher microwave frequencies (UHF and above). *PCH*

Ref.: IEEE (1993), p. 392; Skolnik (1980), pp. 450–456.

Atmospheric emission. By virtue of its properties as a radiation-absorbing medium, and in accordance with the law of the conservation of energy and Boltzmann's black-body radiation theory, the atmosphere must radiate the same amount of energy as it absorbs for the system to be in the state of thermal equilibrium. If T_a is the ambient temperature of the atmosphere, the atmospheric absorption noise power available in a bandwidth B_n is equal to $kT_aB_n(1 - 1/L)$, where k = Boltzmann's constant, and L is the loss due to attenuation of the energy in passing (one way) through the atmosphere. Because of the term $(1 - 1/L)$, this noise is greatest when the radar antenna is oriented along the horizon (exposure to the maximum amount of atmosphere), and least when the antenna is pointed straight up (minimum amount of atmosphere).

Lightning strokes from storms in the atmosphere produce an additional source of atmospheric RF emission, called *atmospheric noise*. A commonly occurring phenomenon, lightning radiates considerably energy at low frequencies and over great distances. The spectrum of atmospheric noise produced by lightning falls off rapidly with frequency, however, and is of little consequence for frequencies above about 50 MHz (see **ATMOSPHERICS**).

Atmospheric emissions in the form of atmospheric absorption noise (thermal) and atmospheric noise (lightning) are two of the environmental noise sources that affect a radar or other RF receiver. Other sources include solar (or galactic) noise, ground noise, and man-made sources of interference. (See **NOISE**). *PCH*

Ref.: Lawson (1950), pp.103–108; Skolnik (1962), pp. 368–369.

The **exponential reference atmosphere** is described by an exponential approximation for the refractive index n of the troposphere as a function of altitude. For all radar frequencies this can be expressed in terms of a refractivity:

$$N(h) = n(h) - 1 = 313\exp(-0.1439h) = 313\exp(-h/7)$$

where h is the altitude above sea level, in kilometers, and the sea level value $N(0)$ = 313. See **atmospheric refraction; PROPAGATION.** *PCH*

Ref.: Bean, B. R., and Thayer, G. D., "On Models of the Atmospheric Refractive Index," *Proc. IRE* ***47***, no. 5, May 1959, pp. 740–755; Blake (1980), p. 183.

Atmospheric irregularities. Earth's atmosphere affects the transmission and reception of radar (and communications) signals in several ways, most importantly: attenuation of the signal, bending of the radar wave from a straight path (through refraction and diffraction), and corruption of the signal with additive noise. The concept of a standard atmosphere has been developed to describe the principal physical characteristics of the atmosphere (temperature, pressure, humidity, wind speed, etc.) as a function of altitude, for more or less "average" conditions. Using this model, the atmospheric effects on specific radar and communications systems are predictable.

We also know, however, that the troposphere, the weather-producing part of the atmosphere, is in a state of continual change. Significant departures from the conditions defined for the standard atmosphere can be termed *atmospheric irregularities*; one example of which is a temperature inversion, where the earth's surface is cool compared with the air above it. The temperature inversion may create conditions suitable for the formation of a superrefracting duct, whereby the energy from radars radiating into the duct at very shallow angles may be refracted along the earth's curvature to great distances. Weather effects, such as hurricanes, tornados, and other storms that can create very high rates of rainfall, are other examples of atmospheric irregularities that, in this case, cause severe radar attenuation and clutter at microwave frequencies. Strong winds and turbulence in any localized or isolated pattern, such as that produced by the jetstream, can cause anomalies in the index of refraction that fit the category of atmospheric irregularities. (See also **PROPAGATION, ATTENUATION, atmospheric turbulence.**) *PCH*

atmospheric loss (see **ATTENUATION**; **LOSS, atmospheric**).

atmospheric noise (see **NOISE**).

Atmospheric refraction is the term describing change in the direction of travel of radiation passing obliquely from one medium to another. Refraction is "the change in direction of propagation resulting from the spatial variation in refractive index of the medium." In empty space and in uniform propagation media the ray paths are straight lines, while in most media (e.g., the atmosphere) the paths deviate from straight lines due to variation in refractive index. In radar applications, atmospheric refraction occurs in the troposphere and in the ionosphere, chiefly as a result of variation in refractive index with altitude. In the troposphere the variation results from changing density of atmospheric gases and is essentially independent of frequency, while in the ionosphere it is from varying electron density and is strongly frequency-dependent (see **atmospheric refractive index**). In most radar applications it is only the tropospheric effect that need be considered, but when the path extends to altitudes above about 100 km it may be necessary to consider also the ionosphere, especially for radars at UHF and lower frequencies.

In Fig. A93, if i is the angle of incidence (the incoming wave) and r the angle of refraction (the outgoing wave), the refraction is determined from Snell's Law: $\sin i = n \sin r$, in which n is the index of refraction. Physically, n is the ratio of the velocity of the disturbance in the first medium to that in the second.

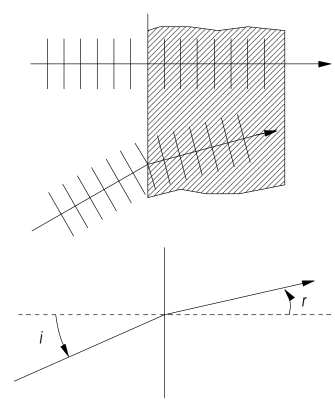


Figure A93 Law of refraction.

Radar waves passing through the earth's atmosphere are bent downward by the changing refractive index of the troposphere and then by the ionosphere. This produces an error in elevation angle measurement, the ray at the antenna having a somewhat larger angle than the direct geometric path to the target (Fig. A94).

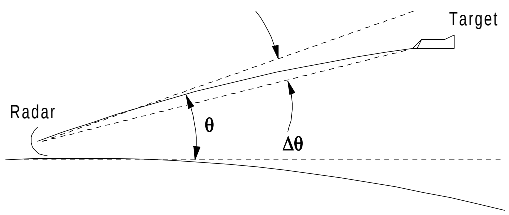


Figure A94 Elevation measurement error due to refraction.

At the same time, an extra time delay is produced, giving a larger range reading than the true range. Random variations in refractive index produce smaller, random errors in measured coordinates. In the troposphere, three effects must be considered:

(1) Regular refraction, resulting in the gradual reduction in the refractive index with altitude, causing elevation and range bias errors.

(2) Tropospheric fluctuations, resulting from random variations in local refractive index and causing slowly varying errors in all measurement coordinates.

(3) Ducting, resulting from steep gradients in refractive index, usually near the surface, creating low-loss propagation paths to low-altitude targets beyond the normal radar horizon and leaving gaps in the coverage for targets just above the duct.

Numerically, atmospheric refraction is calculated using models of the variation of refractive index $n(h)$. The curved ray path length for a specified elevation angle θ_0 to a height h_0 above the surface can be found as

$$R(\theta_0, h_0) = \int_0^{h_0} \frac{n(h)}{\sqrt{1 - \left[\dfrac{n_0 \cos\theta_0}{n(h)(1 + h/r_0)}\right]}} dh$$

where $n_0 = n(0)$ and r_0 is the radial distance of the initial point from the center of the earth.

Depending on the gradient of the refractive index, the following cases can be distinguished:

(1) Normal [regular] refraction.
(2) Superrefraction.
(3) Ducting.
(4) Subrefraction. (See **PROPAGATION**).

In the first three cases the refractive index decreases with height, but in (4), which is rare, it increases (Fig. A95).

The effects of refraction on radar operation are

(1) To change the radar coverage (accounted for through range-height-angle charts for normal condition. (See **CHART**.)

(2) Sometimes to extend coverage beyond the normal horizon (See **DUCTING**; **PROPAGATION, anomalous**).

(3) To introduce errors in angular and range measurements (See **ERROR, propagation**).

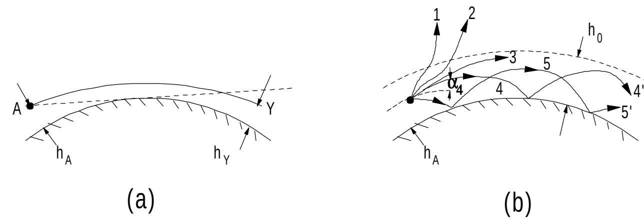


Figure A95 Propagation of radio waves under condition of refraction (a) normal refraction, (b) superrefraction (beams 3, 4, 4', 5, and 5') and subrefraction (beams 1 and 2).

While refraction may be troublesome in radar operations near the horizon, and must be accounted for in accurate tracking applications, it can often be neglected at elevation angles greater than 3 to 5°. *PCH, SAL*

Ref.: Blake (1980), Ch. 5; Skolnik (1980), p. 447; Barton (1988); Van Nostrand (1983).

The **atmospheric refractive index** n is the ratio of the velocity of electromagnetic waves in empty space c to that in a medium:

$$n = c/v$$

In empty space $n = 1$, and in the lower troposphere $n \approx 1.0003$. It varies throughout the atmosphere, the major variation being an exponential change with altitude in the troposphere. In an atmosphere that contains water vapor, the refractive index for radio and microwave frequencies is expressed by

$$(n-1) \times 10^6 = \frac{77.6p}{T} + \frac{3.73 \times 10^5 \cdot e}{T^2}$$

where p is the barometric pressure (mbar), e is the partial pressure of water vapor (mbar), and T is absolute temperature (K). In the ionosphere, n depends on the electron density, N_e and the radar frequency f according to

$$n = \sqrt{1 - \frac{81 N_e \times 10^{-12}}{f_{\mathrm{MHz}}^2}}$$

The refractive index is also called the *index of refraction*.

The refractive index may be modeled, for radar applications as the function of altitude h. Two basic models are used: the exponential model (which is often referred to as exponential reference atmosphere) and the linear model. The exponential model represents the refractive index as:

$$n(h) = 1 + (n_0 - 1) \cdot \exp(-c_e \cdot h)$$

where n_0 is the surface value of refractive index ($h = 0$) and c_e is a constant:

$$c_e = -\frac{1}{n_0 - 1}\left(\frac{dn}{dh}\right)_{h=0}$$

Values of n_0 and c_e vary at different times and places. The linear model assumes a constant negative gradient k of the index of refraction:

$$n(h) = n_0 - k \cdot h$$

that results in the four-thirds approximation widely used in radar range calculations. The ray bending for the exponential model is slightly greater at $h = 0$, but rapidly becomes less than for linear model as altitude increases. The linear model is realistic at altitudes up to about 10,000 ft. (3 km), but predicts excessive ray bending at higher altitudes. The main purpose of atmospheric models is to provide an instrument to evaluate radar performance under the conditions of refraction in atmosphere.

A scaled-up parameter called refractivity is defined as

$$N_s = (n_0 - 1) \cdot 10^6$$

The values N_s = 313 and c_e = 0.1439/km are average for the U.S. at sea level and typically regarded as a standard atmosphere model. *SAL, PCH*

Ref.: Van Vleck, J. H., The Absorption of Microwaves by Oxygen, Physical Review ***71***, April 1947, pp. 413–424; Currie (1987); Blake (1982), p. 182; (1980), p. 178; Skolnik (1980), p. 448.

Atmospheric turbulence. The motion of air in the atmosphere is usually accompanied by turbulence: the presence of eddies of various sizes that migrate to become part of the main airstream or of other eddies or that dissipate shortly after their formation. The result of turbulent flow is the transport of moisture, heat, momentum, particulates, and atmospheric pollutants.

Turbulent flow created by friction between the land mass and the air is primarily responsible for the wind-speed profile between the location near ground level and about 500m altitude, which can be expressed as $V = 10\log h$, where V is the wind velocity in knots, and h is the altitude in meters. This relationship describes a speed profile that increases rapidly with altitude, to about 200m, with a much slower rate of increase above this altitude. Turbulent airflow also occurs around large hills, mountains, canyons, and large man-made structures and is a characteristic of large cyclonic storms and other natural weather phenomena.

Turbulence in the atmosphere can have several consequences for radar:

(1) It can spread the spectrum of rain and chaff clutter, wherein both the mean and standard deviation of velocity are such that the ability of a doppler radar to reject these clutter sources, without also rejecting valid targets, is compromised.

(2) Turbulence may increase the region of heavy precipitation, causing significant radar signal attenuation, particularly at microwave frequencies.

(3) Turbulence can cause irregular perturbations in the index of refraction of the atmosphere along the radar line of sight, potentially introducing angle errors as well as signal depolarization effects.

The consequences of (3) are generally negligible for radars operating below the millimeter-wave frequencies (35 GHz to 140 GHz). (See **ATTENUATION**; **PROPAGATION**; **RADAR, doppler**; **RADAR, millimeter wave**). *PCH*

Ref.: Van Nostrand (1983); Currie (1987), pp. 9, 46.

An **atmospheric window** is a narrow portion of the electromagnetic spectrum for which the propagation losses due to signal attenuation are relatively low compared with that of adjacent regions. For example, at millimeter-wave frequencies, radar operation is usually restricted to "windows" at 35, 95, 140, 220, and 440 GHz. (See also **ATTENUATION**). *PCH*

Ref.: Currie (1987), p. 8; Brussard (1995).

ATMOSPHERICS are "transient bursts of electromagnetic radiation arising from natural electrical disturbances in the lower atmosphere." The most powerful atmospherics arise from thunderstorm electric discharges. When the wavelength is less than 20m, atmospherics practically have no effect upon electronics hardware. *SAL*

Ref.: IEEE (1993), p. 61; Popov (1980), p. 43; Jordan (1985), p. 34.2.

ATTENUATION is the reduction in power resulting from absorption along an atmospheric path or in a circuit. The circuit element intended to reduce power is an **ATTENUATOR**. The following sections discuss unintentional attenuation in paths through the atmosphere and other environments.

Attenuation by clear air, for microwave frequencies, is defined primarily by the troposphere and depends on frequency and the density of atmospheric gases, including water vapor. Figure A96 shows the value of the two-way attenua-

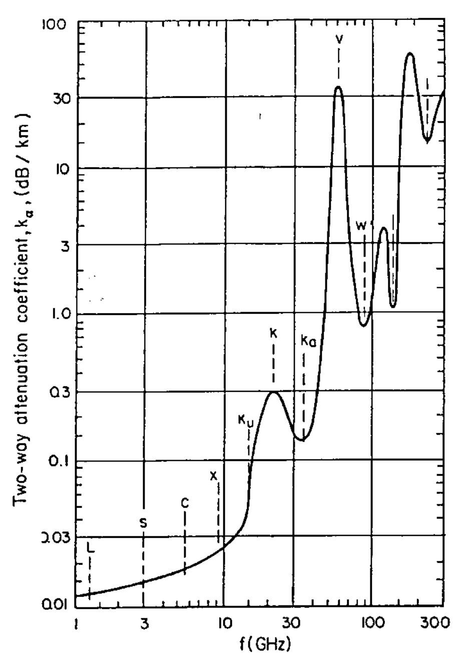


Figure A96 Attenuation coefficient vs. frequency for clear air at sea level (from Barton, 1988, Fig. 6.1.1, p. 279).

tion coefficient k_α for the average sea-level atmosphere as a function of frequency. *DKB*

Ref.: Barton (1988), p. 279.

Attenuation by clouds is a function of frequency and cloud density, as measured by condensed water density in g/m^3, or approximately by the visibility in meters. Figure A97 shows the two-way attenuation coefficient for different cloud or fog conditions. *DKB*

Ref.: Barton (1988), p. 285.

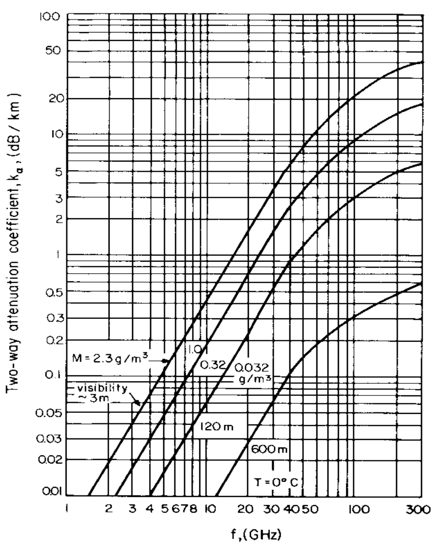


Figure A97 Attenuation coefficient of clouds and fog (from Barton, 1988, Fig. 6.1.6, p. 285).

Attenuation by chaff is generally negligible. A chaff reflectivity η_v implies that this fraction of power entering a 1m cube will be scattered by the chaff, with $1 - \eta_v$ transmitted through the cube. Thus, in chaff with the relatively high density $\eta_v = 10^{-6}$, the fraction scattered in passage through 1 km will be 0.001, leaving 0.999 transmission, giving an effective attenuation coefficient $k_\alpha = 0.004$ dB/km for the transmitted wave. Even if the chaff were to absorb, rather than scatter, the incident wave, the same attenuation coefficient would apply. Only in the immediate vicinity of chaff-dispensing apparatus does the attenuation become significant. *DKB*

Ref.: Barton (1988), p. 285.

clutter attenuation (see **MOVING TARGET INDICATION**).

Clutter attenuation refers to the rejection of clutter in an MTI or doppler processor. The normalized clutter attenuation is defined as the ratio of clutter-to-noise ratio at the processor input to that at the output: CA $\equiv (C/N)_i/(C/N)_o$. (See **MOVING TARGET INDICATION**; **RADAR, doppler**.) *DKB*

Ref.: Barton (1988), p. 244.

The **attenuation coefficient** is the attenuation per unit distance along the path in a given medium, usually expressed in dB/km. It is expressed either as a one-way or two-way value, the latter applicable to the radar case where the wave traverses the path in both directions.

attenuation by fog (see **attenuation by clouds**).

Attenuation by foliage is a severe problem in radar systems that must observe targets through even a thin line of trees. Attenuation coefficients are measured in dB/m, rather than dB/km. Typical values for different frequencies are shown in Table A10 and Fig. A98.

Table A10
Attenuation in Foliage.

Frequency (MHz)	Two-way k_α (dB/m)	Reference
82	0.05	Jakes
210	0.08	Jakes
9,400	2.2	Currie and Brown
35,000	3.5	Currie and Brown
95,000	4.5	Currie and Brown

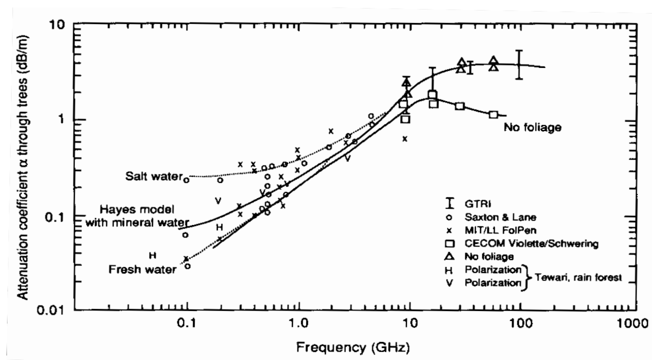


Figure A98 Measured and calculated attenuation coefficient for trees vs. frequency (from Currie, 1992, Fig. 2.19, p. 81).

The calculated values represented by continuous curves in Fig. A98 represent the results when the dielectric constant of the leaves is matched to fresh or salt water, or to an intermediate mineral water model. At frequencies above UHF, the attenuation is such that a typical treeline may be regarded as at impenetrable obstacle, beyond which the field strength may be calculated by assuming knife-edge diffraction. *DKB*

Ref.: Currie (1987), pp. 170–174; Jakes (1974), pp. 107–110; Currie (1992), pp. 77–82

attenuation by gases (see **attenuation by clear air**).

Attenuation in ground penetration refers to the ability of the transmitted signals to penetrate through the surface and into the depths of the ground or other medium, such penetration being quite limited. Penetration can be characterized by the penetration depth for a given attenuation and by the attenuation coefficient, which is the attenuation per unit depth (see Table A11). An estimate of the penetrating capability is needed when the radar targets are located beneath an attenuating medium. *IAM*

Ref.: Mel'nik (1980), p. 71

Table A11

Penetration Capability of Radar in Various Media.

Medium	Wave-length (cm)	Attenuation coefficient, (dB/m)	Penetration depth for 20 dB attenuation, (m)
Snow	10	0.3	67
	100	0.036	520
Frozen soil	300	4.2	4.5
Dry soil	500	0.8	25
Dry sandy soil with 3% moisture	3	300	0.07
	60	3	6.7
Dry clay soil	3	300	0.07
	60	14	1.4

Attenuation by hail is critically dependent on the size distribution and surface condition of the hailstones. Presence of a water film greatly increases the attenuation. Values calculated for typical hail with diameter of 2.9 cm are shown in Table A12. *DKB*

Ref.: Sauvageot (1992), pp. 109–110.

Table A12
Attenuation in Hail

Frequency (GHz)	Two-way k_α (dB/km)	
	Dry	Wet
9.4	3.3	7.6
5.5	0.67	5.2
3.0	0.07	2.4

attenuation by ice (see **attenuation by hail**).

Attenuation by the ionosphere is negligible above the UHF band, and small even at VHF. Figure A99 shows the total one-way attenuation through the ionosphere, as a function of frequency, for paths at 0 and 90° elevation. *DKB*

Ref.: Berkowitz (1965), p. 376.

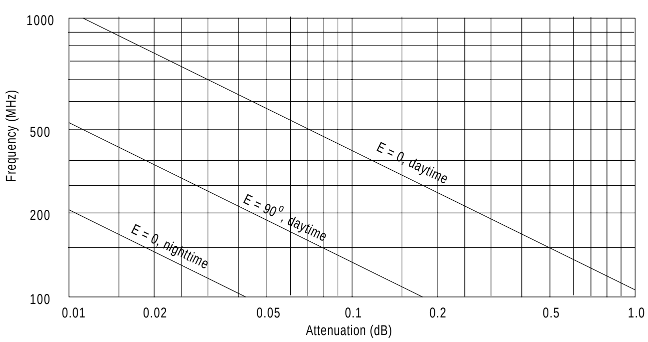


Figure A99 Ionospheric attenuation as a function of frequency, one-way transmission path (after Berkowitz, 1965, Fig. 1.32, p. 376).

Attenuation for laser radar is sensitive to the wavelength and atmospheric conditions. Atmospheric windows exist in the bands 8–14 μm, 3–5 μm, 0.7–2.5 μm, and in the visible spectrum. Within these windows, strong absorption regions, due to water vapor and CO_2, are present near 1.4, 1.9, 2.7, and 4.3 μm. The attenuation coefficients for other atmospheric components are shown as a function of wavelength in Fig. A100.

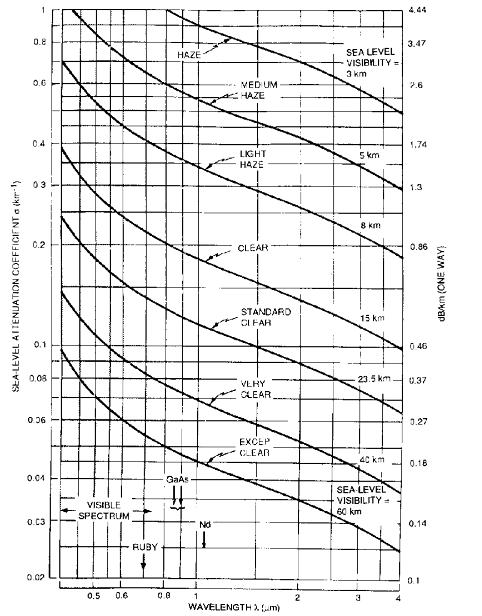


Figure A100 Approximate variation of attenuation coefficients with laser wavelength, at sea level for various atmospheric conditions, excluding H_2O and CO_2 components (from Jelalian, 1992, Fig. 2.3, p. 65).

Attenuation in fog, rain, and snow has little dependence on wavelength but is dependent on water droplet density or precipitation rate, as shown in Fig. A101. *DKB*

Ref.: Jelalian, 1992, pp. 72–74.

Attenuation by particulates, smoke, and aerosol has been calculated under various assumptions, the results generally indicating that there will be little effect on radars even in the millimeter-wave bands. *DKB*

Ref.: Currie (1992), pp. 82–90.

attenuation by precipitation (see **attenuation by hail**, **snow**, or **rain**).

Attenuation by rain is a severe problem for radars in the upper microwave and millimeter-wave frequencies. Figure A102 shows the two-way attenuation coefficient k_α as a function of frequency for different rainfall rates. *DKB*

Ref.: Barton (1988), p. 283.

Attenuation by snow is much less than that for rain with the same water content. Figure A103 shows the one-way attenuation coefficient as a function of precipitation rate (in mm/h of water) for dry snow. *DKB*

Ref.: Sauvageot (1992), p. 110; Blake (1982), p. 214.

attenuation by trees (see **attenuation by foliage**).

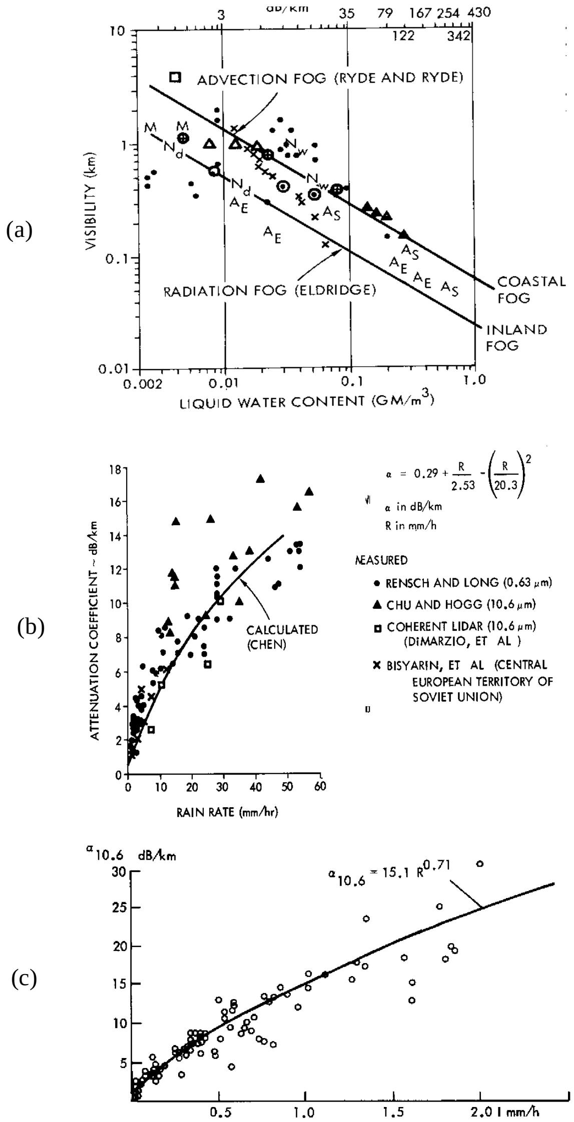


Figure A101 Attenuation coefficients for laser propagation (a) in fog; (b) in rain; (c) in snow (from Jelalian, 1992, pp. 72–74).

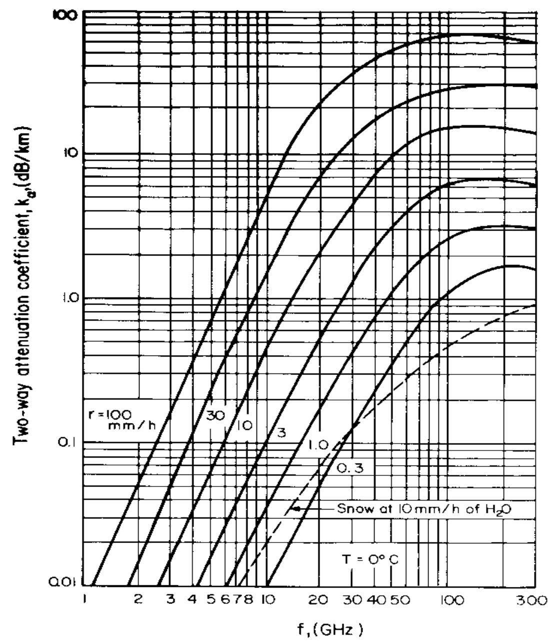


Figure A102 Attenuation coefficient vs. frequency for different rainfall rates (from Barton, 1988, Fig. 6.1.5, p. 283).

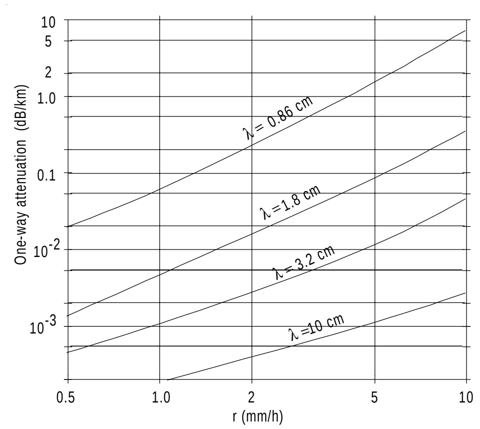


Figure A103 Attenuation coefficient of snow vs. precipitation rate for different frequencies (after Sauvageot, 1992, Fig. 2.17, p. 110).

attenuation by troposphere (see **attenuation by clear air**).

ATTENUATOR, microwave. Microwave attenuators are circuit elements intended to provide attenuation, either in fixed or variable amounts. Several types of microwave attenuators are described below.

An **absorptive attenuator** is based on the absorption of microwave power in the conductance of the controlling element. When a PIN diode is used as the controlling element, the attenuator is a section of transmission line, shunted by several unhoused diodes. Such an attenuator reacts quickly, within microseconds. It may be used as an amplitude modulator and for automatic gain control.

The conductance may also be supplied by a dielectric or ferrite plate covered with a layer of graphite or other absorbing material. *IAM*

Ref.: Gassanov (1988), p. 141; Lavrov (1974), p. 353.

A **bridged attenuator** is a reflection attenuator using a bridge circuit to tap the reflected power from a balanced load. In comparison with simpler reflection attenuators, bridge attenuators provide excellent matching and a large dynamic range for the attenuation. *IAM*

Ref.: Gassanov (1988), p. 140.

A **calibration attenuator** is a device which sequentially connects passive two-ports, each creating a precisely determined attenuation of the output voltage of a power amplifier. This is achieved by using low-resistance noninductive nonwire resistors for the two-ports. *IAM*

Ref.: Popov (1980), p. 169.

A **capacitance [capacitive] attenuator** employs the capacitances at the output of a microwave oscillator circuit. Such an attenuator is used to increase the stability of microwave tubes by reducing the influence of the tube capacitance, which is achieved by using capacitance bridge circuits. *IAM*

Ref.: Popov (1980), p. 124.

A **cutoff attenuator** is based on a transmission line, the cross section of which is less than is critical for the propagating wave. It is usually a portion of cutoff waveguide of a regulated length with a capacitive probe or a coupling loop at the end of the waveguide to receive the attenuated wave. The maximum attenuation in decibels is directly proportional to the length of the waveguide section. To widen the passband in cutoff attenuator circuits, the inputs are matched with absorbing elements. *IAM*

Ref.: Sazonov (1988), p. 153; Lavrov (1974), p. 353.

A **reflection-type attenuator** uses the reflection of a wave from a controlled surface. The simplest reflection attenuator is a p-i-n diode shunting the transmission line, or a diode module placed across the waveguide. A PIN diode attenuator controls power up to hundreds of watts, introducing attenuation due to reflection of up to 20 dB.

Drawbacks to such simple reflection attenuators are the large standing wave ratio and the change in phase of the reflected wave due to the complex conductivity of the diode. *IAM*

Ref.: Gassanov (1988), p. 140.

A **waveguide attenuator** is "a waveguide component that reduces the output power relative to the input, by any means, including absorption and reflection" and is typically used in waveguide transmission lines. The construction of the attenuator depends on the type of waveguide. For metal waveguides with various cross sections, the attenuator is usually a dielectric plate covered with an absorbing material, placed within the waveguide. For dielectric waveguides, the attenuator is often a dielectric controlled plate on its surface, or a portion of stripline waveguide placed along the substrate of a dielectric stripline waveguide. *IAM*

Ref.: IEEE (1993), p. 64; Sazonov (1988), p. 153.

AUTOCORRELATOR. An autocorrelator is a device that produces an output signal proportional to the autocorrelation function of the input signal (see **CORRELATION**). It contains an input filter, a signal delay circuit, a multiplier and an integrator. The functions of an autocorrelator are similar to those of a frequency discriminator for signals with sufficiently narrow bandwidth. Autocorrelators are used in CW radars in automatic frequency-tracking circuits. *IAM*

Ref.: Vinitskiy (1961), p. 281; DiFranco (1968), pp. 214-219.

An **AUTODYNE** is "a system of heterodyne reception through the use of a device that is both oscillator and detector." The main functions of an autodyne are the generation of oscillations, signal reception, and modulation and detection of oscillations resulting in extraction of the useful signal.

Autodyne theory treats the influence of the target as a change in resistance of the antenna-target-antenna system, due to the reflected signal received by the antenna. The main characteristic of an autodyne is its sensitivity S, which is given by the formula

$$S = \frac{U}{\sqrt{P_f/P_t}}\sqrt{\eta_t\eta_r}$$

where U is the amplitude of the reaction voltage in the autodyne; P_t and P_r are the transmitted and received powers; and η_t and η_r are the transmit and receive antenna efficiencies. The radar range equation for a point target with radar cross section σ at wavelength λ takes the form

$$R^2 = \frac{S\lambda DF^2\sqrt{\sigma}}{4\pi\sqrt{\pi}U}$$

where D and F are the directivity and antenna pattern in the direction of the target.

A specific feature of coherent autodynes is their low range capability, due to which they are used primarily in short-range radar systems.

The active elements in an autodyne may be practically any of the known electronic devices: microwave tubes, field-effect and bipolar transistors, and semiconductor diodes (tunnel, IMPATT, and Gunn diodes, etc.). *IAM*

Ref.: IEEE (1993), p. 66; Kogan (1973), p. 70.

A **doppler autodyne** is one that measures the doppler shift of the signal reflected by the target. It is an oscillator, the energy from which is radiated into space and, after reflection by a moving target, returns shifted in frequency by the doppler frequency. After frequency conversion, the low-frequency doppler signal is extracted and processed. An example is an autodyne used to control the motion of an automatic transport based on a reflex klystron, having a horn antenna with 10° beamwidth, 25 mW of power, and an error of 2 to 5% at speeds less than 50 km/h and less than 1% at higher speeds. There are also Gunn diode doppler autodynes that can handle up to 100 mW; other electronic devices are used as well.

Doppler autodynes find use in miniature receive-transmit modules in short-range radar systems used to support the approach or prevent the collision of two objects and to determine their relative distance and speed. *IAM*

Ref.: Khotuntsev (1982), pp. 10, 150.

AUTOMATIC FREQUENCY CONTROL (AFC) (see **FREQUENCY**).

AUTOMATIC GAIN CONTROL (AGC) (see **GAIN**).

AVAILABILITY. Availability is the probability that a radar will be able to perform its designated operational function when required for use. Typically, it is the probability that being in a standby mode the radar will be ready to be placed in operational mode in any arbitrary moment and following that moment will operate without failure during the entire specified period. *SAL*

Ref.: IEEE (1993), p. 76; Aleksandrov (1976), p. 12; Jordan (1985), p. 45.2.

AXIS

The **antenna axis** can be the electrical reference axis of the antenna or the mechanical (boresight) axis. The first is the axis going from the antenna phase center through the peak of the mainlobe, while the latter is the axis of symmetry. When electrical reference axis and boresight axis do not coincide, the angular difference is termed *boresight error* and must be accounted for in the measurement of target directions. *SAL*

Ref.: Skolnik (1990), p. 6.5.

The **null axis** is the term denoting equisignal direction in amplitude-comparison monopulse systems (Fig. A104). *SAL*

Ref.: Leonov (1986), p. 2.

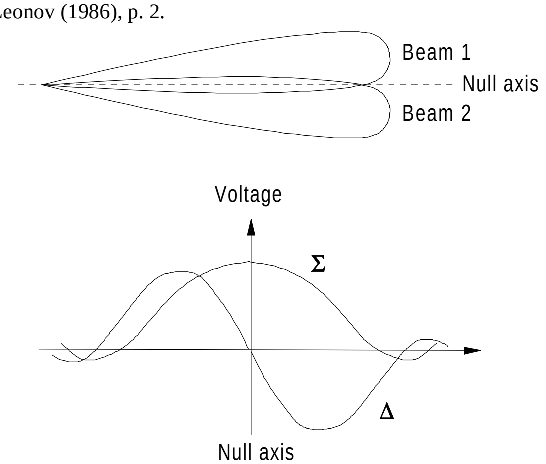


Figure A104 Null axis of amplitude comparison monopulse system.

AZIMUTH (see **ANGLE, azimuth**).

B

BACKSCATTER, BACKSCATTERING. Backscatter is "the energy reflected in a direction opposite to that of incident wave," (i.e., redirected toward the radar). If the signal formed by backscatter is undesired, it is called *clutter*. Backscattering is the process by which backscatter is formed.

Ref.: IEEE (1993), p. 85.

The **backscatter coefficient** is the normalized measure of radar return from a distributed target. The usual notation is σ^0 or η_v. It is defined as $\sigma^0 = \sigma/A_c$ for area targets and as $\eta_v = \sigma/V_c$ for volume targets, where

σ is the average monostatic RCS.

A_c is the surface area.

V_c is the clutter volume.

In radar applications this term typically applies to clutter and to targets of scatterometers and imaging radars. When applied to clutter, σ^0 and η_v are termed *clutter reflectivity.* Another term used interchangeably is *scattering coefficient*. (See **CLUTTER reflectivity.**) *SAL*

Ref.: IEEE (1990), p. 6; (1993), p. 85; Nathanson (1990), pp. 17, 199.

The **backscattering cross section** is "the scattering cross section in the direction towards the source." (see **RADAR CROSS SECTION**).

Ref.: IEEE (1993), p. 85

Bragg backscattering resonance conditions are the conditions defining when the resultant radar backscatter can be interpreted as that obtained from the component of the sea spectrum that is "resonant" with the radar wavelength

$$\lambda_r = 2\lambda_w \cos\psi$$

where λ_r is the radar wavelength, λ_w is the water wavelength, and ψ is grazing angle.

The Bragg scattering resonance condition is termed so because of its similarity to the x-ray scattering in crystals observed by Bragg. In radar applications it is used in the theory of sea clutter. *SAL*

Ref.: Skolnik (1980), p. 480.

BACKWARD-WAVE TUBE. A backward-wave tube is a microwave electronic device using a long interaction between a bunched electron current, moving in longitudinal or perpendicular electrical and magnetic fields, and a backward harmonic wave propagating along a slow-wave circuit. Depending on the directions of the electrical and magnetic waves, a tube may be described as an M-type or O-type backward-wave tube. The physical processes underlying the interaction between the electrons and the wave in a backward-wave tube are the same as those in a travelling-wave tube. The difference is that in a backward-wave tube, the electron current interacts with a backward spatial harmonic. The electron velocity vector v_π and the backward wave group velocity vector v_ϕ are opposed in direction to the group velocity vector v_g for the main wave, for O-type backward-wave tube.

Particular features of the tube depend on whether it is M- or O-type. *IAM*

Ref.: IEEE (1993), p. 86; Dulin (1972), pp. 58, 71; Fink (1982), p. 9.54.

An **M-type backward-wave tube** is one in which the electron current moves in crossed electric and magnetic fields, and is similar to a magnetron amplifier. The slow-wave circuit is an open coil. The electron current is formed with the help of a cathode and auxiliary anode (Fig. B1). In contrast with a

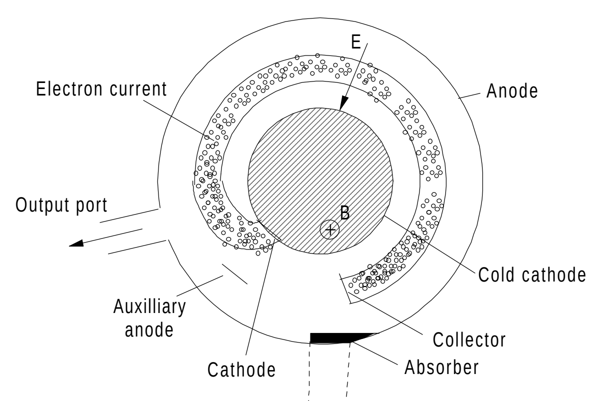


Figure B1 M-type backward-wave tube.

magnetron amplifier, the power is tapped from an output line located near the cathode. In an amplifying backward-wave tube, the input line is located near the collector, without an absorber. The slow-wave circuit consists of built-in pins. The tube is generally cylindrical.

M-type backward-wave tubes may be amplifiers or oscillators, operating at medium to high power levels from 1 to 90 GHz (see **OSCILLATOR, backward-wave tube**; and **AMPLIFIER, backward-wave tube**). *IAM*

Ref.: Andrushko (1981), p. 77; Gilmour (1986), p. 407.

An **O-type backward-wave tube** has a straight-line electron bunch moving in longitudinal electric and magnetic fields. The tube is constructed in the form of a pipe containing an electron gun, slow-wave circuit, and collector (see Fig. B2). In contrast with a traveling-wave tube, the slow-wave structure is usually a two-entry spiral or opposing rods. To focus the electron bundle as it moves along the slow-wave system, the device usually employs a magnetic solenoid, and a set of series-connected coils or fixed magnets.

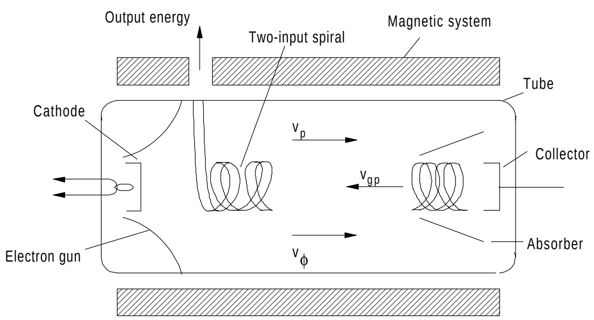


Figure B2 O-type backward-wave tube.

O-type backward-wave tubes are used as amplifiers and oscillators at low and average power from 4 to 470 GHz. (see **AMPLIFIER, backward-wave tube**; and **OSCILLATOR, backward-wave tube**). *IAM*

Ref.: ITT (1975), p. 17.22; Dulin (1972), p. 71.

A **resonator backward-wave tube** is one with increased frequency stability when operating at certain frequencies due to the introduction of strong feedback between the start and end of the slow-wave circuit. The feedback is obtained with the help of a wave reflected from a diaphragm. In the slow-wave circuit, the final absorber is replaced with a short-circuit that fully reflects the incident microwave energy. At the point where the output energy is taken from a resonator backward-wave tube is a diaphragm that partially reflects the microwaves. These devices make it possible to reduce the electron tuning slope while significantly increasing the frequency stability. *IAM*

Ref.: Popov (1980), p. 371.

BAND

radar bands (see **FREQUENCIES, radar band**).

Sidebands are components of the modulated signal spectrum, which are typically located symmetrically about the carrier frequency and separated by the modulation frequency and its harmonics. For example, if the carrier frequency ω_0 is modulated by a sinusoidal waveform with frequency Ω, where $\Omega << \omega_0$, the output spectrum will have sidebands at $\omega_0 \pm \Omega$. *AIL*

Ref.: Popov (1980), p. 54; Schwartz (1959), pp. 93, 121.

BANDWIDTH. The term bandwidth is defined generally to refer to the frequency interval occupied by a signal or passed by a filter or other device. The conventional symbol is B. Several different bandwidths have been defined and used in the literature of radar waveforms and signal processing. *SAL*

The **bandwidth correction factor**, C_B, is the factor accounting for receiver noise bandwidth B_n differing from the optimum. It is applicable primarily to the case of visual detection, and is based on a series of experiments conducted at the MIT Radiation Laboratory, in which observers evaluated the visibility of pulses on cathode-ray-tube displays. The equation for this factor is

$$C_B = \frac{B_n\tau}{4.8}\left(1 + \frac{1.2}{B_n\tau}\right)^2$$

where τ is the pulsewidth and the optimum bandwidth is assumed to be 1.2/t.

The bandwidth correction factor is used as a multiplier to the visibility factor to arrive at the required single-pulse signal-to-noise ratio required for detection. In this it is analogous to the use of the matching factor M as a multiplier to the detectability factor for automatic detection (see **LOSS, filter matching**). *DKB, SAL*

Ref.: Skolnik (1990), pp. 2.14–2.16.

The **effective (noise) bandwidth** is the width of an assumed rectangular bandpass filter having the same transfer ratio at a reference frequency and passing the same power from a white-noise spectrum. For a bandpass filter whose frequency response is $H(f)$ and whose maximum response level is H_0, the effective noise bandwidth is given by

$$B_n = \frac{1}{|H_0|^2} \cdot \int_0^\infty |H(f)|^2 df$$

The **half-power [3-dB] bandwidth** is the width of the signal spectrum or device response at the level 3 dB below the peak. The conventional symbol is B_3. Table B1 compares the half-power and effective noise bandwidths of common types of bandpass filter with center frequency f_0. *DKB*

The **narrow bandwidth assumption** is used in describing a waveform whose spectral density falls to zero at zero frequency, and for which the waveform $s(t)$ can be represented as $s(t) = Re[\psi(t)]$, where $\psi(t)$ is the complex waveform. The thermal noise in a narrowband system can be described in terms of two quadrature components:

$$e_n = e_{ni}\cos\omega t + e_{nq}\sin\omega t$$

where e_n is the noise voltage in the narrowband circuit, e_{ni} and e_{nq} are independent, normally distributed noise components with equal variance, and ω is the angular frequency at the center of the passband. *DKB*

Ref.: Cook (1967), p. 61.

Table B1
Filter Bandwidth Characteristics

Filter type and equation	Half-power bandwidth	Noise bandwidth
Rectangular: $H(f) = H_0$, $\lvert f - f_0 \rvert < B/2$	$B_3 = B$	$B_n = B$
Single-pole: $H^2(f) = H_0^2 f_a^2 / [f_a^2 + (f - f_0)^2]$	$B_3 = 2f_a$	$B_n = \pi f_a = (\pi/2)B_3$
Gaussian: $H(f) = H_0 \exp[-(f - f_0)^2 / 2\sigma_h^2]$	$B_3 = 1.66\sigma_h$	$B_n = 1.77\sigma_h = 1.06B_3$
Cosine: $H(f) = H_0 \cos[\pi(f - f_0)/B]$, $\lvert f - f_0 \rvert < B/2$	$B_3 = B/2$	$B_n = B/2 = B_3$

optimum bandwidth (see **RECEIVER bandwidth**).

quasioptimum bandwidth (see **RECEIVER bandwidth**).

receiver bandwidth (see **RECEIVER bandwidth**).

The **root-mean-square (rms) bandwidth** is the rms deviation of the power spectrum of a signal relative to zero frequency or the spectral center, measured in radians per second. For a narrowband signal centered at carrier frequency f_0, the square of this bandwidth is defined by

$$\beta^2 = \frac{\int_{-\infty}^{\infty} [2\pi(f - f_0)]^2 |S(f)|^2 df}{\int_{-\infty}^{\infty} |S(f)|^2 df}$$

where $S(f)$ is the Fourier transform of the signal waveform, $s(t - \tau_0)$, with true time delay τ_0. The rms bandwidth is proportional to the second derivative of the waveform, and hence β is a measure of the accuracy with which the time delay (or range) of an echo signal can be estimated. *DKB*

Ref.: IEEE (1988), p. 81; Woodward (1993), p. 101.

The **bandwidth-time product** is the product of time duration τ of a pulse and its bandwidth B. For a pulse compression waveform, this product gives the compression ratio of output pulse width (from a matched filter) to transmitted pulse width. *DKB*

Ref.: Nathanson (1969), p. 293.

The **video bandwidth** of a radar system is defined for the one-sided response (from near zero frequency) of stages following the envelope detector. This bandwidth, B_v, is normally set wide enough to pass all significant signal components from a receiver of bandwidth B_n: $B_v >> B_n$. However, when this is not the case, an effective noise bandwidth for the cascaded IF and video filters may be found as

$$B_n' = \left(\frac{1}{B_n^2} + \frac{1}{4B_v^2}\right)^{-\frac{1}{2}}$$

and B_n' may then be used to calculate the system matching loss or collapsing loss to a given signal. *DKB*

Barton (1993), p. 102.

BANG, main. The main bang is “the firing period of the transmitter in a high-power pulse radar.” Typically, this term is applied to any transmitted radar pulse. *SAL*

Ref.: IEEE (1993), p. 767.

BASEBAND is a “band of frequencies occupied by the signal before it modulates the carrier (or subcarrier) frequency to form the transmitted line or radio signal.” *SAL*

Ref.: IEEE (1993), p. 94.

BASELINE is “the line joining two points between which electrical phase or time is compared in determining navigational coordinates.” The concept of baseline is often used in multistatic radar and interferometers. *SAL*

Ref.: IEEE (1993), p. 95.

BAYES CRITERION (see **DETECTION, Bayes criterion**).

BEACON, radar (see **RADAR, secondary**).

BEAM, antenna. The antenna beam is “the main lobe of the (antenna) radiation pattern.”

Ref. IEEE (1993), p. 100.

The **beam broadening effect** is the “spreading of radial velocity components in the doppler spectra of radar echoes due to finite width of the radar beam. Typically used in describing radar returns from the clutter or distributed targets.” The doppler spectrum from precipitation or clouds is broadened by the radial velocity component of a wind blowing across a radar beam of nonzero width. The standard deviation of this spectrum component is

$$\sigma_b = 0.3 v_w \theta_e \sin\beta$$

where θ_e is the one-way, half-power beamwidth in elevation (radians), v_w is the wind velocity at the beam axis, and β is the azimuth of the wind vector relative to the beam axis. In his original analysis, Nathanson used a constant of 0.42 and the two-way, half-power beamwidth. *SAL*

Ref.: Nathanson (1990), pp. 205, 209.

A **cosecant-squared beam** is a fan beam in which the elevation pattern above the main lobe (or below it, for airborne mapping applications) follows the gain relationship

$$G(\theta) = G(\theta_1)\left(\frac{\csc\theta}{\csc\theta_1}\right)^2 , \quad \theta_1 \le \theta \le \theta_2$$

where θ_1 is the elevation (or depression) of the half-power point on the main lobe (see Fig. B3). The objective of this pattern is to maintain a constant signal level on targets having a constant altitude over ranges corresponding to the angular region $\theta_1 \le \theta \le \theta_2$.

The gain G_c of a $\csc^2$ antenna is related to the gain G_m of the fan-beam antenna having the same half-power beamwidths by

$$G_c = \frac{G_m}{2 - \theta_1 \cot\theta_2} \geq \frac{G_m}{2}$$

Thus, the provision of $\csc^2$ coverage introduces a loss not greater than 3 dB in the gain of a fan-beam antenna having given half-power widths. *DKB, SAL*

Ref.: Skolnik (1980), p.259; Barton (1988), p. 26.

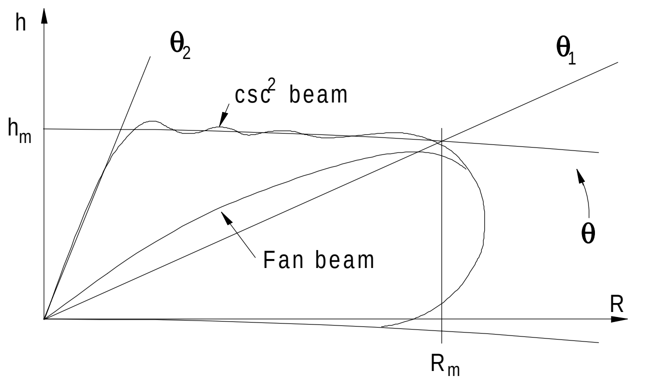


Figure B3 Cosecant-squared beam coverage for air search radar.

difference beam (see **PATTERN, difference**).

beam-forming (see **FEED, antenna**).

A **fan beam** is one that is narrow in one coordinate and wide in the other. As opposed to a pencil beam it provides greater scan coverage for a given scan time, at the expense of reduced gain (see **ANTENNA, fan-beam**). *SAL*

Ref.: Johnson (1993), p. 1.13; Skolnik (1980), p. 55.

interrogation beam (see **RADAR, secondary surveillance**).

A **pencil beam** is a narrow antenna beam, usually symmetrical in azimuth and elevation dimensions. Pencil beams are characteristic of precision-tracking radars and are typically formed by circular reflector antennas. Multifunction phased-array radars form pencil beams for both search and track. *PCH*

Ref.: Johnson (1993), p. 1.13; Skolnik (1980), p. 55.

beam pattern (see **PATTERN**).

beamshape loss (see **LOSS, beamshape**).

doppler beam sharpening (see DOPPLER beam sharpening).

Beam splitting refers to the process of estimating, through interpolation, the exact position of the target within the radar beam. For example, the location of a target in azimuth is nominally defined by the radar antenna's 3-dB beamwidth, but this estimate can be improved by a combination of longer dwell time (providing more samples) and a high signal-to-noise ratio.

The most common beam-splitting techniques are amplitude weighting, sequential lobing, and monopulse (amplitude or phase). Amplitude weighting is the simplest and least accurate technique. Sequential lobing has the disadvantages of introducing target amplitude scintillation as an additional source of tracking error and reducing the subclutter visibility in MTI or pulse-doppler radars. The most frequently used technique is amplitude monopulse, which gives minimum ambiguities and sidelobe levels.

Beam splitting can be accomplished manually, e.g., through visual interpolation of the target displayed on a PPI, or automatically through digital detecting the output of a binary integrator. Beam splitting in this sense then amounts to determining the center of a group of *n* pulses. This process can be applied to the measurement of target range and doppler as well, which for historical reasons, is still referred to as *beam-splitting*. *PCH, SAL*

Ref.: Skolnik (1962), pp.448, 449.

Stacked beams are the simultaneous beams formed at different elevation angles in a 3D surveillance radar. Among other advantages, the technique provides simultaneous lobing for target elevation-angle estimation. The beams are usually contiguous or partly overlapping, and each beam in a stack feeds an independent receiver. The fundamental accuracy performance of a pair of uniformly illuminated stacked beams is presented in Fig. B4, in terms of a normalized sensitivity factor $k = K\lambda/L$ versus normalized sine-space angle-of-arrival $u = L/\lambda \sin\theta$, where

$$K = \frac{|\dot{x}||g_1|}{\sqrt{1+f^2}}$$

and $f = f(\theta) = G_2(\theta)/G_1(\theta)$ is the ratio of the two-way elevation beam power patterns, $\dot{f} = df/d\theta$, g_1 is the two-way normalized voltage pattern in beam position one, L is the aperture dimension, λ is wavelength, and θ is elevation angle. In the figure, Δu is u-space beam separation.

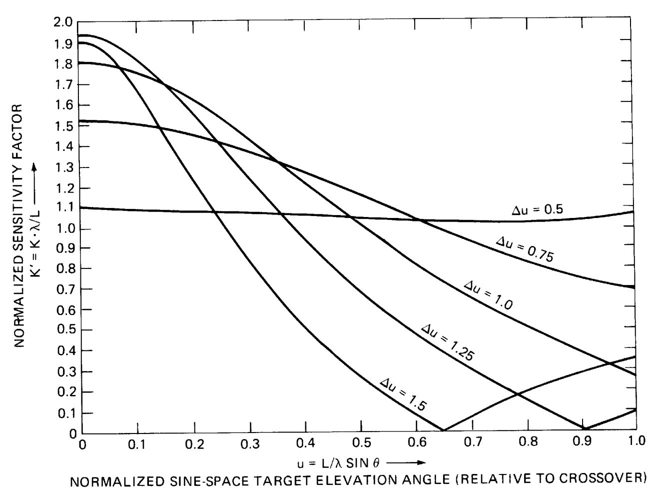


Figure B4 Fundamental accuracy of a stacked-beam elevation estimator (from Murrow, Fig. 20.8, p. 20.32 in Skolnik, 1990, reprinted by permission of McGraw-Hill).

Antenna systems that form stacked beams are termed *stacked-beam antennas* and corresponding radars using this technique are called *stacked-beam radars*. *SAL*

Ref.: IEEE (1993), p. 1272; Skolnik (1990), p. 20.31.

Beam steering is changing the direction of the mainlobe of the antenna pattern, often in accordance with a prescribed scanning pattern. The radar subsystem that controls beam steering is termed a *beam-steering unit*. In mechanically scanned radars the beam steering unit may be a servo system or a constant-speed motor. In electronically scanned radars, the beam steering unit accepts beam position commands from a computer and computes control commands needed by phase shifters in the antenna to direct the beam. *SAL*

Ref.: IEEE (1993), p. 102; Bogush, (1989), p. 46; Mailloux (1994). p. 16.

sum beam (see **PATTERN, sum**).

Beamwidth is the width of antenna mainlobe at some specified level. Typically it is the level at which radiated power density is one-half the maximum value on the beam axis. Beamwidth at this level is termed *half-power* or *–3-dB beamwidth*. *SAL*

Ref.: IEEE (1993) p. 581.

BEL. A bel is "the fundamental division of a logarithmic scale for expressing the ratio of two amounts of power, the number of bels denoting such a ratio being the logarithm to the base 10 of this ratio." If P_1 and P_2 are the two amounts of power, then the number of bels describing the ratio is

$$N = \log_{10}\left(\frac{P_1}{P_2}\right)$$

In radar calculations one-tenth of a bel is widely used (see **DECIBEL**). *SAL*

IEEE (1993), p. 103.

A **BIMATRON** is a nonreentrant forward-wave amplifier related to injected-beam crossed-field amplifiers. It may operate either in a CW or a pulsed mode. The output peak power in the pulsed mode is up to 5 MW, while in the CW mode it is a few hundred watts. The gain is from 20 to 30 dB, and efficiency is from 20 to 35%. *SAL*

Ref.: Fink (1982), p. 9.59.

A **BITERMITRON** is a nonreentrant backward-wave amplifier related to injected-beam crossed-field amplifiers. It is a narrow-bandwidth voltage-tunable device with amplification of 15 to 20 dB and 30 to 35% efficiency. *SAL*

Ref.: Fink (1982), p. 9.59.

BLANKING, BLANKER. "Blanking is the process of making a channel or device noneffective for a desired space of time." In radar, blanking is typically used to turn off the receiver for the interval when the transmitter is turned on or to eliminate undesired interference (like clutter or jamming) received usually through the sidelobes. A blanker is the circuit that performs blanking. *SAL*

Ref.: IEEE (1993), p. 115; Johnston (1979), p. 56.

The **blanking period** is the space of time when the receiver is switched off to ensure transmitter-receiver decoupling or rejection of pulsed interference in pulsed radars. Also used in interrupted continuous-wave radars to eliminate transients, spillover, and close-in clutter. *SAL*

Ref.: Barton (1988), p. 270; Nathanson (1990), p. 368.

Sidelobe blanking is the technique employed to remove the signals received through the sidelobes of the receiving antenna by means of sidelobe blankers. It is one of several techniques used to reduce sensitivity to ECM or any interference. This technique is effective only for low-duty-factor interference (where the duty cycle of removed interference signals is less than about 50%). The system performing sidelobe blanking is a sidelobe blanker. Typically, this is a subsystem that employs an auxiliary wide-angle antenna and receiver to sense whether a received pulse originates in the sidelobe region of the main antenna and to gate it from the output signal if it does. The block diagram of a sidelobe blanker is given in Fig. B5. The output A from the main radar antenna is inhibited when signal B in the guard channel exceeds A. The guard receiver gain is adjusted so that the channel gain for signal B is slightly more than the channel gain for signal A when derived from the peak of the largest sidelobe of the main antenna, and so any signal received through a sidelobe is rejected. *SAL*

Ref.: IEEE (1993) p. 1217; Barton (1991), p. 12.12; Long (1992) p. 248; Johnston (1979), p. 66.

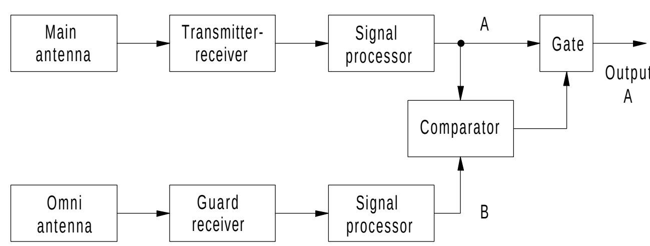


Figure B5 Sidelobe blanker in which the output A is inhibited if signal B exceeds A (after Long, 1992, Fig. 6.13, p. 248).

Wide-pulse blanking is the process that eliminates clutter and noise pulses from the video display in much the same manner as the clutter eliminator circuit. *SAL*

Ref.: Johnston (1979), p. 68.

BLIND

blind phase (see **PHASE, blind**).

blind range (see **RANGE, blind**).

blind speed (see **SPEED, blind**).

BLINKING is an ECM technique employed by two or more movable targets (i.e., aircraft) when they alternately jam the radar. As the result radar antenna beam oscillates from one target to the other, making an accurate solution of fire control problem impossible. Sometimes the term blinking is also related to a method of providing information of the radar display by modifying the signal at its source so it alternately appears and disappears, indicating, for example, that some subsystem or the station is malfunctioning. *SAL*

Ref.: Johnston (1979), p. 56.

A **BLIP** is "a deflection or a spot of contrasting luminescence on a radar display caused by the presence of a target."

Ref.: IEEE (1993), p. 116.

The **blip-scan ratio** is the "fraction of scans for which a blip is observed at a given range." (See also **DETECTION probability**).

Ref.: IEEE (1993), p. 116.

BLOCKING is "the process of obscuring guidance signals by active jamming."

Ref.: Johnston (1979), p. 56.

BLOOMING is "an increase in the blip size on the display as a result of an increase in signal intensity or duration."

Ref.: IEEE (1993), p. 119.

BOLTZMANN'S CONSTANT is a universal physical constant of proportionality that defines the Kelvin temperature scale and relates the energy of thermally generated noise to absolute temperature. Boltzmann's constant *k*, named for its discoverer, Ludwig Boltzmann (1844–1906), is equal to 1.38×10^{-23} J/K, or W/Hz/K. *PCH*

Ref.: Van Nostrand (1983), pp. 401, 402; Jordan (1985), p. 3.16.

BORESIGHTING is "the process of aligning the electrical and mechanical axes of a directional antenna system."

Ref.: IEEE (1993), p. 123.

BOXCAR (see CIRCUIT, sample and hold).

BRAGG

A **Bragg cell** is the basic element of the acoustic-optical modulator. It is an optically transparent crystal of lithium niobate or tellurium dioxide, on one side of which a piezoelectric converter is mounted. When the radio signals being analyzed arrive at the converter, traveling acoustic waves are excited. These waves modulate the refraction coefficient of the crystal and for a traveling diffraction grating. The Bragg cell is illuminated by a parallel coherent laser beam that is incidental to the Bragg angle in relation to the direction of the propagation of the acoustic wave. As a result of the acoustic-optical interaction in the cell the laser beam is deflected, whereby the magnitude of the angle of deflection is directly proportional to the frequency of the radio signal being analyzed. (See **Bragg effect.**)

Thanks to the broad momentary frequency band (all the way up to an octave in the 50-MHz to 2-GHz band) and the sufficiently high-resolution capability in the acoustic-optical receiver, several simultaneously existing analog and sampled signals of various types can be processed. Therefore, the Bragg cell is widely used in the construction of the intercept receivers used in electronic warfare means. *IAM*

Ref.: Zmuda (1994), Ch. 7.

The **Bragg effect** is the phenomenon of the interference amplification of the x-rays reflected by the various atomic planes of a crystal. It is observed when fulfilling the Bragg-Wolf, condition which links the distance between the planes *d*, the wave length λ, and the grazing angle, θ:

$$2d\sin\theta = m\lambda,\ m = 0, \pm 1, \pm 2...$$

The Bragg effect explains the diffraction of x-rays on crystals as well as the diffusion of light on inhomogeneities of the dielectric constant. In the latter case, the Bragg effect is used when describing the modulation of light in acousto-optical modulators (see **Bragg cell**).

By analogy to the x-ray phenomenon, the Bragg effect is used to describe the scattering of radar waves from the pattern of regularly spaced waves on water (see **CLUTTER, sea**). *IAM*

Ref.: Sivukhin (1985), p. 390, 610; Skolnik (1980), p. 480.

Bragg scattering in its original sense describes the propagation of an optical beam when the incident beam is inclined to the normal at the Bragg angle. The Bragg scattering effect is used in the Bragg-cell receiver. Bragg scattering also provides the mechanism by which sea clutter is returned to over-the-horizon radars. (See also **Bragg effect.**) *SAL*

Ref.: Long (1975), p. 84; Neri (1991), p. 297.

BRIDGE, microwave. A microwave bridge is a directional coupler in which the output power is equal in the two output branches, and there is a constant phase difference between the voltages in them over the operating band.

A microwave bridge is characterized by its standing-wave ratio, the coefficient of decoupling between the branches, and the attenuation (see **COUPLER, directional**). A microwave bridge may be constructed from different types of transmission line (e.g., waveguide, coaxial, or microstrip). Most widely used in radar are waveguide bridges of the slotted, circular, or T-types. These are used in microwave amplifiers and modulators, balanced mixers, monopulse radar sum-and-difference networks, and antenna commutators. *IAM*

Ref.: Sazonov (1988), P. 96; Veselov (1988), p. 57.

A **balanced bridge** is "a network with a minimum of two ports or terminal pairs capable of being operated in such a manner that when power is fed into one port, by suitable adjustment of elements connected to one or more other ports, zero output can be obtained at another port." *SAL*

IEEE (1993), p. 129.

A **circular bridge** consists of a ring that is 1.5 wavelengths in circumference, and four branches located a quarter-wavelength from one another (Fig. B6). A circular bridge may be constructed from any type of transmission line.

A circular bridge is characterized by the decoupling of the inputs 1 and 2. When a signal is presented at branches 1 and 2, the sum of these signals is obtained at branch 4, and the sum of the first signal and the second signal, shifted by 180° relative to the first, results at branch 3. This property is used to create balanced mixers, microwave amplifiers and modulators, and multichannel phased-array radar receivers. The band of operating frequencies in a circular bridge usually does not exceed 20%. *IAM*

Ref.: Sokolov (1984), p. 136; Gardiol (1984), p. 284.

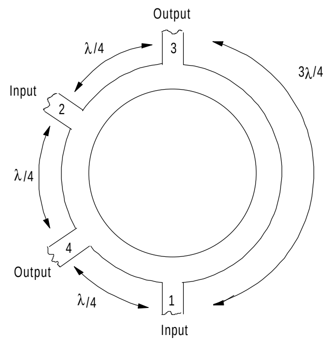


Figure B6 Circular bridge.

A **Marsh and Wiltshire bridge** is a microwave bridge using a balancing element to match the reflections from a resonant cavity in such a way as to cancel the applied carrier, thereby avoiding saturation of a mixer (Fig. B7). It is used in continuous-wave radar in a noise cancellation circuit to recover the noise sidebands of the input carrier so that they may be removed from the transmitted signal. *SAL*

Ref.: Skolnik (1990), p. 14.10.

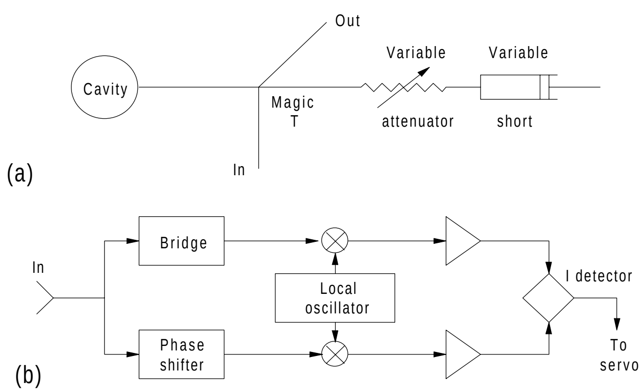


Figure B7 The Marsh and Wiltshire bridge: (a) bridge circuit, (b) noise cancellation circuit using the bridge (after Skolnik, 1990, Fig. 14.4, p. 14.10)

A **slotted-waveguide bridge** consists of two rectangular waveguides sharing a narrow wall, a portion of which has been removed, forming a slot (Fig. B8). In the center of the slot is a capacitive pin, which compensates for wave reflections. The length of the slot is given by the formula

$$l = \frac{\lambda}{4}\left[\sqrt{1-\left(\frac{\lambda}{4a}\right)^2} - \sqrt{1-\left(\frac{\lambda}{2a}\right)^2}\right]^{-1}$$

where λ is the wavelength. The isolation between branches 1 and 2 reaches 30 to 35 dB. The operating bandwidth is 10 to 15%. One advantage over the circular bridge is the simplicity of its construction. *IAM*

Ref.: Sazonov (1988), p. 104; Druzhinin (1967), p. 142.

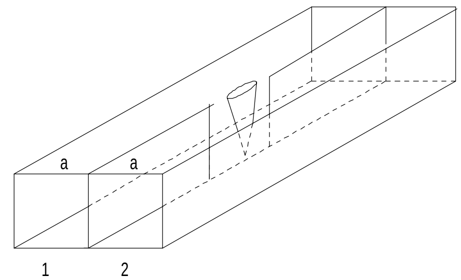


Figure B8 Slotted-waveguide bridge.

The **transmission-cavity bridge** is the simplest microwave bridge used for noise cancellation in continuous-wave radar. It employs the properties of transmission and reflection cavities (Fig. B9). The disadvantage of this configuration compared with the Marsh and Wiltshire bridge is that the carrier frequency is not suppressed before reaching the mixer. *SAL*

Ref.: Skolnik (1990), p. 14.9.

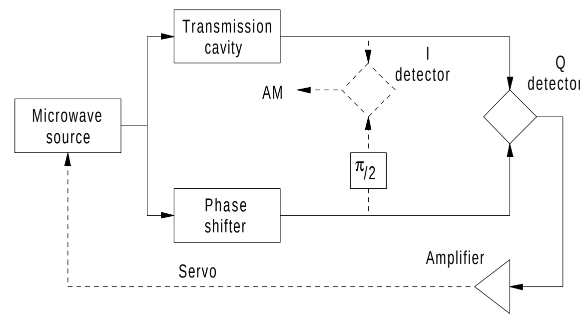


Figure B9 Transmission bridge, video version (after Skolnik, 1990, Fig. 14.3, p. 14.9).

A **waveguide bridge** is a device for splitting electromagnetic energy in a waveguide. The simplest type of waveguide bridge is the double waveguide tee (magic T), which comprises two waveguide tees (Fig. B10). This design features

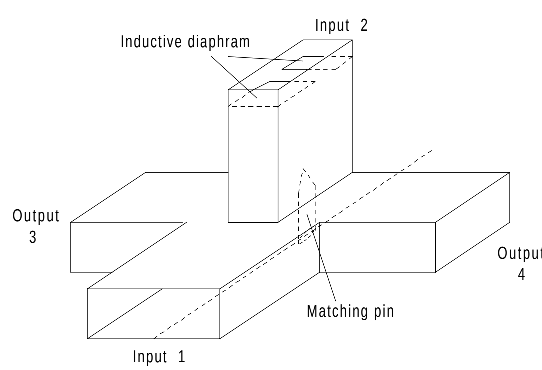


Figure B10 T-type waveguide bridge.

excellent isolation between inputs 1 and 2, the matching of which is achieved with the tuning pin and inductive diaphragm, respectively. The device may be used as a mismatch indicator by exploiting the fact that when the bridge is fed at input 2, there will be energy at branch 1 only if one of the loads connected to branches 3 and 4 are mismatched. The maximum bandwidth is 10 to 15% for a standing-wave ratio of 1.2.

Other configurations include a waveguide formed into a circle (circular bridge) and the slotted-waveguide bridge.

Waveguide bridges are used to decouple generators that operate at different frequencies through a common load, as sum-and-difference devices in monopulse radars, to measure mismatch, and as elements of antenna commutators and balanced mixers. *IAM*

Ref.: Popov (1980), p. 67; Druzhinin (1967), p. 141; Gardiol (1984), p. 282; Sazonov (1988), p. 95.

BURNTHROUGH is the ECCM tactic of increasing the radar energy in the direction of the jammer in hopes of increasing the radar power echo above the jamming noise. Typically it may be accomplished with adding reserve transmitter power or by dwelling longer in the direction of the jammer. The longer dwell reduces the date rate, so it has to be taken into account when such a tactic is supposed to be used. *SAL*

Ref.: Skolnik (1980), p. 549.

burnthrough range (see **JAMMING, barrage**)

BUOY, radar. A radar buoy is a type of anchored marine buoy on which are mounted special radar reflectors. Because of large reflectivity, it can be observed at long distances by shipborne radars. The main application is for ship navigation, marking the places dangerous for ship sailing and indication of ship paths in seas, rivers, lakes, and channels. *SAL*

Ref.: Popov (1980), p. 345.

BURST (see **PULSE**).

C

CALIBRATION is "the process of adjustment to match the designed operating characteristics with the subsequent marking of the positions of the adjusting means." In radar applications, calibrations of different devices in radar subsystems are made at the stages of hardware tuning, testing and service. The calibration of radar reflectivity measurement is very important phase to ensure accurate acquisition of noncoordinate data in the process of radar operation.

Calibration of radar reflectivity measurement is the process of relating signals that can be extracted from radar measurement to the physical characteristics of a target such as its radar cross section or its polarization properties. Accurate calibration is one of the most important phases of radar reflectivity measurement and it employs several techniques. Generally they are amplitude, phase and polarization calibrations. Also absolute, direct, indirect, mixed and relative calibrations are distinguished. *SAL*

Ref.: IEEE (1993), p. 149; Currie (1987), pp. 760-778; Currie (1990), pp. 64-75.

Absolute calibration is the technique in calibration of radar reflectivity measurements in which the target under test is replaced with a known calibration target at the same location, and direct comparison of the target returns is performed. Typically a sphere is used as the reference target. *SAL*

Ref.: Currie (1987), p. 74.

Amplitude calibration is the technique in calibration of radar reflectivity measurements of comparing target RCS with a calibration standard (direct calibration) or using the radar equation with careful measurements of radar and propagation path parameters to predict the RCS of a target from its received power (indirect calibration). Since the errors associated with the two methods of amplitude calibration are for the most part independent, the calibration errors can be estimated by performing both techniques and comparing the results (this is called *closure*). *SAL*

Ref.: Currie (1989), p. 65.

Calibration closure is the criterion of correspondence of reflectivity measurement results obtained using different calibration techniques. Typically 3 dB difference is considered to be acceptable and 1 dB is excellent. *SAL*

Ref.: Currie and Brown (1987), p. 763.

Direct calibration is the technique in radar reflectivity measurement when a calibration standard is used for comparing its return with the signal from an unknown target. Typically a standard is used whose RCS can be accurately inferred from its physical dimensions (i.e., a sphere or a corner reflector). The RCS of the unknown target σ_u can be determined from

$$\sigma_u = P_r\left(\frac{R_u}{R_r}\right)^4\left(\frac{F_r}{F_u}\right)^4\sigma_u$$

where P_r = ratio of power received from the unknown target to that from the reference target, R_u and R_r are ranges to unknown and reference targets, respectively, F_u and F_r are propagation factors for unknown and reference targets, respectively, and σ_r is the RCS of the reference target. The accuracy of this calibration method depends of the precision of σ_r and on the propagation factor being well known. *SAL*

Ref.: Currie (1989), p. 65.

Indirect calibration is the technique in radar reflectivity measurement involving measurement of radar characteristics (such as transmitted power, antenna gain, etc.), measurement of received power from an unknown target, and calculation of the unknown RCS by using the radar range equation. *SAL*

Ref.: Currie (1989), p. 66.

Mixed calibration is a combination of the absolute and relative calibration techniques, which uses the absolute method to establish the value of one point of the relative receiver transfer curve. The RCS of an unknown target at any range R_{un} can be determined from the equation

$$\sigma_{\text{un}} = 20\log(v_{\text{un}}) - 20\log(v_{\text{cal}}) + \sigma_{\text{cal}} - 40\log\left(\frac{R_{\text{cal}}}{R_{\text{un}}}\right)$$

where v_{un} and v_{cal} are the voltages out of the receiver due to the returns from the unknown and the calibration targets correspondingly, R_{un} and R_{cal} are ranges to the unknown target and the calibration reflector correspondingly, and σ_{cal} is the RCS of the calibration reflector. *SAL*

Ref.: Currie and Brown (1987), p. 701.

Phase calibration is the technique in radar reflectivity measurement that is required for coherent systems to ensure phase linearity and stability. *SAL*

Ref.: Currie and Brown (1987), p. 763.

Polarization calibration is the technique in radar reflectivity measurement involving the calibration of the polarization matrix elements. Typically, the process involves the measurement of polarization isolation of the system and the use of several types of calibration targets to calibrate the various components of the polarization matrix. One of the best techniques for fully polarimetric system is the polarization distortion method. *SAL*

Ref.: Currie (1987), pp. 70–72; Currie (1990), pp. 765–768.

Relative calibration is the technique in radar reflectivity measurement that defines the relationship between receiver input power and output voltage (analog or converted to digital form). This relationship is known as the *receiver transfer function*. The process of relative calibration typically is performed by injecting a signal at RF or IF and varying the power in known ratios (often in 5-dB steps), or by illuminating a calibration target at a fixed range and varying an RF or IF attenuator in known ratios. *SAL*

Ref.: Currie (1990), p. 66.

calibration target (see **TARGET, calibration**).

CAMOUFLAGE, radar. Radar camouflage is the concealment of various ground installations, aerospace weapons and combat equipment from radar observation. Radar camouflage is achieved through narrowband interference radar-absorbing coatings, wideband radar-absorbing coatings, low-reflection shapes of an object in (low-reflective in the direction of the radar), as well as through the use of various distortion means (attached distortion means or special distorting elements of the design) and decoy targets for creating false blips on the radar. (see **ABSORBER**; **STEALTH**). *IAM*

Ref.: Druzhinin (1967), p. 507.

CANCELLATION, CANCELER. cancellation is the process of suppression of unwanted signals (clutter, fixed targets and other interference). A canceler is "that portion of the system in which unwanted signals . . . are suppressed."

In radar applications typically clutter and jamming cancellation are distinguished. Clutter cancellation usually implies the suppression of various stationary echoes using doppler processing (see **MOVING TARGET DETECTOR**), or MTI (delay-line cancelers or their modifications, such as double-delay cancelers, double-delay feedback cancelers, four-pulse cancelers, multiple-delay cancelers, shaped delay cancelers, three-pulse cancelers, triple cancelers). In Russian literature, delay cancelers are usually termed *interperiod cancelers*.

Jamming cancellation is usually aimed at suppression of interfering signals received through the antenna sidelobes and, thus, is the sidelobe cancellation (SLC) procedure utilizing sidelobe cancellation algorithms. The methods of such cancellation can be either incoherent (amplitude) or coherent. Of the many sidelobe cancellation methods, Gram-Schmidt and Howells-Applebaum cancelers are the classic ones. *SAL*

Ref.: IEEE (1993), p. 153; Skolnik (1962), p. 119; Maksimov (1979), pp. 172–220.

A **band-partitioned canceler** is a technique for compensation of channel frequency response between the main and auxiliary channels used in sidelobe cancellation. *SAL*

Ref.: Cantafio (1989), p. 469.

cascaded canceler (see **multiple-delay canceler**).

A **delay-line canceler** is the device used in MTI and pulse doppler radars to reject clutter echoes. Functionally, a canceler delays a copy of the input pulse train by one interpulse period and then subtracts the copy from the original. Clutter (stationary) echoes are canceled because they do not vary from pulse to pulse, and moving targets are not canceled. Thus, the delay-line canceler performs as a filter. Its frequency response is shown in Fig. C1. Delay-line cancelers

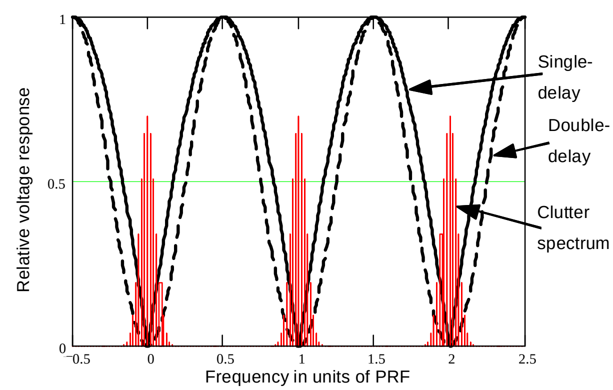


Figure C1 Relative frequency response of the single-delay canceler (solid curve) and the double-delay canceler (dashed curve). (after Skolnik, 1980).

can be typically categorized as single-delay cancelers, double-delay cancelers and multiple-delay cancelers employing or not employing feedback and feed-forward paths. *SAL*

Ref.: Barton (1991), p. 7.33.

A **double-delay canceler** is an MTI delay-line canceler employing the two-delay-line configuration to improve the performance by widening the clutter-rejection notches, as compared with single-delay cancelers. This canceler is also called a double canceler, dual-delay canceler, or three-pulse canceler. The weights for the three pulses are 1, –2, 1. Its relative response, compared with that of a single-delay canceler, is shown in Fig. C1. The improvement factor for a double-delay canceler, when clutter is centered in the notch, is

$$I_2 = 2\left(\frac{f_r}{2\pi\sigma_c}\right)^4 = 2\left(\frac{v_b}{2\pi\sigma_v}\right)^4$$

where f_r is pulse repetition frequency, σ_c is rms clutter spread in frequency, v_b is the blind speed, and σ_v is the rms clutter spread in velocity. *SAL*

Ref.: Barton (1964), p. 219; Skolnik (1980), p. 109.

A **double-delay feedback canceler** is a double-delay canceler in which the feedback is added to adjust the shape of the filter passband, as shown in Fig. C2. The addition of feedback around the canceler makes it possible to reduce the width of the notch, and, to a large extent, to preserve the cancellation characteristics of the double-delay system in the immediate vicinity of the blind speed. *SAL*

Ref.: Barton (1964), p. 220.

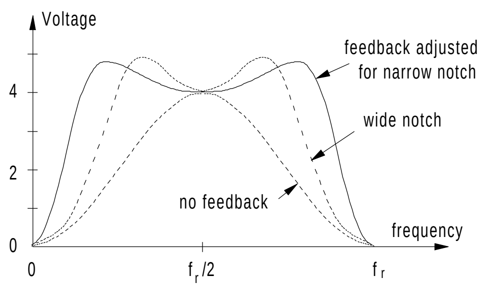


Figure C2 Frequency response of a double-delay canceler with and without feedback (after Barton, 1964, p. 220).

A **four-pulse canceler** is the configuration of a delay-line canceler employing three delay lines connected as a transversal filter (see Fig. C3). The weights are 1, –3, 3, –1. *SAL*

Ref.: Skolnik (1980), p. 110.

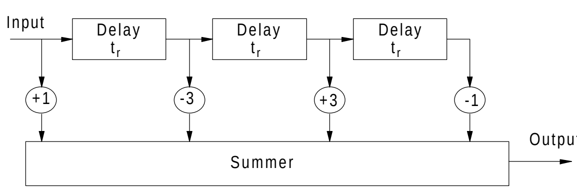


Figure C3 Block diagram of four-pulse canceler.

A **Gram-Schmidt canceler** is a sidelobe canceler implemented with an open-loop technique. Typically, it uses series-connected adaptive loops employing decorrelated auxiliary signals. One of the configurations is shown in Fig. C4, where an adaptive loop decorrelates the signal picked up on auxiliary antenna 1, another loop decorrelates the signal picked up on auxiliary antenna 2, and the result is used by a second series loop in the radar channel, and so forth. Since the auxiliary signals have been decorrelated prior to use, none of the series loops in the radar channels can reinsert interference that has been removed by the previous loop. Such a configuration has better speed of convergence and stability than closed-loop configurations, and the fact that the loops do not interact simplifies the analysis. *SAL*

Ref.: Lewis (1986), p. 122; Farina (1992), p. 122.

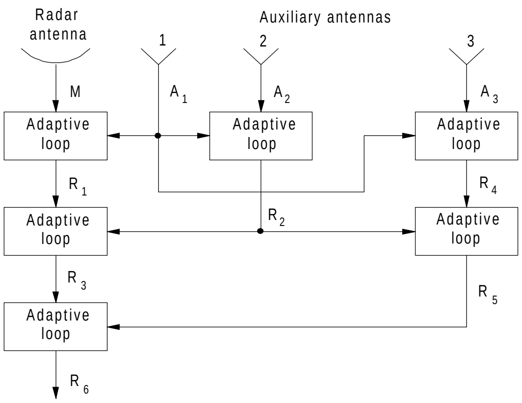


Figure C4 Gram-Schmidt canceler (after Lewis, 1986, Fig. 3.7, p. 122).

The **Howells-Applebaum canceler** is a coherent sidelobe canceler implemented with a closed-loop technique. Originally this canceler was invented by P. Howells, and S. Applebaum was the first to analyze such systems mathematically. The conventional analog Howells-Applebaum canceler for a single auxiliary antenna case is shown in Fig. C5. The cancellation ratio of Howells-Applebaum loop is give by formula

$$\mathrm{CR} = (1 - 2|\rho|^2 + |\rho|^2 k^2)^{-1}$$

where ρ is the normalized correlation coefficient between the main and auxiliary signals and k is the closed loop gain.

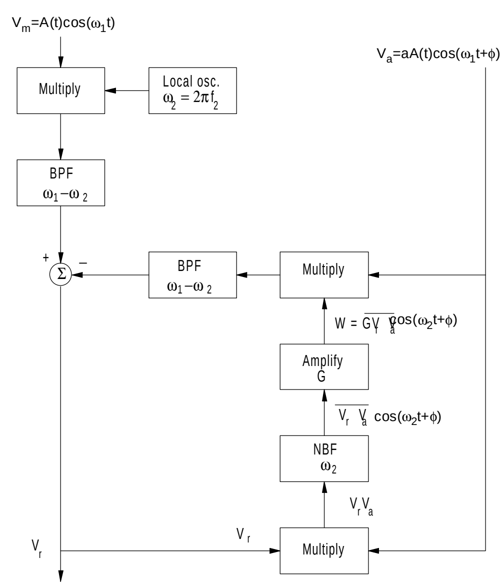


Figure C5 Howells-Applebaum canceler (after Lewis, 1986, Fig. 3-2, p. 118.

The original Howells-Applebaum canceler was implemented at IF by using analog devices. More recently, the SLC has been implemented with digital techniques. *SAL*

Ref.: Farina (1992), pp. 104–111; Lewis (1986), pp. 117–122.

A **multiple-delay(-line) canceler** is a delay-line canceler composed of sections of single-canceler circuits. One form of multiple delay canceler is the *cascaded cancele*r (Fig. C6).

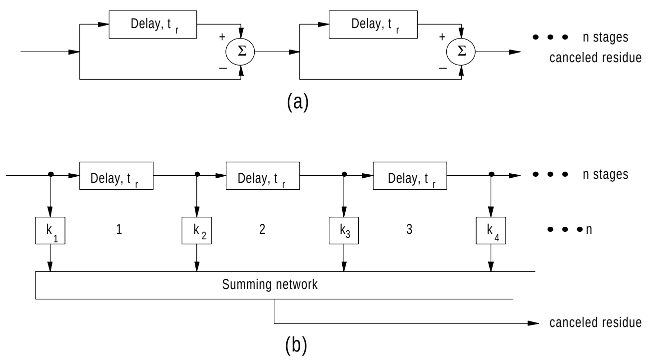


Figure C6 Multiple-delay cancelers. (a) Cascaded canceler sections; (b) weighted summer (after Nathanson, 1969, Fig. 9.7, p. 328).

The configuration using n cascaded sections is equivalent to the weighted sum of the $n + r$ pulses. Multiple-delay cancelers provide greater clutter rejection than the single-delay cancelers, but the rejection of target signals near the blind frequencies becomes more severe as n is increased. This disadvantage can be minimized using various feed-back and feed-forward circuits (Fig. C6), with which the clutter notch may be shaped while preserving a very flat signal response between blind speeds. *SAL*

Ref.: Nathanson (1990), pp. 328–330.

A **multiple-sidelobe canceler** is one using more than one auxiliary channel to cope with multiple-jammer environments.

Ref.: Farina (1992), pp. 110–111; Nitzberg (1992), pp. 55–86.

The **cancellation ratio** is a measure of canceler operation effectiveness. The term is applied to moving target indication, sidelobe cancellation, and polarization cancellation. The usual notation is CR. For MTI, CR is defined as "the ratio of canceler voltage amplification for fixed target echoes received with a fixed antenna to the gain for a single pulse passing through the unprocessed channel of the canceler." For sidelobe cancellation, CR defined as the ratio of the output interference power without and with the auxiliary antenna.

The term *integrated cancellation ratio* (ICR) is applicable to polarization cancellation. It is "the ratio of radar power received by a radar utilizing a normally circularly polarized antenna (common for transmission and reception) to the radar power received by the same radar utilizing a linear polarized (but otherwise identical to the above) antenna (again, common for transmission and reception) when the antenna, in both instances, is completely surrounded by a large number of randomly distributed small symmetrical targets." The integration is over the entire two-way antenna pattern. Such circular polarized antennas are typically used to suppress precipitation clutter returns (see **CLUTTER, rain**), and achieve ICR of 10 to 20 dB. *SAL*

Ref.: IEEE (1993), p. 153; Johnston (1984), pp. 23–28; Schleher (1991), p. 118; Farina (1992), p. 101.

A **shaped-delay-line canceler** is a delay-line canceler employing both feed-forward and feed-back paths, making it possible to shape the passband of the MTI filter (Fig. C2). The main disadvantage of such a canceler is poor transient response resulting from the feedback paths that allow interfering pulses to recirculate through the filter for many interpulse periods. *SAL*

Ref.: Schleher (1991), p. 82.

Sidelobe cancellation (SLC) is the process of the cancellation of the interfering signals if they are sensed as originating in the sidelobes of the main antenna. Typically, sidelobe cancellation is implemented by using one or more auxiliary channels, incorporating receivers connected to broad beamwidth antennas. The methods used in sidelobe cancellation can be categorized as incoherent (amplitude) and coherent. The configuration of the incoherent canceler is given in Fig. C7.

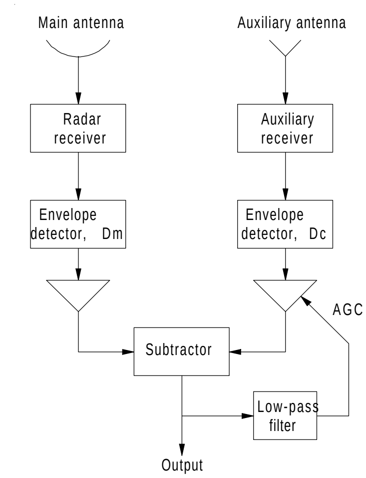


Figure C7 Noncoherent (video) sidelobe canceler (after Fielding, et al., 1977).

Jamming received through the sidelobes can be canceled in the subtracting system if the signals generated by detectors Dm and Dc begin to act at the same time and have identical durations and envelopes. Since this condition is very difficult to achieve in practice, the effectiveness of incoherent cancellation is rather poor in most of the practical interference environments. The canceling technique for coherent sidelobe cancellation is shown in Fig. C8.

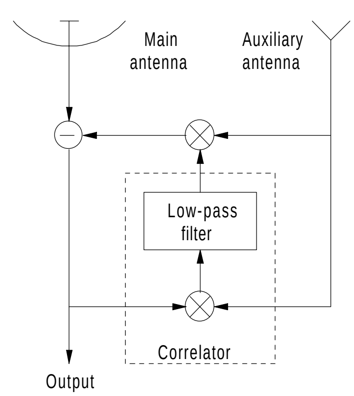


Figure C8 Coherent sidelobe canceler.

A portion of the signal from the auxiliary antenna is coherently subtracted from the main channel. The correlator is used to look for a component of the main channel output, which correlates with the output of the auxiliary channel, and the weight of the coupling from the auxiliary to the main channel is changed until there is no longer any correlated output. Then the combined pattern of the antennas has a null at the jammer location, which moves automatically as the main antenna scans its sidelobes through the jammer position (Fig. C9). Two distinct ways of forming the auxiliary channels are called element space SLC and beam space SLC.

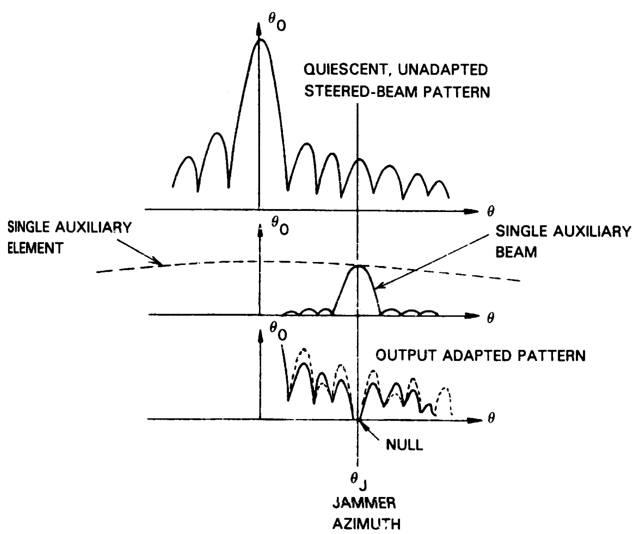


Figure C9 Patterns of radar antenna with sidelobe canceler (from Cantafio, 1989, p. 457).

Sidelobe cancellation with beam space implementation refers to a way of forming the auxiliary channel when the auxiliary antenna (antennas) form a beam (beams) in the direction of the individual jammers and the beam-formed outputs are sent into individual receivers, where SLC auxiliary outputs are formed. The beam space SLC implementation perturbs the main antenna's quiescent sidelobes less than does an element space SLC, but its complexity and cost are higher.

Sidelobe cancellation with element space implementation refers to a way of forming the auxiliary channel when individual elements or a cluster of elements are sent through a receiver, and, thereafter, this output forms an auxiliary channel into whichever SLC scheme is implemented. *SAL*

Ref.: Fielding, J. G., et al., *Radar-77*, London, 25-28 Oct. 1977, pp. 212–217; Skolnik (1990), p. 9.11.

A **sidelobe canceler** is the device that employs one or more auxiliary antennas and receivers to allow linear subtraction of interfering signals from the desired output if they are sensed to originate in the sidelobes of the main antenna. The technique uses the same antenna-and-receiver configuration as the sidelobe blanker, except that a gain-matching and -canceling process takes place. A sidelobe canceler using more than one auxiliary channel is termed a *multiple-sidelobe canceler.* Implementation schemes for sidelobe cancelers are usually categorized as closed-loop techniques (e.g., the Howells-Applebaum canceler) and open-loop (or direct solution) techniques (e.g., the Gram-Schmidt canceler). The closed-loop technique is simpler and less costly than the open-loop one, but the latter has better speed of convergence and stability. Recently a third group of methods (data-domain or voltage-domain methods) has been developed. They are similar to the open-loop techniques but do not require estimating and inverting the covariance matrix for data. *SAL*

Ref.: IEEE (1993), p. 1217; Farina (1992), pp. 95–216; Nitzberg (1992), pp. 41–105.

The **sidelobe canceler-blanker system** is a combination of a sidelobe blanker, for jamming of low duty factor, and a sidelobe canceler, for jamming of high duty factor, using the same auxiliary antenna and receiving preamplifier. A block diagram of combined SLC-SLB system is shown in Fig. C10. *SAL*

Ref.: Barton (1988), p. 369; (1991), p. 12-10; 172–189; Nitzberg (1992), pp. 46–53; Maksimov (1979), pp. 172–189; Cantafio (1989), p. 456.

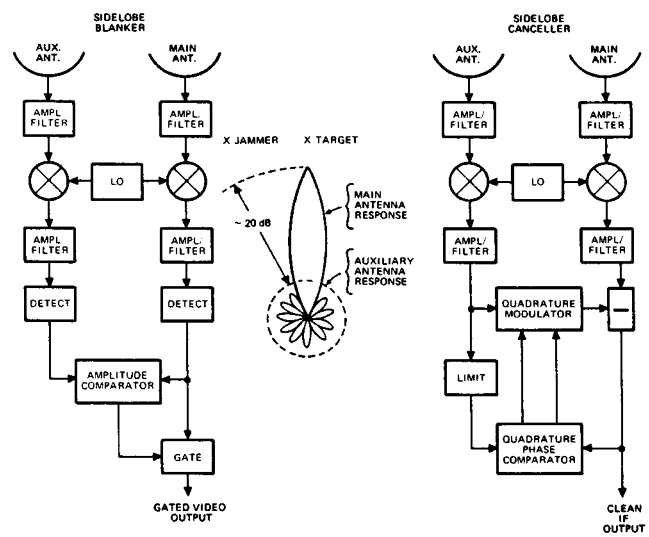


Figure C10 Combined sidelobe canceler and blanker (from Barton, 1988, Fig. 7.5.4, p. 371).

A **single-delay(-line) canceler** employs only one delay line (Fig. C11). Its improvement factor, for clutter centered in the notch, is given by the formula

$$I_2 = 2\left(\frac{f_r}{2\pi\sigma_c}\right)^2 = 2\left(\frac{v_b}{2\pi\sigma_v}\right)^2$$

where f_r is the pulse repetition frequency, σ_c is the rms clutter spread in frequency, v_b is the blind speed, and σ_v is the rms

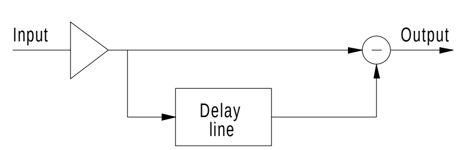


Figure C11 Single-delay MTI canceler.

velocity spread of the clutter. (See **CLUTTER spectrum.**)
Ref.: Barton (1964), p. 212; Currie (1987), pp. 640, 739.

three-pulse canceler (see **double-delay canceler**).

A **triple canceler** is a cascaded configuration of delay-line cancelers employing three delay lines, each connected as a single canceler. It is also called a *four-pulse canceler. SAL*
Ref.: Skolnik (1980), pp. 110, 135.

CAPACIVITY (see **DIELECTRIC constant**).

CAPTURE EFFECT. The capture effect is "the tendency of a receiver to suppress the weaker of two time-coincident signals within its passband."
Ref.: IEEE (1993), p. 158.

A **CARCINOTRON** is a type of backward-wave tube using a nonreentrant injected beam that interacts with a traveling wave in a dispersive backward-wave circuit. M-carcinotron and O-carcinotron types are distinguished. The typical frequency range is between P- and K_u-band, CW power is greater than 100W. The primary use is for noise-jamming signal sources in electronic countermeasures equipment. *SAL*
Ref.: Skolnik (1970), p. 7.19; Popov (1980), p. 170.

A **CATHODE-RAY TUBE (CRT)** is a cathode-ray electronic device having the shape of a tube elongated in the electron-beam direction (Fig. C12). It is used in radar as a "display device, in which controlled electron beams are used to present alphanumeric or graphical data." It produces visible radiation by bombardment of a thin layer of a phosphor material by an energetic beam of electrons, and typically consists of an electron-beam-forming system, electron-beam-deflecting system, phosphor screen and evacuated envelope.

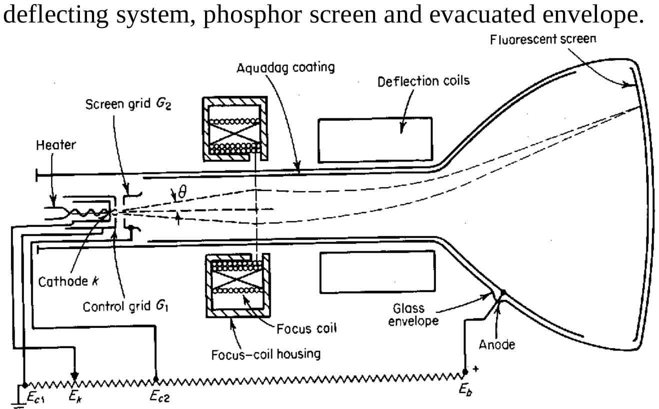


Figure C12 Cathode-ray tube (from Fink, 1982, p. 11.49, reprinted by permission of McGraw-Hill).

CRTs usually are classified, with respect to the methods of electron-beam control, into electrostatic or electromagnetic focusing CRTs. They are classified, with respect to the methods of blip formation, into deflection-modulated or intensity-modulated CRTs. Because of its flexibility of performance, resolution, dynamic range, and simplicity of hardware, the CRT is the most common display device used for radar indicators. (See also **DISPLAY, radar**.) *SAL*
Ref.: IEEE (1993), p. 166; Skolnik (1970), Ch. 6, (1980), pp. 353–359; Fink (1982), pp. 11.3–11.45, 25.84.

A **color CRT** is a one for the display of information in color. As applied to radar, target altitude, cross section, clutter intensity, and other parameters might be color coded to provide an additional dimension for the display of target and environmental information. One configuration is a penetration color CRT using a multilayer screen. The color is controlled by the anode voltage, which determines the depth of penetration of the electron beam into a series of phosphor layers. This method of color excitation permits better resolution than that obtained with conventional TV color tubes, which explains its interest for radar applications. *SAL*
Ref.: Skolnik (1980), pp. 353–359.

A **deflection-modulated CRT** displays echo signals by deflection of the electron beam (e.g., the A-scope display). See **DISPLAY, radar**.
Ref.: Skolnik (1970), Ch. 6, (1980), p. 353.

direct-view storage CRT (see **cathode-ray storage tube**).

A **cathode-ray tube with dual deflection** uses simultaneous magnetic and electrostatic deflection systems. Magnetic deflection shifts the beam to a specific point on the screen, and electrostatic provides small deflections at small angles. Thanks to this design, the CRT makes it possible to write symbols at high speed at a point to which the beam of the magnetic deflection system is directed. The distance between the deflecting plates may be quite small to obtain high deflection sensitivity. The electron gun and the deflecting plates can be made in the form of a single unit with connecting wires run through the base.

With such a CRT design with a high voltage in the focusing electrode, the deflecting plates can be under the potential of the focusing electrode. *IAM*
Ref.: Skolnik (1970), Ch. 6.

A **CRT with electromagnetic focusing** uses an electromagnet around the neck of the CRT to provide an axial magnetic field to accomplish the focusing and deflection of the electron beam. The sine of the deflection angle of the beam is directly proportional to the intensity and extent of the magnetic field and inversely proportional to the square root of the accelerating voltage. As a result of the weaker dependence on the accelerating voltage (in comparison with CRTs with electrostatic deflection), in a CRT with electromagnetic (magnetic) deflection, usually one accelerating anode is used. A CRT with electromagnetic focusing provides good beam focusing, and such tubes are typically used for large, intensity-modulated CRTs (e.g., plan position indicators). CRTs with electromagnetic focusing are relatively insensitive and require more drive power than CRTs with electrostatic focusing. *IAM*
Ref.: Skolnik (1970), Ch. 6.

A **CRT with electrostatic focusing** uses two pairs of deflecting electrodes or plates to provide an electric field to accomplish both focusing and deflection of the electron beam. Equal voltages of opposite polarity are sent to the paired plates. The tangent of the angle of deflection of a beam is directly proportional to the intensity of a homogeneous electrical field between the deflecting plates and the extent of the deflecting field, and inversely proportional to the accelerating voltage. For this reason, to increase the brightness of the spot on the screen with small deviating voltages and short tube length, CRTs with postacceleration are used. The beam is deflected at some intermediate value of the accelerating voltage, after which the accelerating voltage reaches its full value.

CRTs with electrostatic deflection provide wider bandwidth compared with CRTs with electromagnetic deflection. This determines the primary use for displaying rapidly changing processes. Such tubes are typically used for the smaller, deflection-modulated CRTs (e.g., A-scopes). (See **DISPLAY, radar**.) They usually have a greater length-to-diameter ratio than CRTs with magnetic focusing, but overall size, weight, and power dissipation are less. *IAM*

Ref.: Skolnik (1970), Ch. 6.

An **intensity-modulated CRT** is one in which echo strength is indicated by intensifying the electron beam and presenting a luminous spot on the face of the CRT. An example is the plan-position indicator (PPI). (See **DISPLAY, radar**.) *IAM*

Ref.: Skolnik (1970), Ch. 6, (1980), p. 353.

penetration color CRT (see **color CRT**).

A **cathode-ray storage tube** is designed for writing signals on a dielectric with their subsequent reproduction in the form of an optical or energy signal. Storage CRTs may or may not have a visible image. The former are analogous to oscilloscope tubes, but the image on the screen can be saved without change for a specific time. The latter are distinguished by the fact that the image written on the storage surface in half-tones can be saved for a long time and "read" at any time (i.e., input in the form of electrical signals). These tubes can be used, for example, for converting a radar image to a television image or for creating bright displays. In the latter case, a direct-view storage CRT can be used (see **DISPLAY, radar**). *IAM*

Ref.: IEEE (1993), p. 166; Popov (1980), pp. 97, 130, 312; Skolnik (1980), p. 357.

CELL, radar resolution. The radar resolution cell (or resolution element) is "a spatial and velocity region contributing echo energy which can be separated from that of adjacent regions by action of the antenna or receiving system. In conventional radar its dimensions are given by the beamwidths of the antenna, the transmitted pulsewidth, and the receiver bandwidth; its dimensions may be increased by the presence of spurious regions (sidelobes), or decreased by use of specially coded transmissions and appropriate processing techniques." If the volume contains a target (though it may occupy only a portion of the cell), it is termed a *target cell*, and if it contains undesired scatters, it is a *clutter cell*. *SAL*

Ref.: IEEE (1993), p. 1,133.

The **target cell** is the radar resolution cell within which the target is located.

The **test cell**, in a constant-false-alarm-rate detection system, is the radar resolution cell in which target detection is attempted, based on a threshold established by the surrounding reference cells.

The **volume cell** V_c is the volume within which the targets are not resolved. For specified azimuth beamwidth θ_a elevation beamwidth θ_e and processed pulse width, τ_n:

$$V_c = \frac{R\theta_a}{L_p} \times \frac{R\theta_e}{L_p} \times \frac{\tau_n c}{2}$$

where R is the target range, c is the velocity of light and $L_p = 1.33$ is the beamshape loss. *SAL*

Ref.: Nathanson (1969), p. 69; Barton (1988), p. 38.

CENTROIDING, in detection, is the process of determining the location of the target within a cluster of detections. Usually angle centroiding and range centroiding are distinguished.

Angle centroiding is a process for determining the angle location of the target when signals from the target are detected in more than one angle (resolution) cell. For targets of a large RCS, particularly when the targets are very close, detection can occur when the target is within the main beam but outside the 3-dB beamwidth. When beam scans across the target, detections occur over a sector much larger than the antenna beamwidth. The appearance is as if there were multiple targets at the same range, spread over a wide angular sector. One of the algorithms, as shown in Fig. C13, maintains a set of counters for each range bin. When the number of adjacent hits (detections) reaches a number corresponding to the angular region in which the main antenna gain exceeds the guard antenna gain, or when two successive misses occur, the count is terminated and the detection is displayed at an angle midway between the current detection and first detection.

Range centroiding is a process for determining range location of the target when signals are detected in more than one range (resolution) cell. One approach is similar to the angle centroiding scheme of Fig. C13. An alternative approach to ranging, not using centroiding, examines the set of magnitudes that produce adjacent hits and chooses the local maximum magnitude as true target location. *SAL*

Ref.: Morris (1988), p. 266.

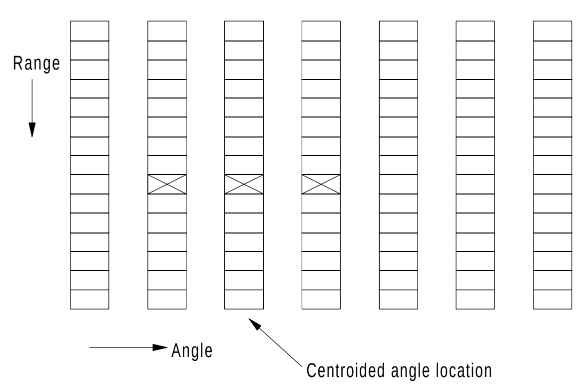


Figure C13 Angle centroiding algorithm (after Morris, 1988, Fig. 13.11, p. 266).

The **CEPSTRUM** is a function used in nonlinear signal processing. The term is a "half-inverse" of the word *spectrum*. The most common forms are the cepstrum power and the complex cepstrum, related to the Fourier transform F of the signal by the relations

$$K_p(\tau) = |F\{ln|F[s(t)]|^2\}|^2$$
$$K(\tau) = F\{ln[F\{s(t)\}]\}$$

respectively. The argument of the cepstrum has the units of time.

A feature of a cepstrum analyzer is its ability to perform the deconvolution of signals (i.e., the operation which is the inverse of the convolution of two functions and makes it possible to extract the convolved signals).

Cepstral processing is performed using the fast Fourier transform (FFT) in digital computers, charge-coupled devices, or Fourier processors based on surface acoustic wave techniques.

The cepstrum power is used to detect and classify signals, to determine pulse duration, and to measure the pulse repetition frequency. The complex cepstrum, on the other hand, makes it possible to reconstruct signals distorted by multipath propagation and to process radar images. *IAM*

Ref.: Childers, D. G., Skinner, D. P., and Merait, R. C., "The Cepstrum: A Guide to Processing," *Proc. IEEE* ***65***, no. 10, 1977, pp. 1428–1443.

CHAFF is "strips of lightweight metal or metallic material that are dispersed in large numbers (bundles) into surveillance or observation volume of a radar to reflect impinging signals and simulate a true target." It typically consists of strips of aluminum foil or metal-coated fibers. Each bundle of chaff may contain thousands of individual reflectors whose lengths are related to the wavelength of the radar. Chaff for use at HF and VHF bands is called *rope* (or *rope chaff*).

As a passive electronic countermeasures technique, chaff may be used in three distinct ways:

(1) A localized chaff puff or burst can simulate a true target, serving as a decoy for radars not having doppler clutter rejection capability, and contributing to overload in the data leaving the radar.

(2) A chaff corridor can be laid behind the lead aircraft in a raid, or fired in rockets forward from the lead aircraft, masking aircraft within the corridor.

(3) A chaff cloud may be dispensed over a large region, masking the actual attack corridor of a subsequent raid. *SAL, DKB*

Ref.: IEEE (1993), p. 173; Schleher (1986), p. 133; Barton (1988), p. 194; Nathanson (1990), p. 295.

Chaff dipoles are chaff elements, typically lightweight metal (e.g., aluminum) or aluminum-plated glass or Mylar dipoles, which are dispensed in the atmosphere to reflect radar energy imitating the real target with large RCS. The chaff dipoles are intended to resonate at the frequencies of victim radars, requiring that the length of the dipole be about one-half radar wavelength. Often the dipoles in a package are cut to different lengths to cover an entire radar band or several radar bands. The overall RCS of dipoles randomly oriented after dispersal is about

$$\sigma = 0.18\lambda^2 N$$

where λ = wavelength and N is the total numbers of dipoles. *SAL*

Ref.: Nathanson (1969), p. 223.

Delayed opening chaff is chaff that blooms at a specific elapsed time after it is dispensed.

Ref.: Johnston (1979), p. 58.

jammer-illuminated chaff (see **JAFF**).

Rope chaff is a type of chaff consisting of one or more rope elements, which are long rolls of metallic foil or wire designed for broadband, low-frequency response. *SAL*

Ref.: Johnston (1979), pp. 58, 66; Neri (1991), p. 396.

CHANNEL

A **difference channel** is "a part of monopulse receiver dedicated to the amplification, filtering, and other processing of a difference signal, which is generated by comparison of signals received by two (or two sets of) antenna beams, and indicating the departure of the target from the boresight axis. Also a signal path through a monopulse receiver for processing a difference signal that is commonly generated by comparison of two or more signals received by two antenna beams." (See **MONOPULSE**.) *SAL*

Ref.: IEEE (1993), p. 343.

A **guard channel** is "one or more auxiliary processing channels to control the main processing channel in order to reject interference that is partly in, but not centered on, the main channel. Guard channels may be displaced in time (range), Doppler frequency, carrier frequency, or angle. Sometimes called 'guard gates,' 'guard bands,' or 'sidelobe blanking.' Used against range gate stealers, velocity gate stealers, chaff, sidelobe jamming, and to enhance apparent angle resolution in IFF. May use auxiliary displays." *SAL*

Ref.: IEEE (1993), p. 576.

An **image channel** is the path of image-frequency signals (those displaced by the intermediate frequency on the opposite side of the local oscillator signal from the intended response) into the mixer of the superheterodyne receiver. *IAM*

Ref.: Van Voorhis (1948), p. 17; Popov (1980).

The **in-phase channel** is the channel with its signal in phase with the coherent (reference) oscillator, in a quadrature-channel processor. *IAM*

Ref.: Skolnik (1980), p. 119.

The **quadrature channel** is the channel with 90° phase relatively to the coherent (reference) oscillator in a quadrature-channel processor. *SAL*

Ref.: Skolnik (1980), p. 119

A **radar channel** is a channel in which the generation, transmission, reception, and processing of radar information is performed. The radar channel includes the target, the propagation path, and the radar itself. The radar data is formed through the interaction of electromagnetic waves radiated by the antenna with the radar target. The information is transported by an electromagnetic field (space-time process) to the radar receiver, in which it is transformed into a radar signal (time process).

The radar channel may be analyzed through a description of the transformation of the signal characteristics along the transmitter-target-receiver path, using quantitative estimates in various information metrics.

Such an information-based description of the radar channel is concerned not with the signals themselves, but with their content: the quantity and quality of the information, giving a quantitative estimate in information metrics of the effectiveness of the operation of the radar as a whole and, if possible, its components.

An information-based description makes it possible to examine the functioning of the radar from the most general viewpoint, to optimize performance of the radar, and to simplify estimation of radar characteristics during preliminary design and experimentation. *IAM*

Ref.: Tuchkov (1985), p. 11.

sum channel (see **MONOPULSE**).

CHART

A **Blake chart** is the pulsed-radar-range-calculation worksheet displaying the terms and basic steps for radar range equation calculations (Fig. C15). It is based on logarithmic-decibel form of the range equation and was devised by L. V. Blake in 1969 for manual computation. Today, for computer-aided computation, the output form of range calculation worksheet can differ from Blake's original form but typically is also termed Blake chart. *SAL*

Ref.: Blake (1980), p. 383.

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PULSE-RADAR RANGE CALCULATION WORKSHEET

Detection probability P_d =				False-alarm probability P_{fa} =		Target case:		Hits n =	
Radar antenna height h_r =			m			Target elevation angle θ_t =		deg	
A. Computation of T_s: $T_s = T_a + T_r + L_r T_e$				B. Range Factors		C. Decibel Values		Plus (+)	Minus (–)
				P_t or P_{av} (W)		10 log P (dBW)			
(a)	Compute T_a			τ or t_f (s)		10 log τ (dBs)			
	For $T_g = T_{ta}$ = 290K, T_g = 36K			G_t		$G_{t(dB)}$			
	$T_a = (0.876T_a' - 254) + 290K$			G_r		$G_{r(dB)}$			
	$L_{\alpha(dB)}$:		L_α:	σ (m^2)		10 log σ (dBm^2)			
	T_a' =	K		λ (m)		20 log λ (dBm^2)			
		T_a:	K	T_s (K)		–10 log T_s (dBK)			
(b)	Compute $T_r = T_{tr}(L_r - 1)$			D		$-D_{(dB)}$			
	$L_{r(dB)}$:		L_r:	M		$-M_{(db)}$			
		T_r:	K	L_p		$-L_{p(dB)}$			
(c)	Compute $T_e = T_0(F_n - 1)$			L_x		$-L_{x(dB)}$			
	$F_{n(dB)}$		F_n:	L_t		$-L_{t(dB)}$			
	T_e: K		L_r:	Range-equation constant (dBK·s/K)				**+75.62**	
		L_rT_e =	K	5. Obtain column totals					
(d)	Add:	T_s:	K	6. Enter the smaller below the larger					
7.	Subtract to obtain net decibels $X = 40 \log R_{km}$ (dBm^4)								

8.	Calculate R_0(km) = antilog (X/40)	$R_{0(km)}$ =	
9.	Calculate the pattern-propagation factor $F = \sqrt{F_t F_r}$	F =	
10.	Multiply R_0 by the pattern-propagation factor to obtain $R' = R_0 \times F$	R' =	
11.	From curves of atmospheric attenuation, determine the loss $L_{\alpha(dB)}$, corresponding to R'. This is $L_{\alpha(dB)(1)}$.	$L_{\alpha(dB)(1)}$=	
12.	Find the range factor δ_1 = antilog ($-L_{\alpha(dB)(1)}$/40)	δ_1 =	
13.	Multiply R' by δ_1. This is a first approximation of range, R_1.	R' =	
14.	If R_1 differs appreciably from R', find a new value of $L_{\alpha(dB)}$ corresponding to R_1.	$L_{\alpha(dB)(2)}$=	
15.	Find the range increase factor δ_2 corresponding to the difference between $L_{\alpha(dB)(1)}$ and $L_{\alpha(dB)(2)}$.	δ_2 =	
16.	Multiply R_1 by δ_2 to obtain the maximum radar detection range R_m in km	$R_{m(km)}$ =	

Figure C15 Blake chart for pulse-radar-range calculation, modified as in Barton, 1988, Fig. 1.2.4, p. 21.

A **range-height-angle chart** is used for graphic description of detection coverage of a search radar (or the acquisition envelope of a tracking radar), using an earth surface contour that has been curved so that ray paths appear as straight lines (Fig. C16). For most practical combinations of detection range and altitude, the program VCCALC can be used to draw these charts. Range-height-angle charts for exponential atmosphere for low, high, and space altitudes are given in Morchin (1993). (See also **atmospheric refractive index.**) *SAL*

Ref. Fielding (1988); Skolnik (1990), p. 2.45; Barton (1991), pp. 8–17; Morchin (1993), pp. 307–312.

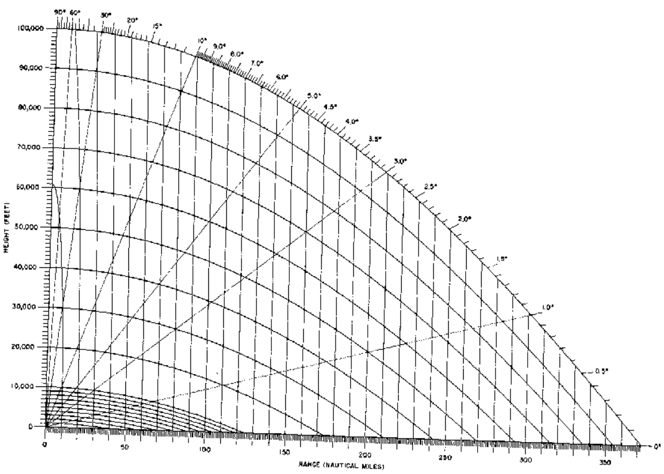


Figure C16 Radar range-height-angle chart (from Skolnik, 1990, Fig. 2.18, p. 2.45)

CHIRP is "a technique for pulse compression that uses frequency modulation (usually linear) during the pulse" (see also **WAVEFORMS, radar.)**

Ref.: IEEE (1993), p. 184; Johnston (1979), p. 56; Wehner (1987), p. 127.

CHOKE, MICROWAVE. A microwave choke is a microwave element of the transmission line used as a large inductive resistor. The choke is constructed in the form of multiple-turn pellicular spirals of various shapes (with an inductance of up to 100 nH) in integrated circuits. In waveguides, chokes are typically used to prevent the energy in a specified frequency range from taking an undesired path. They are also used in contactless couplings (usually rotary joints) of both waveguide and coaxial transmission lines. Such chokes are constructed in the form of a short-circuiting half-wavelength stub, or in the form of grooves and depressions at the joint (flange). Inside the choke a standing wave is formed such that the voltage drop at the resistance of the contacts is equal to zero, there are no microwave power losses, and the electrical characteristics of the rotating coupling at the operating frequency are not dependent on the quality of the friction contacts. With correct values of the groove wave resistance, the traveling-wave ratio of the coupling exceeds 0.9 in the 50-to-70% frequency band of the operating frequency. *IAM*

Ref.: IEEE (1993), p. 184; Ridenour (1947), p. 396; Gassanov (1988) p. 65; Sazonov (1988) p. 53.

CIRCUIT

An **anticlutter circuit** is one "that attenuates undesired reflections to permit detection of targets otherwise obscured by such reflections."

Ref.: IEEE (1993), p. 45.

A **balanced circuit** is one "in which two branches are electrically alike and symmetrical with respect to a common reference point, usually ground." Balanced circuits are widely used in various devices: balanced amplifiers, balanced mixers, and so forth. *IAM*

Ref.: IEEE (1993), p. 87; Popov (1980), p. 45.

beam-forming circuit (see **FEED, beam-forming**).

bridge circuit (see **BRIDGE, microwave**).

Circuit damping is "the temporal decay of the amplitude of oscillations in a tuned circuit associated with the loss of energy."

Ref.: IEEE (1993), p. 302; Popov (1980), p. 131.

A **differentiating circuit** is one in which the output signal is nearly proportional to the rate of change of the input signal. The differentiating circuit consists of series-connected resistors and capacitors or resistors and inductors. They are used to measure the rate of change of a voltage, to select pulse edges, and to perform other voltage transformations. *IAM*

Ref.: IEEE (1993). p. 347; Popov (1980), p. 116.

A **distributed circuit** is one for which circuit elements exist continuously along the network; that is, any elementary small portion of the circuit has its resistance, capacitance, and inductance. In this circuit the areas occupied by electromagnetic fields and the areas of energy loss overlap. Typically, a circuit can be considered distributed if its dimension are more than some specified fraction of a wavelength. Microwave transmission lines (e.g., waveguide) are one example of distributed circuits. *SAL*

Ref.: IEEE (1993), p. 375.

circuit, integrated (see **INTEGRATED circuit**).

An **integrating circuit** has an output signal that varies in proportion to the integral of the input signal. Integrating circuits are used to extend pulses, to smooth (filter) signals, to obtain linearly varying voltages, to create delay in the triggering of electronic circuits, and for other purposes. *IAM*

Ref.: IEEE (1993), p. 663; Popov (1980), p. 158.

A **linear circuit** is a circuit performing linear transformations of the input signals $\vec{x}(t)$. Linear circuits can be of the inertia-type and inertialess-type. For the latter, the vector of output signals $\vec{y}(t) = \vec{k}(t) \cdot \vec{x}(t)$, where $\vec{k}(t)$ is the time-functions matrix, and output signals in the specified time moment t depend not only on the input signals in the same time moment, but also on the input signals in other moments of time. This dependence can be determined as the superposition of the input values in different time moments multiplied by the impulse response $h(t)$

$$y_i(t) = \int_{-\infty}^{t} h_{ij}(t,\tau)x_i(\tau)d\tau, \quad i, j = 1, N$$

When the circuit is of inertialess-type configuration (the parameters of the circuit do not change with time), $h_{ij}(t,\tau) = h_{ij}(t-\tau)$ and the circuit characteristics depend only on one parameter: time shift, τ. A linear circuit does not change the probability distribution of the input signal (i.e., if the input is Gaussian the output will also be Gaussian). Examples of linear lumped circuits are oscillating circuits, filters, differentiating, and integrating circuits. Linear distributed circuits are represented by different types of transmission lines (e.g., coaxial lines). *IAM*

Ref.: Levin (1974), p. 214; Oppenheim (1983), Ch. 3; Gonorovskiy (1986), p. 12.

A **lumped circuit** is one in which electromagnetic fields are concentrated in discrete elements (capacitors, inductors, resistors). The elements of the circuit are connected by wires having fields that can be neglected. *SAL*

Ref.: Terman (1955), Ch. 3.

A **nonlinear circuit** (or network) is one performing nonlinear transformations of input signals. The generic methods of analysis are well designed primarily for inertialess nonlinear circuits, in which the input *s*(*t*) and output *y*(*t*) at any moment are related by a single-valued function:

$$\vec{y}(t) = f[\vec{x}(t)]$$

This dependence is a good approximation for describing many types of radio circuits: limiters, mixers, detectors, and so forth. A characteristic of nonlinear circuits is the presence in the output spectrum of harmonics not present in the input spectrum. (See also **filter, nonlinear.)** IAM

Ref.: IEEE (1993), p. 856; Levin (1974), vol. 1, p. 214; Gonorovskiy (1986), p. 14.

An **oscillatory circuit** is one "containing inductance and capacitance so arranged that when shock excited it will produce a current or voltage that reverses at least once." Typically, oscillatory circuits have one or several resonators and additional circuits to control and maintain oscillations. According to the character of the electrical and magnetic fields distribution, they are divided into lumped circuits, based on lumped resonators with capacitance *C* and inductance *L*, and distributed circuits, based on the sections of different transmission lines (coaxial lines, waveguide lines, strip lines, etc.). According to the operating band, oscillatory circuits are divided into narrowband circuits with no capabilities for frequency agility (only limited tuning capabilities), typically based on a single resonator, and wideband circuits employing frequency agility capabilities and based on a set of resonators. Depending on the type of resonator, the most widely used are oscillatory circuits with coaxial, waveguide, and strip line resonators, which are used in microwave filters, antenna radiating elements, and vacuum tubes. *AIL*

Ref.: IEEE (1993), p. 896; Gassanov (1988), p. 46; Kovalev (1972), p. 195.

A **parametric circuit** is one in which at least one of the parameters (resistance, capacitance, inductance) is a function of time. If the circuit parameters do not depend upon the operating mode, then the circuit is linear. Circuits with variable active impedance are used for detection, amplitude modulation, and frequency conversion. Circuits with variable reactive elements are suitable for storing and releasing energy and are used in parametric generators and amplifiers. *IAM*

Ref.: Popov (1980), p. 276.

A **printed circuit** is "a pattern comprising printed wiring formed in a predetermined design in, or attached to, the surface or surfaces of a common base." It is a wiring assembly of electronic components on a printed board, using printed wiring, while the elements themselves can also be produced by printing (using attached elements) on the boards using assembly openings or by placing planar leads on the contact areas of the boards. In this case such an assembly is termed a *printed-circuit assembly. IAM*

Ref.: IEEE (1993), p. 1,007; Popov (1980), p. 287.

A **sample-and-hold circuit** is a sampling circuit whose output is held at the level of the last sample until the next one arrives. Typically, it consists of an electric circuit that clamps the potential of a storage element (e.g., capacitor) to the video pulse amplitude each time the new sample is received. Obsolete terms for this circuit are the *boxcar detector* and *boxcar generator. SAL*

Ref.: Skolnik (1980), p. 156.

A **sampling circuit** is one "whose output is a series of discrete values representative of the values of the input at a series of points in time." (See **SAMPLING.**)

Ref.: IEEE (1993), p. 1,165.

CIRCULATOR, microwave. A circulator is a decoupling multichannel structure, in which the electromagnetic waves propagate from one channel to another only in a particular order. It is typically a "passive waveguide junction of three or more arms in which the arms can be listed in such an order that when power is applied to any arm it is transferred to the next on the list, the first arm being counted as following the last in order." Applied to a radar duplexer, it allows a single antenna to be used for both transmission and reception. It can be used in parametric amplifiers, circuits that add power from several generators, and so on.

Circulators are categorized as being either wave rotation or phase-shift circulators. They are constructed from various types of transmission line (Y-junction circulators). Phase circulators are usually used in high-power circuits, while low-power circuits usually employ Y-junction circulators. *IAM*

Ref.: IEEE (1993), p. 190; Skolnik (1970), pp. 8.20–8.24; Lavrov (1974), p. 355.

A **ferrite circulator** is one based on the interaction of electromagnetic waves with a magnetized ferromagnetic substance. Virtually all circulators used in radar systems contain ferrite elements (pins, disks, or plates) and are thus ferrite circulators.

Ferrite circulators are distinguished by the orientation of the magnetization. In wave-rotation circulators the ferrite is magnetized along the direction of propagation, while in phase-shift circulators the magnetization is perpendicular to that direction. This determines the various requirements for the degree of magnetization and heat transfer. *IAM*

Ref.: Skolnik (1970), pp. 8.20–8.24; Lavrov (1974), p. 355; Popov (1980), p. 455.

A **phase-shift circulator** is based on the creation of a nonreciprocal phase shift. This occurs in a rectangular waveguide with a perpendicularly magnetized plate, in which waves propagating in opposite directions have different phase velocities. The common element in various designs of such a circulator is the presence of nonreciprocal phase shifters in rectangular waveguides that share a common wall. *IAM*

Ref.: Skolnik (1970), p. 8.23; Popov (1980), p. 453.

A **wave-rotation circulator** operates by rotating the polarization plane of a wave in a waveguide with a magnetized ferrite rod. A wave entering at one of the branches is rotated as it passes the ferrite, which is magnetized along the direction of propagation, resulting in the required transmission and blocking at the other branches in the circulator. When the direction of magnetization is reversed, the sequence of transfers within the circulator is also reversed. A wave-rotation circulator is distinguished by requiring a relatively low magnetization level (on the order of several tens or hundreds of oersteds for circulators in the 3-cm waveband). Among its drawbacks are the difficulties associated with transferring heat from the ferrite. *IAM*

Ref.: Skolnik (1970), p. 8.21; Lavrov (1974), p. 357.

A **Y-junction circulator** is constructed with three rectangular waveguides joined at 120° angle relative to one another, with a magnetized ferrite rod at the joint. With a certain magnetization applied to the ferrite, it will reradiate a secondary wave with which the input wave from one of the branches is in phase at a second branch, and out of phase at the third branch. If the magnetization is reversed, the sequence of branches is reversed.

Y-junction circulators may also be constructed from coaxial and strip lines. They are distinguished by their simplicity and low size and weight. The bandwidth for a waveguide Y-junction circulator may reach 30%, and an octave for a strip line device. *IAM*

Ref.: Lavrov (1974), p. 358; Gardiol (1984), p. 266.

CLASSIFICATION OF TARGETS (see **TARGET RECOGNITION AND IDENTIFICATION.**)

CLIPPING refers to circuit operation in which the maximum amplitude of a signal is limited to a predetermined value. In digital signal processing, it causes quantization error and occurs when a quantized signal exceeds the saturation level of the A/D converter or register. Clipping results in a broadening of the signal spectrum and a narrowing of its correlation function. *SAL*

Ref.: Nathanson (1990), p. 556.

CLUTTER. Radar clutter is defined as "unwanted echoes, typically from the ground, sea, rain or other precipitation, chaff, birds, insects, or aurora." There is no single definition because one user's target is another's clutter: to the radar meteorologist, precipitation is the target and aircraft are clutter.

The effects of clutter on target detection are to mask some targets as they pass through regions occupied by clutter echoes and generate false alarms that may distract attention and draw resources from real targets. The effects on tracking and measurement are to generate false tracks and increase error in data on real targets. To reduce these effects, radars rely on reducing the size of the spatial resolution cell to minimize clutter input, providing velocity (doppler frequency) resolution to filter clutter from the signal processor output and applying constant-false-alarm-rate (CFAR) techniques to the detection circuitry. In some cases the proper choice of frequency or polarization can minimize clutter input to the radar.

Even if all practical measures are taken to reject clutter, it is usually necessary to allow the CFAR circuits to raise the detection threshold in regions of heavy clutter, compromising target detection.

Clutter sources are of three types: (1) discrete (or point) clutter, described by specific locations and radar cross sections; (2) surface clutter, described by a dimensionless reflectivity σ^0 (m^2 of RCS per m^2 area of surface); and (3) volume clutter, described by a reflectivity η_v (m^2 of clutter per m^3 of volume). Other descriptive parameters include the probability density function (pdf) of echo amplitude or power (see **clutter (amplitude) distribution**), the spectral distribution or temporal correlation function (see **clutter spectrum**), the spatial correlation function, the polarization and frequency dependence of reflectivity, the spatial distribution of clutter sources, and dependence on weather and diurnal or seasonal factors.

The clutter power entering the radar depends not only on the RCS, σ_c, of the clutter within the spatial resolution cell but also on the pattern-propagation factor, F_c, applicable to the path between radar and clutter. This factor describes the transmission properties of the radar-clutter path and the antenna gains along that path. Most measurements and tabulated values of surface-clutter RCS are actually values of $\sigma_c F_c^4$, dependent on the measuring radar antenna and its location as well as on the stated parameters of frequency, polarization, grazing angle, and type of clutter. Especially at low grazing angles, the propagation term in F_c^4 may be the dominant source of frequency dependence and broadened pdf. *DKB*

Ref.: IEEE (1990), p. 8; Barton (1964), pp. 95–108, (1988), pp. 123–139, (1991), pp. 5.14–5.21; Nathanson (1969), pp. 192–275; Skolnik (1980), pp. 470–512, (1990), pp. 12.1–13.40; Blake (1980), pp. 293–330; Long (1983); Schleher (1991), pp. 19–54, 171–277; Currie (1992); Morchin (1993), pp. 55–85.

Atmospheric clutter is defined as echoes from natural sources within some volume of the atmosphere, principally in the lower troposphere, and is a form of volume clutter. The usual sources include precipitation (rain, snow, hail), aurora, and refractive index anomalies sometimes called "angels" because of their uncertain origin. Chaff, the man-made equivalent of rain clutter, is discussed separately, although its model parameters are similar to those of atmospheric clutter. When a meteorological radar is considered, the precipitation actually constitutes the target, and the RCS calculations will yield target, rather than clutter, RCS. However, the following discussion applies the term clutter without regard to the radar objective.

The RCS of this type of clutter is calculated using the volume of the clutter cell V_c and the volume reflectivity η_v (see **volume clutter**). The reflectivity of precipitation depends on radar wavelength and the dielectric constant of the precipitation, according to the relationship

$$\eta_v = \frac{\pi^5}{\lambda^4}|K|^2 Z$$

where λ is radar wavelength, $K = (m^2 - 1)\,/\,(m^2 + 2)$ is the refractive index factor), m is the complex index of refraction, and Z is the average value of the sixth power of droplet diameter (called the radar reflectivity factor). Values of $|K|^2$ are near 0.93 for raindrops and 0.2 for ice crystals and snow. A typical model giving a value for Z in mm^6/m^3 is

$$Z = ar^b$$

where a and b are empirically determined constants and r is the precipitation rate. Values used in this relationship are given below for different types of precipitation. Note that these values, expressed in the conventional radar meteorological units, must be multiplied by 10^{-18} to find the reflectivity in m^2/m^3. The result, when using values of a and b predicting z in mm^6/m^3, is

$$\eta_v = \frac{\pi^5}{\lambda^4}|K|^2 \cdot a \cdot 10^{-18} \cdot r^b \quad \mathrm{m}^2/\mathrm{m}^3$$

(See also **ANGEL ECHO**, **aurora clutter**, **cloud clutter**, **rain clutter**, **snow clutter**.) *DKB*

Ref.: Atlas (1964), pp. 371–374; Bogush (1989), p. 30; Sauvageot (1992), pp. 111–122.

Clutter attenuation (CA) is defined as the ratio of processor input clutter-to-noise ratio $(C/N)_i$ to output ratio $(C/N)_o$. In a pulsed doppler processor, the input ratio is defined for a filter centered on the clutter doppler frequency, and the output ratio for a filter centered on the target doppler frequency. Hence, for pulsed doppler radar, CA will vary with target velocity. For MTI radar, the definition given above implies normalization to the noise gain of the processor. An older definition of CA in MTI radar using a delay-line canceler is "the ratio of clutter power at the canceler input to clutter residue at the canceler output, normalized to the attenuation for a single pulse passing through the unprocessed channel of the canceler." (See also **MTI**.) *DKB*

Ref.: Barton (1988), p. 254; Schleher (1991), p. 108; IEEE (1993), p. 157.

Aurora clutter. The aurora consists of ionized gases in the E region of the atmosphere, excited by charged particles from the sun. Radar echoes are observed primarily in the VHF and low UHF bands and can take either of two forms. Discrete echoes, observed at night, are extended vertically at a specific range, while diffuse echoes are observed during daylight and are extended horizontally along the E-layer. Both types have the fluctuating characteristics of complex targets and are displaced and spread in velocity. *DKB*

Ref.: Skolnik (1962), pp. 621–624.

Bistatic clutter is clutter illuminated from the direction of the transmitter and scattering in a different direction to a receiver. The bistatic angle is defined as the angle between the transmitting and receiving paths, varying from zero for the monostatic radar to π for inline forward scattering. The bistatic surface or volume reflectivity at small bistatic angles for clutter composed of scattering elements smaller than one wavelength is approximately the same as the monostatic reflectivity. Except at angles near forward scatter, bistatic clutter reflectivity can be approximated by the monostatic reflectivity at the bisector of the bistatic angle. For surface clutter in a vertical plane containing the two paths, bistatic reflectivity can be evaluated (Barton, 1985) as

$$\sigma_b^0 = \gamma \sin\left(\frac{\psi_1 + \psi_2}{2}\right) + \sigma_f^0$$

where ψ_1 and ψ_2 are the grazing angles of the two paths, and

$$\sigma_f^0 = \left(\frac{\rho_0}{\beta_0}\right)^2 \exp\left(-\frac{(\psi_1 + \psi_2)^2}{4\beta_0^2}\right)$$

is the specular reflectivity for a surface having reflection coefficient ρ_0 and rms slope $\beta_0/\sqrt{2}$. *DKB*

Ref.: Barton, D. K., "Land Clutter Models for Radar Design and Analysis," *Proc. IEEE* **73**, no. 2, Feb. 1985, pp. 198–204; Willis (1991), Ch. 9; Nathanson (1991), pp. 342–349.

chaff clutter (see **CHAFF**.)

Cloud clutter at microwave and lower frequencies is usually produced by precipitation droplets, although these may be suspended by updrafts, producing no precipitation at the surface. In millimeter-wave bands, droplet sizes normally associated with nonprecipitating clouds are more detectable. The reflectivity for the average cloud conditions is shown in Fig. C17, as a function of water droplet density, for different radar bands.

The reflectivity as a function of cloud density is most conveniently expressed in terms of the parameter Z, the summation of the sixth power of droplet diameters within one cubic meter of the cloud (see **atmospheric clutter**), but the usual relationships of Z with precipitation rate are invalidated by updrafts. Instead, a relationship with condensed water density, M in g/m^3, may be used:

$Z = 0.048M^2$ (average clouds)
$Z = 8.2M^2$ (advection fog)
$Z = 391M^{1.96}$ (cumulus at onset of rain)

These may be compared with

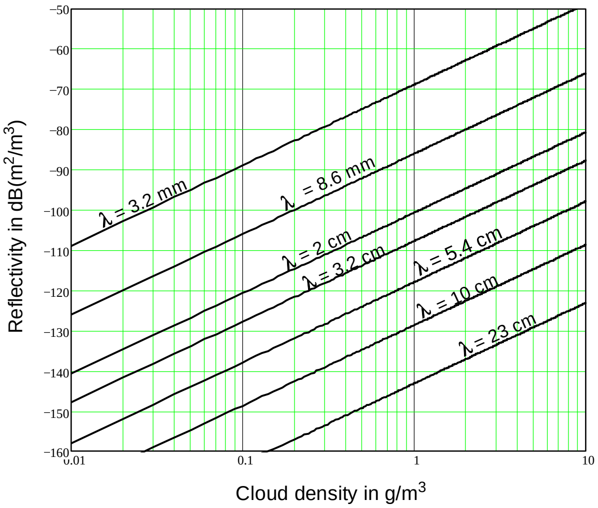


Figure C17 Cloud reflectivity vs. droplet density for different radar bands.

$Z = 24{,}000M^{1.82}$ (stratiform rainfall).

Note that the reflectivity increase by four orders of magnitude from average cloud conditions to onset of rain, and by another two orders of magnitude after rain begins, indicating that cloud reflectivity estimates are subject to wide variation.

Cloud particles are essentially spherical, and hence the reflectivity is reduced substantially through the use of same-sense circular polarization or orthogonal linear polarizations for transmitting and receiving. *DKB*

Ref.: Atlas (1964), pp. 376–378; Sauvageot (1992), pp. 119–123.

Clutter correlation functions in time, frequency, or space are statistical descriptions of relationships between clutter amplitudes observed at different times, carrier frequencies, or locations. The temporal correlation function $R(\tau)$ is defined and related to the Fourier transform of the clutter frequency power spectrum $S(f)$:

$$R(\tau) = \mathrm{E}[x(t)x^*(t-\tau)] = \int_{-\infty}^{\infty} S(f)\exp(\mathrm{j}2\pi f\tau)df$$

where E[·] denotes expected value and τ is the time difference between clutter samples. For a Gaussian clutter spectrum, the correlation time is $t_c = \sqrt{2\pi}\sigma_t = 1/(\sqrt{2\pi}\sigma_f)$.

The frequency correlation function is related to the impulse response of the clutter $h(t)$:

$$R(\upsilon) = \mathrm{E}[x(f)x^*(f-\upsilon)] = \int_{-\infty}^{\infty} h^2(t)\exp(\mathrm{j}2\pi\upsilon t)dt$$

The spatial correlation function is similarly defined in terms of a spatial displacement.

Clutter models often include correlation intervals in time and space, representing the widths of these correlation functions at some level below their peak values (for $\tau = 0$ in the case of correlation time). The correlation time interval t_c should be defined such that the number of statistically independent clutter samples n_c obtained within a given observation time t_o will be

$$n_c = 1 + \frac{t_o}{t_c}$$

This relationship corresponds to $t_c = 1/\beta_n$, where β_n is the noise bandwidth of the clutter spectrum. A similar relationship applies to samples available over an observation frequency interval Δf when the frequency correlation interval is f_c:

$$n_c = 1 + \frac{\Delta f}{f_c}$$

The correlation distance interval is similarly defined in space and is often controlled by the resolution of the radar, since the correlation function of observed clutter in a given coordinate is the convolution of the radar resolution in that coordinate and the inherent correlation function of the clutter scatterers. For example, in atmospheric clutter, where the individual scattering sources are small and closely spaced, the correlation distances in range and angle are equal to the widths of the range cell and the beamwidths. *DKB*

Ref.: Nathanson (1969), pp. 212–213, 249–253.

clutter detectability factor (see **DETECTABILITY FACTOR**).

Discrete clutter refers to echoes from objects whose dimensions are smaller than the clutter cell of the observing radar and exceeds by a significant ratio the amplitudes of the surrounding cells. Examples are echoes from radio or power-line towers, watertanks or other tall structures, cliffs, steep ridges, or similar natural features. Each discrete clutter source is characterized by its RCS in m^2, and is also termed *point clutter.* DKB

Ref.: Barton (1988), pp. 136–139; Schleher (1991), pp. 172–174, 272, 532;

distributed clutter (see **homogeneous clutter**).

The **clutter (amplitude) distribution** refers to the relative frequency of occurrence of different echo amplitudes and is described by the probability density function (pdf). The distribution can be Rayleigh (typical of uniform precipitation or surface clutter viewed at high grazing angles), or non-Rayleigh (surface clutter viewed at low grazing angles, especially with high-resolution radar). The central limit theorem of statistics dictates that the resultant of many independent, random vectors will follow the Rayleigh distribution, representing the sum of two independent components with Gaussian distribution and quadrature-phase relationship. This applies generally to precipitation and to surface clutter when no single scattering source is dominant. The detection threshold may be set with respect to the average value of Rayleigh clutter as for random noise, taking into account possible differences in integration gain when clutter is correlated from pulse to pulse.

For non-Rayleigh clutter, the pdf is broader, as shown in Fig. C18 for log-normal and Weibull distributions. The log-normal pdf appears as a normal distribution on a decibel scale, and when the standard distribution σ_y is set to 4 dB, this distribution gives a close match to the Rayleigh for amplitudes near or above the median. The Weibull family of pdfs have shapes such that the *a*th root of the amplitude is Rayleigh distributed, and hence their breadth increases in proportion to the parameter *a*, while the density decreases with this parameter. Another distribution used to describe clutter is the K-distribution, which may be thought of as the pdf of a Gaussian vector whose variance is modulated by a gamma distribution (Farina, et al., 1995). This type of distribution is useful in describing clutter such as waves at sea, rolling terrain, or rain cells, observed with radar resolution capable of separating regions of different mean level.

The large peak amplitudes encountered in non-Rayleigh clutter make it difficult to control false alarms with conventional CFAR techniques. Two-parameter CFAR, in which the spread of the pdf is measured and used to set the threshold higher above the mean than with Rayleigh clutter, causes greater target suppression, which can be expressed in the radar equation as a clutter distribution loss (see **LOSS**). Clutter maps, which permit setting a separate threshold for each resolution cell, can be used to restrict target suppression to the few cells actually containing strong clutter. The ability to detect targets between such strong clutter cells has been termed *interclutter visibility* (Barton and Shrader, 1969). Typical pdfs used to describe clutter amplitudes are given in Table C1 and illustrated in Figs. C18 through C25. *DKB*

Ref.: Barton, D. K., and Shrader, W. W., "Interclutter Visibility in MTI Systems," *IEEE Eascon Record*, 1969, pp. 294–297; Schleher (1991), p. 19; Currie (1992), pp. 14–27; Farina, A., et al., "Coherent Radar Detection of Targets against a Combination of K-Distributed and Gaussian Clutter," *IEEE Radar-95*, May 8-11, 1995, pp. 83–88.

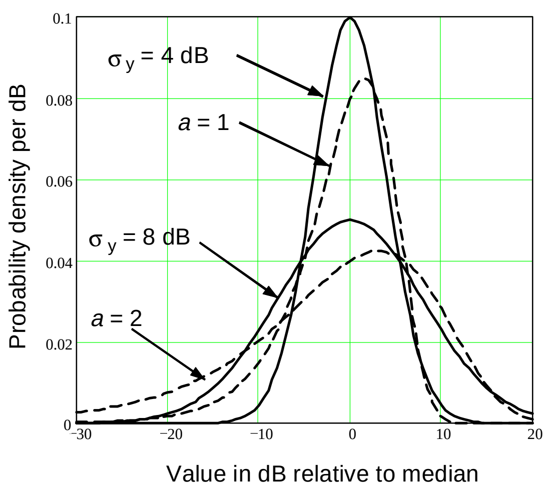


Figure C18 Log-normal (solid) and Weibull (dashed) distributions plotted on logarithmic scale (after Barton, 1988, Fig. 3.6.4, p. 132).

Table C1
Typical PDFs Used to Describe Clutter Amplitudes

Distribution and analytical form	Scattering conditions
Gaussian $W(x) = \frac{1}{\sqrt{2\pi}\sigma_x}\exp\left[\frac{-(x-\bar{x})^2}{2\sigma_x^2}\right]dx$	In-phase or quadrature voltage components of many scattering centers with random phase
Rayleigh (voltage) $W(x) = \frac{x}{\sigma_x^2}\exp\left(\frac{-x^2}{2\sigma_x^2}\right)dp$	Many scattering centers with random phase
Exponential (power) $W(p) = \frac{1}{\bar{p}}\exp\left(-\frac{p}{\bar{p}}\right)dp$	Many scattering centers with random phase
Rician (voltage) $W(x) = \frac{x}{\sigma_x^2}\exp\left(\frac{-(x^2+d^2)}{2\sigma_x^2}\right)I_0\left(\frac{xd}{\sigma_x^2}\right)dx$	Constant-amplitude dominant scatterer plus many smaller with random phase
Weibull (power) $W(p) = \frac{p^{1/a}}{\alpha a p}\exp\left(-\frac{p^{1/a}}{\alpha}\right)dp$	Scattering from a collection of several size classes of scatterers
Log-normal $W(y) = \frac{1}{\sqrt{2\pi}\sigma_y}\exp\left[-\frac{(y-\bar{y})^2}{2\sigma_y^2}\right]dy$	Scattering from a collection of scatterers of different sizes, including a small number of large scatterers

x is a voltage, *p* is a power, *y* is a decibel value, σ_x and σ_y are the standard deviations, $\bar{x}$, $\bar{p}$, and $\bar{y}$ are the mean values, m^2 is the ratio of power of the steady component to the average of the random component, *a* is the Weibull spread parameter, and α is the Weibull power parameter.

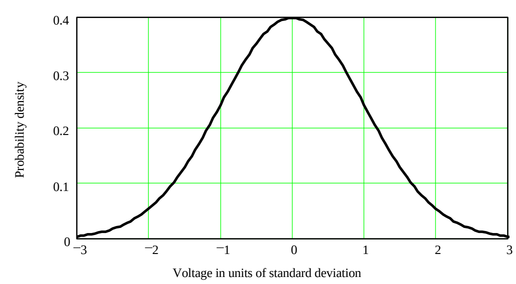


Figure C19 Gaussian distribution.

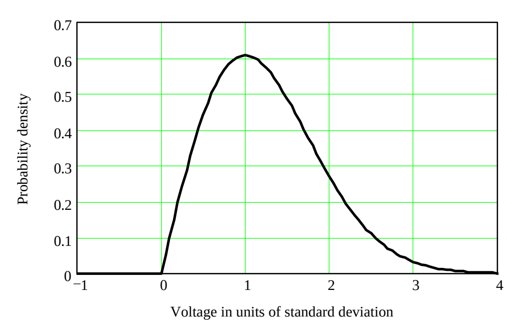


Figure C20 Rayleigh distribution.

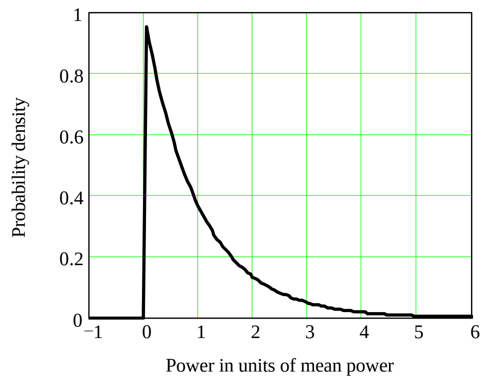


Figure C21 Exponential distribution.

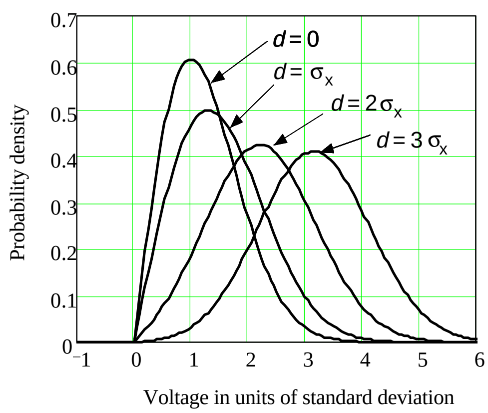


Figure C22 Rician distribution.

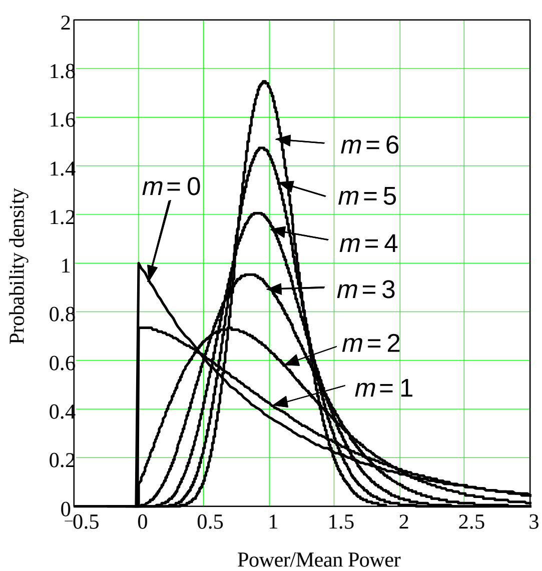


Figure C23 Power of Rician distributed variable.

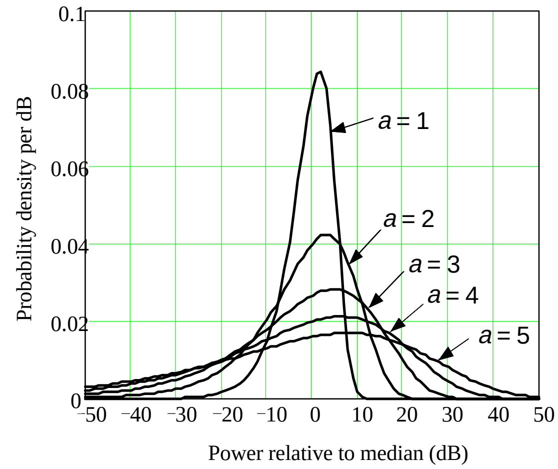


Figure C24 Weibull distribution, plotted as density in dB.

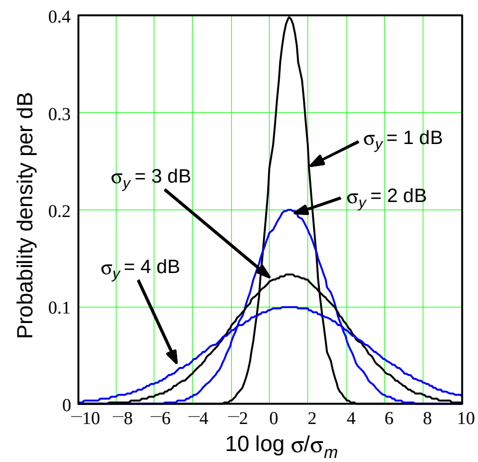


Figure C25 Log-normal distribution. Median value = σ_m.

The **clutter (spatial) distribution** describes the reflectivity of clutter as a function of location, usually in range, azimuth, and elevation radar coordinates. In range, it may be limited by the clutter horizon or (for surface clutter) by the interface between land and sea. In azimuth it may depend on the horizontal extent of a cloud or on the land-sea interface. In elevation (at a given range) it depends on the maximum height of the precipitation or chaff cloud. These parameters are a necessary part of the clutter model. *DKB*

clutter equations (see **RANGE EQUATION with clutter**).

A **clutter filter** is a signal-processing element that is intended to reject the frequency components of clutter while passing target signals. Examples are the MTI canceler, the notch filter in a range-gated MTI or pulsed doppler processor, and the response envelope of the filter bank in which filters containing clutter have been omitted or desensitized. *DKB*

clutter fluctuations (see **clutter spectrum**).

Gaussian clutter (see **clutter (amplitude) distribution**).

ground clutter (see **land clutter**).

Homogeneous clutter is distributed clutter for which the amplitude pdf is approximately Rayleigh with constant mean value over many clutter cells of the observing radar. A threshold set by a cell-averaging CFAR processor can control false alarms from such clutter with minimal losses. *DKB*

Ref.: Nitzberg (1992), pp. 213–216.

The **clutter horizon** refers to the range beyond which clutter is masked by the earth's surface. For surface clutter, the horizon range R_h is set by the height h_r of the radar antenna above the surface:

$$R_h = \sqrt{2kah_r}$$

where ka is the effective earth's radius (8.5×10^6m for standard refractive conditions, using the four-thirds radius approximation, $k = 4/3$). For atmospheric clutter, it is

$$R_h = \sqrt{2ka}(\sqrt{h_r} + \sqrt{h_{mc}})$$

where h_{mc} is the maximum altitude of the clutter source.

When ducting conditions are present ($k >> 4/3$), surface clutter may be visible at ranges more than 400 km from radars at the surface. *DKB*

Ref.: Barton (1988), pp. 127–128.

clutter improvement factor (see **clutter attenuation**, **MTI improvement factor**).

Interclutter visibility (ICV) is "a measure of radar capability to detect targets between points of strong clutter by virtue of the ability of the radar to resolve the areas of strong and weak clutter" (see also **clutter (amplitude) distribution**). *DKB*

Ref.: IEEE (1993), p. 665; Skolnik (1990), p. 15.3

Land clutter is the echo from a land surface, characterized by its surface reflectivity, σ^0 or γ, or by RCS values of discrete clutter sources. Using the constant-γ model (see **surface clutter**), for angles below ψ_s (at which specular reflection becomes significant), values of γ for different land surfaces are approximately independent of radar frequency and are shown in Table C2, along with approximate values for the rms height deviation σ_h.

Table C2
Land Clutter Characteristics

Type of Surface	γ	10 log γ	Rms Height Deviation, σ_h (m)
Mountains	0.32	–5	100
Urban	0.32	–5	10
Wooded hills	0.10	–10	10
Rolling hills	0.063	–12	10
Farmland, desert	0.032	–15	3.0
Flatland	0.01	–20	1.0
Smooth surface	0.0032	–25	0.3

The average reflectivity observed by a radar at low grazing angle, $\psi < \psi_c = \lambda/(4\pi\sigma_h)$, is dominated by the clutter propagation factor F_c:

$$\sigma^0 \overline{F_c^4} = (\gamma \sin\psi)\overline{F_c^4} \approx (\gamma\psi)(\psi/\psi_c)^4$$

which is a strong function of frequency. The strong dependence of F_c^4 for a particular clutter cell on the heights of clutter sources within that cell causes a broad spread of amplitudes from cell to cell, typically described by a Weibull or log-normal distribution. Weibull spread parameters, $a = 4$ to 5, or log-normal standard deviations σ_y = 12 to 15 dB, are normally observed in land clutter near the horizon (see **clutter (amplitude) distribution.**)

Typical values of land clutter reflectivity are shown in Fig. C26, based on parameters of Table C2 with low grazing angle values calculated for S-band.

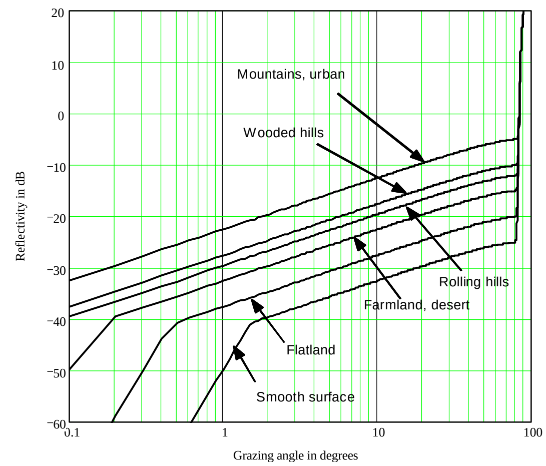


Figure C26 Land clutter reflectivity for different surfaces, with propagation calculated for S-band.

The velocity spectrum of clutter from trees and other vegetation depends on the wind velocity. Early measurements matched the spectrum with a Gaussian function having a standard deviation σ_{vt} = 0.1 to 0.3 m/s. More recent measurements using equipment with linear dynamic range more than 60 dB show exponential spectra:

$$S(v) = \frac{1}{\alpha}\exp\left(-\frac{|v|}{\alpha}\right), \quad |v| \geq 0.2 \text{ m/s}$$

where α is the –4 dB spectral width, ranging from 0.04 m/s in light air (2 to 4 m/s) to 0.06 m/s in breezy air (5 to 6 m/s) and 0.09 m/s under windy conditions (8 to 12 m/s). These spectra are shown in Fig. C27, normalized such that the total power is unity.

Measurements in which spectral density as a function of frequency f was reported to fall off as $1/f^2$ or $1/f^3$, giving very broad spectra at levels from –40 to –60 dB from the peak, can be attributed to receiver nonlinearity (in some cases, log receivers were used). *DKB*

Ref.: Long (1975); Barton (1964), Ch. 3; Skolnik (1970), Ch. 25; Nathanson (1990), Ch. 7; Billingsley, J. B., and Larrabee, J. F., "Measured Spectral Extent of L- and X-Band Radar Reflections from Wind-Blown Trees," *MIT Lincoln Lab. Proj. Rpt. CMT-57*, 6 Feb. 1987.

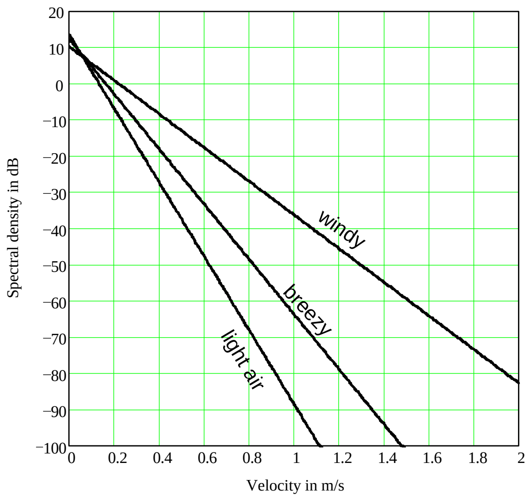


Figure C27 Spectra of windblown tree clutter.

Clutter locking refers to a closed-loop process by which the frequency of a coherent oscillator (COHO) signal is controlled to keep the clutter spectrum centered in the rejection band of the clutter filter. *DKB*

Ref.: Skolnik (1970), pp. 17.32–17.36.

log-normal clutter (see **clutter (amplitude distribution)**).

Main-beam [mainlobe] clutter refers to clutter illuminated and received from within the antenna mainlobe. This will be the dominant source of clutter at the receiver input, but not necessarily at the output of a doppler processor for a moving radar platform or moving field of clutter. The clutter velocity component caused by relative radar-platform motion at velocity v_p is

$$\sigma_{vp} = \frac{\theta_a v_p}{3.34} \sin\alpha$$

where θ_a is azimuth beamwidth and α is the angle between the velocity vector and the beam axis. This component is added in rss fashion to the components from scanning, clutter turbulence, and wind shear to determine the total clutter velocity spread. Given a rejection notch adequate to suppress mainlobe clutter by many tens of decibels, the residue from sidelobe clutter, having velocities from $-v_p$ to $+v_p$, may become significant. *DKB*

Ref.: Barton (1988), p. 245.

A **clutter map** is a stored table of received clutter magnitudes for individual resolution cells or groups of cells, averaged over some number of scans to remove temporal fluctuations. The map may be used to set detection thresholds or filter response parameters, to select processing paths, or to select waveforms appropriate to the clutter environment. Typically, clutter maps are part of the moving target detector configuration. *DKB*

Ref.: Skolnik (1990), p. 15.65; Schleher (1991), p. 46; Nitzberg (1992), pp. 233–236.

A **clutter model** is a mathematical description of clutter reflectivity and other parameters of clutter, as functions of grazing angle, frequency, polarization, and physical parameters of the surface and environment. A complete clutter model will describe the sources affective a given radar, and for each source the following:

(1) Surface reflectivity, σ^0, for given land or sea description, frequency, grazing angle, and polarization;
(2) Volume reflectivity, η_v, for given precipitation type and rate, or cloud or chaff density, for given radar frequency and polarization;
(3) Amplitude probability density function (e.g., parameters of Rayleigh, Weibull, or log-normal distribution);
(4) Spatial extent in range and angles;
(5) Velocity (or doppler) spectrum;
(6) Spatial correlation function; and
(7) Temporal correlation function (which may be derived from the doppler frequency spectrum).

Each of these parameters is defined elsewhere in this section and described for atmospheric clutter, land clutter, or sea clutter. *DKB*

Ref.: Barton (1988), p. 123.

point clutter (see **discrete clutter**).

precipitation clutter (see **rain clutter, snow clutter**).

Rain clutter is significant at all frequencies above VHF and is an important source of interference to radars at X-band and above. The basic parameter controlling the reflectivity of rain clutter is the reflectivity factor Z representing the summation of the sixth power of droplet diameters in millimeters (see **atmospheric clutter**). For rain, this parameter can be expressed as

$$Z = ar^b \text{ (mm}^6\text{/m}^3\text{)}$$

where r is rainfall rate in mm/h and a and b are constants for a given radar wavelength, shown in Table C3. The resulting reflectivity, as a function of rainfall rate, is shown in Fig. C28 for different radar bands. For $\lambda > 0.02$m, the resulting rain reflectivity can be written

Table C3
Rain Reflectivity Constants

Wavelength, λ (m)	Type of Rain	*a*	*b*
>0.02	Orographic	31	1.71
>0.02	Thunderstorm	486	1.37
>0.02	Stratiform	200	1.60
0.02	Stratiform	330	1.54
0.0086	Stratiform	570	1.00
0.006	Stratiform	280	0.80
0.0032	Stratiform	23	0.60

$$\eta_v = \frac{\pi^5}{\lambda^4}|K|^2 Z \times 10^{-18} = \frac{r^{1.6}}{\lambda^4} \times 5.7 \times 10^{-14} \ \mathrm{m}^{-1}$$

where $|K|^2 = 0.93$ is the refractive index factor for water.

In the MMW bands, raindrops are no longer in the Rayleigh scattering region, and the λ^{-4} relationship for reflectivity no longer applies. This is shown in Fig. C28 by the close grouping of lines for $\lambda < 0.02$m (Crane and Burke, 1978).

The fall velocity of raindrops flattens them slightly, leading to depolarization of an incident circularly polarized wave upon reflection, and limiting to about 20 dB, or less for heavy rain, the ability of circularly polarized radars to reduce rain reflectivity. *DKB*

Ref.: Atlas (1964); Crane, R. K., and Burke, H. K., "The Evaluation of Models for Atmospheric Attenuation and Backscatter Characteristic Estimation at 95 GHz," *Environmental Research and Technology* doc. no. P-3606, Feb. 1978.

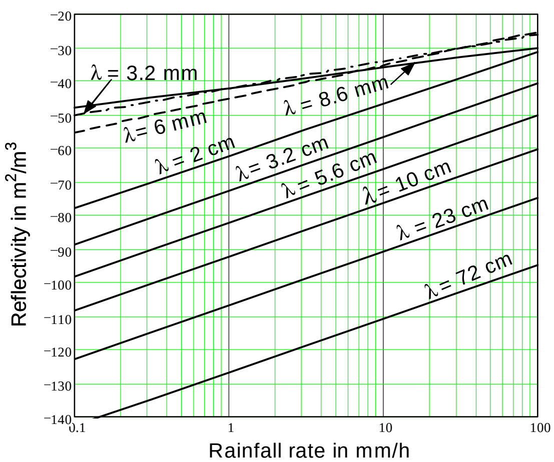


Figure C28 Rain reflectivity vs. rainfall rate for different radar bands.

Rayleigh clutter (see **clutter (amplitude) distribution**).

Clutter reflectivity is "the backscatter coefficient of clutter," describing the clutter RCS per unit area, σ^0, for surface clutter, or per unit volume, η_v, for volume clutter. Typical values are given under discussions of atmospheric clutter, chaff, land clutter, and sea clutter, and are summarized in Table C4. *DKB*

Clutter rejection is the general term referring to use of radar resolution and other properties to reduce the output clutter level relative to target echoes. Other terms used interchangeably are clutter cancellation, clutter reduction, and clutter suppression. Resolution in velocity (doppler) is most commonly used, in which case a more precise term is clutter attenuation or MTI improvement factor. Other techniques that can be used for clutter rejection are (a) high resolution in range or angles, (b) suitable choice of transmitted and received polarizations, (c) choice of pulse repetition and carrier frequencies, and (d) siting or pointing the antenna beam to favor target echoes over clutter. One of the most efficient techniques for clutter rejection is the moving-target detector. *DKB*

Table C4
Typical Clutter Reflectivity Models

Source	Model	Conditions
Land	$\sigma^0 = \gamma\sin\psi$, $\gamma = 0.06$	Wooded hills, ψ is grazing angle
Point clutter	$\sigma = 10^4$ m^2	
Sea	$\sigma^0 = \gamma\sin\psi$, $10\log\gamma = 6K_B + 10\log\lambda - 64$ dB	K_B is Beaufort wind scale number
Rain	$\eta_v = (r^{1.6}/\lambda^4) \times 3 \times 10^{-14}$	$r = 3$ mm/h
Chaff	$\eta_v = 3 \times 10^{-8}\lambda$	Typical density
Birds	$\sigma_{50} = -30$ dBm^2, $\sigma_y = 6$ dB	Log-normal distribution, σ_{50} = median, σ_y = std. deviation

Clutter residue is "the clutter remaining at the output of a moving-target indicator (MTI) system." The term is applied more generally to the output of any clutter filter or rejection device. For possible sources of clutter residue, see **MTI improvement factor**. *DKB*

Ref.: IEEE (1993), p. 199.

Sea clutter is the echo from a water surface, characterized by its surface reflectivity, σ^0 or γ. These parameters are approximately proportional to the square root of the Beaufort wind scale number K_B, and inversely proportional to wavelength:

$$10\log\gamma = 6K_B + 10\log\lambda - 64 \text{ dB}$$

This relationship gives the average overall angles relative to the wind direction. In general, the upwind values are a few decibels higher and crosswind values a few decibels lower than this average.

As with land clutter, the average reflectivity observed by a radar at low grazing angle, $\psi < \psi_c = \lambda/4\pi\sigma_h$, is

$$\sigma^0\overline{F_c^4} = (\gamma\sin\psi)\overline{F_c^4} \approx (\gamma\psi)(\psi/\psi_c)^4$$

which is a strong function of frequency. Here, ψ_c is the critical grazing angle, σ_h is the rms surface-height deviation, and F_c is the clutter pattern-propagation factor (see **surface clutter**). The rms surface height deviation in meters may be estimated as $\sigma_h \approx K_B^3/300$, for a fully developed sea. The strong dependence of F_c^4 for a particular high-resolution clutter cell on the height of the wave surface within that cell broadens the spread of amplitudes from cell to cell, typically described by a Weibull or log-normal distribution. Weibull spread parameters, $a = 1.5$ to 2, or log-normal standard deviations $\sigma_y = 5$ to 7 dB, are often observed in sea clutter near the horizon, when high-resolution radars are used

Figure C29 shows sea clutter reflectivity for different radar bands under medium sea conditions (sea state 4 to 5, wind scale 5 to 6). *DKB*

Ref.: Barton (1988), Ch. 3; Skolnik (1990), Ch. 13; Nathanson (1990), Ch. 7; Currie (1992), pp. 168–180; Morchin (1993), pp. 70–80.

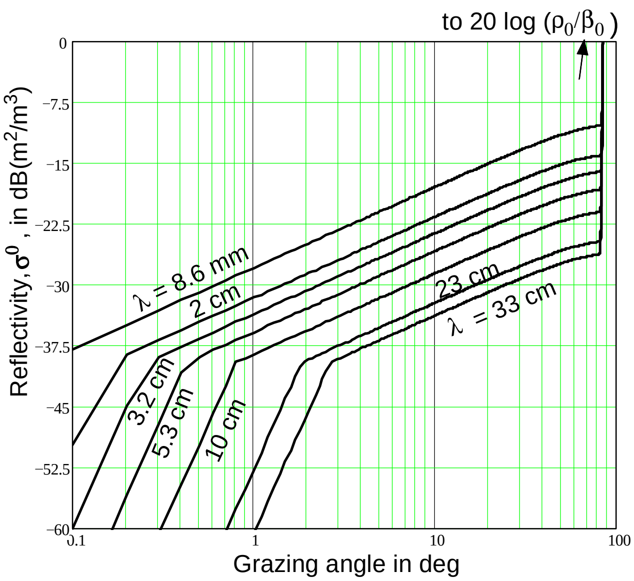


Figure C29 Reflectivity of a medium sea for different radar bands.

Sidelobe clutter refers to clutter illuminated and received from beyond the antenna mainlobe. The total power of sidelobe clutter in a range cell containing mainlobe clutter is normally a small fraction of the mainlobe clutter. However, when there is relative motion between the radar antenna and the clutter sources, the velocity spread of sidelobe sources may exceed the width of the clutter rejection notch, leaving sidelobe clutter as the principal source of clutter residue. This is a critical problem in medium- and high-PRF airborne radars when detecting receding targets. Here, the sidelobe clutter power must be integrated over sidelobes in two-coordinate angle space of the lower hemisphere, producing a clutter spectrum extending from $-2v_p/\lambda$ to $2v_p/\lambda$, where v_p is the velocity of the radar platform and λ is wavelength. A similar problem applies to shipborne radar, with lower values of v_p. *DKB*

Ref.: Skolnik (1990), Ch. 16.

Snow clutter is slightly less intense than rain clutter for the same precipitation rate (when r is defined in terms of the water content of melted snow). For dry snow, the relationships for reflectivity as a function of precipitation rate are

$$\eta_v = \frac{r^{1.6}}{\lambda^4} \times 3 \times 10^{-14}$$

$$Z = 500r^{1.6}$$

For snow of unknown characteristics, values approximately twice those given in these equations appear reasonable, while for aggregate snow the constant changes from 500 to 2,000, and the exponent from 1.6 to 2.0, giving a reflectivity greater than that for rain.

A particularly intense reflection comes from the altitude layer in which snow melts into rain (the "bright band"). Here, values of η_v may increase by several decibels, before dropping to the value for rain. *DKB*

Ref.: Atlas (1964), pp. 371–372.

The **clutter spectrum** in velocity or doppler frequency is the result of relative motion between the radar antenna and the clutter sources. For clutter that fills the radar beam, the mean velocity will be

$$\bar{v} = v_c \cos\alpha - v_p \cos\beta$$

where v_c is the mean clutter velocity relative to inertial space, v_p is the mean radar platform velocity, α is the angle between the radar beam and the clutter velocity vector, and β is the angle between the radar beam and the platform velocity vector. The corresponding doppler shift is $\bar{f_d} = 2\bar{v}/\lambda$, where λ is the wavelength. Airborne clutter moves with the local wind velocity, while sea clutter has an average velocity between 1/8 and 1/4 of the wind velocity.

The spectrum, $W(f)$ in frequency or $W(v)$ in velocity, can usually be approximated as Gaussian in shape:

$$W(f) = W_0 \exp\left(-\frac{f^2}{2\sigma_c^2}\right)$$

$$W(v) = W_0 \exp\left(-\frac{v^2}{2\sigma_v^2}\right)$$

with standard deviation (clutter velocity spread)

$$\sigma_v = \sqrt{\sigma_{va}^2 + \sigma_{vt}^2 + \sigma_{vs}^2 + \sigma_{vp}^2}$$

where σ_{va} is the antenna scanning component, σ_{vt} is the clutter turbulence component, σ_{vs} is the wind shear component, and σ_{vp} is the platform motion component. The corresponding clutter spread in frequency is $\sigma_f = 2\sigma_v/\lambda$. Velocity spectra of five types of clutter are shown in Fig. C30, and reported values of spectral spread are shown in Table C5.

The antenna scanning component is directly proportional to the angular velocity of the beam, or inversely proportional to the observation time t_o the dwell time of the beam:

$$\sigma_{va} = 0.13(\lambda/t_o)$$

The value will apply to mechanically scanning antennas and also to electronically scanned antennas that are moved from pulse to pulse, rather than step-scanned on a burst-to-burst basis.

The clutter turbulence component is dependent on the internal motion of the clutter, induced by the wind. Typical values for precipitation or chaff are 1 to 2 m/s; for sea clutter 1/8 of the wind speed (see **land clutter** for data on the spectrum of trees and foliage). Bird clutter can be considered as having a total velocity spread $\sigma_{vt} = 5$ m/s, corresponding to a 30 m/s band of speeds relative to the air.

The wind shear component of clutter velocity spread, for clutter filling the elevation beam, is given by

$$\sigma_{vs} = 0.3 R_c \theta_e k_{sh} \cos\alpha$$

where R_c is the range to the clutter cell, θ_e is the elevation beamwidth, k_{sh} is the wind shear coefficient, and α is the angle between the beam axis and the wind velocity vector. In addition to this shear-spread term, there may be greater spreading if clutter appears at ambiguous ranges where the beam accepts clutter from altitudes different from that of the target cell.

The platform motion component of velocity spread is

$$\sigma_{vp} = 0.3\theta_a v_p \sin\alpha$$

where θ_a is the azimuth beamwidth, v_p is the platform velocity relative to the clutter, and α is the angle between the velocity vector and the beam axis. *DKB*

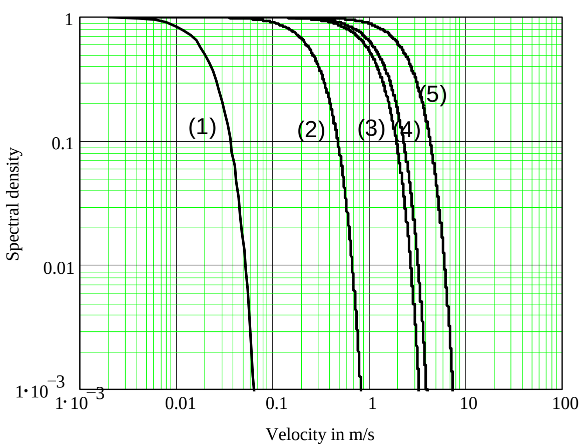


Figure C30 Velocity spectra of five clutter types: in order of increasing spectral spread, (1) sparsely wooded hill, calm day, σ_v = 0.017 m/s, (2) heavily wooded hills, wind 10 m/s, σ_v = 0.22 m/s, (3) sea echo, windy day, σ_v = 0.89 m/s, (4) chaff, σ_v = 1.06 m/s, (5) rain clouds, σ_v = 2.0 m/s (after Skolnik, 1980, Fig. 4.29, p. 132).

Table C5
Reported Values of Clutter Spectral Spread

Clutter source	**Wind speed (m/s)**	**Spread, σ_v (m/s)**
Sparse woods	Calm	0.017
Wooded hills	5	0.04
	10	0.22
	12	0.12
	20	0.32
Sea echo		0.7–1.0
	4–10	0.46–1.1
	Windy	0.44
Chaff		0.37–1.1
	12	1.2
Rain clouds		1.8–4
(after Barton, 1964, Table 3.3, p. 100)		

Specular clutter is viewed at normal incidence to a surface facet, giving substantial increase in reflectivity. For a surface characterized by facets with rms slope $\beta_0/(\sqrt{2})$, the specular clutter reflectivity viewed at an angle $\beta = (\pi/2) - \psi$ from vertical is

$$\sigma_f = \left(\frac{\rho_0}{\beta_0}\right)^2 \exp\left(-\frac{\beta^2}{\beta_0^2}\right)$$

where ρ_0 is the Fresnel reflection coefficient of the surface at vertical incidence. Typical land surfaces have $\beta_0 \approx 0.05$, $\rho_0 \approx 0.5$, giving a maximum $\sigma_f = 100 = +20$ dB (reduced if vegetative cover is present). Water surfaces have $\beta_0 \approx 0.04$, $\rho_0 \approx 0.8$, giving $\sigma_f = 400 = +26$ dB. *DKB*

Ref.: Barton (1988), p. 125.

Subclutter visibility is "the ratio by which the target echo power may be weaker than the coincident clutter echo power and still be detected with specified detection and false alarm probabilities." It describes of the performance of an MTI or pulsed-doppler processor system. It can be expressed as

$$\mathrm{SCV} = \frac{I_m}{D_{xc}}$$

where I_m is the improvement factor of the processor and D_{xc} is the clutter detectability factor. *DKB*

Ref.: IEEE (1993), p. 1,303.

clutter suppression (see **clutter attenuation**).

Surface clutter. When clutter sources are distributed homogeneously within a resolution cell of area A_c, the RCS of the clutter is given by

$$\sigma_c = A_c \sigma^0$$

where

$$A_c = \left(\frac{R_c \theta_a}{L_p}\right)\left(\frac{\tau_n c}{2}\right)$$

and σ^0 is the reflectivity, R_c is the range to the cell, θ_a is the half-power azimuth beamwidth, L_p is the beamshape loss, τ_n is the processed pulse width, and c is the velocity of light. The beamshape loss is defined such that θ_a/L_p is the integrated angle within the two-way pattern of a Gaussian beam.

The surface clutter reflectivity can usually be modeled accurately as a function of the grazing angle, ψ:

$$\sigma^0 = \gamma \sin\psi, \quad \psi < \psi_s$$

where γ is dependent on the surface conditions and ψ_s is an angle near 90° at which specular reflection from the surface facets becomes significant. A typical plot of measured reflectivity (as affected by the pattern-propagation factor F_c) is shown in Fig. C31.

The measured reflectivity drops rapidly in the region below a critical angle, $\psi_c = \lambda/4\pi\sigma_h$, where σ_h is the rms height deviation of the surface. In this region, $F_c \approx \psi/\psi_c$ and hence the measured reflectivity varies inversely with the fifth power of grazing angle (for data on surface reflectivity and

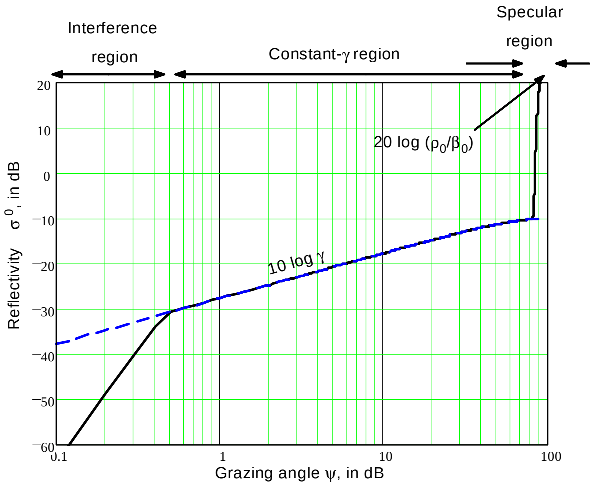


Figure C31 Typical surface clutter reflectivity vs. grazing angle.

other properties of surface clutter, see **land clutter** and **sea clutter**). *DKB*

Ref.: Barton (1988), pp. 124–132.

Transmitted clutter is the reflection of sidebands of the transmitted carrier, which may lie outside the rejection notch of the clutter filter. When these components are reflected from strong clutter, they may appear in the target filter passband, obscuring the target. This effect is most pronounced in CW and high-PRF pulsed-doppler radars operating near the surface or within a cloud of precipitation or chaff, where the nearby clutter has great range advantage over the target. *DKB*

Ref.: Skolnik (1990), pp. 14.3-14.7.

tree clutter (see **land clutter**).

The **clutter visibility factor** is defined as "the predetection signal-to-clutter ratio that provides stated probabilities of detection and false alarm on a display." The equivalent term for radars using automatic detection is the *clutter detectability factor*. *DKB*

Ref.: IEEE (1993), p. 199.

Volume clutter. When clutter sources are distributed homogeneously within a resolution cell of volume V_c, the clutter RCS is given by

$$\sigma_c = V_c \eta_v$$

where

$$V_c = \left(\frac{R_c\theta_a}{L_p}\right)\left(\frac{R_c\theta_e}{L_p}\right)\left(\frac{\tau_n c}{2}\right)$$

and η_v is the volume reflectivity, R_c is the range to the cell, θ_a and θ_e are the half-power beamwidths in azimuth and elevation, L_p is the beamshape loss, τ_n is the processed pulse width, and c is the velocity of light. The beamshape loss, L_p = 1.33, is defined such that $\theta_a\theta_e/L_p^2 = \pi\theta_a\theta_e/(8 \ln 2)$ is the integrated solid angle within the two-way pattern of a Gaussian beam (Probert-Jones, 1962; Doviak and Zrnic, 1984, p. 52). Note that the factor $\pi/4$ relating the area of an ellipse to that of the rectangle is included in this factor.

In cases where the clutter does not fill the beam, the equation for volume uses the actual clutter extent in the appropriate coordinate, weighted by the two-way antenna gain. For example, if precipitation is visible over an altitude interval h_0 to h_m in which antenna gain is constant,

$$V_c = \left(\frac{R_c\theta_a}{L_p}\right)(h_m - h_0)\left(\frac{\tau_n c}{2}\right)$$

(For data on the volume reflectivity, η_v, and other properties of volume clutter, see **atmospheric clutter**, **rain clutter**, **snow clutter**, and **CHAFF**). *DKB*

Ref.: Probert-Jones, J. R., "The Radar Equation in Meteorology," Proc. 9th Weather Radar Conf., Oct. 23–26, 1961; Doviak, 1984; Barton (1988), pp. 133–137; Morchin (1993), p. 80.

weather clutter (see **atmospheric clutter**, **rain clutter**, **snow clutter**).

COAST. The coast mode is a radar memory mode that causes the range- or angle-tracking system to continue to move at a constant rate equal to that of a target that has been under track. It is used to prevent transfer of lock to a stronger target crossing the position of the target being tracked, or to continue a track during a signal fade or momentary burst of interference. *SAL*

Ref.: IEEE (1990), p. 8.

COAXITRON. The coaxitron is a modification of a grid-controlled tube for high-power radar applications in the UHF band. The electrical interaction system and the RF input and output circuits in the coaxitron are integrated within the vacuum envelope, providing better performance and increased bandwidth in comparison with conventional grid-controlled tubes. An example is the RCA A15193 device, 0.76m long, 0.42m diameter, 63.5-kg weight, operating at 406–450 MHz with 1.5 MW peak power. *SAL*

Ref.: Skolnik (1980), p. 213; Gilmour (1986), p. 194.

CODE, CODING. A code is a plan for representing each of a finite number of values or symbols as a particular arrangement or sequence of discrete conditions or events. Coding is the corresponding process of transforming messages or signals in accordance with a definite code. In radar applications there are two primary applications of codes: for generating modulated waveforms and for coding the information in radar data transmission.

The main types of codes used in generating phase-coded waveforms are binary (phase) codes (or biphase codes) using only two sets of phase discretes (0 and 180°) to modulate the carrier, and polyphase codes using smaller increments of phase, in general N sets. An example of the first type is the Barker code, and examples of the second type are the Frank code, P-codes, quadriphase codes. Both biphase and polyphase codes can be formed by the use of complementary codes to cancel the time sidelobes, as in the Golay (biphase) and Welti (polyphase) codes. Codes can be used also for amplitude modulation of the waveform (e.g., Huffman code).

The basic code that is used in coding information for data transmission is the Gray code. Some other codes for specific tasks can be formed from the Gray codes (e.g., Gillham code for transmitting information about the height of aircraft). *SAL*

Ref.: IEEE (1993), p. 202; Cook (1967), Ch. 8.

The **Barker code** is a binary (phase) code with the property that the peak of the autocorrelation function is equal to N, where N is the code length, and all the time sidelobes have unit amplitude. The time duration of a Barker code is equal to $N\delta$, where δ is the chip width. The peak-to-sidelobe ratio is equal to $20\log N$. The only Barker codes possible are listed in Table C6, and the longest of these has length $N = 13$.

The Barker codes have a sidelobe structure containing the minimum energy theoretically possible, uniformly distributed among the sidelobes (Fig. C32). That is why Barker codes are sometimes called *perfect codes*. They can serve as building blocks for much longer codes (see **combined Barker codes**). *SAL*

Ref.: Barton (1988), p. 221; Barton (1991), p. 7.30; Skolnik (1990), p. 10.17; Morris (1988), p. 134.

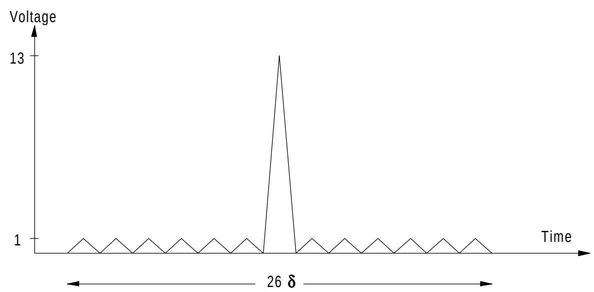


Figure C32 Matched-filter output of Barker code of length 13.

Table C6
The Barker Codes

Code length	Code elements	Sidelobe level (dB)	
		Peak	Integrated
1	+	0.0	–
2	+ –, or – –	–6.0	–3.0
3	+ + –, or + – +	– 9.5	– 6.5
4	+ + – +, or + + + –	– 12.0	– 6.0
5	+ + + – +	– 14.0	– 8.0
7	+ + + – – + –	– 16.9	– 9.1
11	+ + + – – – + – – + –	– 20.8	–11.5
13	+ + + + + – – + + – + – +	– 22.3	–11.5

A **binary (phase) [biphase] code** is the sequence of symbols, each plus or minus, used to form binary-coded waveforms (sometimes called *binary code* or *biphase code*). Allomorphic forms of binary phase codes are binary phase code representations in any one of four forms, all of which have the same correlation characteristics. These forms are the code itself, the inverted code (i.e., written in reverse order), the complemented code (1s changed to 0s and vice versa), and the inverted complemented code. *SAL*

Ref.: Hovanessian (1984), p. 241; Skolnik (1990), p. 10.17.

The **combined [compound] Barker code** is a code using known Barker codes to generate codes longer than 13 bits. For example, either a 5-×-4 or a 4-×-5 combined Barker code can be used to devise a system with 20:1 pulse-compression ratio. *SAL*

Ref.: Morris (1988), pp. 135–138; Skolnik (1980), p. 432.

Complementary codes are the codes in which a pair of equal-length codes have the property that the time sidelobes of one code are the negative of the other, so the autocorrelation function of a complementary pair is equal to zero except for the central peak. Complementary codes are used to form complementary-code phase-coded waveforms with no residues due to sidelobe cancellation. (See **Golay code**, **Welti code.**) *SAL*

Ref.: Skolnik (1980), p. 432; Nathanson (1990), p. 491.

The **Frank code** is a polyphase code composed of N sets of phase sequences that can be described as an $N \times N$ matrix in which the phase in the ith row and jth column is given by

$$\phi_{ij} = (2\pi/N)\,(i-1)\,(j-1).$$

SAL

Ref.: Barton (1991), p. 7.17; Skolnik (1990), p. 10.25.

frequency coding (see **MODULATION, frequency**)

The **Gillham code** is a special digital code based on the Gray code and used in the air traffic control systems with secondary surveillance radars. It is approved by the International Civil Aviation Organization for transmission of data on the height of aircraft. *SAL*

Ref.: Pereverzentsev (1981), p. 71; Stevens (1988), p. 113.

The **Golay code** is a binary phase code using a pair of complementary codes transmitted and received over two channels. The Golay code pair of length 8, their autocorrelation functions, and the zero sidelobe sum of their autocorrelations are shown in Fig. C33. The two channels must preserve phase and amplitude matching so that the sidelobes are canceled. For radar, this implies transmission on alternating pulse repetition intervals (PRIs), with a delay of one PRI introduced in one of the channels before the final summing, and with target scatterers preserving the same phase over that interval t_r. The resulting doppler sensitivity corresponds to $f_d << 1/t_r = f_r$, a difficult requirement that explains the lack of practical application of complementary codes. *SAL*

Ref.: Morris (1988), p. 142; Barton (1991), p. 7.31.

The **Gray code** is a cyclical binary code used as the basis for systems of data transmission. The cyclical codes have an advantage over conventional binary codes when transmitted information continuously varies in time, since cyclical codes do not lead to significant errors when the adjacent digital values intersect. *SAL*

Ref.: Sloka (1970), p. 141; Jordan (1985), p. 43.5.

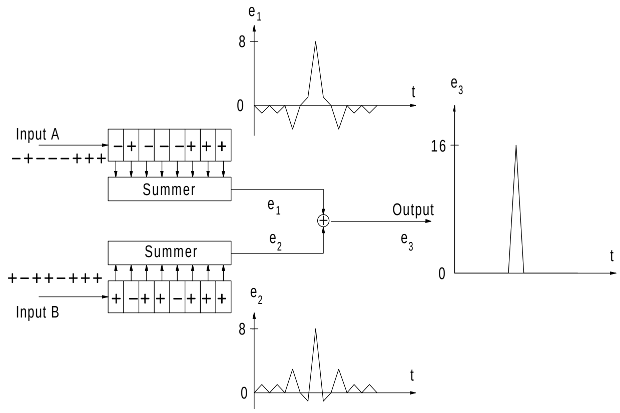


Figure C33 Matched-filter response to Golay code.

A **Huffman code** is the sequence of complex quantities corresponding to each value of modulation used to form a Huffman-coded waveform, theoretically having only a single unavoidable sidelobe at each end of the compressed signal. Huffman codes may have binary phase values (0 and 180°) with varying amplitudes, near-constant amplitudes, with varying phase angles, or combinations of varying amplitude and phase. A limitation of this type of code is its sensitivity to target doppler shift. If the shift is such as to cause a significant fraction of one radian during the transmitted pulse width τ, sidelobes are not canceled to zero value in the region intended to be clear of sidelobe response. Hence the allowable doppler shift is $f_d << 1/\tau$. The compressed 14-element, zero-doppler Huffman code is illustrated in Fig. C34. *SAL*

Ref.: Brookner (1977); Lewis (1986), pp. 96–105.

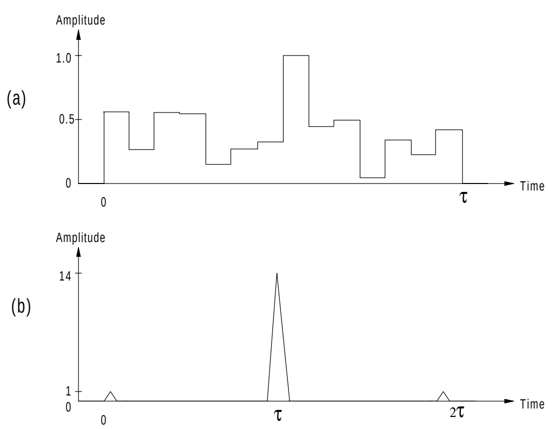


Figure C34 Matched-filter response to 14-element Huffman code: (a) transmitted amplitude; (b) matched-filter output (after Cook, in Brookner, 1977, p. 146).

P-codes are polyphase codes originated at the U.S. Naval Research Laboratory and are related to and are extensions of Frank codes. There are four P-codes: P1, P2, P3, and P4 codes. They have properties similar to linear FM, step FM, and the Frank code and offset some of the problems inherent in the Frank codes and step FM. *SAL*

Ref.: Barton (1991), pp. 7.20–7.23; Lewis (1986), pp. 10–12.

A **perfect code** is a code having a sidelobe structure containing the minimum energy theoretically possible, with this energy uniformly distributed among the sidelobes. The Barker code is an example. *SAL*

Ref.: Morris (1988), p. 135.

phase coding (see **MODULATION, phase**).

polarization coding (see **MODULATION, polarization**).

A **polyphase code** is a sequence of symbols with more than two discrete values, used to form polyphase coded waveforms. Instead of only two phase shift values, 0 and 180 δεγ, as in the binary phase code, smaller increments can be applied. A quadriphase code is a polyphase code with four phase values: 0, 90, 180, and 270°. This code can be generated from a binary phase code using conversion formula

$$(V_k = j^{s(k-1)})W_k \ , k = 1, 2, \dots N$$

where V_k is the phase of the quadriphase code, W_k is the phase of biphase code, s is either +1 or −1, and N is the code length. *SAL*

Ref.: Skolnik (1980), p. 432, (1990), p. 10.25; Nathanson (1969), p. 491; Morchin (1993), p. 379.

pseudorandom code (see **SEQUENCE**).

A **quadriphase code** is a sequence of symbols denoting 0, 90, 180, and 270° phase values used to form a quadriphase-coded waveform. This code can be generated from a binary phase code using conversion formula

$$(V_k = j^{s(k-1)})W_k \ , k = 1, 2, \dots N$$

where V_k is the phase of the quadriphase code, W_k is the phase of biphase code, s is either +1 or −1, and N is the code length. *SAL*

Ref.: Morchin (1993), p. 379; Skolnik (1980), p. 432; (1990), p. 10.25; Nathanson (1980), p. 491.

A **ternary code** is the sequence of symbols where the coding includes +, – (as in the Barker code), and 0, corresponding to the absence of a segment. *SAL*

Ref.: Nathanson (1990), p. 462.

The **Welti code** is a polyphase code forming complementary codes. *SAL*

Ref.: Morris (1988), p. 141; Barton (1991), p. 7.31.

COHERENCE is the concept generally applied to harmonic oscillations:

$$u(t) = V_0 \sin(wt + y)$$

Two or more harmonic oscillations are termed coherent at the interval T_c if the phase shift between them is constant for the whole interval T_c. In radar, coherence is considered in a broader sense, and typically the signals are considered to be coherent if their phase structure is linked and the character of this linkage is known. *SAL*

Ref.: Ridenour (1947), p. 631; Leonov (1988), p. 21; Vinitskiy (1961), p. 23.

collapsing loss (see **LOSS, collapsing**).

COMPRESSION (see **PULSE compression**).

Compression ratio, in a radar that uses pulse-compression techniques, is the ratio of the uncompressed-pulse width to the compressed-pulse width (see also **PULSE compression**). *PCH*

CONDUCTANCE is "that physical property of the element, device, branch, network, or system that is the factor by which the mean square voltage must be multiplied to give the corresponding power lost by dissipation as heat or as other permanent radiation or loss of electromagnetic energy from the circuit." *SAL*

Ref.: IEEE (1993), p. 236.

Eigenconductance is the conductance of each input of a network determined when there is a short circuit of all the other inputs in the form of the ratio of the current to the voltage exciting the given input.

Eigenconductances are the diagonal elements of the matrix of conductances. This matrix is most often used in the designs of multielement antennas for the calculation of the mutual influence of individual radiators on one another. *SAL*

Ref.: Sazonov (1988), p. 79.

Negative conductance is a falling portion of the dynamic or static voltage-current curve, in which the voltage and current increments have opposite signs. Negative conductance can be used for generation of electromagnetic energy in such devices as tunnel diodes, Gunn diodes, IMPATT diodes, and also traditional active devices such as electron tubes and transistors. *IAM*

Ref.: IEEE (1993), p. 839; Popov (1980), p. 273.

CONFUSION AREA. The effective confusion area is the chaff cross section that equals the radar cross section of a particular aircraft at a particular frequency. *SAL*

Ref.: Johnston (1979), p. 58.

CONNECTOR, microwave. A microwave connector is "a coupling device employed to connect conductors of one circuit or transmission elements with those of another circuit as transmission elements," usually based on pin and socket contacts. The basic requirements for the connectors are retention of matching and power-handling capability of the circuit with minimum induced attenuation of power and the absence of parasitic radiation. In high-frequency joints for flexible coaxial cables (see coaxial transmission lines), the contacts are produced using spring-clips and plugs held in the joint by external threaded connections or other locking devices. Matching in the connectors depends greatly on the connection of the cable, and as a standing-wave ratio of 1.05 to 1.15 when it is done correctly. Connectors for rigid coaxial lines are made without supporting dielectric washers using pin and jack-type contacts of internal and external conductors. In a waveguide transmission line it is typically a mechanical device (excluding an adapter) that electrically joins separable parts of a waveguide.

The connector usually does not allow turning of the connected sections of the circuit. For these purposes, and also for connection of transmission lines of other types, connections of circuits in the form of joints are used. *IAM*

Ref.: IEEE (1993), p. 244; Sazonov (1988), p. 53.

CONOPULSE is a hybrid angle-tracking system combining monopulse and conical scan techniques. Two squinted beams are rotated or nutated in space in a conical scan manner. The received signals are processed either with monopulse processing followed by conical scan or vice versa. The advantage claimed for the conopulse technique is that, like monopulse, amplitude fluctuations do not affect the angular accuracy, while only two receivers are required instead of three used in a conventional monopulse tracker. With modern solid-state technology, it can be easier to realize the third receiver than to arrange proper scanning of a pair of squinted beams. Sometimes conopulse is called *scan with compensation*. *SAL*

Ref.: IEEE (1990), p. 9; Barton (1988), p. 420; Skolnik (1980), p. 165.

CONSTANT FALSE-ALARM RATE (CFAR) is "a property of threshold or gain control devices that maintain an approximately constant rate of false target detections when the noise, and/or clutter levels, and/or ECM (electronic countermeasures) into the detector are variable." CFAR techniques are used in reception and signal processing to avoid increased false-alarm rates in the presence of jamming, clutter residue, or other interference sources. The two fundamental approaches are to adapt the detection threshold of a given test cell to the environment as determined by (1) statistics of reference cells surrounding the test cell (e.g., cell-averaging CFAR or nonparametric CFAR) or (2) time statistics of past observations in the test cell itself (e.g., clutter mapping). In the first method, statistical parameters of the amplitudes in the reference cells (e.g., mean, standard deviation) may be used to set a threshold, or a nonparametric method (e.g., rank order) may be used. The reference cells may be separated from the test cell in time delay (range), doppler shift, angle, or some combination of these coordinates. Nonlinear processing of reference cells (e.g., clipping or editing the largest samples) may precede the estimation of statistical parameters.

The use of CFAR circuits protects against excessive numbers of false alarms, but at the cost of suppression of targets lying within regions of strong interference. *DKB*

Ref.: IEEE (1993), p. 247; Skolnik (1980), pp. 392–395, (1990), pp. 3.46–3.54; Barton (1988), pp. 88–94; Nitzberg (1992), Ch. 9.

amplitude-discrimination CFAR (see **cell-averaging CFAR**).

Cell-averaging CFAR (CACFAR) is a technique in which the threshold is controlled by the average of received amplitudes in cells surrounding the test (target detection) cell, as a means of controlling the false-alarm rate. A typical method uses the integrated outputs of several range cells (Fig. C35), which are summed and multiplied by a CFAR constant K to establish the detection threshold. In a doppler radar, the reference cells in range may be replaced by filters adjacent in frequency to the target filter. In pulsed doppler radar, range-gated filters surrounding the target cell in both range and doppler may be used (Fig. C36).

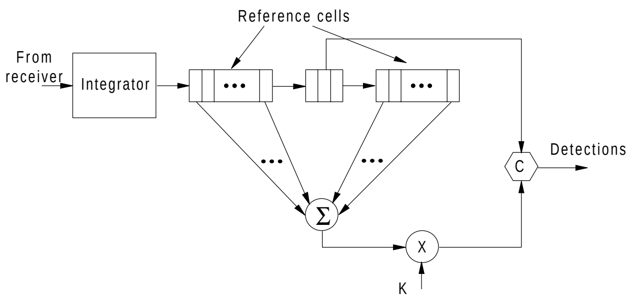


Figure C35 Cell-averaging CFAR (after Skolnik, 1990, Fig. 8.12, p. 8.13).

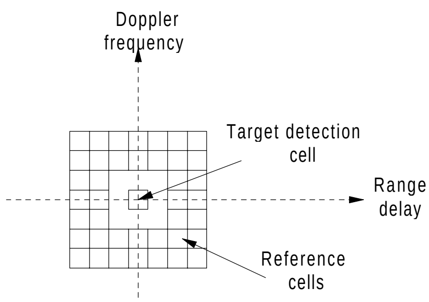


Figure C36 Use of reference cells from combined range and doppler regions surrounding the target cell.

The CFAR constant K is set to produce the desired false-alarm probability in the output of the threshold detector C. The performance of cell-averaging CFAR depends on the number of reference cells and their extent in the selected radar coordinate. A CFAR loss is introduced by the fact that the estimate of interference level in the reference cells is subject to random error, which varies inversely with the number of cells, forcing the user to use K values higher than would be used if the estimate were exact. On the other hand, if the number of cells is increased to reduce this loss, the circuit fails to respond to rapid changes in interference level, permitting a burst of false alarms to occur when this level increases rapidly. Also, the greater the number of reference cells the greater the probability that an adjacent target will fall into the reference cells, and distort the threshold setting (see **multiple-target CFAR**). Use of combined range and doppler regions around the target cell is one means of avoiding excessive delay in CFAR response in either coordinate. *DKB*

Ref.: Skolnik (1990); Nitzberg (1992), Ch. 9.

A **clutter-map CFAR** sets the detection threshold in a given test cell based on the average interference observed in that cell (and often several adjacent cells) over several scans of the antenna. The time average may be determined recursively, using an α-β filter. For fluctuating interference, the time averaging increases the precision of the interference estimate without extending the spatial extent of the reference region. By reducing the size of the reference region (in the limit to the test cell itself), the clutter-map CFAR preserves the inter-clutter visibility inherent in the use of a high-resolution search radar in non-Rayleigh clutter. Output of the clutter map may also be used to control the input dynamic range of the receiver, to select processing paths appropriate to each type of interference, and to blank cells containing interference so strong that no other processing can prevent false alarms. Clutter-map CFAR is often called *temporal* [time-averaging] *CFAR* or simply a *clutter map*. *DKB*

Ref.: Skolnik (1990); Nitzberg (1992), pp. 233–236.

Dicke-fix CFAR is a cell-averaging CFAR technique in the frequency domain, in which a broadband IF amplifier is followed by a limiter and a narrowband amplifier (Fig. C37).

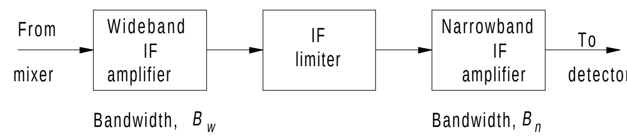


Figure C37 Dicke-fix CFAR.

The effective number of reference cells is equal to the bandwidth ratio, $M = B_w/B_n$. When used with pulse compression, the number of cells becomes $B\tau M$, making CFAR possible with $M = 1$ (see **phase-discrimination CFAR**). *DKB*

Ref.: Skolnik (1980), p. 394.

distribution-free CFAR (see **nonparametric CFAR**).

Greater-of cell-averaging CFAR is a technique in which the reference cells are divided in leading and lagging cells, with separate averages being taken (Fig. C38). The greater of the

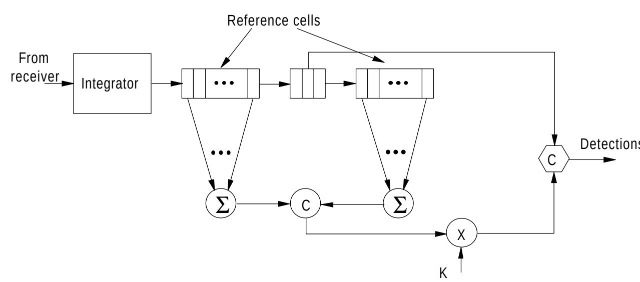


Figure C38 Greater-of cell-averaging CFAR.

two averages is used to control the threshold, permitting more rapid adaptivity to changes in the interference environment at some expense in CFAR loss. The increase in CFAR loss for this configuration, with respect to conventional cell-averaging CFAR, is typically about 0.2 dB. *DKB*

Ref.: Nitzberg (1992), pp. 226–229.

Hierarchal CFAR techniques perform separate calculations of the threshold using two methods (e.g., CACFAR with parameters chosen for an environment of noise, and an adaptive alternative suitable for clutter or jamming). The normal CACFAR threshold T is used, with minimum CFAR loss, unless the second method yields a threshold, $T_a > \alpha T$, where $\alpha > 1$. *DKB*

Ref.: Nitzberg (1992), pp. 245–249.

Log(arithmic) CFAR uses the output of a logarithmic receiver as input to a cell-averaging CFAR circuit (similar to Fig. C39), to avoid excessive target suppression from multiple targets or non-Rayleigh clutter. *DKB*

Log-FTC CFAR is a technique in which a logarithmic receiver is used, followed by a fast-time-constant (high-pass) video circuit. The log receiver compresses the fluctuations of strong clutter to the same peak-to-peak amplitude as thermal noise (Fig. C39), and the FTC circuit removes the DC level to create a uniform interference output, above which a fixed threshold can establish the desired false-alarm rate. Log-FTC CFAR proceeding a visual display can protect against saturation and total loss of target detection in regions of strong clutter or other interference. *DKB*

Ref.: Skolnik (1980), pp. 394, 506.

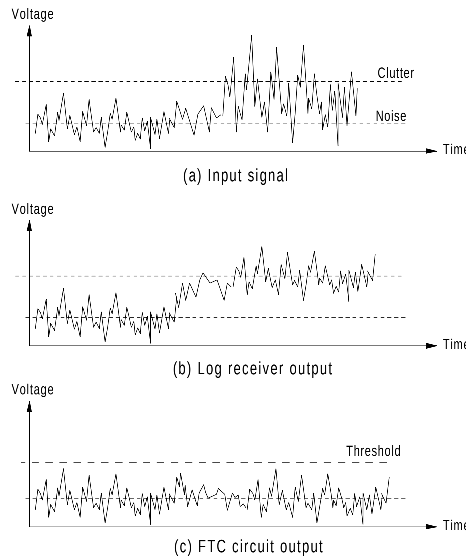


Figure C39 Log-FTC CFAR.

Multiple-carrier-frequency CFAR techniques are applied when frequency diversity is used with multiple channels in parallel. Two methods are (1) to sum both the test cell outputs and reference cell outputs several channels before comparison of the summed target with the summed threshold, or (2) to test each channel separately for detection, summing those outputs that have passed their individual thresholds. *DKB*

Ref.: Nitzberg (1992), pp. 240–245.

Multiple-target CFAR techniques are used to avoid the undesirable suppression of each target by its neighbors within a multiple-target region. One approach is to censor the strongest one, two, ... cells from the reference region, reducing the number of reference cells and increasing the CFAR loss but avoiding effects from multiple targets within the reference region. Another approach, involving somewhat greater CFAR loss, is to use the median level of reference cells, rather than the average, as the basis for calculating the threshold. *DKB*

Ref.: Nitzberg (1992), pp. 229–233.

Nonparametric CFAR techniques use statistical properties other than the average and high-order moments of the reference cells to establish the detection threshold. Examples are the median detector and the more general rank-order detector, or ordered-statistic CFAR. Ranking means arranging the m samples from the smallest to the largest and assigning a numerical value equal to the position in the rank order, from 0 to m–1. The target cell amplitude is compared with a threshold based on the amplitude of the cell of rank X, where $0 < X \leq m - 1$, and detection is declared if this threshold is exceeded. The largest amplitudes in the reference cells may be ignored in this process, making this CFAR immune to target suppression from multiple targets.

When interference samples are correlated over the integration time of the signal, a modified rank detector known as the modified generalized sign test (MGST) is used, in which the ranker outputs are integrated before being applied to a threshold.

In the median detector, the median value of the target cell, over the n samples gathered during the integration time, is applied to the threshold. An alternative procedure uses a binary integrator with the second threshold set to $(n - 1)/2$. *DKB*

Ref.: Skolnik (1970), pp. 8.19 to 8.21; (1980), p. 486.

ordered-statistic CFAR. (see **nonparametric CFAR**).

Phase-discrimination CFAR refers to the technique in which a phase-coded waveform is subject to hard limiting before matched filtering for compression. The technique is sometimes called the *coded-pulse anti-clutter system* (*CPACS*). The number of reference cell samples is equal to the number of subpulses (the pulse-compression ratio). *DKB*

Ref.: Skolnik (1990), p. 3.49.

spatial CFAR (see **cell-averaging CFAR**).

temporal [time-averaging] CFAR (see **clutter-map CFAR**).

Two-parameter CFAR is a technique in which the CFAR constant K (see **cell-averaging CFAR**) is varied to adapt the system to non-Rayleigh clutter having different spreading of its probability density function (see **clutter (amplitude) distribution**). Both the mean and standard deviation of the reference cell amplitudes are measured, and the threshold setting is proportional to the product of these two values. In most cases, the resulting threshold is so high, when the spread is significantly greater than that of the Rayleigh distribution, that excessive target suppression (CFAR loss) results. *DKB*

Ref.: Barton (1988), pp. 92–94.

CONTRAST, radar. Radar contrast is the degree to which objects observed by a radar are discriminated from the background, making possible their detection, identification, and interception with seekers, such as missiles and torpedoes. Radar contrast is measured by the signal-to-noise ratio. *IAM*

Ref.: Popov (1980), p. 188; Mel'nik (1980), p. 19; Finkel'shteyn (1983), p. 373.

CONTROL

automatic frequency control (see **FREQUENCY**).

automatic gain control (AGC) (see **GAIN**).

automatic noise-level control (see **CFAR**).

control, radar. In modern multifunction radars, operations are, to a large extent, automated. The role of the human operator is one of monitoring, selection of scenario-dependent software, and override in case of radar malfunction or at the direction of higher authority. Virtually all radar functions, such as transmitted waveform selection and scheduling, and antenna beam formation and steering, are controlled by a set or sets of computer programs or preestablished templates. Surveillance, multiple-target tracking, target identification, and, if required, weapon guidance and control, are all performed automatically. Radar action scheduling conflicts are typically resolved by special algorithms that establish priorities and provide override responses. Radar control can extend to the radar receiver and signal processor, which may be adaptively reconfigured in response to the sensed environment of weather, surface clutter, and electronic countermeasures.

Whether the radar system control is implemented by a central computer or with multiple distributed "processors" is a matter of radar architectural preference and economics, but with either approach, redundancy is generally built in to enhance the control system's reliability.

The preceding discussion is most applicable to a multifunction phased-array radar (e.g., the U.S. Army's Patriot), where computer control is a necessity. For example, the actions required to control the radar beam alone are clearly beyond the response capability of any human operator. The importance of the phased-array antenna to multifunction radar has led some radar developers to limit use of the term "radar controller" to the computer and software associated with its control. In its more general interpretation, however, a "radar controller" includes the entire ensemble of computers, data and signal processors, and all the software required to operate the radar in its various modes.

A radar controller can also direct the operation of nonphased-array radars, including reflector antenna radars, those with planar arrays, as well as various hybrid configurations. The Westinghouse-developed ARSR-4, a long-range surveillance radar shared by the FAA and the U.S. Air Force, is an example of an array-fed reflector antenna radar that is designed to operate under automatic control and is completely unmanned. *PCH*

sensitivity time control (see **SENSITIVITY**).

CONVERSION, CONVERTER

An **angle-to-code (digital) converter** is a device that changes the angular coordinates of a target into digital code. In a surveillance radar, when the antenna rotates at constant speed, the converter converts the time interval proportional to the angle of rotation of the antenna into digital code. Because of insufficient stability of the speed of rotation of the antenna, indirect converters of angle into digital code are used. These are subdivided in to cumulative and positional types.

Positional converters contain a transparent disk with an axis, which is rigidly connected to the antenna drive, with cyclic Gray code marked on it in the form of alternating transparent and opaque sectors. The sectors are arranged in rings, the number of which is equal to the number of bit positions of the count. The angular dimension of each sector is equal to the reference angle. On one side of the disk is a slotted pulse source of light, and on the other, photo-detectors, one for each bit position. The code is read when a target pulse is applied to the light source. The signal of the photo-detectors is converted to binary code.

Cumulative converters were used with older radars. In them the count is made with angle marks uniformly arranged about the circumference of a disk or drum (for example, illuminated openings in an optical converter). They have a drawback associated with the necessity to input the initial values, and the accumulation of errors when malfunctions occur. *IAM*

Ref.: Vasin (1977), p. 216; Kazarinov (1990), p. 414.

An **analog-to-digital converter (ADC)** is "a device that converts a signal that is a function of a continuous variable into a representative number sequence." Typically, an ADC consists of a sampling unit, a quantization unit, and a coding unit (Fig. C40).

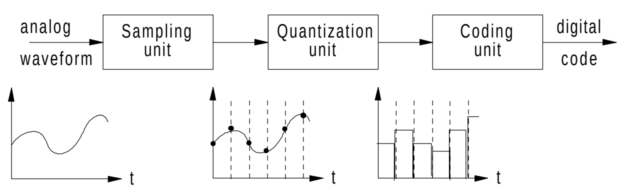


Figure C40 Analog-to-digital converter.

The basic configurations of ADCs are the simultaneous converter (Fig. C41), and the sequential converter (Fig. C42). The simultaneous converter performs all operations simultaneously. In this case, an array of 2:1 comparators independently convert the analog voltage into quantized magnitudes, and 2:1 input logic converts the quantized amplitude into an N-bit parallel digital code. Such a converter has the highest conversion speed. The sequential converter can use N identical cascaded stages, and a delay line is often introduced to compensate for the time required for each quantization operation. This converter is slower than the simultaneous converter but is often simpler and cheaper. Other configurations are modifications of these two basic types (e.g., successive-approximation and ripple-through converters).

Modern ADCs typically use integrated circuits for high-speed simultaneous conversion. One advanced type is the optical-electronic converter using optical-electric interaction. Laser light goes to the input of an electrical-optical modulator, whose other input receives the analog signal. The change in the voltage of the analog signal is converted into a distribution of characteristics of the optical signal (into amplitude or phase distribution). A binary representation of the analog value is formed, usually by means of simultaneous electronic comparison of the signal amplitude received from the photo-

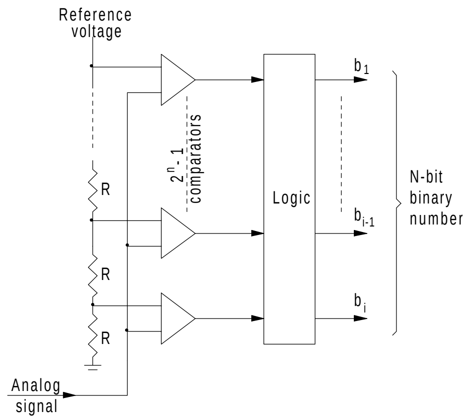


Figure C41 Simultaneous analog-to-digital converter.

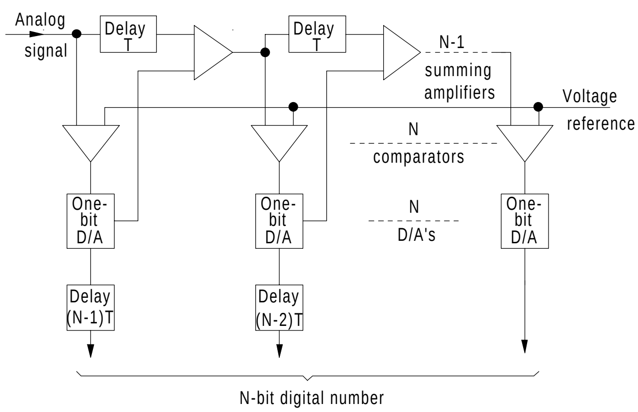


Figure C42 Sequential analog-to-digital converter.

receiver with the threshold amplitude in each of the channels, the number of which is equal to the number of bit positions. *IAM*

Ref.: IEEE (1993), p. 36; Skolnik (1970), pp. 5.43–5.49; Gol'denberg (1985), p. 6; Demler (1991).

Coordinate conversion refers to the computational process of changing data from one coordinate system (e.g., radar spherical coordinates), to another (e.g., rectangular coordinates) (see **COORDINATES, radar**). This conversion is typically carried out in digital computers. *SAL*

A **digital-to-analog converter (DAC)** is "a device that converts an input number sequence into a function of a continuous variable." The input digital code is first converted into a signal of continuous level, constant within the bounds of one time increment, and then is smoothed by a filter (Fig. C43).

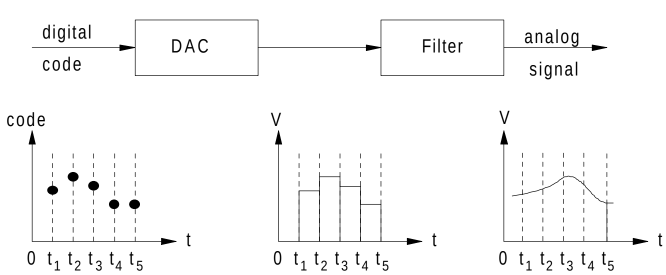


Figure C43 Digital-to-analog converter.

Conversion of parallel digital code into an output current consists of switching and summing currents of current generators, the number of current generators being equal to the number of bits of the code. The current of each is doubled from one bit position to another. Usually the current generators are made in the form of transistorized current stabilizers (frequently based on multiemitter transistors).

Another method of producing such currents is to use a resistor circuit (matrix) formed by serial and parallel connected resistors with two values (R, $2R$) and switching stages. Elements of the matrix R–$2R$ are produced, usually with thin-film technology, and the switch stages are integrated circuits (based for example on MOS transistors).

The DAC parameters are analogous to those of ADCs. DACs are used mainly to convert digital waveforms produced by digital waveform generators into analog waveforms, and in the interfaces of digital signal processing circuits with analog data users (e.g., radar displays). *IAM*

Ref.: IEEE (1993), p. 351; Fink (1982), p. 8.70; Erofeev (1989), p. 493.

downconverter (see frequency converter).

A **frequency converter** is a device that transfers the spectrum of a signal from one region of the frequency band to another without changing the signal structure. Typically it consists of a mixer and a local oscillator. If the frequency of the output signal is less than that at the input, it is termed a *downconverter*, while in the other direction it is an *upconverter*. Downconverters are used in radar receivers to shift the input signal from the carrier frequency to intermediate frequency. Upconverters are used in exciters to shift the transmitted waveform from intermediate frequency to the carrier frequency. *IAM*

Ref.: Druzhinin (1967), p. 368; Fink (1982), p. 14.57; Gassanov (1988), p. 112.

A **range converter** changes the time delay of the echo signal into digital code. It contains a generator of reference pulses which pass through a coincidence circuit to a counter using a set of triggers. The repetition frequency of the reference pulses is selected to obtain a given accuracy and resolution. The counter moves to the initial zero state with the arrival of a reset pulse at the start of range reading. The counter records the number of reference pulses that have passed, in binary representation. The return signals control the counting circuit. If a second target appears, then a second read pulse connects the counter to a second range cell and so forth. *IAM*

Ref.: Vasin (1977), p. 214.

upconversion (see **frequency converter**).

A **velocity converter** changes doppler frequency into digital code. The operation of the converter is based on measurements of the doppler frequency in a specific interval of time. The voltage of the doppler frequency is applied from the output of a mixer to a beat counter through a coincidence circuit that opens for a fixed measurement time. The value of the radial velocity is proportional to the number of counts, calculated with a digital counter. *IAM*

Ref.: Vasin (1977), p. 218.

CONVOLUTION is defined as a function $f(t)$ obtained from two other functions, $f_1(t)$ and $f_2(t)$, by the following rule:

$$f(t) = \int_{-\infty}^{\infty} f_1(\tau) \cdot f_2(t-\tau) d\tau$$

The typical symbolic notation is

$$f(t) = f_1(t) \otimes f_2(t)$$

The concept of convolution is widely used in the theory of spectral analysis of radar waveforms and digital signal processing. The convolution theorem states that the Fourier transform of the convolution of two functions is the product of their individual Fourier transforms. This means that convolution in the time domain can be carried out by multiplication in the frequency domain. In terms of a signal passing through a filter it means that convolution of the input signal and the filter impulse response is equivalent to forming the product of the signal spectrum and the filter transfer function. Digital convolution employing fast Fourier transform algorithms is sometimes termed fast convolution. *SAL*

Ref.: Wehner (1987), p. 138.

COORDINATES, radar. Radar coordinates may refer to antenna coordinates or to the radar target coordinates relative to the radar system location. Radar-centered coordinate systems include rectangular and spherical, as shown in Figs. C44 and C45.

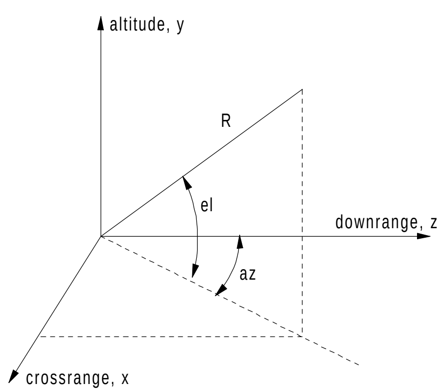


Figure C44 Rectangular coordinate system.

It can be seen that in the spherical coordinate system, the azimuth angle in Fig. C45 is equal to (θ cos ϕ) in Fig. C45, and the elevation angle in Fig. C44 is equal to (θ sin ϕ). Figure C45 may be used to represent the coordinate system for a radar antenna. Here the aperture is located in the x–y plane, with the antenna broadside beam directed along the z-axis. The angle θ is the deviation from broadside, while ϕ indicates the direction of this deviation.

For planar array antennas, which are capable of steering a beam or beams in two dimensions, an alternate coordinate system called “sine space” is often convenient for visualizing the effects of scanning. In a spherical coordinate system, the two angles θ and ϕ define a point on the surface of a unit hemisphere (radius = 1), as in Fig. C46.

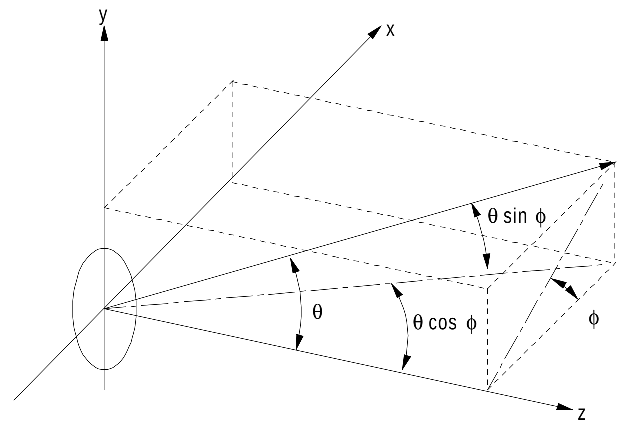


Figure C45 Spherical coordinate system.

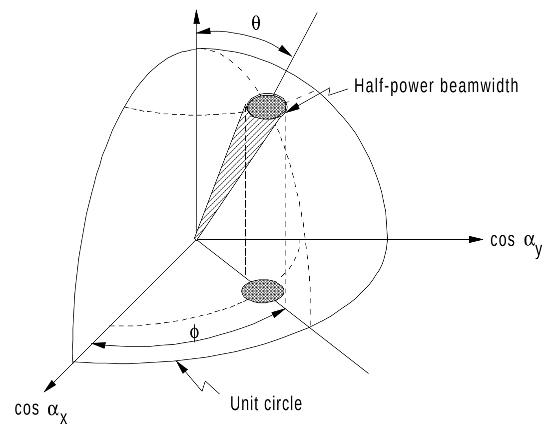


Figure C46 Array antenna coordinate system in sine space (after Skolnik, 1970, Fig. 10, p. 11.16, reprinted by permission of McGraw-Hill).

If the point on the surface of the hemisphere is projected onto a plane, the axes of the plane are the direction cosines cos α_x, cos α_y. The direction of the scan is given by the direction cosines:

$$\cos\alpha_{xscan} = \sin\theta_{scan}\cos\phi_{scan}$$

$$\cos\alpha_{yscan} = \sin\theta_{scan}\sin\phi_{scan}$$

and the plane of the scan is found from:

$$\phi = \operatorname{atan}\left(\frac{\cos\alpha_{yscan}}{\cos\alpha_{xscan}}\right)$$

The scan angle θ is determined by the distance of the point (cos α_{xscan}, cos α_{yscan}) from the origin on the plane of the array, which, because we are dealing with a unit hemisphere, is simply equal to sinθ

Advantages of using the sine-space coordinate system for array antennas are (1) the antenna pattern shape is invariant with the direction of the scan and (2) calculations required for setting individual array element steering phases are greatly simplified. *PCH*

Ref.: Barton (1988), p.146, Skolnik (1970), pp. 11.15–11.17.

CORRELATION function. The correlation function is the average value of the product of two functions, $\xi_i(q)$, for two different values of the arguments

$$K_{\xi_1\xi_2}(q_1, q_2) = \langle \xi_1(q_1)\xi_2(q_2)\rangle$$

where <y> represents the statistical average of *y*. In radar applications the concept of the correlation function is typically applied to random processes, $\xi(t)$, describing radar, signals $s_i(t)$ as functions of time:

$$K_{s_1s_2}(t_1, t_2) = \langle s_1(t)\cdot s_2(t)\rangle$$

$$= \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} s_1(t_1)s_2(t_2)f_{s_1s_2}(x_1, x_2; t_1, t_2)dx_1dx_2$$

where $f_{s_1s_2}$ is the second-order probability density functions of the joint distribution of $s_1(t)$ and $s_1(t)$.

The function $K_{s_1s_2}$ is sometimes termed the *cross-correlation function* of the two signals $s_1(t)$ and $s_2(t)$, and it is the autocorrelation function if $s_1(t) = s_2(t) = s(t)$. In many practical cases it is considered that the correlation function varies only with the time difference, $|t_1 - t_2| = \tau$:

$$K_{s_1s_2}(t_1, t_2) = K_{s_1s_2}(\tau)$$

which is valid for the stationary random process. In this case the variance of the signal $\sigma_s{}^2$ does not depend on time and the correlation function is a function of one argument only:

$$K_s(\tau) = \sigma_s^2\cdot R_s(\tau)$$

where $0 \le |R_s| \le 1$ is the normalized correlation function or correlation coefficient.

The main approximations used for correlation functions of stationary random processes and corresponding power spectra are given in Table C7. The space in the argument of the correlation function within which the function falls below some specified value (e.g., $|R(\tau)| \le 0.1$) is called the *correlation interval*. If this argument is time, the correlation interval is called the *correlation time*.

In radar applications the concept of the correlation function is used primarily for describing radar signals and measurement errors. *SAL*

Ref.: Barkat (1991), p. 67

The **autocorrelation function** of a signal is a function that determines the interrelationship between the signal $u(t)$ and its time shifted copy $u(t - t)$. If the signal is described by a steady random process then the auto-correlation function is determined by the equation

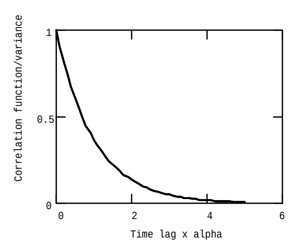
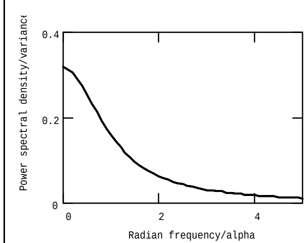
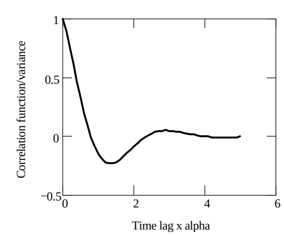
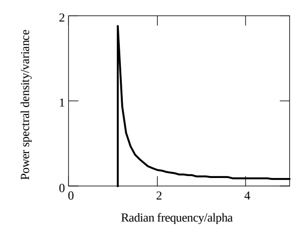
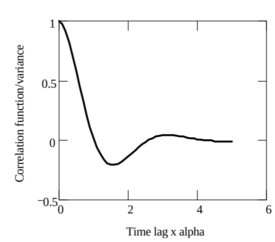
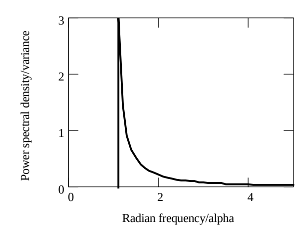
$$K(t) = \frac{1}{E_s}\int u(t)u(t-\tau)d\tau$$

where E_s is the signal energy.

The correlation function is an important concept in signal theory. It is explained by the fact that, in radar to extract useful information, one can use only the voltage at the output of a matched filter, because to a considerable degree it is free of interference. But filter response is a correlation function of the signal. If the normalized voltage at output of the matched filter is described by the correlation function $k(t)$, then at the output of the detector, located after the filter, it is determined by the absolute value of the complex envelope of the correlation function $|K(t)|$. In this way it allows one to determine measurement accuracy of useful parameters of signals. *AIL*

Ref.: DiFranco (1968), p. 111; Varakin (1970), pp. 47–72.

Table C7
Typical Correlation Functions and Power Spectra

<table>
<tr><th>Correlation function</th><th>Power spectrum</th></tr>
<tr><td>$K(\tau) = \sigma^2\exp-\alpha|\tau|)$
$\alpha > 0$</td><td>$G(\omega) = \frac{\sigma^2}{\pi}\cdot\frac{\alpha}{\omega^2+\alpha^2}$</td></tr>
<tr><td>
</td><td>
</td></tr>
<tr><td>$K(\tau) = \sigma^2\exp(-\alpha|\tau|)\cos\beta\tau$
$\alpha > 0$</td><td>$G(\omega) = \frac{\sigma^2\alpha}{\pi}\cdot\frac{\omega^2+\alpha^2+\beta^2}{(\omega^2-\alpha^2-\beta^2)+4\alpha^2\omega^2}$</td></tr>
<tr><td>
</td><td>
</td></tr>
<tr><td>$K(\tau) = \sigma^2\exp(-\alpha|\tau|)\Big(\cos\beta|\tau|$
$+\frac{\alpha}{\beta}\sin\beta|\tau|\Big)$ $\alpha > 0$</td><td>$G(\omega) = \frac{2\sigma^2\alpha}{\pi}\cdot\frac{\alpha^2+\beta^2}{(\omega^2-\alpha^2-\beta^2)+4\alpha^2\omega^2}$</td></tr>
<tr><td>
</td><td>
</td></tr>
</table>

CORRELATOR. A correlator is a device that calculates the correlation function of random processes. It comprises a device to cross-multiply the input signals and an integrator in which the products from the multiplier are summed. The existence of a statistical relationship between the input signals results in the appearance of constant components at the output of the multiplier, fluctuations in which are reduced by the integrator.

Depending upon the type of multiplier, correlators are categorized as being either direct or indirect. Direct correlators use physical effects that generate quantities proportional to the product of the inputs: various types of modulation, the Hall effect, and so on. Indirect correlators use operations other than time-domain multiplication; for example, multiplication of the signals' Fourier transforms with the inverse transform (as in a functional correlator). Some types of correlators are described below. *IAM*

Ref.: Vinokurov (1972), p. 51; Zmuda (1994), Ch. 15.

An **acousto-optical correlator** is an optoelectronic device in which correlation is performed on the basis of the diffraction of coherent optical radiation by two acoustic waves, created by the input signals and propagating in opposite directions. There are two types of acousto-optical correlator: those that perform spatial integration of the output light distribution and those that perform temporal integration of the output signal with a photoreceiver. These two types of correlator differ in the construction of the optics and the existence in the spatial correlator of a strip of photodetectors the output currents of which are integrated, in place of a single photodetector with an integrating filter. The duration of signals processed in a spatial correlator is on the order of microseconds or tens of microseconds, while a temporal correlator may process signals up to tens of milliseconds in duration.

The main elements of an acousto-optical correlator are a pair of acousto-optical modulators, which determine the basic properties of the correlator: the carrier frequency (on the order of 1 GHz), and the passband (several hundred megahertz). *IAM*

Ref.: Kulikov (1989), pp. 110, 113; Zmuda (1994), p. 405.

A **charge-coupled device correlator** is a correlator constructed with charge-coupled devices, generally containing two delay lines, from the output taps of which samples of the input signals are taken to be multiplied and summed. A distinguishing feature of a charge-coupled device delay line is its invariance to the delay of one signal relative to another. A large portion of the correlator is occupied by the multiplier, which is usually implemented using field-effect transistor technology.

The advantages of charge-coupled device correlators are realized at video frequencies. The maximum clock frequency at which the correlator may operate is inversely proportional to the length of the multiplier, and, for an accuracy exceeding 1%, is 5 to 8 MHz.

Charge-coupled device correlators are used in synchronizers and optimum filters, as functional transform devices, and also for coding and decoding. Sometimes this correlator is called a *surface-charge correlator. IAM*

Ref.: Gassanov (1988), p. 227.

A **functional correlator** is one in which the multiplication of the analyzed processes is replaced by their functional transform $f(x,y)$. An example of a functional correlator is the so-called sign correlator, or polarity coincidence correlator, for which $f(x,y) = f_1(x)\, f_2(y) = \mathrm{sign}(x)\mathrm{sign}(y)$.

Correlators based on an annular balanced mixer are often used at radio and video frequencies. The mixer is operated as a nonlinear transform device, in the "switch" mode, with one of the voltages significantly greater than the other. When there is a correlation between the analyzed signals, there appears at the output a signal at the difference frequency with amplitude proportional to the correlation coefficient. *IAM*

Ref.: Vinokurov (1972), p. 51.

An **optical correlator** is one that uses optical devices to perform the required multiplication and integration operations. This relies on the change in amplitude and phase of a light wave as it passes through an inhomogeneous medium and on the focusing action of a lens. Depending on whether the illumination is coherent, such correlators are categorized as being either *coherent-optical* or *incoherent-optical* correlators.

Common to all optical correlators are a pair of imprinted transparencies that modify the amplitude and phase of the signal in accordance with the Fourier transform of the imprinted signals; one of several types of optical lens; and a photodetector at the output (e.g., a photoelectronic multiplier).

Coherent optical correlators use coherent illumination and possess a number of advantages relative to noncoherent optical correlators. They are suitable for operation with complex functions, the accuracy of the correlation calculation does not depend upon diffraction effects within the optics, and it is possible to form filters with given transfer characteristics with great flexibility and simplicity.

The advantages to incoherent optical correlators include their simple construction and the fact that they may be used with objects that emit their own (generally incoherent) light.

In accordance with the manner in which the correlation function is calculated, optical correlators are further classified as being either *spatially independent*, in which case all terms of the correlation function are calculated simultaneously, producing the correlation function as the output field, and *spatially dependent*. A spatially dependent correlator calculates one term of the correlation function for a given spatial position of the transparencies, thus greatly increasing the calculation time and the required number of samples of the signals being analyzed. *IAM*

Ref.: Baklitskiy (1986), p. 179; Nathanson (1990), p. 311, Brookner (1977), pp. 235–237, 244; Zmuda (1994), p. 403.

COTTON-MOUTON EFFECT. The Cotton-Mouton effect is a phenomenon entailing the conversion of the linear polarization of waves into elliptical polarization as a result of the transverse propagation (perpendicular to the direction of the magnetic field) of the electromagnetic wave through a gyrotropic medium. This is explained by the separation of the falling wave into two normal linearly polarized waves with different phase speeds. The vector of one of these waves is directed along the external field.

The Cotton-Mouton effect occurs when radio waves with frequencies less than 300 MHz are propagated in the ionosphere. *IAM*

Ref.: Kravtsov (1983) p. 82; Nikol'skiy (1964) p. 189.

COUNTER-COUNTERMEASURES (see **ELECTRONIC COUNTER-COUNTERMEASURES**)

COUNTERMEASURES (see **ELECTRONIC COUNTERMEASURES**)

COUPLER, directional. A directional coupler is a multiport device providing directional coupling of energy. The device contains a main line and a single or several auxiliary lines (coupling lines). When one branch of the main line is excited, a portion of the power is transferred to the other branch of the main line, and a portion to one of the branches of the coupling line. The second branch of the auxiliary line is not coupled to the excited branch of the main line, so that no power is transferred to it.

One of the main parameters of a directional coupler is the attenuation, equal to the ratio of the coupled power to the power in the forward wave. The directivity is the ratio of the power in the coupled forward wave to the power in the coupled back wave. The decoupling ratio is the ratio of the power in the forward wave to the power in the decoupled output of the auxiliary line. Typically, a directional coupler comprises two transmission lines (waveguide, coaxial, or stripline), connected with coupling elements. Depending upon the method by which the auxiliary line is excited (in-phase or antiphase excitation), the direction of propagation in the auxiliary line will either be the same or opposite to that in the main line. Directional couplers are also distinguished on the basis of the type of coupling element (e.g., aperture, slot, or loop). Some types of directional couplers are described below.

A common radar application of directional couplers is to provide a test point for the transmitter and receiver, after their combination in the duplexer. Power and receiver sensitivity test equipment can be connected to this test point, which normally has a directivity of −30 dB or more. *IAM*

Ref.: Rakov (1970), p. 255; Gardiol (1984), p. 272; Veselov (1988), p. 57; Sazonov (1988), p. 106.

A **hybrid [3-dB] coupler** is one in which there is equal power in the output branches. A hybrid coupler in which the electrical fields in the output branches have identical voltages and a constant phase shift is called a *bridge*. *IAM*

Ref.: Sokolov (1984), p. 190; Gardiol (1984), p. 279.

A **ring coupler** is a closed configuration type of directional coupler. Typically it is a stub ("square") coupler or hybrid ring. A stub coupler is one constructed from two lengths of transmission line, comprising between them two or more stubs. The length of the stubs and the distance between them is a quarter-wavelength. The band of operating frequencies increases with the number of stubs. In practice, the number of stubs never exceeds three, due to the sharp increase in the wave impedance at the edges of the stubs as their number increases, and an increase in the active loss.

The hybrid ring coupler usually differs from the stub coupler in that the length of transmission line between adjacent inputs is increased to three-quarters of the wavelength. In the shortwave portion of the microwave band cavity hybrid ring couplers are used (see Fig. C47), while at larger wavelengths the stub coupler in the form of a meander line is employed.

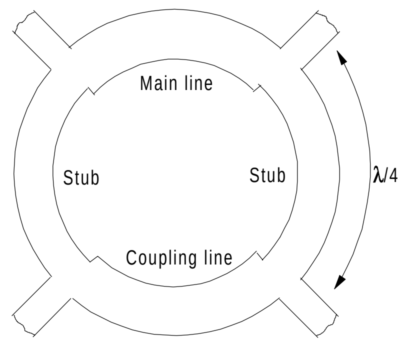


Figure C47 Two-stub coupler.

Ring couplers are widely used in waveguide, coaxial, and stripline transmission lines. *IAM*

Ref.: Gardiol (1984), p. 284; Veselov (1988), p. 61.

A **waveguide coupler** is a directional coupler constructed from waveguide transmission lines. The coupling element may either be a set of stubs, or an aperture or slot (Fig. C48). In a waveguide coupler that has several coupling elements along the center of the wide wall of the waveguide, the excited wave is in phase with the input wave in the main line, and therefore travels in the same direction. An example of a waveguide coupler is the Bethe hole coupler, which is the simplest type of directional waveguide coupler, with a single aperture as the coupling element. The opening is usually located in the wide wall of the waveguide. The phase and direction of propagation of the field in the coupling line are opposite to those of the wave in the main line. *IAM*

Ref.: Montgomery (1947), p. 313; Druzhinin (1967), p. 140; Rakov (1970), vol. 2, p. 258.

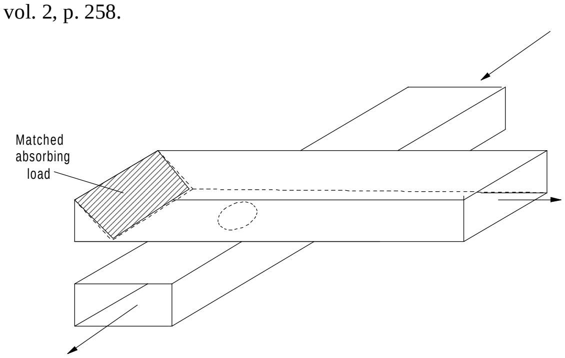


Figure C48 Waveguide (Bethe) coupler.

COVERAGE, radar. Radar coverage is the general term for the three-dimensional boundary describing the volume in space within which radar operational capabilities meet the specified requirements. These requirements depend on the type of radar and can be described in terms of detection and false-alarm probabilities if the task is detection only. For detection and tracking tasks, resolution and errors in measurement are added to the requirements, while for target recognition the probabilities of correct and incorrect classification become important.

The volumetric coverage of a given radar can be described in terms of its maximum (and minimum) detection ranges as a function of target azimuth and elevation angle, if the detection criteria and the environment are adequately specified. For doppler-measuring (CW, MTI, and PD) radars, "coverage" also includes the range of target radial velocities over which the radar can detect and process target data. Required detection criteria include probability of detection P_d (per scan or cumulative), probability of false alarm P_{fa}, or false-alarm time t_{fa} per scan, target radial velocity, target radar cross section σ, and statistical RCS fluctuation model assumed. For comparing the performance of different radar designs for the same mission, it is common practice to assume a standard set of detection criteria (e.g., for a volume search radar, $P_d = 0.5$ to 0.9, $P_{fa} = 10^{-6}$, $\sigma = 1\ \mathrm{m}^2$, Swerling case 1 fluctuation model) and a standard environment, such as free-space.

The graphical representation of coverage is called a *coverage chart* (or diagram). A complete description of radar coverage includes charts for (a) free-space coverage; (b) coverage with lobing (multipath propagation), if applicable; (c) coverage with ground clutter; (d) coverage with sea clutter; (e) coverage with weather clutter; (f) coverage with combined clutter (e.g., sea plus weather clutter); (g) coverage with jamming; (h) coverage with combinations of clutter and jamming; and (i) coverage under certain special conditions (e.g., anomalous propagation, fail-soft operation for solid-state radar, etc.) To obtain an accurate assessment of radar coverage under such conditions, an appropriate description of the conditions (corresponding to models of clutter and jamming) must be included (see **CLUTTER**, **JAMMING**).

Figure C49 is an example of radar coverage in the elevation plane for a 2D ground-based, horizon-search radar. The lobing effects due to interference from surface reflections are clearly visible in the figure. (See **PROPAGATION, wave.**) Figure C50 is an example of the elevation versus range coverage provided by a 3D, stacked-beam search radar. Complete coverage of the desired search volume would be covered by scanning the radar antenna 360° in azimuth.

Airborne or spaceborne radars are often employed to provide surveillance or mapping of the surface. Figure C51 illustrates the ground coverage, in the elevation plane, of an airborne radar in its air-to-ground mode, and Fig. C52 shows the wide-area coverage typical of a satellite-based ocean surveillance radar.

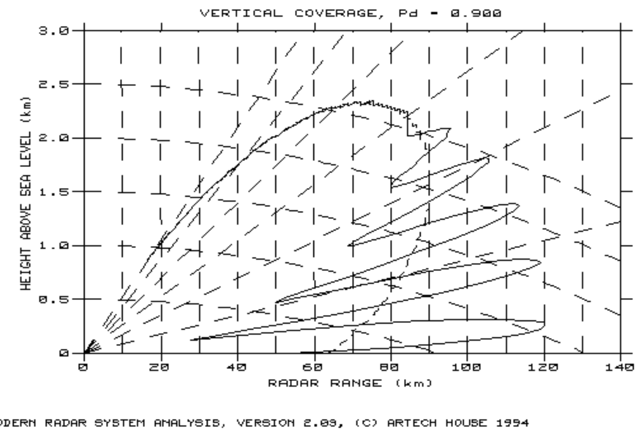


Figure C49 2D search radar coverage (from Barton, 1993, Fig. 2.36, p. 69).

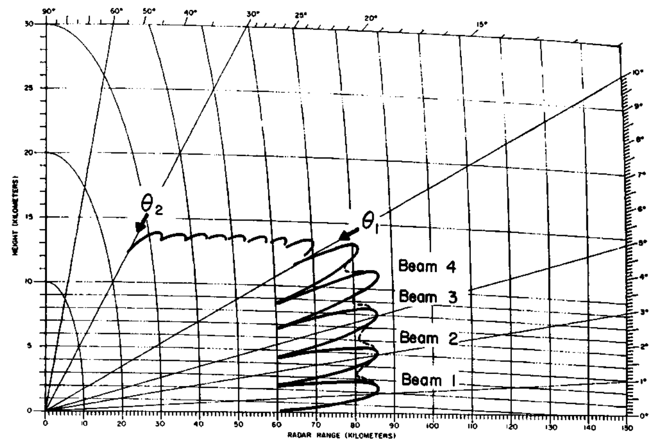


Figure C50 3D, stacked-beam radar coverage (from Barton, 1988, Fig. 7.3.1, p. 348).

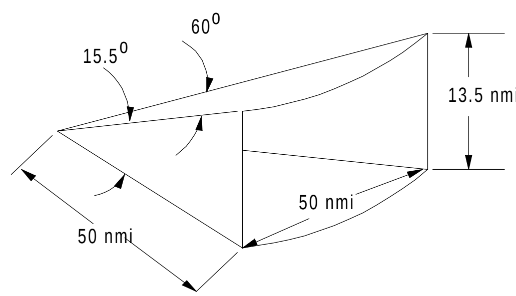


Figure C51 Airborne radar ground coverage.

Other types of radar coverage charts can be prepared to portray the combined coverage of an air defense network, air traffic control system, or spaced-based ocean surveillance system. In general, the broader the scope of the radar coverage information desired, the more difficult and expensive it is to provide accurate, detailed, and timely information.

Radar coverage charts can be prepared to show target detection limits in other-than-spatial dimensions. Figure C53 is a range-doppler map for a multiple-PRF radar, indicating blind regions within the coverage due to eclipsing effects, mainlobe clutter, and range-doppler ambiguities.

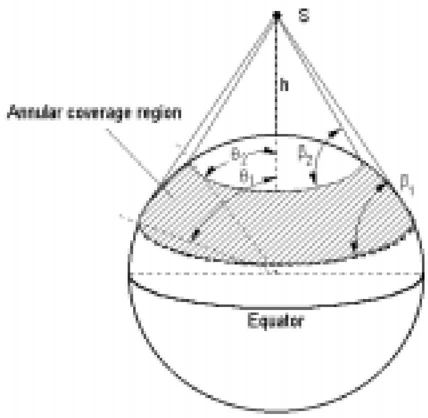


Figure C52 Typical spaced-based radar coverage.

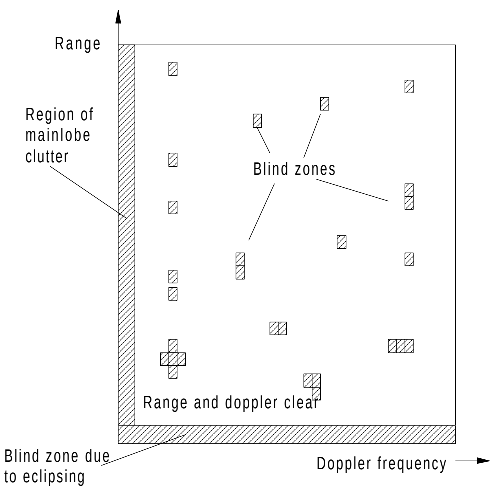


Figure C53 Range-doppler coverage.

The coverage zone of bistatic radar is described, for given signal-to-noise ratio, by the ovals of Cassini (Fig. C54). Three distinct regions are defined: (1) the receiver-centered region, (2) the transmitter-centered region, and (3) cosite region, which envelopes both receiver and transmitter. The signal-to-noise ratio can be expressed as

$$\frac{S}{N} = \frac{\kappa^2 (S/N)_{min}}{(r^2 + L^2/4)^2 - r^2 L^2 (\cos\theta)^2}$$

where κ includes the terms of the bistatic radar equation, excluding the range terms, $r^2 = R_t R_r$. *PCH, SAL*

Ref.: Blake (1982); Hovanessian (1984); Barton (1988); Cantafio (1989); Willis (1991), Ch. 4.

COVERING, antiradar (see **ABSORBER, radar**).

CRITERIA FOR DETECTION (see **DETECTION criteria**).

CROSSED-FIELD AMPLIFIER (CFA). A crossed-field amplifier is a microwave tube in which the output amplification results from the interaction of the electromagnetic wave propagating along the slow-wave circuit and the electron beam moving in crossed electric and magnetic fields. All

(a)

(b)

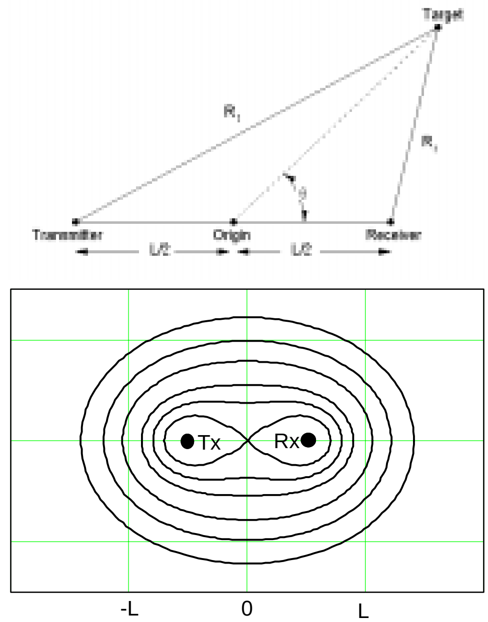


Figure C54 Bistatic geometry and coverage: (a) projection in vertical plane; (b) contours of constant SNR (ovals of Cassini), for $\kappa = L^4$ (after Willis, 1991, p. 72).

crossed-field amplifiers are typically divided into the following groups: from the point of view of electrons emission mode as distributed-emission CFAs (terms *emitting-sole CFA* and *continuous-cathode CFA* are also used interchangeably) and injected-beam CFAs; from the point of view of operation mode as pulsed and CW CFAs; from the point of view of using reentrance of the electron beam as reentrant and non-reentrant CFAs; from the point of view of interaction with the traveling wave as forward-wave and backward-wave CFAs; and from the standpoint of the format used as linear-format and circular-format CFAs.

The CFA falls in the same class of crossed-field tubes as the magnetron, so it has much resemblance with the latter in characteristics and even in physical appearance. Although there are different types of CFAs, all of them employ cathode, input and output ports, and a slow-wave circuit as the basic elements. The dominant types of CFAs used in radar are pulsed, reentrant, distributed-emission CFAs. The main assets of CFAs are: high efficiency (typical figures are 40 to 60%, and even 80 to 90% were reported); relatively low operating voltage (in comparison with linear-beam tubes); rather broad bandwidth (10 to 25% in forward-wave CFA and about 10% in backward-wave CFA); good phase stability; compatibility with pulse-compression waveforms; long life; and relatively low weight and small size. The main disadvantages associated with this device are: relatively low gain (in comparison with linear-beam tubes); worse noise performance (in comparison with linear-beam tube), and some problems identified with all crossed-field devices (e.g., spurious RF output; see **MAGNETRON**).

CFAs are widely used in all types of radars. They can serve as a power booster following a magnetron oscillator, as the high-power stage in amplifier chains, or as the individual transmitter in phased-array radars. When used in an amplifier chain, the output-stage CFA is often preceded by a medium-power traveling-wave tube, which makes it possible to combine the best qualities of both tubes. The traveling-wave tube provides high gain, and CFA ensures high resultant power with high efficiency and good phase stability. In Russian radar literature, the crossed-field amplifier is called a *magnetron amplifier*. Several types of commercial available CFAs are listed in Table C8. *SAL*

Ref.: Ewell (1981), pp. 37–54; Skolnik (1980), pp. 208–212, (1990), pp. 4.12–4.14; Brookner (1988), p. 317–324; Leonov (1988), pp. 49–51.

A **backward-wave CFA** is a crossed-field amplifier in which electron beam interacts with backward traveling wave as the phase and group velocities of the propagating signal are in the opposite direction. In backward-wave CFAs, the voltage required for a given peak current is proportional to frequency, and the operating voltage for constant power output depends on the frequency amplified; so this device is the voltage-tunable amplifier. The backward-wave structure was developed and applied first in comparison with forward-wave CFAs. (See **AMPLITRON.**) *SAL*

Ref. Ewell (1981), p. 38; Skolnik (1990), p. 4.13.

A **cathode-driven CFA** is a crossed-field amplifier employing a cold cathode and started by the RF drive. Such a design permits achievement of about 30 dB of gain in a pulsed mode in comparison to 10–17 dB that is typical for conventional pulsed CFAs. Sometimes this type of CFA is termed a *high-gain CFA*. *SAL*

Ref. Skolnik (1980), p. 212.

The circular-format CFA is the structure of a distributed-emission crossed-field amplifier in which the electrons from the output are isolated from the input forming the nonreentrant configuration (Fig. C55). *SAL*

Ref.: Ewell (1981), p. 37.

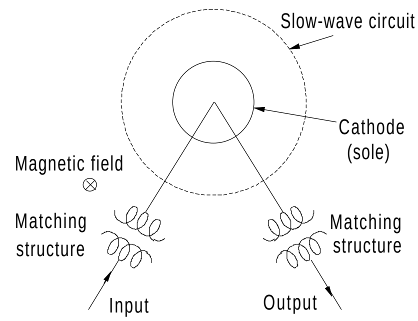


Figure C55 Simplified representation of circular-format CFA (after Ewell, 1981, Fig. 2-12, p. 38).

The **distributed-emission [continuous-cathode] CFA** is a crossed-field amplifier in which electron current is obtained from the cathode in the interaction space by electron beam back bombardment (the cathode in CFAs is also known as the *sole*). A schematic diagram of distributed-emission CFA is cited in Fig. C56. Electrons start from the cylindrical cathode which is coaxial to the RF slow-wave circuit that acts as the anode. The slow-wave structure is designed in a manner to make RF signal propagate at a velocity near that of the electron beam to exchange energy from the electron beam to the RF field to produce amplification. The main frequency range for such tubes is from VHF to K_u band, peak powers up to several megawatts. This device is also called interchangeably *emitting-sole CFA* and *continuous-cathode CFA*. It is the main type of CFAs used in radar applications. *SAL*

Ref.: Fink (1975), pp. 9.55–9.60; Skolnik (1980), p. 209; Ewell (1981), p. 37.

emitting-sole CFA (see **distributed-emission CFA**).

Table C8
Some Commercially Available High-Power Pulsed CFAs

Tube type	Center frequency	Peak P_o	Frequency range	Maximum duty cycle	Peak		Gain
	(GHz)	(MW)	(GHz)		Voltage (kV)	Current (A)	(dB)
1AM10	1.288	1.8	1.225–1.350	0.02	46	50	9.2
QKS1452	2.998	3.0	2.994–3.002	0.0015	47	100	–
SFD222	5.65	1.0	5.4–5.9	0.001	35	60	18
SFD237	5.65	1.0	5.4–5.96	0.01	35	60	13
QKS506	9.05	1.0	8.7–9.4	–	40	45	7
SFD236	16.5	0.1	16–17	0.001	14	23	17

(from Ewell, 1981, p. 55).

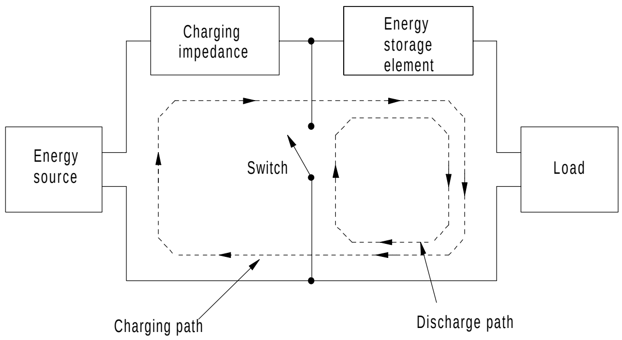


Figure C56 Simple representation of a distributed-emission crossed-field amplifier (after Skolnik, 1980, Fig. 6.12, p. 209).

A **forward-wave CFA** is a crossed-field amplifier in which the electron beam interacts with a forward-traveling wave as the phase and group velocity of the signal propagating along the slow-wave circuit are in the same direction. In comparison with the backward-wave CFA, it can operate over a broad range of frequencies (typical bandwidth is 10 to 20%), but it has lower power efficiency than the latter. This structure was invented in France in 1950. *SAL*

Ref.: Skolnik (1980), p. 209; Brookner (1988), p. 263.

high-gain CFA (see **cathode-driven CFA**).

An **injected-beam CFA** is a crossed-field amplifier in which the electrons are not emitted by the cathode, as in distributed-emission CFAs, but injected into the interaction region by an electron gun. This type of CFA generally is not suited for high powers. So despite its inherent capability for achieving greater gain than distributed-emission CFA (more than 20 dB), they have not found much application in practical radar and are basically ECM devices if used. (See **BIMATRON**, **BITERMITRON**.) *SAL*

Ref.: Ewell (1981), p. 37; Brookner (1988), pp. 317, 318.

A **linear-format CFA** is a form of crossed-field amplifier having nonreentrant configuration, in which the electrons are terminated by the collector after the interaction with RF field (Fig. C57). *SAL*

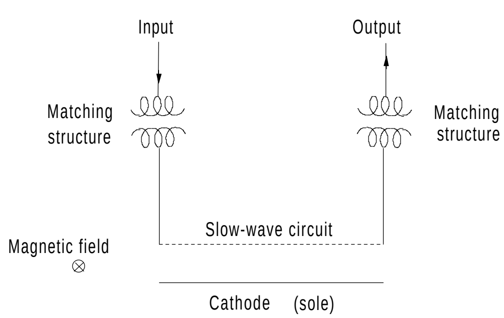


Figure C57 Simplified representation of linear-format CFA (after Ewell, 1981, Fig. 2-12, p. 38).

Ref.: Ewell (1981), p. 37.

A **nonreentrant CFA** is a crossed-field amplifier using a special collector electrode to terminate the electron stream. This mode of operation is typical to linear-format CFAs, although it can be implemented also in circular-format CFAs. *SAL*

Ref.: Ewell (1981), p. 37; Skolnik (1980), p. 209.

A **pulsed CFA** operates in the pulsed mode. Depending on the modulation techniques, the main types of pulsed distributed-emission CFAs are cathode-pulsed CFAs, dc-operated CFAs with the combination of dc voltage and pulsed turn-off voltage, and dc-operated CFAs with only dc voltages applied. Most CFAs used in radar applications are pulsed ones. They are capable of several megawatts of peak power, up to 20 dB of gain, and 40 to 60% efficiency (even figures of 80 to 90% efficiency were reported). *SAL*

Ref.: Fink (1975), p. 9.55; Skolnik (1980), p. 210.

A **reentrant CFA** is a crossed-field amplifier in which the electrons that are not collected after energy is extracted are permitted to reenter interaction area at the input. This improves the efficiency of the tube in comparison with nonreentrant CFA structure, but RF feedback, like any feedback, opens the possibility of oscillations as the reentering electrons might contain modulation that will be amplified when the next pass through the circuit occurs. Some special measures, including debunching of electron stream, are employed to eliminate the possibility of oscillation. *SAL*

Ref.: Ewell (1981), p. 41; Skolnik (1980), p. 209.

CROSS SECTION (see **RADAR CROSS SECTION**).

CROWBAR. A crowbar is a device used to discharge energy from a circuit as a result of onset of an electrical discharge. The basic types of crowbars in radar systems are thyratrons, ignitrons, gas and vacuum crowbars, and ball and multigap spark crowbars.

The diatron and ignitron were historically the first gas-filled discharge tubes with discharge in hydrogen (thyratron) and in mercury vapors (ignitron). They are less convenient due to the presence of a thermal cathode in a thyratron and the complex predischarge circuit of the conventional ignitron, and may be damaged from a high rate of growth of the current, on the order of 10 kA/s. Gas crowbars are distinguished by their compactness, but they do no permit exceeding of the nominal powers. Vacuum discharges with plasma control have the capability of ignition at any operating voltage, but they have limitations with respect to rate of current increase. The ball-type protective spark crowbar consists of two balls with a needle igniting electrode between them. Because of the need to connect additional inductance coils, they remove thermal energy less effectively from the load than crowbars of other types. Multigap crowbars are the most widely used. They have practically unlimited power and the capacity of self-ignition.

Crowbars are used as protective devices for active modulators in radar transmitters and as protective devices at the input of radar receivers. The basic parameters of some common crowbars are given in Table C9. They are also called *dischargers* or *energy diverters*. *IAM*

Ref.: Skolnik (1990), p. 4.40.

Table C9
Crowbar Devices

	Hydrogen thyratron	Mercury pool ignitron	EG&G triggered gas gap	Energy Systems ball gap	Multigap crowbar	Plasma triggered vacuum gap
Voltage rating (kV, max)	40–125	50	70	No limit	No limit	75
Current and joule rating	5–10 kA	50 kA	100 kA, 4 kJ	No limit	No limit	70 kA
Firing range	10:1	50:1	3:1	"infinite"*	"infinite"	"infinite"
Self-firing	No	No	No	No	Yes	No
Triggered end	Negative	Negative	Either	Either	Either	Negative
Trigger *V* as a fraction of voltage rating	1/10	1/50	1/3	1/2	1/2	5 kV (all sizes)
Size	Small	Small	Small	Large	Large	Small
Cost	Low	Low	Medium	Medium	High	Medium

*With some inductance in series with load, which limits effectiveness.
(from Skolnik, 1990, Table 4.4, p. 4.40, reprinted by permission of McGraw-Hill).

D

DATA, radar. The term *radar data* generally refers to the information gathered about the target by a radar or a set of radars. Typically, this information makes possible the determination of radar target coordinates (range, azimuth, elevation, radial velocity) and reflective characteristics of the target (e.g., radar cross section). The radar data are collected, processed, stored, and displayed in the radar channel. Various information measures can be used for theoretical description of the quantity and quality of information circulating in the radar channel. *SAL*

Ref.: Skolnik (1962), p. 453; Tuchkov (1985), p. 11.

Data association is the procedure of assigning a set of estimate of the dynamic states of the targets (e.g. tracks) to a set of measurements generated by a radar or a set of radars. This is sometimes referred to as the *correlation problem*, which is a serious problem in data fusion and multiple-target tracking tasks. In this case, a decision must be made about the correspondence of a given set of measurements with a given set of tracks and estimates made of correctness of the decision. Typically, these estimates are the probability of correct association and the probability of misassociation. These probabilities depend primarily on the ratio between radar resolution (accuracy) and the density of the target within radar coverage. The simple analytic models for predicting association performance can be derived only for some specific cases. In the general case, a computer simulation using the Monte Carlo approach is the most appropriate.

The two basic algorithms used to implement data association are *nearest neighbor* and *matrix assignment* (all-neighbors) association. In the first case, the total summed distance from observations (reports) to the assigned tracks is minimized. The state estimate (position and velocity in each coordinate) for each track is projected forward to the next scan, and when the signal processor provides the next set of reports the algorithm assigns to each track the plot that is closest to its prediction. In the all-neighbors approach the association problem is considered a matrix, each row representing an existing track and each column a number of reports for the current scan. The *ij*th element of the matrix indicates the likelihood (or probability) that the *j*th report is associated with the *i*th track. The optimum solution for the report-to-track assignment requires selecting the set of report-track pairs that maximizes the sum of the matrix entries. The nearest neighbor approach is much simpler, but target-tracking performance can be severely degraded by misassociation. The all-neighbors approach provides considerable improvement in association performance but at the expense of complexity, often requiring considerable computation resources. *SAL*

Ref.: Hovanessian (1988), p. 253; Bar-Shalom (1992), v. II, pp. 183–227.

Data differentiation is part of the process of radar measurement. Differentiating is used primarily to determine the target velocity using derivatives obtained from radar position data output. Velocity in three coordinates can be obtained from a single radar only by differentiation. The process is performed by differentiating circuits or algorithms, typically results in an increase in noise, and must always be accompanied by smoothing. *SAL*

Ref.: Barton (1964), pp. 422–428.

Data encoding is the process by which information is transformed to a sequence of standard signals, so that the information may be transmitted between a processing system and a digital computer. This coding is generally performed using a binary code in which the only symbols are 0 and 1.

Ref.: Sloka (1970), p. 140; Jordan (1985), Ch. 24.

Data fusion is the combining of information obtained from several sources relating to the same target. It is employed in radar networks, multistatic radar systems and multiple-sensor systems (e.g., those that combine radar and infrared sensors). The general methods of data fusion include combining radio or intermediate frequency signals in the linear portion of the receiver, combining video signals, combining detections and individual reports, and combining trajectories. When radar signals from different sensors (e.g., different sites in multistatic radar) are combined within the processing system, all the information about the target and interference sources is transmitted, requiring wideband communications links and special data-compression methods. The advantage to such a system is that it permits coherent processing in which information about the spatial structure of the electromagnetic waves is used.

When the processing system combines trajectory data, the data processed at each site are transmitted using more simple transmission techniques. The greater the level of data collection at the processing site, the lower the information loss at the individual radars and the greater the capabilities of the system as a whole. The primary problems associated with combining individual reports and trajectories are the extraction and association of data as belonging to a given target, and the calculation of smoothed estimates of the parameters describing the trajectory and the reflection characteristics.

The conditions necessary to combine radar data at the RF signal level are defined by the ability to control and synchronize the various radar sites, and to form reference signals with sufficient stability to permit coherent processing to be performed at each site.

Data fusion from multiple radars is used in air traffic control, antimissile defense, sea-surface surveillance, and other systems.

IAM

Ref.: Chernyak (1993), p. 9; Milne, K., *IEE Int. Conf. Radar-77*, p. 46; Buchner, M. R., *IEE Int. Conf. Radar-77*, p. 72; Hovanessian (1988), pp. 260–267; Antony (1995); Hall (1992); Waltz (1990); Tech Reach (1996).

data quantization (see **QUANTIZATION**).

Data recording is the process of registration and storage of information acquired during radar operation. From the point of view of performance the recording systems can be classified as real-time and non-real-time systems, and from the point of view of practical implementation as analog and digital. The main analog recording medium that is still widely used is frequency modulation (FM) recording. Modern recording techniques are primarily digital, and the main digital recording media are 9-track digital tape, very large data store (VLDS), high-density digital recording (HDDR), magnetic disks (fixed and removable), and optical disks.

SAL

Ref.: Currie (1989), pp. 189–201.

data sampling (see **SAMPLING**).

data smoothing (see **SMOOTHING**).

Data transmission permits data to be combined and processed at points other than the radar site. In addition to radar data, the transmission channel is used to transmit data necessary for the effective operation of the radar system as a whole: control commands, synchronization, and reference signals. The transmission lines may be either digital and analog (e.g., cable, fiber-optic, and radio lines). The composition of the transmitted radar data and the requirements placed on the transmission network are determined by the extent to which data must be combined within the radar system.

The transmission of individual reports and trajectory data is distinguished from the transmission of video and radio frequency signals obtained from the various radar receivers. The former may usually be transmitted over common telephone channels (cable and radio-relay links) using modems, while transmission of video and radio signals requires wideband channels.

An important problem is that of transmitting synchronization signals with the accuracy required for multistatic radar systems, for which special navigational systems and intersite communications links are used, including those based on the reception of a direct signal from the transmitting position. The availability of accurate timing and frequency references through the Global Positioning System (GPS) offers a new set of solutions to these problems.

IAM

Ref.: Barton (1964), pp. 413–415; Salah, J. E., and Morriello, J. E., *IEEE Int. Conf. Radar-80*, p. 88; Retzer, G., *IEEE Int. Conf. Radar-80*, p., 288.

DECEPTION, radar. Radar deception is the term used as an abridged variant of deception electronic countermeasures (see **ELECTRONIC COUNTERMEASURES**). Typically, angle, range, and velocity deception are distinguished (see **JAMMING, deception**). Other categories of deception are manipulative (the introduction of radiations into enemy channel that imitates its own emissions) and imitative (the alteration and simulation of friendly electromagnetic radiations to accomplish deception). A radar deceiver is ECM equipment that attempts to deceive or mislead radar by emitting a pulse-like signal similar to the radar signal.

SAL

Ref.: Johnston (1979), pp. 57, 61, 63, 65; Schleher (1986), pp. 9, 138.

DECIBEL. The decibel (dB) is one-tenth of a bel. The number of decibels denotes the ratio of the two power or voltage levels:

$$\mathrm{dB} = 10\log\frac{P}{P_0} = 20\log\frac{U}{U_0}$$

where P is a power level to be related to the reference level P_0, and U is a voltage level to be related to the reference level

U_0. When the gain or loss of a network or device is given, the reference power P_0 is the input power and P is the output. When P_0 is taken as 1W, P is expressed in dBW (decibels relative to 1W) and when P_0 is 1 mW, P is expressed in dBm.

In radar applications, decibel notation is used to describe not only power but other quantities that are proportional to power (e.g., radar cross section of the target in dBm^2). Quantities in decibel notation are often used to describe signal-to-noise ratio, losses, sidelobe levels, and so forth.

A **decibel conversion chart** is used to translate power (P/P_0) and voltage (U/U_0) ratios from decibels to dimensionless units and vice versa (Table D1). With the widespread use of pocket calculators with scientific notation, reference to such charts is not often required.

Example: 3 + 0.7 = 3.7 dB = 1.995262 × 1.174898 = 2.344299 as a power ratio, or 1.412538 × 1.083927 = 1.531088 as a voltage ratio. *SAL*

Ref.: IEEE (1993), p. 315; Leonov (1998), p. 175.

Table D1
Decibel to Ratio Conversion

dB	$X_a = P/P_0$	$X_a = U/U_0$
0	1.000000	1.000000
0.1	1.023293	1.011579
0.2	1.047129	1.023293
0.3	1.071519	1.035142
0.4	1.096478	1.047129
0.5	1.122018	1.059254
0.6	1.148154	1.071519
0.7	1.174898	1.083927
0.8	1.202264	1.096478
0.9	1.230269	1.109175
1.0	1.258925	1.122018
2.0	1.584893	1.258925
3.0	1.995262	1.412538
4.0	2.511886	1.584893
5.0	3.162278	1.778279
6.0	3.981072	1.995262
7.0	5.011872	2.238721
8.0	6.309573	2.511886
9.0	7.943282	2.818383
10	10	3.162278
20	10^2	10.00000
30	10^3	31.622780
40	10^4	10^2
50	10^5	316.2278
60	10^6	10^3
70	10^7	3162.278
80	10^8	10^4
90	10^9	31622.78
100	10^{10}	10^5

DECODER. A decoder is a device for decoding a series of coded signals and transferring them into the initial form of representation. In radar applications decoders are usually used in secondary radars receiving the signals from transponders and in radars employing phase-coded signals. *SAL*

Ref.: IEEE (1993), p. 317.

DECORRELATION is the reduction of correlation that is usually expressed in a reduction of correlation coefficient during the (de)correlation time.

DECOY, radar. A radar decoy is a device used to divert or mislead radars and to prevent them from performing their functions with required performance. It may take the form of a radiating (active) or reflecting (passive) decoy and is considered a form of electronic countermeasure. *SAL*

An **active decoy** is a decoy operating at locations separate from real targets, and providing beacon signals to suppress the real target echoes. Typically it is the miniaturized *repeater jammer* deployed on board or towed by military vehicles (airborne platforms, naval platforms, etc.). *SAL*

Ref.: Barton (1991), p. 12.7; Neri (1991), p. 407.

An **expendable decoy** is a minimissile-shaped object that is ejected from a military vehicle and whose payload generates deceptive signals. To increase the effectiveness, systems based on forward-fired active decoys have been developed. The main problem in using these decoys is the timing of ejection. Before launching the decoy, one should be sure that the sensor to be confused, for example a missile seeker, is already approaching and is at the right range. In the absence of this information, expendable decoys should be launched at regular intervals when the threat is detected. That is not always possible because of a limited supply of such decoys. *SAL*

Ref.: Neri (1991), p. 410; Chrzanowski (1990), p. 156.

A **passive decoy** is a nonradiating, reflecting object that is typically used to increase the number of apparent targets seen by radar. The most common types of passive decoys are radar reflectors (e.g., corner reflectors and Luneburg lens reflectors) and special false targets (e.g., reentry vehicles), typically used to mask the warhead and to mislead an ABM radar. *SAL*

Ref.: Barton (1991), p. 12.9; Neri (1991), p. 399.

A **towed decoy** is a decoy that is tied to the military platform (airborne, spaceborne, naval, etc.). Unless it includes an active deception repeater, it produces an echo at the same doppler as the true target. Used against attacks from the front and rear sectors, it may fail to provide a satisfactory defence. *SAL*

Ref.: Neri (1991), p. 408.

A **DEFRUITER** is the equipment that deletes random nonsynchronous unintentional returns in a beacon system. (See **FRUIT**.)

Ref.: IEEE (1993), p. 320; Stevens (1988), p. 114.

DELAY

Group [envelope, instantaneous] delay is the delay of the energy passing through a network, associated with a narrow frequency band of signals. The usual notation is t_g. For example, if the spectrum of the pulse going through a dispersive network is relatively wide, the output pulse envelope will differ from that of the input one because the delay of the lower frequency components will be different than the delay of the higher frequency components. The phase delay and group delay are equal only when there is no dispersion (i.e., when the carrier phase delay is linearly related to frequency). *SAL*

Ref.: Wehner (1987), p. 52.

Phase delay is the delay of the signal phase when it passes through a medium or network. For example, an input signal of phase $2\pi f\, t + \psi_i$, where f is the frequency and ψ_i is the input phase, exits the network with a phase of $2\pi f[t - \tau_p(f)] + \psi_i$, where $\tau_p(f)$ is the frequency-dependent phase delay through the network. The phase delay through the network is

$$\tau_p(f) = -\frac{\phi(f)}{2\pi f}$$

where $\phi(f)$ is the frequency-dependent insertion phase in radians, the difference between output and input phase. *SAL*

Ref.: Wehner (1987), p. 51.

Delay time is the time t_d between transmission of a radar pulse and reception of the corresponding echo. Target range, R, is estimated as

$$R = \frac{2t_d}{c}$$

where c is the velocity of light. *SAL*

Ref.: IEEE (1993), p. 322; Skolnik (1962), p. 2; Leonov (1988), p. 13.

DELAY LINE. A delay line is a device that provides temporal delay of signals. Such devices are used for temporal selection (gating), pulse measurements, matching the operation of pulse devices, the separation of channels, in systems using pulse compression, and in frequency-scanned arrays. The basic parameters of a delay line are its time delay, passband, electromagnetic transmission coefficient (a loss for a passive delay line, or gain for an active delay line), and the spurious signal level. Delay lines may be categorized as analog or digital, the former being either electrical or acoustic. The greatest time delay is achieved with ultrasonic delay lines (up to 100 ms), and the least with coaxial cable delay lines (up to 1 µs). The middle ground is occupied by electrical delay lines using the charge migration effect (up to 60 µs).

A particularly descriptive characteristic of a delay line is the product of its time delay and bandwidth. The highest product, up to 25×10^4, is found in beam-type ultrasonic delay lines, manufactured from piezoelectric monocrystals. The main characteristics of the most common delay lines are given in Table D2. *IAM*

Ref.: IEEE (1993), p. 323; Skolnik (1970), pp. 20.3–20.15; Sloka (1970), p. 178; Fink (1975), p. 3-31; Brookner (1977), pp. 182, 185–186, 191–195, 384; Lukoshkin (1983), p. 217.

An **acoustic-wave delay line** is based on the excitation and propagation of acoustic waves within elastic materials. There are surface-wave delay lines and volume-wave delay lines. Surface acoustic waves propagate in a relatively narrow surface layer of elastic solids and are accompanied by the periodic motion of the nodes of the crystal array, either in the direction of wave propagation or perpendicular to it. Volume waves propagate throughout all a thicker body.

Table D2
Characteristics of Delay Lines

	Bandwidth (MHz)	Delay time (µs)	Typical operating frequency (MHz)	Typical loss (dB)	Typical spurious level (dB)
Electrical					
All-pass time delay network	40	1,000	25	25	−40
Folded-type meander	1,000	1.5	2,000	25	−40
Waveguide	1,000	3	5,000	60	−25
Acoustic wave					
Strip:					
aluminum	1	500	5	15	−60
steel	20	350	45	70	−55
Diffraction grating:					
perpendicular	40	75	100	30	−45
wedge	250	65	500	50	−50
Acoustic surface wave	40	50	100	70	−50
Yttrium-iron-garnet (YIG)	1,000	10	2,000	70	−20

The most common delay lines use hypersonic oscillations, up to 10^{10} Hz (see **ultrasonic delay line**, **surface-acoustic-wave delay line**). The main advantage of acoustic-wave delay lines is that longer delays can be achieved than with electrical lines of comparable size, as the wave propagates at sonic speeds. The main disadvantage is that transducers are required to convert the waves, introducing insertion losses. *IAM*

Ref.: Skolnik (1970), p. 20.6; Gassanov (1988), p. 213.

An **all-pass [time] delay network** is an electrical delay line, typically a four-terminal lattice network. Ideally it provides constant gain over its operating band, and the phase shift varies with the square of frequency to provide a constant delay slope. Bridged networks are often used in place of lattice networks for more convenient implementation. Several networks can be cascaded to increase differential delay. The network can be used in linear frequency-modulated waveform generation. *SAL*

Ref.: Skolnik (1970), p. 20.9.

An **analog delay line** uses continuous energy transfer processes. They are implemented with devices in which electromagnetic waves (electromagnetic delay lines), acoustic waves (ultrasonic delay lines), and spin waves (YIG delay lines) propagate. In addition to analog delay lines, wide use is made of discrete-analog delay lines, which rely on the discrete nature of charge movement in charge-coupled devices.

Analog and discrete-analog delay lines are used to process analog waveforms. The main disadvantage of analog as compared with digital delay lines is that their performance is not stable and repeatable in time, depending on environmental factors, and they are bulkier and less reliable. *IAM*

Ref.: Skolnik (1962), p. 119; Sloka (1970), p. 178; Gassanov (1988), p. 227.

A **beam-type delay line** is an ultrasonic delay line in which ordinary divergent propagation of energy takes place in an elastic wave. The trajectory of the ultrasonic beam may be a straight line or a segmented (broken) line, the latter due to multiple reflections from the surface of the device, which increase the time delay. A device that supports such multiple reflections is referred to as a *multiple-entry delay line*, and the waves are *volume waves* (as opposed to surface waves). To minimize the diffraction losses due to beam divergence, the area of the transducer is much less than the square of the wavelength of the acoustic wave, and the acoustic waveguide is constructed from metals or monocrystals in which the acoustic waves travel slowly.

Ultrasonic beam-type delay lines are made from the following materials: sapphire (Al_2O_3), rutile (TiO_2), lithium niobate ($LiNbO_3$), iron-sodium garnet (ISG), and aluminum-sodium garnet (ASG). Transducers are constructed from a film of piezoelectric semiconductor, such as cadmium sulfide (CdS).

The maximum time delay in beam-type delay lines is 5×10^3 μs (a monocrystal of Kbr in the form of a polyhedron), and the maximum bandwidth is 100% (a rod of fused quartz SiO_2 with a diffused cadmium sulfide transducer, at a frequency of 50 MHz). A typical value for the total loss is 50 to 60 dB (70 dB for frequencies above 1 GHz). *IAM*

Ref.: Skolnik (1970), p. 20.6; Sloka (1970), p. 186.

A **charge-coupled delay line** uses a charge-coupled device as the fundamental component. Such a delay line usually comprises several (from 2 to 8) multielement registers that have surface channels, a common timing bus, and a master timing generator.

The advantage of a charge-coupled delay line is its complete compatibility with integrated semiconductor microcircuits. These devices provide a delay up to 60 μs, a voltage gain between 1.0 and 2.0, and a harmonic coefficient of 3 to 4%. The passband extends from dc to half the repetition frequency of the timing pulses and does not exceed 10 MHz. *IAM*

Ref.: Gassanov (1988), p. 227.

A **coaxial cable delay line** uses a coaxial transmission line. The properties of the delay line are completely determined by the length of the cable. The propagation speed of the electromagnetic waves over the coaxial cable is two-thirds of its propagation speed in a vacuum, so that the linear delay is 0.0005 μs/m. Therefore, in practice it requires that great lengths be used.

Typical losses in the line are on the order of 40 dB. Coaxial cable delay lines are distinguished by a large bandwidth-duration product: 50 to 500 for a conventional cable and 100 to 1,000 for a superconducting cable, which limits its basic use to the processing of broadband signals of short duration. *IAM*

Ref.: Lukoshkin (1983), p. 217.

A **diffraction-grating delay line** is an acoustic delay line using a diffraction grating to achieve the desired transfer function. The gratings direct the sound wave across the different paths and so different frequencies acquire different delays. The grating spacing may be varied so that different portions of the grating are resonant to different frequencies. The general shape of the wedge and perpendicular diffraction grating configuration is shown in Fig. D1. *SAL*

Ref.: Brookner (1977), p. 138.

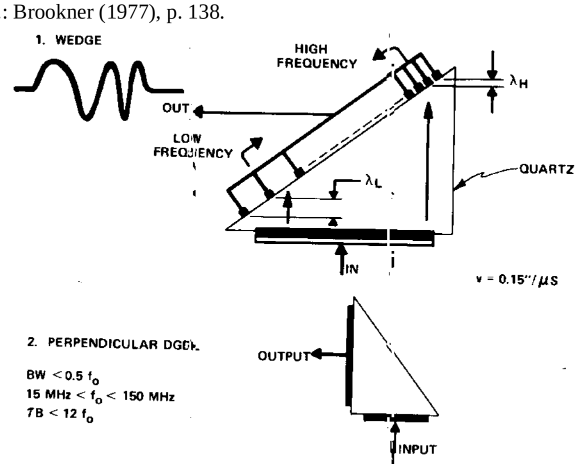


Figure D1 Diffraction-grating delay line (from Brookner, 1970, Fig. 30, p. 138).

A **digital delay line** is implemented in digital circuitry and is used in the processing of digital waveforms. Arbitrarily long delays may be achieved with digital delay lines, subject only to the limitations imposed by the cost of the associated circuitry. The minimum delay is determined by the speed of the digital components.

The digital delay line is the fundamental element of digital filters, which are widely used in digital signal processing. Digital delay lines have considerable advantage over analog lines, principally in flexibility, stability, repeatability, and reliability of performance. *IAM*

Ref.: Kuz'min (1986), p. 44.

A **dispersive delay line** has a frequency-dependent delay. The dispersion can be natural or artificial. Natural dispersion exists in ultrasonic delay lines and yttrium-iron-garnet (YIG) delay lines. Delay lines with artificial dispersion are surface-wave delay lines and charge-coupled delay lines. *IAM*

Ref.: Skolnik (1970), p. 20.6; Sloka (1970), p. 190; Shirman (1974), p. 92.

An **electrical [electromagnetic] delay line** is implemented with electromagnetic wave transmission lines. They are classified on the basis of the type of transmission line used. The most common types of electrical delay lines are the all-pass time-delay network, the folded-type meander delay line, the coaxial cable delay line, the waveguide delay line, and the charge-coupled delay line. Electrical delay lines have less insertion loss than acoustic (-wave) delay lines, but the delay is less for a given size. *IAM, SAL*

Ref.: Skolnik (1970), p. 20.6; Sloka (1970), p. 180.

electromechanical delay line (See **ultrasonic delay line**).

A **fiber-optic delay line** is a delay line based on a fiber-optic transmission line. The simplest fiber-optic delay line includes a laser with a modulator to apply the electrical signal, a fiber circuit of given length, determining the amount of delay of the signal, and a photodetector to recover the signal. More complex fiber-optic delay lines with variable delay use integrated optical switches for discrete changing of the fiber-optic length. For weighted processing of radar signals, tapped delay lines are widely used. These consist of an optical fiber with taps along its length, or a group of fibers with progressively increasing lengths. Weighting of the optical signals is done either by a frequency shift of the signal in each of the taps, or by the introduction of losses in each of the fibers. The introduction of sections with variable delay into fiber-optic delay lines makes it possible to control the point of connection of the taps in accordance with a set program. Such delay lines are called *transversal filters* and allow rapid switching both of weight functions and tap positions.

The basic application of fiber-optic delay lines is for highly stable precision delay of wideband signals. In comparison with delay lines of other types, fiber-optic delay lines ensure lower transducer losses, lower losses per unit of delay, wider bandwidth, higher signal-to-noise ratio, higher temperature stability, and greater flexibility. *IAM*

Ref.: Montgomery, J. D., and Nixon, F. W., *Microwave J.*, 1985, vol. 28, no. 4; Zmuda (1994), p. 481.

The **folded-tape meander delay line** consists of a thin conducting tape extending back and forth between two ground planes. The space between tape and ground planes is filled with dielectric material (Fig. D2). The time delay per meander is a function of the dimensions of the loop and the distance from the ground plane. This configuration is the microwave analog of the all-pass time-delay network and can be used for generation of frequency-modulated waveforms. *SAL*

Ref.: Skolnik (1970), p. 20.12.

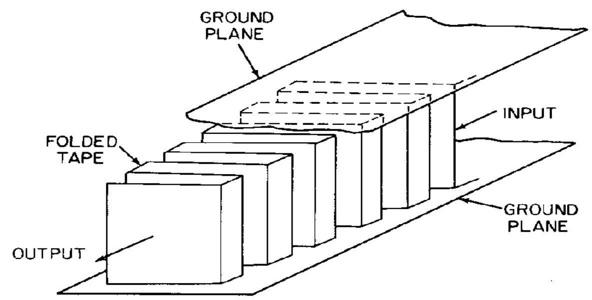


Figure D2 Tapered folded-tape meander line (from Skolnik, 1970, Fig. 10, p. 20.12, reprinted by permission of McGraw-Hill).

A **magnetostriction delay line** is an acoustic delay line that transforms the electrical signal into an acoustic (ultrasonic) wave using the phenomenon of magnetostriction, which is a change in the geometric dimensions of a sample of ferromagnetic material placed in an alternating magnetic field. The acoustic waveguide is usually constructed from a thin iron-nickel alloy.

The magnetostriction delay line is distinguished by loose coupling between the transducer and the acoustic waveguide, which makes them suitable for tapped delay lines. *IAM*

Ref.: Sloka (1970), p. 187.

A **piezoelectric delay line** is based on the propagation of surface acoustic waves along a piezoelectric crystal. They are distinguished by the simplicity of their transducers, which are small metallic electrodes (with dimensions on the order of microns) placed against the surface of the crystal. Due to this feature they are used as tapped delay lines. *IAM*

Ref.: Lukoshkin (1983), p. 218.

A **programmable delay line** is a delay line with electronically controlled coherent delay. The typical approach to build such a device is the tapped delay line and switch-matrix combination (Fig. D3). The main range of usage of programmable delay lines is in ECM techniques to delay in a coherent manner radar pulses for the repeater jamming mode. *SAL*

Ref.: Wiegand (1991), p. 83.

A **quartz delay line** is an ultrasonic delay line, using a quartz (SiO_2) acoustic waveguide. Despite the fact that quartz is not distinguished by low acoustic absorption, large crystals of synthetic quartz may be produced industrially, and are easily processed. Quartz is usually used in volume-wave delay lines, which exploit multiple reflections from the crystal walls. This

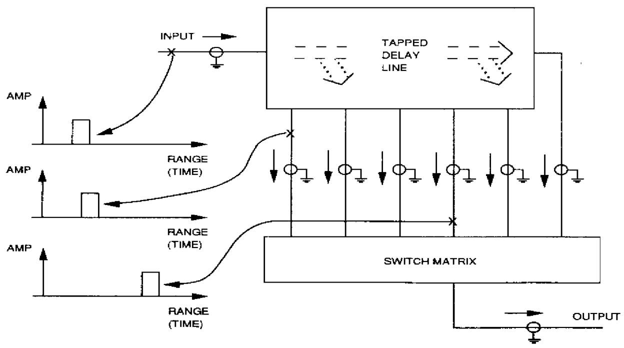


Figure D3 A programmable delay line (from Wiegand, 1991, Fig. 2.20, p. 84).

makes it possible to obtain long delays (1 to 3 ms) at frequencies of 10 to 80 MHz, with a bandwidth on the order of 30%. To reduce the absorption of acoustical energy significantly, monocrystal quartz delay lines may be cooled to the temperature of liquid helium, which also results in highly stable parameters. The total loss in a quartz delay line is 40 to 60 dB. *IAM*

Ref.: Skolnik (1962), p. 120; Lukoshkin (1983), pp. 247, 254.

A **strip delay line** is an ultrasonic delay line made from a long, thin strip of metallic material (typically, aluminum or steel) with transducers at opposite ends (Fig. D4). Aluminum lines operate at lower frequencies than steel, with lower losses. Strip delay lines have linear delay versus frequency characteristics and are used in generation and processing of linear frequency-modulated waveforms. *SAL*

Ref.: Skolnik (1970), p. 20.6; Brookner (1977), p. 138.

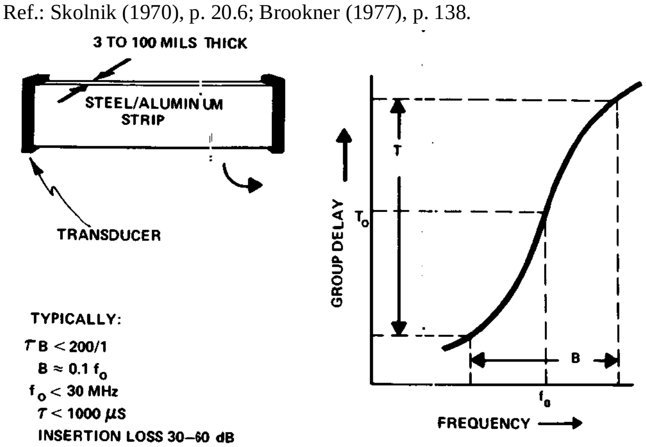


Figure D4 Metallic strip dispersive delay line (from Brookner, 1970, Fig. 29, p. 138).

A **surface-acoustic-wave (SAW) delay line** is an ultrasonic delay line based on the effect of the propagation of acoustic wave in the surface layer of the solid body. Typically, the acoustic-wave guide is made from a piezoelectric monocrystal with electrodes deposited as metal components through photocoating technology. When ac voltage is applied to these electrodes, the acoustic wave arises as the result of piezoelectric effect. When the quantity of electrodes is large, the tapped delay lines can be formed.

The maximum delay time is usually restricted by the substrate length and typically is about 50 µs, and the maximum frequency about 1.5 GHz. Surface-acoustic-wave delay lines are widely used in dispersive filters because of the simplicity in getting the required frequency responses. *IAM*

Ref.: Skolnik (1970), p. 20.6; Morgan (1985), pp. 16, 175.

A **tapped delay line** is one in which required signal-processing function is achieved through proper choice of the location, frequency response, and coupling of the taps. In surface-acoustic-wave delay lines, interdigital transducers are employed as taps. Modern analog tapped delay lines also can be built using charge-coupled devices and fiber-optics. (See **charge-coupled delay line** and **fiber-optic delay line**.)

Tapped delay lines feature a large number of taps offering a choice of delays and are used to process complex signals. One type of tapped delay line is based on the propagation of longitudinal and twisting waves in a wire waveguide, usually with magnetostriction transducers. A tapped delay line may also be based on the propagation of surface waves along a piezoelectric crystal. (See **waveguide delay line**.)

A tapped magnetostriction delay line can provide delays up to 100 ms with relatively small dimensions. An advantage to this type of delay line is the ability to tune the phase and gain of each tap continuously and independently. They are used when filtering narrowband signals, with bandwidth less than 1 MHz and delay up to 1 ms.

Due to the simplicity of the electrodes, a piezoelectric tapped delay line may have up to 10,000 taps. An advantage to this type of delay line is the ability to construct them in an integrated fashion. They are used to process complex signals with bandwidths on the order of several hundred megahertz and delays up to 100 µs.

Tapped electrical delay lines, usually of coaxial cable, are used to process wideband signals of short duration. It is necessary, however, to consider the large size and weight of such devices. *SAL, IAM*

Ref.: Lukoshkin (1983), p. 217; Fink (1977), p. 9.73; Brookner (1977), pp. 163, 172, 182, 186–194, 384.

An **ultrasonic delay line** exploits the low propagation speed of ultrasonic waves along an acoustic waveguide. Such a delay line consists of an acoustic waveguide made from a hard elastic material and two or more transducers that convert radio signals to acoustic waves and vice-versa. The acoustic waveguide may be a piezoelectric crystal, a metal with low ultrasonic propagation speed, or a ferromagnetic sample. The transducers are placed on the surface of the delay line, at a separation corresponding to the desired delay. Piezoelectric and magnetostriction delay lines differ in the way in which the waves are excited, relying on the coupling between mechanical deformation and electric and magnetic fields, respectively.

The most common ultrasonic delay lines are strip delay lines, diffraction-grating delay lines, and yttrium-iron-garnet (YIG) delay lines. The frequency and bandwidth performance of ultrasonic delay lines depend on the characteristics of the transducers used. *SAL, IAM*

Ref.: Skolnik (1970), p. 20.6; Sloka (1970), p. 178.

A **waveguide delay line** is an ultrasonic delay line in which an acoustic wave propagates with virtually no divergence along the driving surface of a mechanical waveguide. In waveguide delay lines the acoustic wave always propagates in a straight line. The most common types of waveguide delay lines are strip and wire delay lines. Strip delay lines usually use shear waves, propagating along the line with multiple reflections from the driving (wide) surfaces. Piezoelectric finger transducers are attached to the facing surfaces of a strip delay line. The most suitable materials for the delay line itself are magnesium-aluminum, aluminum and iron-nickel alloys. The delay may reach 5 ms, the bandwidth 30% at a frequency of 1 MHz, and the typical loss is 6 to 16 dB.

In wire delay lines, ring- or rod-shaped transducers are used to excite twisting or longitudinal waves. The wire is usually of an aluminum or iron-nickel alloy. Wire delay lines are used to obtain long delays of between 10 and 100 ms, providing a bandwidth of 10 to 20% for signal frequencies in the range 0.5 to 1 MHz, and exhibit typical losses in the range 30 to 50 dB. *IAM*

Ref.: Skolnik (1970), p. 20.7; Sloka (1970), p. 184; Lukoshkin (1983), p. 247.

An **yttrium-iron-garnet (YIG) delay line** is manufactured from crystals of yttrium iron garnet ($3Y_2O_3{\cdot}5Fe_2O_3$). It is used to create delays of long duration, relying on multiple reflections (see **beam-type delay line**), and also as delay lines with artificial dispersion, which arises in the presence of an external magnetic field (Fig. D5). The latter are used for

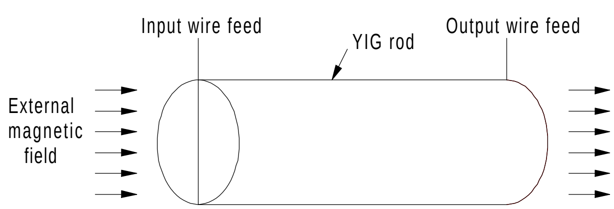


Figure D5 YIG delay line (after Skolnik, 1970, Fig. 6, p. 20.10).

matched filtering of frequency-modulated waveforms. YIG delay lines do not produce linear delay versus frequency, but their delay characteristics are very repeatable. *IAM*

Ref.:Skolnik (1970), p. 20.9; Shirman (1974), p. 150; Sloka (1970), p. 186.

DELTA [DIRAC] Function. The Dirac delta function is an impulse of infinitesimal duration $\delta(t)$ that has infinite amplitude at time $t = t_0$ and unity area:

$$\delta(t-t_0) = \begin{cases} \infty, & t = t_0 \\ 0, & t \neq t_0 \end{cases}$$

The concept of the delta function is widely used in the theory

$$\int_{-\infty}^{\infty} \delta(t-t_0) = 1$$

of radar signal analysis and is a basic model for the correlation function of completely random noise. The random process that is uncorrelated at any two moments of time is often termed *delta-correlated*. *SAL*

Ref.: Levin (1974), p. 522; Urkowitz (1983), p. 6.

The **DEMATRON** is a trade name introduced by Litton Industries for distributed-emission, linear-format, forward-wave interaction crossed-field amplifier. The dematron (distributed-emission magnetron amplifier) is a microwave amplifier in which the slow-wave circuit and electron current are open circuits, and the emitting surface of the cathode is distributed along the interaction space. This results in a very high-power magnetron amplifier. The dematron is distinguished by the absence of a controlling electrode, so it may be operated with very complex pulsed modulation.

The basic characteristics of a dematron are its pulse power (up to 1 MW at a duty factor of 0.005), its normalized bandwidth (up to 10%), gain (up to 15 dB), and efficiency (typically near 30%).

A dematron has a comparatively simple construction, in the form of a long, thin housing with a relatively small cross section. This makes it well suited to close packing within an active phased-array antenna. *IAM*

Ref.: Skolnik (1970), pp. 7.21, 7.6; Ewell (1981), p. 37.

DEMODULATION is the process inverse to modulation and has as its purpose restoration of the modulating wave. A demodulator is the device to effect the process of demodulation. (See **DETECTOR.**)

Ref.: IEEE (1993), p. 326.

DEMULTIPLEXING is the separation of individual signals from a multiplexed channel. (See **MULTIPLEXING**.)

DEPOLARIZATION, DEPOLARIZER. Depolarization is the phenomenon when the polarization of an electromagnetic wave is changed with respect to that transmitted. A depolarizer is an element that causes depolarization. In radar applications the main sources of depolarization are the propagation medium and reflection from radar targets or underlying surfaces (terrain or sea). A natural low-frequency depolarizer is the ionosphere, due to the Faraday and Cotton-Mouton effects. Significant depolarization may also be caused by complex targets, the earth's surface, or vegetation. The only reflectors that do not distort polarization are spheres, conducting plates that are large relative to the wavelength, and bodies having radius of curvature much greater than the wavelength. Regardless of transmitted polarization, the reflected signal from a real target is elliptically polarized, and parameters of the ellipse varies randomly. In this case, if the radar antenna receives only one polarization, there is a loss caused by its inability to receive the cross-polarized signal component.

The fact that bodies of different shape introduce differing degrees of depolarization can be exploited to reduce unwanted reflections relative to desired targets. The primary example of this is the use of circular polarization to discriminate against rain clutter. Because raindrops are nearly spherical, they reflect circularly polarized waves with a sense of rotation opposite to that incident on them. The spherical symmetry of rain is not perfect, and the antenna does not radiate and receive perfectly circular polarizations, but the rain return is reduced by the integrated cancellation ratio (over the two-way pattern of the antenna in angle space). A complex target such as an aircraft introduces significant depolarization, giving nearly equal power in the two senses of rotation. The loss in target power relative to use of a linearly polarized signal is approximately 3 dB. *SAL, IAM*

Ref.: IEEE (1993), p. 328; Skolnik (1980), p. 504; Vasin (1977), p. 93.

DEPTH OF FOCUS is "the range distance over which the cross-range dimensions of the impulse response at a fixed focal range is essentially constant. Applies principally to synthetic aperture radar."

Ref.: IEEE (1990), p. 10.

DESIGNATION is "a selection of a particular target and transmission of its approximate coordinates from some external source to a radar, usually to initiate tracking."

Ref.: IEEE (1990), p. 10.

Jammer strobe designation is the designation in angle of the jamming vehicle as a target. Then a collocated tracker can scan at the angle (or fix its beam up the strobe, for a designation by a 3D search radar). Jammer designation from a network of search radars, using triangulation, provides range and azimuth data to the tracker, permitting decisions on the degree of threat and whether it is within engagement range. *DKB, SAL*

Ref.: Barton (1991), p. 9.18.

Optical designation refers to the designation of targets by a collocated optical instrument, eliminating the need for acquisition scan by antenna. The acquisition range of the radar is then found from the basic radar equation, setting the observation time equal to the time in which a given probability of acquisition is to be obtained. *DKB, SAL*

Ref.: Barton (1991), p. 9.18.

Search radar designation is the designation by a collocated or remotely located search radar. If this is a 2D radar, then it can provide range and azimuth designation data to the tracker, but target elevation (and usually radial velocity) will be unknown. The tracker acquisition scan will cover the elevation sector on which targets might be located, and some azimuth scanning may also be required to cover the 3σ error of the search radar data. For 3D radar, the designation includes target elevation or height. This reduces the sector that the tracker acquisition scan must cover. *DKB, SAL*

Ref.: Barton (1991), p. 9.18.

DETECTABILITY FACTOR is the ratio of input signal energy to noise spectral density required to achieve a given detection performance. For pulsed radar it is defined as "the ratio of single-pulse energy to noise power per unit bandwidth that provides stated probabilities of detection and false alarm, measured in the IF amplifier." It is used in the radar range equation as the minimum, or required, signal-to-noise ratio for calculation of maximum range for the given detection performance. The conventional symbol is D. Depending on whether beamshape and signal-processing losses are included in D or as separate terms, different detailed definitions apply, as given below (see also **DETECTION curves**). *DKB*

Ref.: IEEE (1993), p. 336.

The **basic detectability factor** is the signal-to-noise power ratio given by detection theory for:

(1) A steady target signal, of which n samples (pulses) are available for integration, $D_0(n)$.

(2) Swerling Case 1, 2, 3, or 4 target signals, where $D_1(n)$, ... , $D_4(n)$ represent the required average SNR.

(3) A generalized fluctuating target signal of which n_e independent amplitudes are observed within the n samples integrated, $D_e(n,n_e)$.

This basic detectability factor is the value required at the input to the envelope detector, and it represents the input energy ratio for samples received with full antenna gains in a receiver matched to the single sample (or pulse), assuming optimum video integration with no other signal-processing losses. It can be expressed, in the general case, in terms of the basic steady-target detectability factor for a single sample, $D_0(1)$ as

$$D_e(n, n_e) = \frac{D_0(1) L_i(n) L_f(n_e)}{n}$$

where $L_i(n)$ is the video integration loss and $L_f(n_e)$ is the target fluctuation loss when n_e independent target samples are integrated. *DKB*

Ref.: Barton (1993), p. 15.

The **clutter detectability factor** D_{xc} is the counterpart of the detectability factor D_x when the interfering background is dominated by clutter. If the clutter probability density function has a Rayleigh distribution, and the correlation time is less than the pulse repetition frequency, $t_c < t_r$, then $D_{xc} = D_x$. Clutter which is correlated between pulses, $t_c > t_r$, leads to $D_{xc} > D_x$, because the integration gain will be reduced. Non-Rayleigh clutter requires a higher threshold to control false alarms, resulting in $D_{xc} > D_x$ (see **LOSS, clutter distribution**). The loss may be from a few decibels to as much as 15 or 20 dB in the case of land clutter observed at low grazing angles. *DKB*

Ref.: Barton (1988), pp. 61–85.

The **coherent detectability factor** is the input signal-to-noise energy ratio required on a steady target when ideal coherent detection is used: a synchronous detector with its reference oscillator input matched to the known phase of the signal, which is received with full antenna gains and a matched-filter receiver. This detectability factor can be calculated exactly, using the inverse of the integral of the normal distribution:

$$D_c(n) = \frac{1}{2n}[Q^{-1}(P_{fa}) - Q^{-1}(P_d)]$$

where the inverse function Q^{-1} is defined by

$$Q(E) = \frac{1}{\sqrt{2\pi}}\int_E^\infty \exp\left(-\frac{v^2}{2}\right) dv = P \quad , \quad Q^{-1}(P) = E$$

where P_d and P_{fa} are probabilities of detection and false alarm. *DKB*

Ref.: Barton (1988), p. 63.

The **effective detectability factor** D_x is the actual energy ratio required of the central sample (pulse) at the receiver input, given a matching factor M a beamshape loss L_p, and a miscellaneous signal-processing loss, L_x:

$$D_x(n, n_e) = D_e(n, n_e)ML_pL_x$$

Since this factor includes all terms that depend on the probability of detection, its use permits all other terms in the radar range equation to be entered as constant parameters. *DKB*

Ref.: Barton (1988), pp. 19–20.

DETECTION [of radar targets]. In radio applications, the term *detection* has two distinct meanings. The first meaning is sensing the presence of electromagnetic fields, or in radar applications sensing the presence of a radar target. This is termed *radar detection* or *target detection* and is covered immediately below. The second meaning is *demodulation* of signals, to be covered after target detection under the heading of **DETECTOR**.

Target detection is the process of reaching a decision on whether a target is present in the specified volume of space. The decision is made between two mutually exclusive conditions:

A_1 = target is present; A_0 = target is absent.

These conditions are unknown at the moment the decision is to be made. Because interference is present in the receiver, along with the signal, either input condition may lead to either of two decisions:

B_1 = target is present; B_0 = target is absent.

As a result there are four possible results:

B_1A_1 = true detection; B_0A_1 = target missed;

B_1A_0 = false alarm; B_0A_0 = true nondetection.

The sum of the probabilities of these four results is equal to unity:

$$P(B_1A_1) + P(B_0A_1) + P(B_1A_0) + (B_0A_0) = 1$$

The values describing the quality of detection performance under conditions of target presence (A_1) are the (true) detection probability:

$$P_d = P(B_1|A_1) = \frac{P(B_1A_1)}{P(A_1)}$$

and the target missed probability:

$$P_{tm} = P(B_0|A_1) = \frac{P(B_0A_1)}{P(A_1)}$$

Under conditions of target absence (A_0) are the false alarm probability:

$$P_{fa} = P(B_1|A_0) = \frac{P(B_1A_0)}{P(A_0)}$$

and the true nondetection probability:

$$P_{tn} = P(B_0|A_0) = \frac{P(B_0A_0)}{P(A_0)}$$

from which

$$P_d + P_{tm} = 1; P_{fa} + P_{tn} = 1$$

The basic values used in practical applications are detection probability and false-alarm probability. The decision as to target presence is made based on analysis of the received radar return to determine whether the signal plus interference or only interference are present. This is done by comparing the intensity (amplitude) of the received return with a threshold level and making the decision based on the chosen detection criterion. Since interference is generally a random function of time, the process of radar detection is a statistical procedure using the mathematical approaches of probability theory. The most used theory is developed for detection of weak signals in a background of white noise with a Gaussian distribution. Many recent works have been devoted to algorithms for radar detection in other types of interference (colored noise, nonstationary clutter, etc.), but the resulting algorithms for nonstationary, non-Gaussian interference are largely empirical rather than theoretical.

Radar circuits performing the process of target detection are called radar detectors. Radar detectors fall into several basic categories, depending on the fundamental principles underlying the detection process, as shown in Table D3.

Table D3
Types of Radar Detection

Basis	Options
Number of hypotheses tested	Binary or M-ary [multialternative]
Structure of algorithms	Optimum or Nonoptimum (quasioptimum)
Principle of integration	Coherent or Noncoherent
Threshold	Automatic or Visual
Knowledge of statistics	Parametric (distribution dependent) or Nonparametric (distribution-free)

Typically, modern radar detection systems are binary, quasioptimum, automatic, distribution-free devices using either or both coherent and noncoherent integration. M-out-of-n detection (binary integration) is widely used. *SAL*

Ref.: IEEE (1993), p. 337; Helstrom (1968); Whalen (1971); Meyer (1973); Blake (1980), Ch. 2; DiFranco (1968); Skolnik (1980), Ch. 10; Bakut (1984); Barton (1988), Ch. 2; Barkat (1991); Haykin (1992); Morchin (1993), Ch. 5; Shirman (1970), Ch. 3; Kolosov (1989).

Accumulator detection uses an accumulator (an adder with a memory) containing a delay element and amplification logic (Fig. D6). The amplification logic assigns a weight α to each signal crossing the first threshold T_1, and a weight β to each failed detection. To make a decision about the presence of the target, the instantaneous sum at the accumulator output is compared with a second threshold T_2. *SAL*

Ref.: Neri (1991), p. 119.

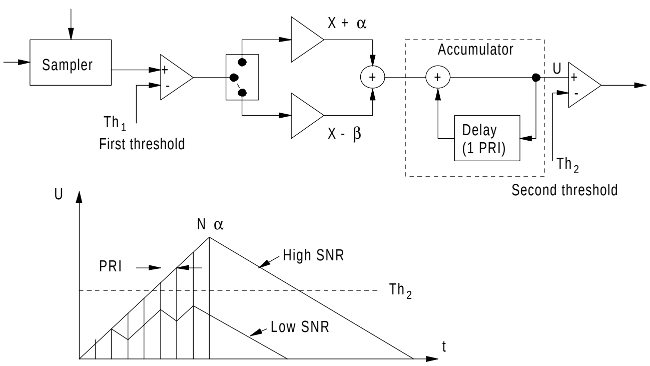


Figure D6 Accumulator detection (after Neri, 1991, Fig. 2.55, p. 120).

In **adaptive detection** the parameters and even the structure of a detector can change depending on the signal and interference levels, to maintain some performance characteristics at a specified level or to respond in a specified way to variation in input data. An example of an adaptive detector is the CFAR receiver, when such a parameter as the probability of false alarm is maintained at the constant level depending on interference environment characteristics. The block diagram of the simplest adaptive detector is cited at Fig. D7. This detector

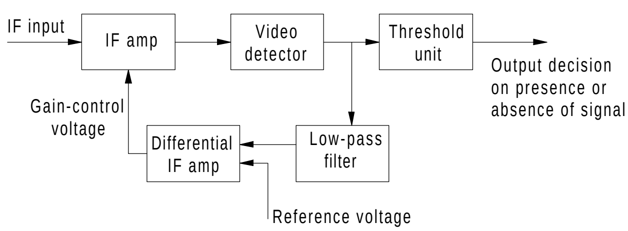


Figure D7 Simplified adaptive detection.

can adapt to slow fluctuations of noise level by means of an IF amplifier with an automatic gain control circuit. The abrupt increase in level because of the signal presence, makes the threshold unit work before the AGC circuit will make up for the level change. More complicated self-adjustable adaptive detectors combined with the adaptation of surveillance period and waveform structures are typically used in phased array radars. *AIL*

Ref.: Skolnik (1970), pp. 15–29; Shirman (1981), pp. 313–316; Sosulin (1978), p. 292-306.

In ***a posteriori*** **detection** a decision is made on the basis that the estimated amplitude will be larger than some minimum value chosen for reliable detection. This approach is closely related to maximum likelihood detection. *SAL*

Ref.: DiFranco (1980), pp. 229–233.

In **automatic detection** the decision about target presence is done by electronic circuitry without the involvement of a radar operator. The basic aspects of automatic detection include:

(1) Integration of received pulses when more than one is expected to be received.

(2) Decision-making as to target presence.

Frequently the means of maintaining a constant false-alarm rate (CFAR) are also included in automatic detection circuitry, in which case the detector is called a CFAR detector. The functions of target detection are often combined with measurement of target location in range and angular coordinates (e.g., azimuth). In this case the automatic detector is called a *plot extractor* or *data extractor. SAL, AIL*

Ref.: Skolnik (1970), Ch. 15, (1980), p. 388; Kuz'min (1974), pp. 5–10; Schleher (1980).

Automatic detection and tracking is the automatization of the target detection and tracking process when the functions of target detection, track initiation, track association, track update, track smoothing, and track termination are performed in an automatic mode. *SAL*

Ref.: Skolnik (1980), p.183.

The **Bayes criterion of detection** is the optimum detection criterion producing the minimum average risk when the decision as to target detection is made under the conditions of known *a priori* probability of a signal presence.

To apply the criterion, one has to know *a priori* probability P_a of the fact that a signal $u(t)$ is present in the received realization $y(t)$, and the average cost for risk r has to be assigned. If the cost of a false alarm is r_{fa}, and the cost of missing the target is r_{tm}, then in general, when many detections are performed during radar operation, the average cost of false alarms is $r_{fa}q_aP_a$, and the average cost of missing targets is $r_{tm}P_aP_{tm}$, where, $q_a = 1 - P_a$, and P_{fa}, P_{tm} are the probabilities of false alarm and missing the target, respectively. The best detector under this criterion will be that which requires minimum average risk for both erroneous decisions:

$$r = r_{fa}q_aP_{fa} + r_{tm}P_aP_{tm} = \text{min.}$$

The intelligent choice of the coefficients r_{fa} and r_{tm} is not a trivial task and is based on the assumption of how dangerous and undesirable each kind of error is. The Bayes criterion is generic and many other detection criteria are special cases of it. Another term for this criterion is the *minimum average risk criterion*. *AIL*

Ref.: Barkat (1991), p. 118; Kazarinov (1990), p. 24

In **binary detection** a decision is made between one of two possible outcomes: noise alone or signal plus noise. The decision is made only about the existence of a target, and target parameters such as range, velocity, and others are not obtained without further processing. In a binary detection problem, it is typically considered that the transmitter may send a deterministic signal $s_0(t)$ under the null hypothesis H_0, or deterministic signal $s_1(t)$ under the alternative hypothesis H_1. When the signal is received, it is corrupted with an additive noise process (white or colored), and the goal is to decide whether hypothesis H_0 or H_1 is true. Such a problem typically is termed *general binary detection*, and if $s_0(t)$ is equal to 0 the problem is called *simple binary detection*. In this sense, binary detection is opposed to multialternative detection, and circuitry implementing this type of detection is termed a *binary detector*. Sometimes that term is used to denote a detector whose input process is binary. *SAL*

Ref.: DiFranco (1980), p. 263; Barkat (1991), pp. 344–370.

Blob detection is the term sometimes applied to radars that measure only the location of target in range and angle (i.e., they recognize a target only as a "blob" located somewhere in space and do not extract any additional information such as the target type or RCS). *SAL*

Ref.: Skolnik (1980), p. 434.

Closed-loop automatic detection is used in an ESM system where information already learned by the detector is deleted from the flow of new pulses but is used to improve the information about detected emitters (Fig. D8). *SAL*

Ref.: Neri (1991), p. 312.

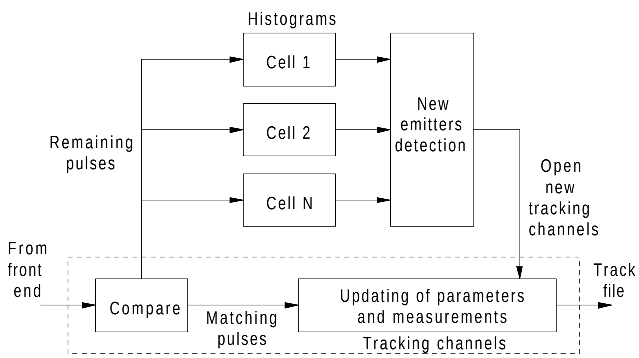


Figure D8 Closed-loop automatic detection (after Neri, 1991, Fig. 4.28, p. 312).

Coherent detection is based on addition of all received pulses before envelope detection [demodulation]. This is the most efficient technique for combining received signal information but is also the most complicated. (See **INTEGRATION, coherent.**) The basic coherent detection configurations are the correlation detector, the filter detector, and the correlation-filter detector. In the correlation detector, the correlation integral is evaluated, typically using two channels (Fig. D9). In this case the output does not depend on the random initial phase of the signal. The signals form the output

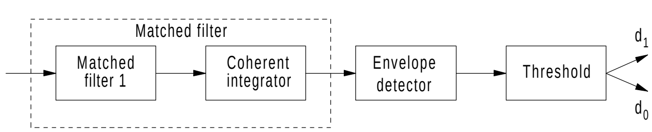


Figure D9 Correlation detector.

of the matched filter are combined with two reference signals offset from one another by 90° in phase detectors. The coherent integrators compute the correlation integrals z_1 and z_2, which are summed before passing to the threshold. The advantage of the correlation detector is that it is invariant to time delay, while the disadvantage is that two channels and complicated integrators using tapped delay lines are required.

The filter detector consists of a filter matched to the single-pulse waveform, a coherent integrator, an envelope detector, and a threshold (Fig. D10).

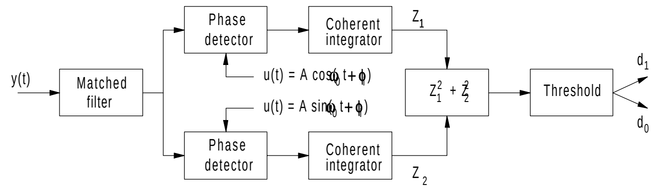


Figure D10 Filter detector.

In the correlation-filter detector, the received and reference signals pass to a multiplier, the output of which passes through a low-pass filter to the envelope detector and threshold (Fig. D11). The advantage of the correlation-filter detector is the absence of the I and Q channels and complicated integrators, but it is sensitive to the time delay of the signal. When this delay is unknown *a priori*, multichannel detection is required.

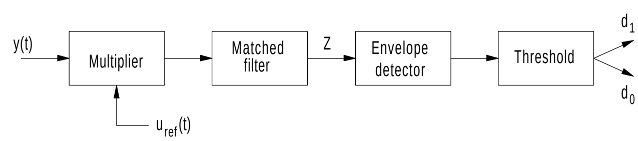


Figure D11 Correlation-filter detector.

Ideal coherent integration requires that the signal phase not change from pulse to pulse during reception of a pulse train. In this case, if there are m pulses in a train, the resultant amplitude of a received signal is m times that for a single pulse, increasing the signal power by m^2. Receiver noise has random amplitude and phase from one pulse to the next, resulting in an average noise power m times that of a single pulse, so the signal-to-noise ratio increases by the factor $m^2/m = m$. It is impossible to achieve ideal coherent integration in practice, even when the radar system is completely stable, because of the phase changes resulting from doppler effect on the target. Since target velocity is unknown *a priori*, channels must be provided matched to all doppler frequencies, leading to the pulsed-doppler radar implementation. If m pulses are integrated, there will be m parallel doppler filters, centered at $\pm f_r/2m$, where f_r is the pulse repetition frequency. The processing is then matched only for doppler frequencies at the center of each filter, and a filter straddling loss is introduced for all other frequencies.

In MTI radar using delay-line cancelers two or more pulses are added coherently with 180° between adjacent pulses, producing an optimum response at doppler frequencies midway between multiples of f_r. While this produces coherent integration gain at those frequencies, the average response for all targets is equal to the noise response, and hence there is no gain in SNR. To produce coherent integration gain for targets of unknown doppler shift, there must be multiple filters covering all or most of the ambiguous doppler interval f_r. Such operation is characteristic of the pulse doppler signal processor, the low-PRF version of which is known as the *moving-target detector* (MTD). When MTI or MTD radars are operated with clutter rejection notches (blind speeds) in the response near-zero doppler, they will inevitably lose targets in that region (and often at ambiguous blind speeds as well), incurring a velocity response loss that can be large when high values of detection probability are required. *SAL*

Ref.: Skolnik (1980), p. 385; Shirman (1970), pp. 100–109; Blake (1980), p. 55; Barton (1988), p. 69; Sosulin (1992), pp. 59–63; Scheer (1993).

Coincidence detection requires *m* or more consecutive pairs, or triples of detections, to occur in *n* consecutive trials, to declare a target present. The application of such a principle to a double-threshold detection system is illustrated in Fig. D12.

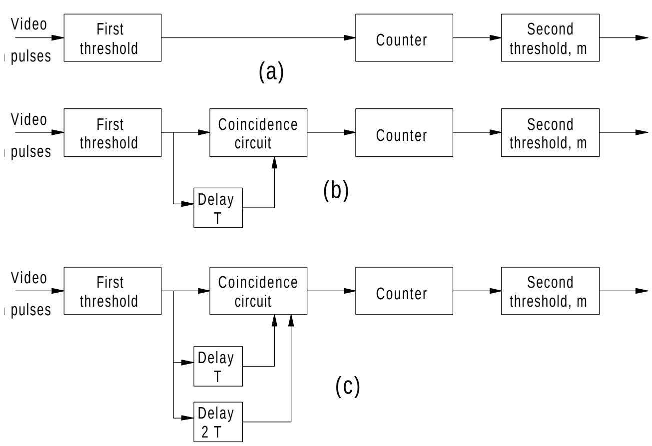


Figure D12 Principle of coincidence detection: (a) no coincidence; (b) double coincidence; (c) triple coincidence (after DiFranco, 1980, Fig. 14.5-1, p. 515).

The first variant in the figure is a conventional double-threshold system without coincidence; the second and the third variants require that two or three consecutive pulses with the same range delay exceed the first threshold before an output can pass to the counter. Coincidence detection has the improved performance in the presence of random-pulse interference.

Sometimes digital moving-window detection is termed *coincidence detection*, and is classified on the basis of the number of bits *m* used for amplitude quantization. For $m = 1$, the detection is called *binomial* or *coincidence detection*; for $m > 1$, it is called *multiple-coincidence* or *multinomial detection*. *SAL*

Ref.: Skolnik (1970), p. 15.14; DiFranco (1980), pp. 515–519.

CFAR detection is the detection process in which a constant-false-alarm rate technique is used. (See **CONSTANT FALSE ALARM RATE**, **automatic detection**.)

Detection criteria are the mathematical rules used to make a decision as to signal presence or absence in the radar return. See **Bayes detection criterion** (minimum average risk), **ideal observer detection criterion**, **likelihood ratio detection criterion**, **minimax detection criterion**, **Neyman-Pearson detection criterion**, **sequential observer detection criterion**, and **weighting detection criterion**. From the point of view of practicality, the Neyman-Pearson and weighting detection criteria are the most convenient. *AIL*

Ref.: DiFranco (1968), Ch. 8; Kazarinov (1990), pp. 24–26.

correlation detection (see **coherent detection**).

Cumulative detection is based on making an independent decision on each of *n* signal samples and is equivalent to *m-out-of-n* detection (binary integration) with $m = 1$. The cumulative probability P_c of detecting the target on at least one of *n* successive scans is

$$P_c = 1 - (1 - P_d)^n$$

where P_d is the single-sample probability of detection. This is the least efficient technique for combining multiple samples, but it may be the only one available when the target moves through more than one radar resolution cell between samples (as in the case of detection over *n* scans of a search radar).

Ref.: Barton (1964), p. 141, (1988), p. 74; DiFranco (1980), pp. 476–494.

Detection curves show the dependencies among signal-to-noise ratio, probability of detection P_d, and probability of false alarm P_{fa} for specified integration and target conditions. They are used to find the detectability factor when P_d and P_{fa} are given (see **detectability factor**). Families of detection curves have been calculated by Blake and are reproduced below. Figures D13 through D18 show the curves for detectability factor for a single pulse on a steady signal, denoted by $D_0(1)$ in the discussion of detectability factor, for the factor $D_0(n)$ as a result of *n*-pulse integration, for the four Swerling models of fluctuating targets, $D_1(n)$, $D_2(n)$, $D_3(n)$, and $D_4(n)$, all for $P_d = 0.9$. *SAL, DKB*

Ref.: Blake (1980), Ch. 2.

Distribution-dependent detection uses prior knowledge of probability density functions of signals and of interference to achieve the required performance. If the statistical parameters of the input coincide with the design assumptions, optimum performance can be obtained (and optimum detection implies use of a distribution-dependent process, sometimes referred to as *parametric detection*). The more the input statistics differ from those assumed in design, the greater will be the degradation in performance. In situations where the input statistics are uncertain (e.g., operation in clutter), this type of detection is unsatisfactory, suffering excessive false alarms and reduced probability of detection. Distribution-free detection is then the preferred approach. An example of distribution-dependent detection is likelihood-ratio detection. *SAL*

Ref.: Skolnik (1970), p. 15.19.

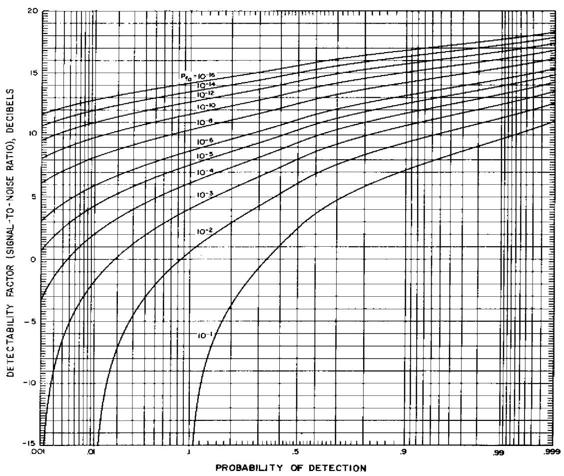


Figure D13 Detectability factor for single-pulse detection on steady signal (from Blake, 1980, Fig. 2-3, p. 39).

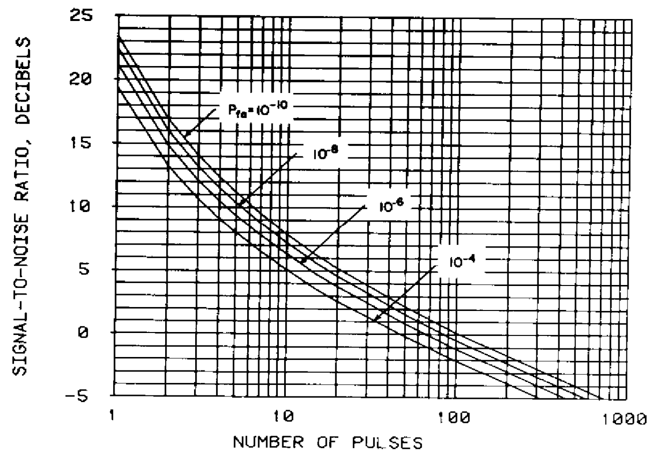


Figure D16 Detectability factor for P_d = 0.9 using n-pulse detection on case 2 fluctuating target (from Blake, 1980, Fig. 2-17, p. 53).

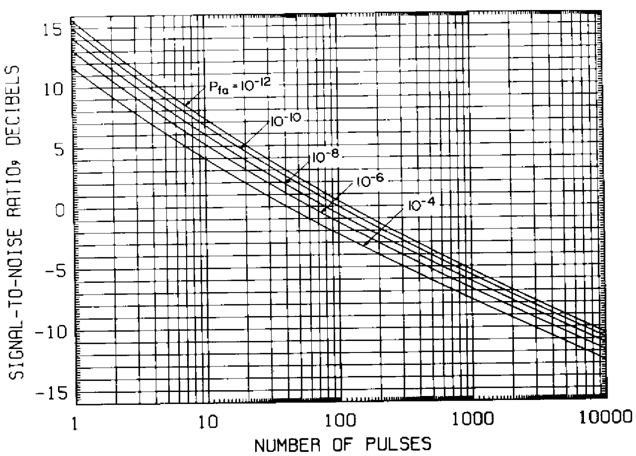


Figure D14 Detectability factor for P_d = 0.9 using n-pulse integration on steady signal (from Blake, 1980, Fig. 2-7, p. 44).

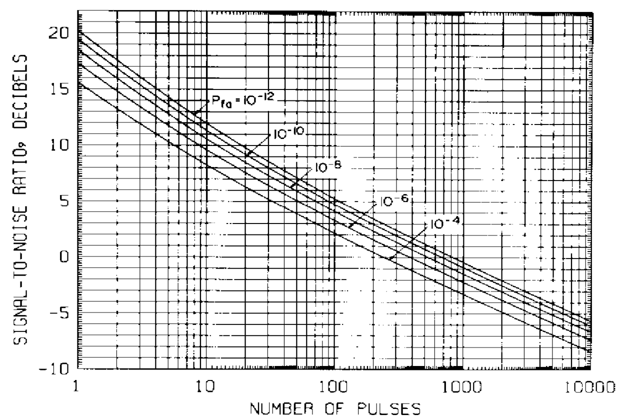


Figure D17 Detectability factor for P_d = 0.9 using n-pulse detection on case 3 fluctuating target (from Blake, 1980, Fig. 2-22, p. 57).

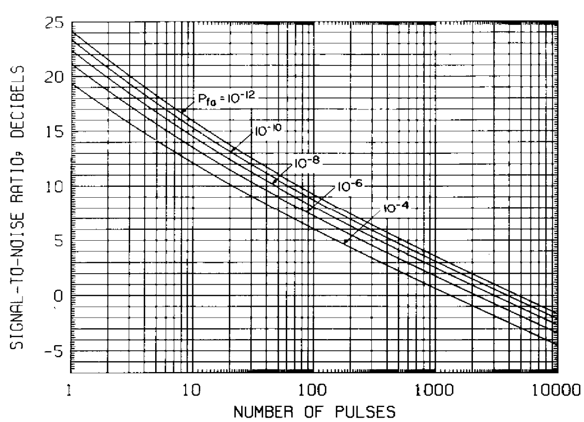


Figure D15 Detectability factor for P_d = 0.9 using n-pulse integration on case 1 fluctuating target (from Blake, 1980, Fig. 2-12, p. 50).

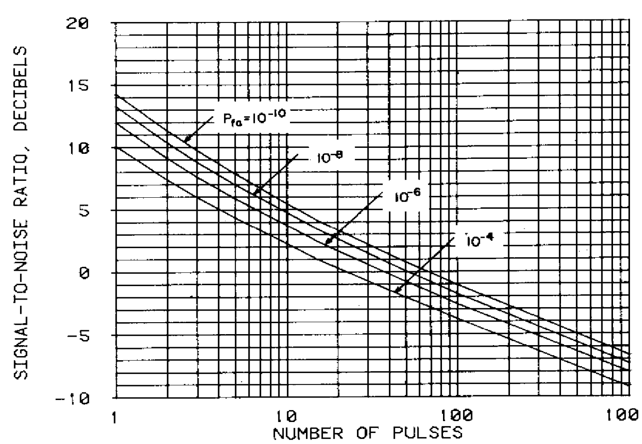


Figure D18 Detectability factor for P_d = 0.9 using n-pulse detection on case 4 fluctuating target (from Blake, 1980, Fig. 2-27, p. 60).

Distribution-free detection does not require prior knowledge of probability density functions for signal or interference. Typically it is described as CFAR detection, because it maintains a constant false-alarm rate. Because it makes no use of input statistical distributions, it has greater loss than optimum detection. One form of distribution-free detection is rank-order detection, which uses the ranks of observed inputs (e.g., their order of amplitudes) and bases detection on some function of those ranks. Distribution-free detection is also called *nonparametric detection*, and it is the preferred procedure when the statistics of input interference are unknown (e.g., when operating in nonhomogeneous and nonstationary clutter). *S*AL

Ref.: Skolnik (1970), p. 15.27, (1980), p. 393.

Double-threshold detection is the process that requires a signal to exceed two thresholds to make a final decision as to signal presence. Typically, the signal is first screened by a binary threshold whose level is low enough to pass very weak echoes and many noise peaks. After storage of the resulting signals, those that originated in the same range element over several successive repetition periods are added to determine whether they exceed the second threshold, which is set high enough to achieve the desired low false-alarm rate. The simplified double-threshold detector (range gating is not shown) is shown in Fig. D19.

Figure D19 A double-threshold system.

Double-threshold detection is often called *m*-out-of-*n* detection or *binary integration*. *SAL*

Ref.: Barton (1976), p. 10; DiFranco (1980), p. 498.

Energy detection is based on the output of an integrator, which is the energy integrated over some time interval. One type of energy detector is the *radiometer*. Ideal energy detection is described by the equation

$$V = \int_{(t-T)}^{t} y^2(r)dr$$

where $y(t)$ is the input voltage, V is the output, and T is the integration time (Fig. D20).

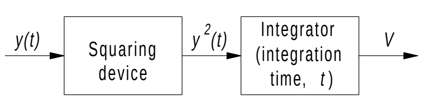


Figure D20 Ideal energy detection.

There are two main kinds of ideal energy detection: continuous integration and integrate-and-dump. *SAL*

Ref.: Dillard (1989), p. 13.

Fixed-sample detection requires a fixed number of observations, regardless of the outcome of each observation (as opposed to the sequential detector). The usual implementation is likelihood-ratio detection using the Neyman-Pearson detection criterion. *SAL*

Ref.: Skolnik (1970), p. 15.4.

The **ideal observer detection criterion** is the optimum criterion, giving the minimum sum of erroneous decision probabilities. To apply the criterion, one must know *a priori* the probability of target presence P_a in the specified area and *a priori* probability of target absence $q_a = 1 - P_a$. Then the probability of erroneous "target present" decisions is q_aP_{fa}, where P_{fa} is the false-alarm probability, and the probability of erroneous "target absent" decisions is P_aP_{tm}, where P_{tm} is the probability of missing the target. According to ideal observer criterion, the best detector gives the minimum sum of the erroneous decisions P_Σ

$$P_\Sigma = q_aP_{fa} + P_aP_m = \min$$

In Russian radar literature, this criterion is often termed the *Kotel'nikov criterion*. *AIL*

Ref.: DiFranco (1968), p. 271; Kazarinov (1990), p. 25.

In-phase/quadrature (I/Q) detection uses two components of the signal, one of which is shifted 90° with respect to the other (Fig. D21). This eliminates the influence of random initial signal phase and provides 3-dB gain in SNR at the output of a doppler filter, in comparison with single-channel detection. *SAL*

Ref.: Morris (1988), p. 42.

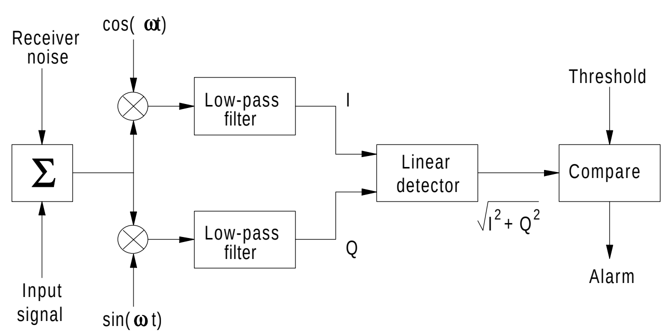


Figure D21 I/Q detection (after Blake, 1980,

Jamming detection is a process to detect the presence of a noise jammer. It is typically used in tracking radar to warn the operator of the threatening situation. *SAL*

Ref.: Neri (1991), p. 442.

The **likelihood-ratio detection criterion** is the basis of an optimum detection process in which the decision as to target presence is made by comparing the likelihood ratio Λ with a threshold value Λ_0. The likelihood ratio determines the ratio of the probabilities that the observed sample $y(t)$ results from presence of the signal, $u(t)$ rather than from its absence:

$$\Lambda(y, u) = \frac{f(y, u)}{f(y, 0)}$$

where $f(y,u)$ is the probability density of the observed signal under the condition that signal $u(t)$ and noise are present, and $f(y,0)$ is that when only noise is present.

The procedures performed in the detector when likelihood ratio $\Lambda(u)$ is calculated are determined by the shape of $f(y,u)$ and $f(y,0)$ and can be worked out in detail for known statistical characteristics of interference $n(t)$ and signal plus interference combination, $y(t) = u(t) + n(t)$. The ratios of *a priori* probabilities of target presence and absence can be taken for the threshold values.

The likelihood-ratio detector based on this criterion is a distribution-dependent detector. The Marcus-Swerling detector is a version of a sequential detector for use in search radar. It operates on multiple detector cells but does not test all 2^k possibilities (where k is the number of resolution cells) but makes a decision as to target presence on the basis of the likelihood ratio. The Wald detector is a sequential detector using the theory of sequential analysis developed by Wald. *AIL*

Ref.: Dulevich (1978), pp. 60–62; Dillard (1989), pp. 139–142.

M-ary [multialternative] detection makes the decision as to signal presence or absence using more than two hypotheses (as opposed to binary detection). The common example of a multialternative detector is the detector making a decision about the target presence for each of the range cells. In this case, the number of possible decisions is $M = 2^k$, where k is the number of range resolution cells. It can consist from the set of binary detectors: one binary detector per one resolution cell. In a more generic case, the problem of range and velocity detection can be solved by means of a multialternative detector, Fig. D22.

There are K_R range channels and $K_u = mK_R$ velocity channels, where m is the number of velocity channels per range channel. From the output of the Kth range channel 1, the signal goes to a velocity channel 2, and then to switch 3 and threshold unit 4. If output voltage of the Kth channel exceeds the threshold voltage, the decision about the target presence is made, and the signal goes to decision-making unit 5, which determines in what cell the target was detected (i.e., determines target range and velocity). Multialternative detectors are typically used in multichannel phased-array radars. *AIL*

Ref.: Skolnik (1970), p. 179; Lukoshkin (1982), p. 42.

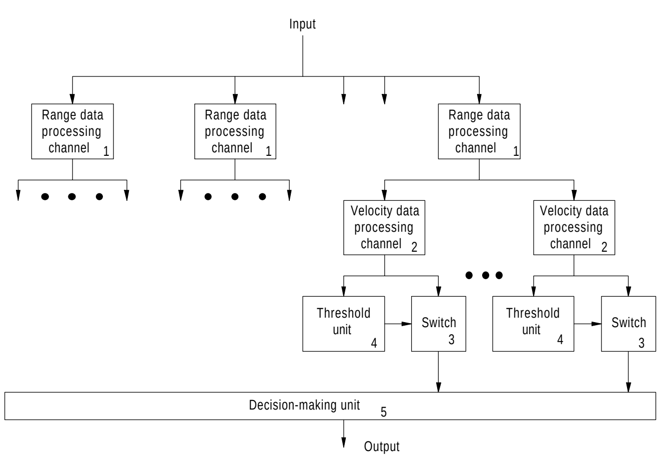


Figure D22 Multialternative detector for range and velocity.

Mean detection compares the mean value of n received pulses to a predetermined threshold. The conventional receiver typically uses mean detection. *SAL*

Ref.: Skolnik (1980), p. 486.

Detection-measurement is detection of a signal in an interference background with simultaneous evaluation of parameters (e.g., range, azimuth, elevation) in case a target-present decision is made. The separation of these tasks is quite arbitrary and concerns only the differences in practical implementation of evaluating signal parameters in detection and measurement modes.

Due to the random nature of interference and parameters of received signals, the tasks of detection and measurement are statistical. In statistical decision theory, a number of criteria for detection of signals have been developed (see **detection criteria**). One can distinguish tasks of detection-measurement of two kinds: (1) when measured parameters do not change during the observation interval and (2) when one cannot disregard such a change. Optimization of detection-measurement of the first kind reduces to single-stage minimization based on the average risk criterion. A rigorous solution to such problems for the total set of targets on the whole is quite complex. If minimization is conducted for each target, then detection-measurement reduces to discovery of excesses of the threshold by the likelihood-ratio logarithm. In practice, for example, if the measured parameter is delay time, this amounts to using the matched filter with an amplitude detector. Optimization of detection-measurement of the second kind requires one to introduce certain models describing the change for the measured parameters in time. In the majority of cases, Markovian discrete or continuous models are applicable. This amounts to multistage tracking detection-measurement; for example, for detection of trajectory. The most common circuits for implementation of detection-measurement are plot extractors. *AIL*

Ref.: Skolnik (1980), p. 388; Shirman (1981), pp. 293–300.

Median detection is a process in which the median value of n received pulses is found and compared against a threshold. The process is robust, as the threshold values and the detection probabilities do not depend on the detailed shape of the clutter probability density function but depends only on the median value. This method of detection has better performance in non-Rayleigh clutter than mean detection. *SAL*

Ref.: Skolnik (1980), p. 486.

m-out-of-n detection (see **double-threshold detection**, **INTEGRATION, binary**).

The **minimax detection criterion** is the optimum detection criterion that ensures minimum average risk for unknown *a priori* probability of occurrence of the signal. This criterion is a particular case of the minimum-average-risk criterion, when *a priori* probabilities P_a of appearance of signals from targets are unknown, but losses r_{lt} and r_{pr} due to selection of an invalid hypothesis (r_{lt} is the loss for a false alarm, r_{pr} is the loss for a target miss). When using this criterion, one calculates the average risk $r(P_{aj})$ for the worst values of *a priori* probabilities P_{aj}, $j = 0, 1, \ldots, k$, for which

$$r(P_{aj}) \geq r(P_a),\ j = 0, 1, \ldots, K,$$

where $r(P_a)$ is the average risk based on the minimum average risk criterion. *AIL*

Ref.:DiFranco (1968), p. 274; Lukoshkin (1983), p. 28.

Missed detection refers to the erroneous decision that a signal is absent when it is actually present. In statistical detection theory it is called a type II error (as opposed to type I error that is a false alarm). *SAL*

Ref.: Skolnik (1980), p. 376.

moving-target detector (see **MOVING-TARGET DETECTOR**).

Moving-window detection examines continuously the last n samples within each quantized range interval and announces the presence of a target if m-out-of-n samples cross a preset threshold. The block diagram of binary moving-window detection is shown in Fig. D23. Such a procedure is less efficient than ideal postdetection integration, but it has the advantage of simplicity.

In binary moving-window detection, the test statistic S_j is computed using

$$S_j = \sum_{k = j - M + 1}^{j} y_k$$

where y_k is the received sample and M is the window length. To compute and threshold the statistics S_j, the binary moving-window detection can use, for example, a digital delay line, counter, and comparator. The detection decision is made when S_j crosses a threshold K_m in the upward direction: in effect, when $S_{j-1} < K_m$ and $S_j = K_m$. This detection process is often termed m-out-of-n detection, *sliding-window detection*, or *binary integration*. *SAL*

Ref.: Skolnik (1980), p. 388; Neri (1991), p. 121.

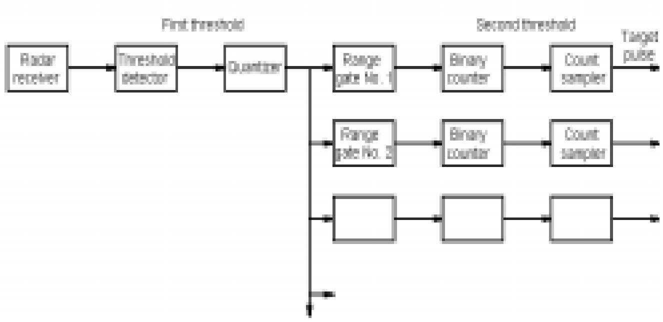


Figure D23 Block diagram of a binary moving-window detection (after Skolnik, 1980, Fig. 10.7, p. 388).

multinomial [multiple-coincidence] detection (see **coincidence detection**).

Multichannel detection is used in a multichannel radar. The block diagram of a multichannel detector when surveillance is performed with a single beam is given on Fig. D24. There are K detection channels, each corresponding to one resolution cell, and employing a binary detector. If the output voltage of the Kth channel exceeds the threshold, then the output voltage from the threshold unit 2 goes to decision-making unit 3, where the decision about the target presence is made and the proper cell where detection took place is determined. Typically, multichannel detection is used in phased-array radars. *AIL*

Ref.: Lukoshkin (1983), pp. 38–43.

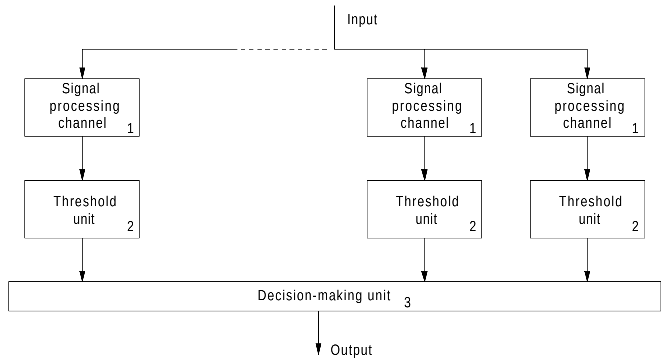


Figure D24 Multichannel detector.

Multiple-pulse detection uses the pulse train received from the target. The technique that improves detection performance in this case is integration. A typical configuration for this type of detector consists of a detector matched to an individual pulse in the train, followed by a coherent or noncoherent integrator, depending on the structure of the transmitted waveform and the radar (see also **INTEGRATOR**). *SAL*

Ref.: Blake (1980), p. 41.

The **Neyman-Pearson (detection) criterion** is the optimum detection criterion giving the maximum probability of detection P_d value. It requires that the acceptable probability of false alarm P_{fa} be assigned. Then the best is considered that detector giving the maximum P_d under the condition that the P_{fa} will exceed the specified value

$$P_d = \max \text{ if } P_{fa} \leq P_{fa0}$$

AIL

Ref.: Druzhinin (1967), p. 532; Barkat (1991), pp. 145–147.

Noncoherent detection is the detection of a noncoherent signal in an interference background. For the noncoherent pulse train, due to the statistical independence of the initial phases of the pulses and the fact that the noise is not correlated from pulse to pulse, the likelihood ratio of the train Λ is equal to the product of likelihood ratios for each pulse. If we take the logarithm of Λ:

$$l = \ln\Lambda = \ln(\Lambda_1 \cdot \Lambda_2 \cdot \ldots \cdot \Lambda_n) = \sum_{i=1}^{n} \ln\Lambda_i$$

the algorithm for optimum noncoherent detection can be cited as

$l \geq z_0$ then target is present.

$l < z_0$ then target is absent.

where z_0 is the threshold defined by the selected detection criterion.

The optimum detector for a noncoherent pulse train typically consists of the matched filter (matched to single pulse in a train), an amplitude detector, a noncoherent integrator, and a threshold unit (Fig. D25). The dependence of the signal-to-noise ratio versus the number of pulses required for detection with specified probabilities of detection and false alarm can be determined from detection curves for chosen target-fluctuation model and detector model. *AIL, SAL*

Ref.: Blake (1980), p. 41; Sosulin (1992), p. 64.

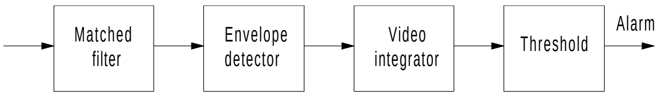


Figure D25 Optimum detector for a noncoherent pulse train.

nonparametric detection (see **distribution-free detection**, **CFAR**).

Open-loop automatic detection is used in an ESM system in which a certain number of many cells are available for incoming pulses (Fig. D26). The available cells are labeled with information about the instantaneous parameters of the emission (i.e., direction-of-arrival or pulsewidth) and when a pulse arrives, a free cell will declare itself available to accept other pulses with the same parameters as those of the first pulse, within a certain tolerance. The agility of the parameters may be taken into account to use the appropriate criterion. *SAL*

Ref.: Neri (1991), p. 309.

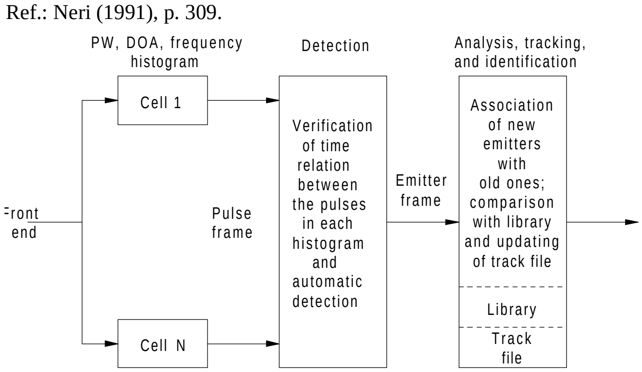


Figure D26 Open-loop automatic detection (from Neri, 1991, Fig. 4.27, p. 310).

Operator detection is the detection of echo signals by the human operator using visual observation of radar displays (sometimes called *visual detection*). Operator detection depends not only on the signal-to-noise ratio but on the visual-mental acuity of the operator, his alertness and experience, and the resolution of the display (which may cause a collapsing loss). In modern radars this method is usually replaced with automatic detection. *SAL*

Ref.: Barton (1964), p. 6; Skolnik (1990), p. 2.24.

Optimum detection is detection of a signal in an interference background using optimum detection criteria. In the optimum detection a decision on presence of a target is made in the case when the likelihood ratio

$$\Lambda(y, u) = \frac{f(y, u)}{f(y, 0)} \geq \Lambda_0$$

where Λ_0 is the detection threshold; $f(y,u)$ and $f(y,0)$ are the probability density functions of the received signal $y(t)$ in interference when the desired signal $u(t)$ is present in the observation interval T and when it is absent, respectively. The likelihood ratio determines by what factor the observed presence of $u(t)$ is more probable in the actual presence of a signal than during its absence.

The decision "there is a signal," given by the detector, is written in the form $R[u(t)] = 1$ and the decision "there is no signal" is given in the form $R[u(t)] = 0$. As a result we can write

$$R[u(t)] = \begin{pmatrix} 1, \Lambda \geq \Lambda_0 \\ 0, \Lambda < \Lambda_0 \end{pmatrix}$$

In place of the likelihood ratio Λ sometimes one uses the logarithmic function of this ratio

$$\lambda = \log\Lambda = \log\left[\frac{f(y, u)}{f(y, 0)}\right]$$

In this case the decision rule for detection is written in the form

$$R[u(t)] = \begin{pmatrix} 1, l \geq \log\Lambda_0 \\ 0, l < \log\Lambda_0 \end{pmatrix}$$

Such detection frequently is called *logarithmic*. There is a large number of variants for detection of different signals in the background of different kinds of interference, and in each case the likelihood ratio or its algorithm can be specified for known characteristics of interference and mixture of a signal with interference to find the optimum detection algorithm. Optimum detection is discussed in radar detection theory for finding maximum achievable signal detection performance in a background of interference. In practice it is difficult to implement completely optimum detection algorithms, and simplified detectors with parameters close to optimum are used (e.g., the binary detector, *m*-out-of-*n* detector, etc.). These detectors experience some detector loss with respect to the true optimum and are termed *quasioptimum detectors*. *AIL*

Ref.: Skolnik (1970), p. 15.5; Dulevich (1973), pp. 60–62; Blake (1980), p. 45.

Detection probability is "the probability that a signal, when actually present at the input to the receiver, will be correctly declared a target signal based on observation of the receiver output." This probability is typically referred to in terms of single-pulse, single-scan, or cumulative probability of detection. Single-scan probability of detection is often called the *blip-scan ratio*, for which the usual notation is P_d. For surveillance radars, the probability that the target is detected at least once in N successive scans is called the *cumulative probability of detection*.

The probability of detection is associated with the probability of false alarm in a way that depends on the detection criterion. For a given receiver configuration if the detection threshold is reduced to increase the probability of detection, there will be an increase in false-alarm probability, and similarly for decreased probabilities if the threshold is raised. The two probabilities can be calculated, for a given threshold, if the statistical properties of the background interference are defined. The simplest case is for random (Gaussian) noise, when the envelope of signal plus noise follows the Rician distribution. In this case the corresponding equations for coherent and noncoherent detection are given in Tables D4 and D5. The North approximation is often used to calculate detection probability, as it is both simple and accurate. *SAL*

Ref.: IEEE (1990), p. 10; Barton (1988), pp. 65–68; Skillman (1991).

Table D4
Equations for Coherent Detection

Probability density function of IF noise (I- or Q-channel)	$dP_v = \frac{1}{\sqrt{2\pi N}}\exp\left(-\frac{v^2}{2N}\right)dv$
False-alarm probability in I-channel	$P_{fa} = \int_{E_t}^{\infty} dP_v = \frac{1}{\sqrt{2\pi N}}\int_{E_t}^{\infty}\exp\left(-\frac{v^2}{2N}\right)dv$ $= Q\left(\frac{E_t}{\sqrt{N}}\right) = \frac{1}{2}\operatorname{erfc}\left(\frac{E_t}{\sqrt{2N}}\right)$
Threshold voltage	$E_t = \sqrt{N}\times Q^{-1}(P_{fa})$ $= \sqrt{2N}\times\operatorname{erfc}(2P_{fa})$
Probability density function of signal plus noise	$dP_s = \frac{1}{\sqrt{2\pi N}}\exp\left(-\frac{v^2+E_s^2}{2N}\right)dv$
Detection probability in I-channel	$P_d = \int_{E_t}^{\infty} dP_s = \frac{1}{\sqrt{2\pi N}}\exp\left(-\frac{v^2+E_s^2}{2N}\right)dv$ $= Q\left(\frac{E_t-\sqrt{2S}}{\sqrt{N}}\right) = \frac{1}{2}\operatorname{erfc}\left(\frac{E_t-\sqrt{2S}}{\sqrt{2N}}\right)$ $= Q[Q^{-1}(P_{fa})-\sqrt{2S/N}]$
Required signal voltage	$E_s = E_t+\sqrt{N}\times Q^{-1}(1-P_d)$ $= \sqrt{N}[Q^{-1}(P_{fa})+Q^{-1}(1-P_d)]$
Detectability factor for coherent process	$D_c(1) = \frac{1}{2}[Q^{-1}(P_{fa})-Q^{-1}(P_d)]^2$ $= [\operatorname{erfc}^{-1}(2P_{fa})-\operatorname{erfc}^{-1}(2P_d)]^2$
Definition of function Q and its inverse:	$Q(E) = \frac{1}{\sqrt{2\pi}}\int_{E}^{\infty}\exp\left(-\frac{v^2}{2}\right) = P$ $Q^{-1}(P) = E$
erf is the error function and erfc is the complementary error function; erf^{-1} and erfc^{-1} are the inverse functions: $\operatorname{erf}(V) = P$, $\operatorname{erf}^{-1}(P) = V$.	
(from Barton, 1988, Table 2.1, p. 65).	

Table D5
Equations and Approximations for Single-Pulse Noncoherent Detection

Probability density function of IF noise envelope	$dP_e = \frac{E_n}{N}\exp\left(-\frac{E_n^2}{2N}\right)dE_n,\quad E_n > 0$
False-alarm probability	$P_{fa} = \int_{E_t}^{\infty}\frac{E_n}{N}\exp\left(-\frac{E_n^2}{2N}\right)dE_n = \exp\left(\frac{-E_t^2}{2N}\right)$
Threshold voltage	$E_t = \sqrt{2N\ln\left(\frac{1}{P_{fa}}\right)}$
Probability density function of IF signal-plus-noise envelope (Rician)	$dP_s = \frac{E_n}{N}\exp\left[-\frac{E_n^2+E_s^2}{2N}\right]I_0\left(\frac{E_nE_s}{N}\right)dE_n$
Detection probability	$P_d = \int_{E_t}^{\infty} dP_s$
North's approximation	$P_d \approx Q\left[\sqrt{2\ln\left(\frac{1}{P_{fa}}\right)}-\sqrt{2\left(\frac{S}{N}\right)+1}\right]$ $= \frac{1}{2}\operatorname{erfc}\left(\sqrt{\ln\left(\frac{1}{P_{fa}}\right)}-\sqrt{\frac{S}{N}+\frac{1}{2}}\right)$ $D_0(1) \approx \left[\sqrt{\ln\left(\frac{1}{P_{fa}}\right)}-\frac{1}{\sqrt{2}}Q^{-1}(P_d)\right]^2-\frac{1}{2}$ $= \left[\sqrt{\ln\left(\frac{1}{P_{fa}}\right)}-\operatorname{erfc}^{-1}(2P_d)\right]^2-\frac{1}{2}$
DiFranco and Rubin (1968)	$P_d = Q\left[\sqrt{2\ln\left(\frac{1}{P_{fa}}\right)}-\sqrt{2\frac{S}{N}}\right]$ $D_0(1) \approx \left[\sqrt{\ln\left(\frac{1}{P_{fa}}\right)}-\frac{1}{\sqrt{2}}Q^{-1}(P_d)\right]^2$ $= \left[\sqrt{\ln\left(\frac{1}{P_{fa}}\right)}-\operatorname{erfc}^{-1}(2P_d)\right]^2$
(from Barton, 1988, Table 2.3, p. 68.	

Radar detection of agitated metals (RADAM) is a technique used for target recognition on the basis of detecting the modulation of the scattered signal caused by the modification of the current distribution on a metal target that results from intermittent contacts on the target. *SAL*

Ref.: Skolnik (1980), p. 437.

detection range (see **RANGE EQUATION**).

rank(-order) detection (see **distribution-free detection**, **CFAR**).

Reliability of detection is the description of detection performance by means of statistical parameters, such as probability of detection: probability that the target will be detected at a given range, within a given time, or similar measures. *SAL*

Ref.: Nathanson (1990), p. 8.

Sequential detection is "a method of automatic detection in two or more steps, normally the first using a high probability of false alarm and the last a low probability of false alarm. In radars with controllable scanning, the first detection can be used to order the scan to return to, stop at, or stay longer at the suspected target position."

In sequential detection the sample size is not fixed (as opposed to fixed-sample detectors). Typically, it uses likelihood ratio detection in applications where an agile-beam antenna is used and the dwell times correspond to the sequence of random sample sizes used by the detector.

Such a detection process can be practically implemented with a phased-array radar, obtaining a power saving of 3 to 4 dB as compared with uniform scanning. The term *sequential detection* sometimes is used synonymously with the sequential observer criterion. *SAL*

Ref.: IEEE (1990), p. 27; Barton (1988), p. 483; Skolnik (1980), p. 381; Barkat (1991), p. 162; DiFranco (1980), p. 547.

The **sequential observer (detection) criterion** gives the minimum time for detection when the parameters of probability density are sequentially checked under the conditions of fixed values of probability of detection P_d and probability of false alarm P_{fa}. The discrete values, $Y_1, ..., Y_n$, of the signal-plus-noise combination, $Y(t) = u(t) + n(t)$, are supplied at the detector input. The likelihood ratio is $\Lambda(y) = [W_1(y)]/W_0(y)]$ (see **likelihood ratio detection criterion**). Sequential detection consists of the comparison of likelihood ratio $\Lambda(Y_n)$ with two fixed thresholds A and B ($A > B$), applied at every step, k. If $\Lambda(Y_n) \geq, B$, then the decision of signal presence is made. If $\Lambda(Y_n) \leq A$, then the decision of signal absence is made. If $B \leq \Lambda(Y_n) \leq A$, the decision is not made and the next $(k + 1)$ step is performed. The threshold values can be determined as

$$A \approx P_d / P_{fa}$$

$$B \approx \frac{(1 - P_d)}{(1 - P_{fa})}$$

Sometimes this criterion is called the *Wald criterion* after the person who first to introduced it. *AIL*

Ref.: Dulevich (1978), p. 62; Barkat (1991), p. 162.

Detection of signals with full polarization reception occurs when two orthogonal polarized components of the reflected electromagnetic waves are received. For conventional radar operating with fixed polarization, the intensity of the received signal depends much on how the antenna polarization coincides with polarization optimum for observed target. This shortcoming can be eliminated in the radars with full polarization reception giving a greater increase in signal detection. For example, for monopulse radar with full polarization reception, the probability of detection can be two to four times that of analogous radar employing horizontal polarization only. The value of this increase depends on the polarization properties of the target. *AIL*

Ref.: Tuchkov (1985), pp. 213–215.

Detection of a signal with known parameters applies when the signal parameters are completely known. If the parameters of signal $u(t)$: amplitude $A(t)$, frequency ω_0, and phase $\phi(t)$, are completely known, one can write

$$u(t) = A(t) \cos [\omega_0 t + \phi(t)]$$

If the white-noise model is applicable to the interference, the output of the detector in time t there will consist of a signal $y(t)$, which is an additive mixture of the useful signal $u(t)$ and white noise $n(t)$, that is

$$y(t) = u(t) + n(t).$$

The likelihood ratio for a completely known signal is

$$\Lambda = \exp\left[-\frac{E_s}{N_0} + \frac{2}{N_0} Z(T)\right]$$

where

$$E_s = \int_0^T u^2(t)dt$$

is the signal energy,

$$Z(T) = \int_0^T u(t)y(t)dt$$

is the correlation integral, and

N_0 is the noise power spectral density.

To derive a solution one must compare a Λ with the threshold Λ_0:

If $\Lambda \geq \Lambda_0$ there is a signal.

If $\Lambda < \Lambda_0$ there is no signal.

This comparison is difficult to implement in practice. Therefore, insofar as E_s = const, and the likelihood ratio Λ depends on the correlation integral $Z(T)$, one must compare $Z(T)$ with the threshold Z_0:

If $Z(T) \geq Z_0$ there is a signal.

If $Z(T) < Z_0$ there is no signal.

An optimum detector, which computes the correlation integral and compares it with the threshold, can be of two kinds. In the first case it consists of a multiplier and integrator (which make up the correlator) and a threshold unit and is called a *correlation detector*. In the second case the detector consists of a matched filter and threshold unit. Such a detector is called a *filter detector*. Signals with known parameters hardly exist in practice and are used mainly in theoretical analysis. *AIL*

Ref.: DiFranco (1968), p. 291; Dymova (1975), pp. 51–57; Sosulin (1992), pp. 48–55.

Detection of a signal with unknown parameters is the usual case where a signal is present in a background of interference. A signal with unknown parameters is one in which either the phase $\phi(t)$ or amplitude $A(t)$ and phase are random. A signal with unknown random initial phase can be written in the form

$$u(t,\psi) = A(t) \cos [\omega_0 t + \phi(t) + \psi]$$

where ψ is a random quantity, uniformly distributed in the interval 0, ... , 2π.

At the output of the detector during time T there will be an additive mixture of the useful signal $u(t,\psi)$ and interference $n(t)$:

$$y(t) = u(t,\psi) + n(t)$$

If the interference is white noise, the likelihood ratio for a signal with random phase has the form

$$\Lambda_1 = \exp\left[-\frac{E_s}{N_0} J_0\left(\frac{2Z(T)}{N_0}\right)\right]$$

where E_s is the signal energy, $Z(T)$ is the correlation integral, $J_0(\cdot)$ is the modified Bessel function of zero order, and N_0 is the noise power spectral density.

The optimum decision rule for the problem of signal detection is calculation of the correlation integral $Z(T)$ and its comparison with the threshold value Z_0:

If $Z(T) \geq Z_0$ there is a signal.

If $Z(T) < Z_0$ there is no signal.

The circuit of an optimum detector in this case will consist of a matched filter, which computes the correlation integral $Z(T)$, an amplitude detector, which removes the envelope of $Z(T)$, and a threshold unit. The threshold Z_0 is selected to achieve the desired low false-alarm probability.

The correlation integral $Z(T)$ can also be represented in the form of quadrature components $Z_1(T)$ and $Z_2(T)$:

$$Z(T) = \sqrt{Z_1^2(T) + Z_2^2(T)}$$

where

$$Z_1(T) = \int_0^T Ay(t)\cos[\omega_0(t) + \phi(t)]dt$$

$$Z_2(T) = \int_0^T Ay(t)\sin[\omega_0(t) + \phi(t)]dt$$

An optimum circuit of a detector in this case should include two channels (I and Q), a summing circuit and a threshold unit. The presence of two channels is due to lack of knowledge of the initial phase of the signal. Such a detector is called a *correlation detector* (see **coherent detection**).

A signal with random amplitude and initial phases can be written in the form

$$U(t,\psi,\beta) = \beta A(t) \cos[\omega_0 t + \phi(t) + \psi]$$

where β and ψ are random quantities. Usually it is assumed that the value of β has a Rayleigh distribution, and ψ has a uniform distribution over 2π. The likelihood ratio for this signal is written in the form

$$\Lambda_2 = \frac{N_0}{E_s + N_0}\exp\left[\frac{Z^2(T)}{N_0(E_s + N_0)}\right]$$

Insofar as $Z_2(T) \geq 0$, the likelihood ratio Λ_2 is a monotonic function of its argument and the circuit of an optimum detector in this case does not differ from the circuit of an optimum filter of a signal detector with random phase. Only the optimum threshold Z_0 changes, which will depend on the distribution law of the random amplitude (quantity β). *AIL*

Ref.: DiFranco (1968), p. 298; Dymova (1975), pp. 57–63; Sosulin (1992), pp. 48–55.

Detection of a random signal is based on an anticipated random process. The problem of detecting random processes occurs in radar astronomy, radio prospecting, and passive radar. In detecting a random process in an observation $y(t)$, in addition to white $n(t)$ there might also be a normal useful process $u(t)$ with zero mean and correlation function $K_u(t, t + \tau)$, which is uncorrelated with the noise $n(t)$, where τ is the correlation time. If the process $u(t)$ is stationary and, consequently, $K_u(t, t + \tau) = k$ then the decision on detection of a random process is made when the quantity Z exceeds the threshold

$$Z = \int_T y^2(t)dt \geq Z_0$$

where Z_0 is the threshold which depends on the selected detection criterion, and T is the random process observation time. In practical implementation, the detector repeats the structure of the correlation receiver with the difference being that the reference signal $u_y(t)$ in it is not generated autonomously but from the observed oscillation $y(t)$, itself, which passes through the linear filter.

During reception of a radio thermal signal the power densities of the detected and interfering signals are unknown, therefore one usually uses the contrast method of detection, which consists of comparing output signals with one another for two different measured components (two positions of the antenna radiation pattern). In this case the decision on detection is assumed based on comparison with the threshold, which will depend on the anticipated temperature contrast, difference $\Delta Q = Q_1 - Q_2$, where Q_1 and Q_2 are the output signals of the thermal receiver (radiometer) for the analyzed components. (See also **detection of a signal with known parameters** and **detection of a signal with unknown parameters**.) *AIL*

Ref.: Kazarinov (1990), pp. 56–60, 432–436.

Single-channel detection occurs in a single-channel radar when the signal detection and processing is done sequentially for each resolution cell. It requires more time to detect the target in comparison with multichannel detection, but with the advantage of simplicity and cost reduction, as the number of hardware components is considerably less in this case. *AIL*

Ref.: Sosulin (1992), p. 25.

Single-pulse [hit, sample] detection is based on processing a single pulse or receiver output. In most radars many pulses are available, and these are best combined by integration before being applied to the detection threshold. If this integration is performed coherently (using a filter matched to the pulse train), the output can be considered a single-sample process. Such processes are also used in some phased-array radars that dwell for only a single pulse repetition interval during search or track. Theoretical evaluation of single-pulse detection probability is widely used to estimate overall performance of radar in target detection. *SAL*

Ref.: Blake (1980), p. 26.

Single-scan [burst] detection is detection in a single scan (observation), obtaining the single-scan probability of detection. It is sometimes called *single-burst detection*, or (when only one pulse per beam is used) single-hit, single-pulse or single-sample detection. *SAL*

sliding-window detection (see **moving-window detection**).

stationary-target detection (see **TARGET RECOGNITION AND IDENTIFICATION**).

Threshold detection is based on establishing a threshold level at the output of the receiver and making the decision that a signal is present if the receiver output exceeds that threshold level. It is the general term defining detection using a threshold to make a decision whether the signal is present or absent. *SAL*

Ref.: Nathanson (1990), p. 124.

Trimmed-mean detection discards the smallest and the largest of the n pulses before taking the mean and performs as a mean detector. It can improve the performance of mean detection but needs to perform the ranking of the pulses, sometimes a rather difficult task to implement in practice. *SAL*

Ref.: Skolnik (1980), p. 486.

Two-stage detection is sequential detection that uses two stages of data collecting and processing. The second stage is used only in the case when the results of the first stage show that the target is "possibly present." During the first stage, n pulses are processed. If the threshold of the first stage is exceeded, then the second group from m pulses is used and integration is performed for the combined common quantity of pulses $m + n$. The accumulated sum is compared with the second threshold, and it is exceeded then "target is present" decision is taken. Two-stage detection reduces by a factor of approximately two the surveillance times in comparison with a conventional detector and provides a gain in the signal-to-noise ratio of about 3 to 6 dB. *AIL*

Ref.: Skolnik (1970), p. 15.24.

visual detection (see **operator detection**).

The **weighting (detection) criterion** is the optimum detection criterion giving the biggest difference ΔP between the probability of detection P_d and probability of false alarm P_{fa}, while taking into account its importance (weight) in the decision-making process:

$$\Delta P = P_d - \eta P_{fa} = \max$$

where η is the weight assigned to false alarms, which are dangerous and undesirable. *AIL*

Ref.: Druzhinin (1967), p. 531.

DETECTOR [DEMODULATION, DEMODULATOR]. This set of articles discusses circuits and devices used to detect signals, in the sense of demodulation, as opposed to the preceding articles dealing with target detection. Detection, in this sense, is the conversion of the modulated voltage of a carrier frequency into the modulating voltage of a video frequency to reveal a transmitted message. Depending on signal modulation type, detectors are classified as amplitude, frequency, phase, or amplitude-phase detectors; all are nonlinear devices.

In the amplitude detector, all phase information is destroyed, and this method is used in noncoherent radars where phase information is not used for signal processing.

The difference between the amplitude-phase (or phase-sensitive) detectors and phase detectors is that in the former both amplitude and phase information are present at the output, while only phase information is present in the latter case. This also distinguishes them from mixers, where amplitude, frequency, and phase information is present (doppler frequency is not taken into consideration in this classification). The phase detector is important in coherent radars that use signal phase in subsequent processing.

Based on the type of nonlinear elements used, detectors are classified as *vacuum-tube* or *solid-state (crystal) detectors*. The latter are most widely used in modern radars. *AIL, SAL*

Ref.: Druzhinin (1967), p. 382; Blake (1980), p. 45; Chistyakov (1986), p. 138; Skolnik (1990), p. 3.32.

An **amplitude detector** extracts the amplitude (voltage) of a signal having modulation. Semiconductor diodes or transistors are used as the nonlinear element in the detector. Main detector characteristics include gain, input conductance, and the degree of signal distortion introduced. Figure D27 provides a diagram of a series diode detector. Under the effect of input voltage, pulses of current, Fig. D27(b), containing a constant component and those with angular frequencies ω, 2ω, and so forth flow through the diode. Detector output voltage is averaged by means of load in the form of a circuit comprising resistor R_n and capacitor C_n. As a result, output voltage U_Ω changes based on the modulation law. This detector is also called an envelope detector. *AIL*

Ref.: Schwartz (1959), p. 107; Chistyakov (1986), p. 140.

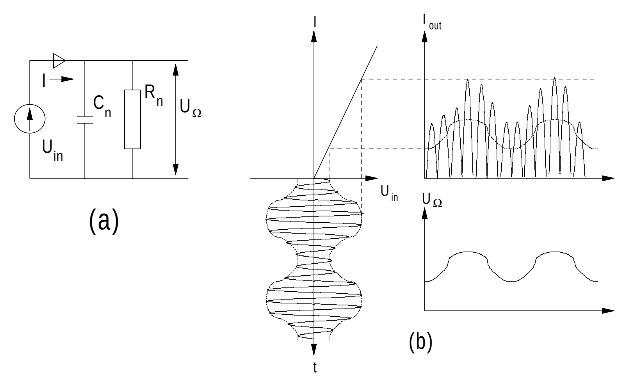


Figure D27 Diode detector: (a) detector circuit; (b) detector operating principle (after Chistyakov, 1986, Fig. 5.2, p. 140).

The **amplitude-phase detector** uses linear circuits (resistors, capacitors, inductors) with varying parameters instead of nonlinear circuits. Examples are synchronous and asynchronous detectors, the first often termed a *phase-sensitive [discriminating] detector*. *SAL*

Ref.: Druzhinin (1967), p. 381; Skolnik (1990), p. 3.32.

An **asynchronous detector** is an amplitude-phase detector not using synchronization of phases of input and reference signals. In the simplest case, an asynchronous detector is a diode amplitude detector, to the input of which two signals are applied: the received signal from the IF amplifier and the local oscillator signal. *AIL*

Ref.: Druzhinin (1967), p. 382.

The **back bias detector** performs much the same function as instantaneous automatic gain control, except that it operates at the diode detector instead of the IF amplifier. *SAL*

Ref.: Johnston (1991), p. 58.

A **balanced(-diode) detector** is a phase detector, the output voltage of which is the difference between the voltages in the loads of each detector arm (Fig. D28). At the detector output, oscillations of phase-modulated signal U_{in} are supplied in antiphase, while reference voltage U_2 is supplied in cophase. Therefore, if signal phase and reference voltage phase do not coincide, then voltages varying in amplitude act upon the diodes and signal polarity may change. The expression for the amplitude-phase response has this form:

$$U_{out} = 2K_d U_{in} \cos\phi$$

where K_d = amplitude detector gain; U_{in} = input signal amplitude; ϕ = angle of the phase shift between the reference and input signals. An advantage of a balanced detector over the conventional phase detector is the better amplitude-phase response curve. Amplifiers are used instead of diodes if there is a requirement to increase gain and input resistance. *AIL*

Ref.: Chistyakov (1986), p. 155; Skolnik (1990), p. 3.36.

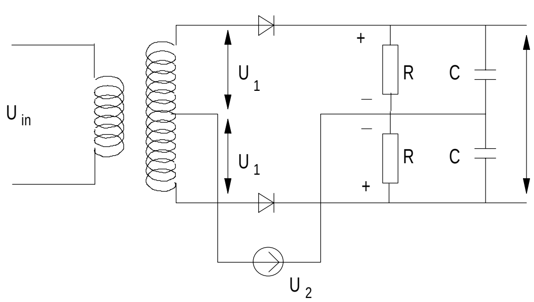


Figure D28 Balanced phase detector (after Chistyakov, 1986, Fig. 5.22,. 155).

A **balanced bias detector** is a controlling circuit used in radar systems for anticlutter purposes.

Ref.: Johnston (1979), p. 58.

boxcar detector (see **CIRCUIT, sample-and-hold**).

A **coherent detector** uses a reference-oscillator signal with the same frequency and phase as the input signal to be detected (Fig. D29). It does not destroy the phase information as does envelope detector, nor does it destroy the amplitude information as does the zero-crossing detector. The signal-to-noise ratio from the coherent detector is better than from the previous two, but because the phase of the received signal is not usually known in radar applications, this type of detector is seldom used. *SAL*

Ref.: Skolnik (1980), p. 385.

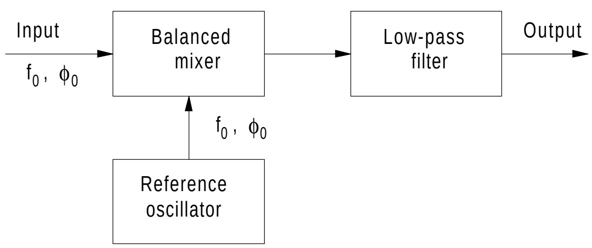


Figure D29 Basic configuration of a coherent detector (after Skolnik, 1980, Fig. 10.4, p. 385).

A **crystal detector** is an amplitude detector in which crystal diodes or transistors are used as the nonlinear element. Crystal diode types used in detectors include germanium diodes, Schottky-barrier diodes, and tunnel diodes with junctions based on the Josephson effect. Bipolar and field-effect transistors usually may also be used. When bipolar transistors are used, the detectors are referred to as base, collector, or emitter detectors depending upon where load is connected. When field-effect transistors are used, load usually is connected to the drain circuits and the detectors are called drain detectors. Advantages of crystal elements in detectors as opposed to vacuum-tube elements include small size, high reliability, and good response curve. Schottky and tunnel diode detectors are used in millimeter-wave receivers. An important property of these detectors is high sensitivity, large signal bandwidth-duration product, sufficient mechanical strength, and resistance to unfavorable climatic effects. *AIL*

Ref.: Van Voorhis (1948), p. 197; Chistyakov (1986), p. 144; Rozonov (1989), pp. 29, 112–141.

Diode detectors are based on diodes of different types as the main rectifying element. The main types of diode detectors are balanced diode detector, Schottky-barrier diode detector, and tunnel diode detector. *SAL*

An **envelope detector** reproduces the amplitude of the carrier envelope and so extracts the modulation and rejects the carrier. In this case all phase information is destroyed and subsequent processing is based on the envelope amplitude. See also **amplitude detector**. *SAL*

Ref.: Skolnik (1980), p. 382.

A **frequency detector** extracts the voltage representing the frequency modulation applied to the carrier. Depending upon operating principle, such detectors are categorized as frequency-amplitude, frequency-phase, or frequency-pulse detectors. In frequency-amplitude detectors, the change in signal frequency is converted to a change in amplitude with subsequent amplitude detection. In frequency-phase detectors, the frequency change is converted into a change in phase shift between the two voltages with further phase detection. In frequency-pulse detectors, the frequency-modulated (FM) oscillations are converted into a pulse train, the pulse repetition rate of which is proportional to the deviation of the input signal frequency from the carrier frequency. An output voltage proportional to the number of pulses per unit time may be formed using a pulse counter.

The frequency detector response is the dependence of output voltage U_{out} out on signal frequency f given constant input voltage amplitude. An important detector parameter is response slope:

$$S_d = \left.\frac{dU_{out}}{df}\right|_{f=f_0}$$

where f_0 = carrier frequency.

Figure D30 depicts frequency detector circuits. In a balanced frequency-amplitude detector (Fig. D30a), one of the circuits is tuned to frequency $f_1 = f_0 + \Delta f$, while the second is tuned to frequency $f_2 = f_0 - \Delta f$. As it increases, the signal frequency approaches f_1 and moves away from f_2. The voltage in

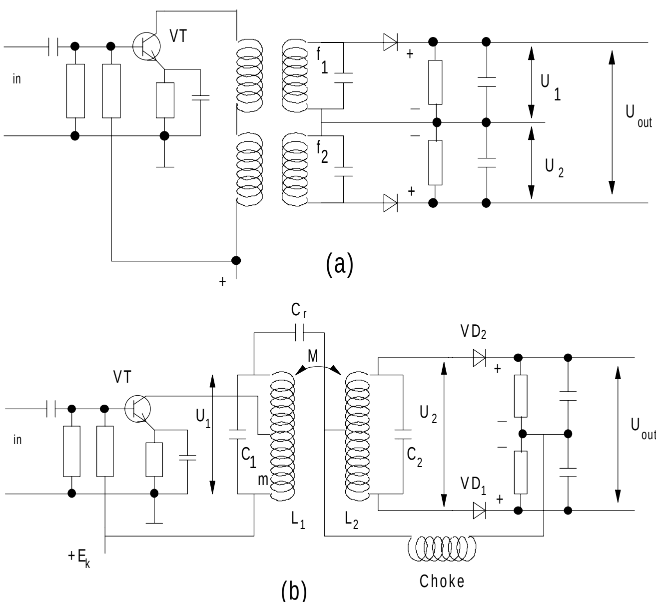


Figure D30 Frequency detectors: (a) frequency-amplitude; (b) frequency-phase (after Chistyakov, 1986, Fig. 5.32, p. 161, Fig. 5.35, p. 163).

the first circuit increases, while in the second it decreases. The opposite is true if the frequency decreases. The FM signal becomes amplitude-frequency modulated. Voltage from the circuits pass to amplitude detectors. The voltage at the frequency-amplitude detector output has this form:

$$U_{out} = U_1 - U_2 = K_d m_k U_{in}|y_{21}|R_d\psi(\xi)$$

where K_d = gain, m_k = coupling coefficient, U_{in} = input voltage, y_{21} = conductance, R_d = equivalent resistance, and $\psi(\xi)$ = detector normalized response.

In a balanced frequency-phase detector (Fig. D30b), in the absence of modulation voltage U_2 is shifted 90° relative to voltage U_1. Given frequency modulation, additional shift proportional to the frequency change appears between U_1 and U_2. Voltages U_1 and U_2 are applied to the diodes of a balanced phase detector, with this voltage being found at the latter's output:

$$U_{out} = K_d m_k U_{in}|y_{21}|R_d\psi(\xi)$$

These frequency detectors require preliminary amplitude limiting of the signal. Therefore, changes are sometimes introduced into detector circuits so they will acquire the additional properties of a limiter. *AIL*

Ref.: Chistyakov (1986), pp. 161–168.

linear detector (see **detector model**).

A **lin(ear)-log(arithmic) detector** is one in which the output signal is linearly proportional to the input signal when its level lies below some required value, and log proportional to the input signal when vice versa. *SAL*

Ref.: Currie (1987), p. 498.

A **logarithmic detector** is a detector with logarithmic dependence between input and output voltages. It usually comprises N amplitude detectors of the series type in which detected output voltages are added (Fig. D31).

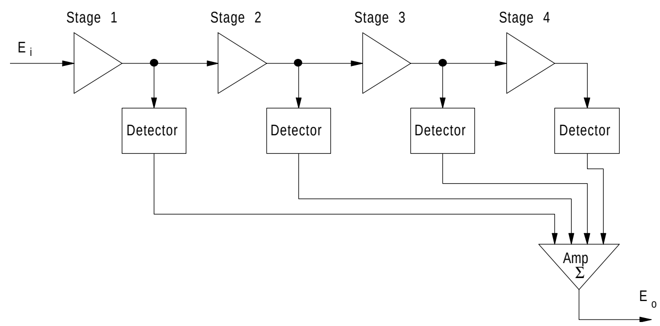


Figure D31 Basic configuration of a logarithmic detector (after Skolnik, 1970, Fig. 29a, p. 5.34).

Here, diode detector voltage-current responses are selected so that a logarithmic dependence is obtained when adding N detector output voltages. To do so, regulation in each stage usually is used. This makes it possible to regulate the slope and length of segments of each response for best coincidence with a logarithmic curve. Logarithmic detectors are widely used in monopulse radars. They are also used in devices suppressing clutter from rain or from the ocean surface, and other variable intensity interference. *AIL*

Ref.: Skolnik (1970), p. 5.34.

A **detector model** describes the ratio between output and input voltages in an amplitude detector. Typically it applies to diode detectors, and if the signal lies on a linear portion of the diode response, the term *linear detector* is used. In this case, the output voltage U_{out} is proportional to the input voltage U_{in}. If the output voltage is proportional to the square of the input, $U_{out} = KU_{in}^2$, the term *square-law detector* is used. The voltage-current response typically follows the square law for weak signals (for diodes, a signal with amplitude $U < 0.25$V is usually considered to be weak), and a linear law for stronger signals.

Detector models are used in the theory of optimum detection, where a square-law detector is shown to be optimum for noncoherent integration at low signal-to-noise ratio (many pulses integrated) and a linear law for high signal-to-noise ratio (few pulses integrated). The linear law is better until approximately 30 to 100 pulses have to be integrated, and since the number of available pulses is usually less than 30, this is the appropriate detector. Theoretically, for a fluctuating target, the square-law detector allows simpler analysis and is often used. The actual difference in performance is only about 0.2 dB in any case. *SAL*

Ref.: Blake (1980), p. 45; Chistyakov (1986), p. 143.

A **multiplier detector** is a phase detector, the output voltage of which results from multiplication of the input signal and a reference voltage. It can be implemented as an integrated circuit based upon control of slope *S* of a differential transistor pair (Fig. D32). The signal is supplied to the detector

$$u_s = U_{ms}\cos(\omega_s t + \phi_s),$$

as is the local oscillator voltage

$$u_0 = U_0 \cos \omega_h t.$$

The signal voltage arrives at transistor bases VT_1 and VT_2 with opposite phases, while the LO voltage is cophasal, causing identical changes of their slope. Therefore, the currents of combined components i_1 and i_2 have opposite phase: $i_1 = i_2 = Su_s$.

As a result of the nonlinearity of slope *S*, the law of its change over time with heterodyne frequency may be represented by Fourier series:

$$S = S_0 + \sum_{k=1}^{\infty} S_{mk}\cos k\omega_h t$$

then

$$i_1 = -i_2 = \left(S_0 + \sum_{k=1}^{\infty} S_{mk}\cos k\omega_h t\right)U_{ms}\cos(\omega_s t + \phi_s)$$

The input voltage is created by the difference of current constant components i_1 and i_2:

$$U_{out} = S_{m1}U_{ms}R\cos\phi,$$

where

$$\phi = (\omega_s - \omega_h)t + \phi_s.$$

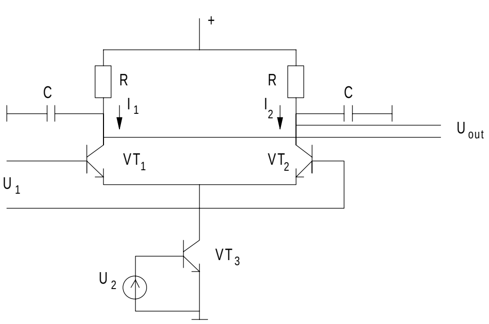


Figure D32 Detector-multiplier (after Chistyakov, 1986, Fig. 5.27a, p. 158,).

Besides this configuration, detector-multipliers based on three differential transistor pairs, which have greater dynamic range of input signal, may be used. An advantage of such detectors is an increase in the slope of the response and gain compared with a diode phase detector. *AIL*

Ref.: Chistyakov (1986), p. 157; Skolnik (1990), p. 3.36.

peak detector (see **video detector**).

A **phase detector** produces an output voltage proportional to the phase difference ϕ between the input and reference signals at a given frequency. A basic phase detector characteristic is the detector response, the dependence of output voltage U_{out} on ϕ.

Basic phase detector parameters include response slope S_{pd} and voltage gain K_{pd}. The response slope is the maximum value of the derivative of output voltage with respect to phase angle:

$$S_{pd} = \left.\frac{dU_{out}}{d\phi}\right|_{\max}; \qquad K_{pd} = \frac{U_{out}}{U_{m1}}$$

where U_{out} = output voltage; U_{m1} = input signal amplitude. Basic phase detector types include balanced detectors, multiplier detectors, and the coincidence phase detector (Fig. D33) that provides a triangular output voltage characteristic. It has a maximum negative output when gates E_s and E_r are in phase (gate B then registers coincidence half the time, and gate A registers no coincidence) and maximum positive output when E_s and E_r are out of phase (then the reverse condition exists).

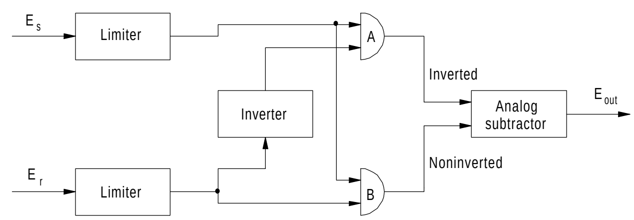


Figure D33 Coincidence phase detector (after Skolnik, 1990, Fig. 3.19, p. 3.36).

Analog-to-digital or fully digital configurations of phase detectors are used in modern radars (Fig. D34). In the first,

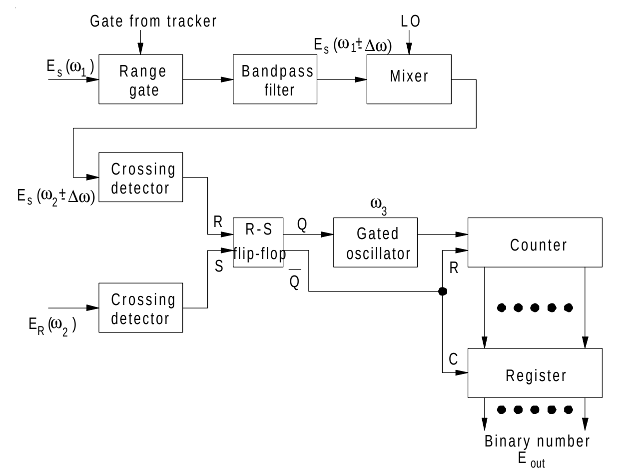


Figure D34 Analog-to-digital phase detector (after Skolnik, 1990, Fig. 3.20, p. 3.37).

the detector measures the time interval between positive or negative zero-crossings of the signal and reference waveforms, which is proportional to phase shift. The resolution of this phase detector is determined by the ratio of the clock frequency to the input frequency (generally the radar intermediate frequency). When digital doppler filters are used for clutter rejection prior to extracting amplitude and phase of the target return, the fully digital phase detector extracts the phase information from the in-phase and quadrature components of input digital codes, and the traditional (analog) phase detectors discussed above are inapplicable. In this case, digital I and Q data are converted into logarithmic format, the polarities of the I and Q signals define the quadrant, and the phase within the quadrant is a function of $\log_2 I^2 - \log_2 Q^2$. *SAL*

Ref.: Skolnik (1990), pp. 3.32–3.38; Chistyakov (1986), p. 155.

square-law detector (see **detector model**).

A **synchronous detector** is an amplitude-phase detector in which output signal polarity and amplitude depend upon the phase difference between input and reference signals. A balanced phase diode detector may be used as a synchronous detector. The reference voltage (a local oscillator) must be synchronized in phase with the input signal carrier. Voltage at the detector output is maximum when the phase difference $\phi = 0$. When $\phi = 90°$, the output voltage is zero, and when $\phi = 180°$, the output voltage polarity reverses. Thanks to the linear dependence of phase detector output voltage on input voltage, the detector may be used for detection of amplitude-modulated signals (amplitude-phase detection). Synchronous detectors were used in older types of radars to implement moving target indication. They are also called *phase-sensitive [discriminating] detectors*. *AIL*

Ref.: Chistyakov (1986), p. 157; Skolnik (1990), p. 3.32.

A **video detector** is an amplitude detector designed for detection of pulse waveforms. Such detectors are divided into two types: the pulse detector and the peak detector. The first is for conversion of RF pulses into direct current pulses (i.e., for reproduction of the envelope of each pulse). A peak detector is for extraction of the envelope of the entire pulse train. Since, during envelope extraction, the output voltage at each moment in time must be proportional to the amplitude (peak value) of the pulses, this is referred to as a *peak detector*. Peak detection may occur by means of single or double detection. In single detection, RF pulses are supplied directly to a peak detector, the most common type being the diode detector (see **amplitude detector**). In double peak detection, RF pulses initially are converted into video pulses by a pulse detector, are amplified, and reach the peak detector, which extracts the pulse train envelope. *AIL*

Ref.:Fink (1982), p. 20.81; Chistyakov (1986), p. 150.

The **zero-crossing detector** is a circuit that produces an output pulse when the input voltage passes through zero. It is widely used in time (range) discriminators and other receiver circuits. *SAL*

Ref.: Skolnik (1990), p. 10.18.

DEVICE, microwave

acoustic-surface-wave device (see **surface-acoustic-wave device**).

An **acoustic-wave device** uses the propagation of acoustic waves in special materials to produce the effect of signal amplification, processing, or delay. The primary advantage of acoustic waves is a relatively low velocity, typically 10^{-5} times that of the electromagnetic waves, which makes possible relatively long signal-delay times in a physically small space. Both bulk-mode propagation and surface waves have been employed in bulk acoustic-wave devices and surface-acoustic-wave (SAW) devices, respectively. The main range of applications for radar purposes is in signal processors and delay lines. *SAL*

Ref.: Fink (1975), p. 9.72.

A **bulk acoustic-wave device** is an acoustic-wave device employing propagation of bulk waves in the solid crystals, primarily to produce amplification in acoustic amplifiers. *SAL*

Ref.: Fink (1975), p. 13.77.

A **charge-coupled device (CCD)** is based on the transfer of a charge in a semiconductor structure in which, owing to the controlling potentials in the electrodes, a potential relief is created in the form of a series of potential holes. The operating principle of the devices is based on the significant time (from hundreds of milliseconds to tens of seconds) necessary to fill the holes with secondary carriers, with the result that it can be viewed as a memory cell. Transfer of charge from one memory cell to an adjacent one is effected with delivery of a more negative voltage to the corresponding electrode. Initially devices with charge transfers were produced in two types: “bucket brigades” and CCDs. In bucket brigades, diffusion p-regions are created in the semiconductor between electrodes located at a greater distance than in charge-coupled devices. Under the action of control voltages in the electrodes, packets of charges are successively moved along a chain from one region to another. Because of the difficulty of forming diffusion regions, bucket brigades are practically never used now.

A basic element of charge-coupled devices is the metal-oxide-semiconductor (MOS) capacitor, which consists of a metal electrode and semiconductor substrate of the n- or p-type, separated from each other by a thin layer of this semiconductor. The role of such a capacitor is frequently performed by more sophisticated surface-charge transistors. The capability of organizing movement of the charge package along the surface in any direction is a specific feature of operation of the CCD. If necessary, the charge can be read from any intermediate electrode through a capacitor.

CCDs are capable of operating at lower frequencies (to 1 GHz) than other functional devices. They are usually used as devices for discrete-analog processing of analog signals. They can also be used for digital processing of signals, combining the high precision of digital systems with the low power consumption and great functional capabilities of analog devices. Analog CCD delay lines and CCD filters of all types are widely used: band-pass, nonrecursive, transversal, recursive, matched, and so forth. *IAM*

Ref.: Gassanov (1988), p. 227; Galati (1993), p. 516.

A **concentrated interaction device** uses the interaction of electron flows with concentrated electromagnetic fields. To achieve the highest effectiveness of interaction of the electron flow with the electromagnetic field, the latter is concentrated in the smallest possible volume, and the electron flow, modulated in density, is passed along the direction of the lines of force of this field. The length of the interaction space must be as limited as possible, and the field intensity as great as possible to compensate for the briefness of interaction of charge and field. The latter is usually achieved by using cavities with concentrated fields.

Devices with concentrated interaction include resonant klystrons and microwave triodes. *IAM*

Ref.: Levitskiy (1986), p. 59.

An **electron quantum device** is based on forced emission of excited molecules (atoms) of an active medium. Depending on the range of the operating frequencies, quantum devices are subdivided into two classes. In the microwave band these are masers, and in the optical band, lasers. Depending on the aggregate state of the active substance, masers are usually subdivided into gas and solid-state, and lasers into gas, solid-state, liquid, and semiconductor. Depending on the operating mode, we distinguish between lasers operating in continuous mode, in pulse mode with a pulse duration of 10^{-3} to 10^{-6} sec, in impulse mode with a duration of 10^{-7} to 10^{-9} sec, and in the synchronization mode, at which the pulse duration can be 10^{-10} to 10^{-12} sec. *IAM*

Ref.: Andrushko (1981), p.12; Zherebtsov (1989), p.176.

An **electron-wave device** is a microwave device with distributed interaction, in which the modulated beam of electrons passes through a nonresisting medium in which the amplitude of the space-charge waves (periodic bunches of electrons) increases owing to the interaction of the beam with the polarization that it excites in the medium. The polarization waves have the same wavelength and phase velocity as the space-charge waves that engender them and are shifted relative to them in phase. Owing to this phase shift, they produce grouping of the electron beam.

Electron-wave devices are used to amplify microwave signals. The removal of energy from an electron beam can be implemented by an actively conducting medium or by a medium that has inductive conductivity. In the latter case, plasma is used as the medium, and an electron-wave amplifier with such a propagation medium is classed a *plasma amplifier*. The medium with inductive conductivity may be a slow-wave circuit in the form of cavities in a specific frequency range, for which propagation of their own waves is no longer possible, but the cavities themselves still possess inductive conductivity.

Because of the difficulty of converting electromagnetic waves to space-charge waves and back, electron-wave devices have not become widely used. *IAM*

Ref.: Levitskiy (1986), p. 105.

An **extended interaction device** uses the interaction of electron flows with traveling waves. The basis of operation of these devices is the physical phenomenon of Vavilov-Cherenkov radiation. In this phenomenon, with movement of the charged particles in some medium at a speed exceeding the speed of electromagnetic waves in the given medium, these waves are radiated. The propagation medium in the devices is a a slow-wave circuit. When the electron speed is close to the phase velocity of the waves in the slow-wave circuit, or somewhat exceeds it, the energy of the electrons is transferred to the wave.

Devices of this class include traveling-wave tubes and backward-wave tubes, and, to some degree, electron-wave devices. Thanks to the extent of interaction, they have a wider bandwidth than concentrated interaction devices. *IAM*

Ref.: Gilmour (1986), p. 15; Levitskiy (1986), p. 83.

In a **fast-wave device**, fast-wave interactions occur in large, smooth, multimode waveguide channels and thus involve higher order, overmoded resonators with large magnetic field requirements. Such a configuration gives much higher efficiency and greater output power compared with conventional slow-wave devices. The main representative of fast-wave device is the gyrotron. *SAL*

Ref.: Currie (1987), p. 448.

Gunn effect device (see **DIODE, Gunn**).

A **gyroresonance electronic microwave device** is an electronic device that use the interaction of a spiral electronic flux with an unslowed wave in a waveguide or cavities. They are also called *cyclotron resonance masers* or *gyrotrons*. Varieties of gyrotron devices include the gyroklystron, the gyro-TWT, the gyro BWT (backward-wave tube), and the gyro-twystron. *IAM*

Ref.: Andrushko (1981), p. 85; Gilmour (1986), p. 431.

IMPATT device (see **DIODE, IMPATT**).

A **MESFET (metal semiconductor field-effect transistor) device** is a field-effect transistor using a metal semiconductor structure.

Ref.: Currie (1987), p. 411; Golio (1991).

A **MITATT (mixed tunneling and transit time) device** is the microwave device using the mixture of tunneling and impact-ionization effects. It may be considered as a cross between the IMPATT and tunnel devices. *SAL*

Ref.: Currie (1987), p. 403.

A **MODFET (modulation-doped field effect transistor) device** is a field-effect transistor using undoped GaAs material and the AlGaAs layer to form a heterojunction which traps a two-dimensional gas between the conduction-band discontinuity and the bending of the conduction band due to charge structure. *SAL*

Ref.: Currie (1987), p. 411.

A **MOSFET (metal oxide semiconductor field effect transistor) device** is a field-effect transistor with a metal-oxide-semiconductor structure. *SAL*

Ref.: Currie (1987), p. 478.

A **negative-resistance device** is a microwave device having the falling portion of voltage-ampere curve (AB in Fig. D35) which is characterized by the negative resistance to an alternating current. The availability of such a characteristic makes possible the generation of continuous waves. The typical example of negative-resistance device is the tunnel diode. *SAL*

Ref.: Leonov (1988), p. 50; Fink (1975), p. 16.43.

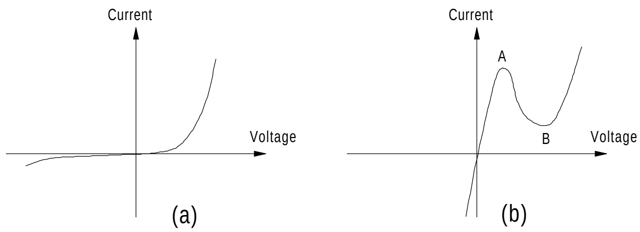


Figure D35 Voltage-ampere characteristic of (a) conventional and (b) tunnel diodes (after Leonov, 1988, Fig. 2.15, p. 52).

A **piezoelectronic device** is based on the piezoelectric effect (the appearance of electrical charges of opposite signs at opposite ends of piezoelectric materials when mechanical pressure is applied). In radar applications these are used primarily in oscillators and as transducers in acoustic-wave delay lines. *IAM*

Ref.: Fink (1982), p. 13.71; Zherebtsov (1989), p. 164.

A **relativistic device** is based on the interaction of a rectilinear relativistic electron flow at the input with rotating electromagnetic fields. Such devices have no mechanism for phase grouping of electrons, which is the basic operating principle of conventional power microwave vacuum-tube devices. The relativistic flux formed in the gyrotron enters along the axis into a round circular-sweep cavity in which the rotating field E_{110} is excited. Under the action of the transverse RF magnetic field, the electron flux is deflected from the axis in accordance with the transit phase of the cavity. The deflected electrons enter the slot of the cavity in the form of a rectangular waveguide convoluted in a ring, where they give up their energy as a result of interaction with the field of the main wave type. There are device circuits that use conversion of the drift velocity of the electrons into oscillating velocity of the relativistic flux, which is first swept in a modulator, in a section of a monotonously growing or periodic magnetostatic field. Pickup of energy from the electrons is effected on the condition of gyro-resonance in the cavity and does not require the creation of high field intensities.

Specific features of devices based on relativistic fluxes include the very high intensity of the excitation field (more than 300 kW), the high currents of electron flux (200 to 1000 A), and the output power of more than 600 kW. The type of interaction used in the devices makes it possible to practically completely convert the kinetic energy into the energy of the electromagnetic field in the decimeter and centimeter bands, which assures an efficiency of the devices of more than 90%.

Relativistic devices can perform amplifier and oscillator functions. *IAM*

Ref.: Kuraev (1986), p. 169.

A **semiconductor device** is based on the use of semiconductor properties. The basic types of microwave semiconductor devices can be divided into nonjunction (Gunn-effect devices), single-junction devices (diodes), and two-junction (bipolar transistors). Field-effect transistors can be single-gate or double-gate (field-effect tetrode). Compared with electron tubes, semiconductor devices have vital advantages: low mass and size, absence of filament energy expenditures, higher reliability and operating life, mechanical strength, efficiency, capability of operation at low supply voltages, and lower cost. For this reason, in most radars semiconductor devices are used primarily in receiver circuits, and partially in transmitting circuits (or entirely, in the case of solid-state radars).

With an increase in frequency, the increasing influence of parasitic inductances, capacitances and inertial processes in electrical junctions have a negative effect on operation of semiconductor devices. For this reason, design and technological measures directed towards reducing the capacitance of junctions and the transit-time of the charge carriers are the chief features of development and production of microwave semiconductor devices. In terms of design and technology of production, semiconductor devices are subdivided into monolithic, which are made in a single technological process with integrated circuits of solid-state instruments (most often gallium-arsenide, field-effect transistors), and discrete (individual), while the latter are subdivided into housed and nonhoused. The design of the housing and leads makes it possible obtain low parasitic inductances and capacitances of the devices and makes them convenient to connect to microwave transmission lines. *IAM*

Ref.: Fink (1982), p. 6.83; Gassanov (1988), p. 71; Zherebtsov (1989), p. 11.

A **slow-wave device** is a microwave device using the_slow-wave structure as a part of its configuration. All main tubes such as the backward-wave tube, klystron, magnetron, traveling-wave tube, and some others are related to slow-wave devices. *SAL*

Ref.: Currie (1987), p. 448.

A **solid-state device** is a microwave device based on solid-state technology. The main solid-state devices used in radar applications are semiconductor devices and integrated circuits. Compared with vacuum-tube devices, solid-state devices are significantly smaller and lighter and use much less power. *SAL*

Ref.: Kraus (1980); Leonov (1988), p. 51; Tarter (1985).

A **surface-acoustic-wave (SAW) device** is based on excitation of surface acoustic waves in piezoelectric materials, propagation of these waves in them, and interaction of the waves with electrons. Conversion of a radio signal into SAW and back is usually effected by interdigital transducers, two rows of interleaved electrodes deposited on the surface of the piezoelectric substrate with constant or variable spacings in the order of the acoustical wavelength.

SAW devices are used in a frequency band of 1 MHz to 1 to 3 GHz as delay lines (see **DELAY LINE, SAW**), filters (see **FILTER, SAW**), oscillators (see **OSCILLATOR, SAW**), and signal-processing devices.

Circuits based on SAWs are quite effective for matched filtering of phase-coded waveforms, including pseudonoise signals. Operation of a SAW convolver, an analog instrument that performs programmed conversion operations and matched filtering of pulse-compression waveforms in real time, is based on nonlinear interaction of acoustic waves. An extended metal plate applied on the surface of a crystal between two interdigital, to which the signals to be convoluted are applied, is the output electrode of the convolver.

The small size and weight, with high operating reliability and relatively low cost, and the combination of high-speed and wide bandwidth in signal-processing circuits, are advantages of SAW devices.

Industrial production of SAW devices requires high precision of manufacture of topological circuits, and uniformity and stability of parameters of the substrate. Shortcomings of the devices include the difficulty of suppression of parasitic signals with consequent difficulties in repeatability of the devices and high losses of SAW-based programmable devices. *IAM*

Ref.: Fink (1982), p. 9.22; Gassanov (1988), p. 213.

A **transferred-electron device** is one employing the Gunn effect in its operation.

A (**vacuum-)tube device** is an electronic and radio device whose operation requires a high vacuum or the atmosphere of a particular gas (or mixture of gases) at a specific pressure. For this reason, the operating volume of a vacuum-tube instrument is insulated from the surrounding space by a gas-tight glass or metal shell (bulb). In operating principle, vacuum-tube devices are subdivided into nondischarge types, in which the current flows only over conductors inside the device (luminescent lights, vacuum thermal elements, etc.), electronic types (electronic tubes, cathode-ray tubes, etc.), and gas-discharge tubes (thyratrons, etc.).

In terms of type of control of the electron flux, vacuum-tube devices are subdivided into devices with electrostatic control (grid-control tubes) and those with dynamic control (see Fig. D36). In the latter, the electrical field formed by the microwave oscillations is used to change the speed of electron flux. In terms of the nature of the energy exchange, electronic devices are subdivided into O- and M-type devices, which are also termed *linear-beam tubes* and *crossed-field tubes*, respectively. In O-type devices, the kinetic energy of the electrons is converted into the energy of a microwave field as a result of slowing of the electrons by this field. The magnetic field either is not used entirely or is used only to focus the electron flux and does not have primary significance for the process. In M-type devices the potential energy of the electrons becomes microwave energy. Gyroresonance devices make up a special group of devices based on the effect of coherent slowing radiation of electromagnetic waves by grouped electrons in a homogeneous magnetic field. In terms of productivity of interaction with the microwave field, the devices are subdivided into electronic devices with distributed and concentrated interaction.

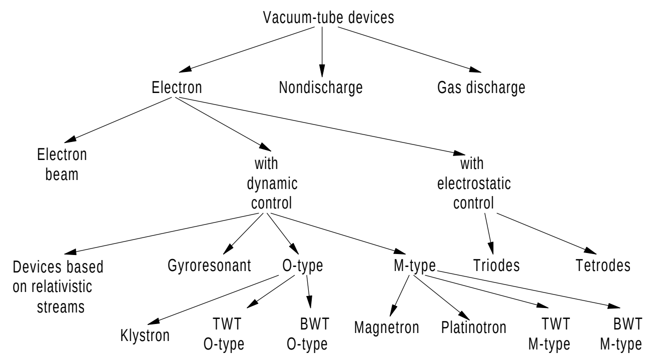


Figure D36 Classification of vacuum-tube devices.

Electronic vacuum microwave devices are still widely used in radar equipment as powerful amplifiers and oscillators in transmitters because of their high power-handling capacity (up to tens of MW), although in many applications they are being replaced by solid-state devices offering less bulk and greater reliability and safety. *IAM*

Ref.: Skolnik (1970), Ch. 7; Popov (1980), p. 490; Andrushko (1981), p. 11; Zherebtsov (1989), p. 206.

DIAGRAM

coverage diagram (see **COVERAGE**).

An **environmental diagram** is the pictorial representation of a clutter model. An example of the environmental diagram for air-defence radar is given in the Fig. D37. *SAL*

Ref.: Nathanson (1990), p. 293.

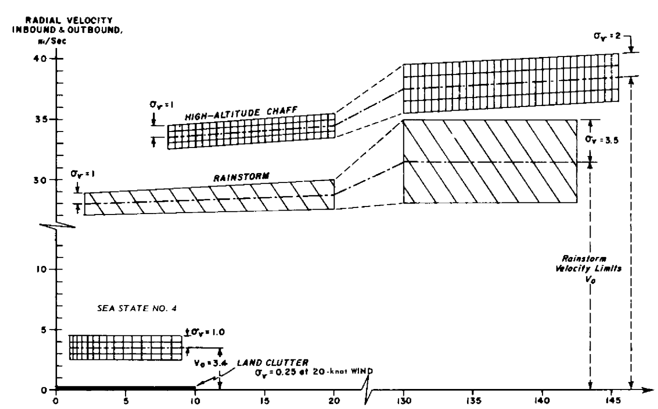


Figure D37 Environmental diagram for air-defense radar (from Nathanson, 1969, Fig. 8.6, p. 294, reprinted by permission of McGraw-Hill).

The **DICKE FIX** is "an ECCM technique that is specifically designed to protect the receiver from jamming. The usual configuration is a broadband amplifier followed by a limiter and then an IF amplifier of optimum radar signal bandwidth." *SAL*

Ref.: IEEE (1990), p. 10; Johnston (1979), p. 58.

DIELECTRIC. A dielectric is "a material that can withstand high electric stress without appreciable conduction." The dielectric properties of a material are characterized by several parameters, basic to which is the dielectric constant. Dielectrics are represented primarily by insulators with low conductance, but semiconductors also have dielectric properties. The only perfect dielectric, with zero conductance, is an absolute vacuum. *SAL*

The **(relative) dielectric constant** is "the ratio of the force between electrical charges in vacuum to the force between them in a specified medium" (a dimensionless constant, usually denoted by ε_r). The word "relative" is usually omitted but is understood. In microwave components it can be considered as the ratio by which the capacitance is increased when a given dielectric material replaces a vacuum between two electrodes. Other terms that are used interchangeably are *relative permittivity* or *relative capacity* (see **PERMITTIVITY**). *SAL*

Ref.: Fink (1982), p. 1.10; Jordan (1985), pp. 4.12, 33.5; Popov (1980), p. 118.

DIFFRACTION is "the deviation of the direction of energy flow of a wave when it passes an obstacle, a restricted aperture, or other inhomogeneities in the medium." It appears as a modification of the free-space electromagnetic field resulting from the interaction of the radar wave with surfaces in the vicinity of the direct path or obstacles in the direct path. The amount of diffraction is a function of the wavelength of the radiation relative to the size of the interfering object; the larger this ratio, the larger the diffraction effect. In considering radar propagation, there are two basic types of diffraction: smooth-sphere diffraction, when waves are diffracted around the curved earth, and knife-edge diffraction, which occurs in the presence of obstacles projecting above the surface. At microwave frequencies, smooth sphere diffraction into the shadow zone is negligible, but very low frequency (VLF); that is, long wavelength, electromagnetic waves are diffracted around the curved earth and provide the basis for worldwide communications. Knife-edge diffraction occurs in the presence of an obstacle such as a hill, tower, or building, which project above the smooth-earth radar horizon in the path of the electromagnetic wave. Behind the mask angle, due to wave-diffraction effects, the radar propagation factor drops to zero more slowly than for the smooth sphere case, enabling some radars to detect targets in the masked, or shadow, region. As with smooth-sphere diffraction, this effect is more pronounced for low-frequency radars, but it has been exploited at microwave frequencies by communications systems. The main radar types that take advantage of diffraction effects are high-frequency surface-wave radars. (See also **PROPAGATION**.) *PCH*

Ref.: IEEE (1993), p. 347; Barton (1988), pp. 297–302; Fink (1982), pp. 18.75–18.82; Jull (1987); Macnamara (1990).

DIODE, microwave. A microwave diode is a two-electrode microwave device. In its solid-state form it uses an electrical junction as a fundamental element of its structure. The junction in this type of diode is formed at the place of contact between metal and semiconductor or at the boundary between two conductors with different types of conductivity (electron conductors are n-type and hole conductors are p-type). These junctions are characterized by the existence of a potential barrier that permits current flow in only one direction. The electrical junction may arise even in the absence of a statistical p-n junction (the Gunn effect). The majority of diodes are made with germanium (Ge), silicon (Si), or gallium arsenide (GaAs) by introducing into a monocrystal a precise quantity of impurities from other elements, thus altering the conductivity. The choice of the semiconductor material, the type of impurity and their distribution is determined by the purpose of the diode and the range of its application. Depending on the application, microwave diodes are typically categorized as belonging to one of the following classes: detector, mixer, oscillator, multiplier, and switching diodes. Each class of diode generally includes diodes of various types. Table D6 shows the main function and limiting conditions of application for various types of microwave diodes.

Table D6
Microwave Diodes

Type of diode	Negative conductivity ?	Application	Max. frequency (GHz)	Max. average power (W)
PN	no	detector	20	0.1
PIN	no	mixer	20	10
Schottky barrier	no	detector, mixer	300	0.1
tunnel	statistical	oscillator (amplifier)	100	10^{-5}
Gunn	dynamic	oscillator (amplifier)	150	1
IMPATT	dynamic	oscillator (amplifier)	200	5
TRAP-ATT	dynamic	oscillator (amplifier)	10	10

The construction of the diode is determined by the operating band and the means by which it is incorporated into the waveguide. There are socketed and coaxial housed diodes and also unhoused diodes. *IAM*

Ref.: Fink (1982), pp. 7.38–7.42; Jordan (1985), p. 18.12; Gorbachev (1968), p. 5; Gassanov (1988), p. 72; Andrushko (1981), pp. 102, 120.

An **avalanche transit-time (ATT) diode** has negative conductivity in the avalanche breakdown mode. A layer of avalanche multiplication of carriers appears at the "breakdown voltage" at the junction between strongly alloyed p- and n-regions. A cluster of electrons formed in this layer drifts through the depleted n-region to the cathode and gives energy to the applied microwave field during its negative half-period. For certain dimensions of the drift region and the wave period the microwave energy is amplified, and the resulting mode of operation is referred to as "IMPact Avalanche and Transmit Time," (IMPATT).

Another mode sometimes used in diodes is "TRapped Plasma and Avalanche Triggered Transit" (TRAPATT), in which the region of impact ionization, moving along the diode, quickly fills the entire depleted region of the electron-hole plasma. The impedance of the diode drops sharply, and a strong current impulse flows through it; the charge is then resorbed and the impedance of the diode increases.

Avalanche transit-time diodes are distinctive in having high noise levels. They are used in medium- and high-power amplifiers and oscillators at frequencies up to 200 GHz for IMPATT diodes, and 10 GHz for TRAPATT diodes. *IAM*

Ref.: Tager (1968), p.17; Fink (1975), p. 8.38; Skolnik (1980), p. 217.

In a **back diode**, the reverse current exceeds the forward current due to a tunnel effect in the presence of a reverse bias on the diode. The impurity concentration is high (about 10^{18}/cm^3), but less than in a tunnel diode. The voltage/current characteristic is analogous to that of a tunnel diode, without the negative impedance portion. Back diodes operate at higher frequencies than normal p-n diodes.

Back diodes can be used as detector diodes, but due to low dielectric strength and difficulty of manufacture, they are rarely used. *IAM*

Ref.: Fink (1982), p. 9.65; Zherebtsov (1989), p. 133.

A **charge storage diode** has a nonuniform distribution of acceptor impurity throughout the semiconductor, leading to an increased concentration of minority carriers near the boundary of the junction when direct current is applied to the diode. When the polarity of the applied voltage is suddenly reversed, there arises a short, strongly nonsinusoidal impulse of negative current, caused by the stored carriers. Such a diode is used in frequency-multiplier circuits and is distinguished by having a high gain. *IAM*

Ref.: ITT (1975), p. 19.5.

A **converter diode** is used in a receiver mixer (see **mixer diode**) or transmitter modulator (see **modulator diode**). These applications usually use the nonlinear capacitance of the diode. *IAM*

A **detector diode** is intended to detect (demodulate) signals, and is characterized by the following basic parameters: its current and/or voltage sensitivity to applied microwave power, which indicates it effectiveness in converting microwave oscillations into direct current; the normalized reverse voltage, at which the reverse current achieves its limiting value; its maximum microwave power dissipation; and the noise ratio, or relative noise temperature, equal to the ratio of the diode's noise power in the operating mode to the noise power in a matched load.

Detector diodes are often Schottky-barrier diodes (sometimes back-diodes). The current sensitivity for point diodes is 1 to 5 A/W. Schottky-barrier diodes have a current sensitivity of 3 to 10 A/W and a voltage sensitivity of 1 to 50 V/W in the millimeter waveband, and 1,000 to 5,000 V/W in the centimeter waveband. The construction of a detector diode depends on the frequency band. For waveguide and coaxial structures used in the decimeter and centimeter wavebands, diodes in a socketed case are used, and in the millimeter waveband, the diodes are in cases which are a type of waveguide insert. In hybrid circuits, unhoused diodes are used. *IAM*

Ref.: Gassanov (1988), p. 76; Fink (1975), p. 9.61.

A **Gunn(-effect) diode** has a section with negative impedance at microwave frequencies and is a semiconductor crystal without a p-n junction. Its operation is based on the Gunn effect. The conductor used, most often gallium arsenide, has two conductivity zones and does not possess rectification properties. It is used to generate and amplify microwave oscillations (see **AMPLIFIER, Gunn diode**).

The operating parameters of a Gunn diode are determined by the origination and migration of domains in accordance with the diode's operating mode. The different modes are the transmit mode, in which the period of the generated oscillations is equal to the transit time of the domain, modes in which the domains are suppressed or delayed, and a mode with limited volume charge buildup, which is the most common mode. The operating frequency in this last mode is several hundred gigahertz, with tuning over an octave band, with continuous-wave power on the order of several watts (at 15 - 20 per cent efficiency) and impulse power on the order of several kilowatts. *IAM*

Ref.: Gassanov (1988), p. 186; Andrushko (1981), p. 120; Skolnik (1990), p. 5.11; Fink (1975), p. 9.70.

IMPATT diode (see **avalanche-transit-time diode**).

An **integrated diode** is the diode implemented on the basis of integrated circuit technology when any one of the semiconductor junctions forming the monolithic circuit structure can be used as a diode. *SAL*

Ref.: Fink (1975), p. 8.37.

A **light-emitting diode (LED)** is a diode using the effect in which the p-n junction can emit visible light when biased into the avalanche-breakdown region. They find the use in visual displays. *SAL*

Ref.: Fink (1975), p. 7.35.

A **mixer diode** utilizes the property of nonlinear resistance. The fundamental parameter is the normalized noise factor, characterizing the noise properties of the receiver containing the mixer with the diode, and the noise ratio, which depends upon the noise factor (see **detector diode**). Other parameters include the normalized reverse voltage and the maximum dissipated power, along with the usual detector diode parameters.

Typical values for the normalized noise factor are 5 to 8 dB in the centimeter waveband and 10 to 12 dB in the millimeter waveband. The basic types and features of diodes used as mixer diodes are the same as for detector diodes. *IAM*

Ref.: Van Voorhis (1948), p. 141; Gassanov (1988), p. 79.

A **modulator diode** is one that is used to change the parameters of a continuous or pulsed signal. In radar modulators, the modulation diodes are usually used for mixing at high power. For these purposes diodes with nonlinear capacitance are used (PN diodes and high-power PIN diodes). *IAM*

Ref.: Gorbachev (1968), p. 6.

A **multiplier diode** is used in a frequency multiplier or divider. In varactor multipliers and frequency-shift mixers that operate at up to several watts, p-n diodes in which the capacitance varies nonlinearly with voltage are usually used. The limiting frequency is 30 to 300 GHz, and the breakdown voltage in silicon or gallium-arsenide p-n multiplier diodes is up to 100 volts. In simple low-power frequency multipliers one finds Schottky-barrier diodes, which have a nonlinear impedance. In devices with large multiplying factors, charge storage diodes are used, due to the ability to obtain a short (0.1-ns) pulse of reverse current, which is rich in higher-order harmonics. Such diodes are limited to 10 to 20 GHz (for silicon) and 200 to 300 GHz (for gallium-arsenide), with power on the order of tens of milliwatts being possible at these frequencies. *IAM*

Ref.: Fink (1982), p. 9.69; Gassanov (1988), p. 82.

An **oscillator diode** is used in a solid-state microwave oscillator. It is differentiated from other devices in that it requires low power and is small in size, but exhibits an internal noise level several times higher than other devices. Oscillator diodes are usually avalanche transit-time diodes, tunnel diodes, or Gunn diodes. *IAM*

Ref.: Gorbachev (1968), p. 7.

A **parametric diode** is a variant of the varactor used in low-noise parametric amplifiers and weak-signal mixers, in low-noise frequency multipliers and dividers, in microwave limiters, and for electronic tuning of semiconductor microwave generators.

The limiting frequency for parametric diodes is 150-750 GHz. The usual operating mode is with a reverse bias and no conductance current, due to which the low noise level is obtained.

Parametric diodes may be Schottky-barrier diodes with power dissipation on the order of tens of milliwatts or p-n diodes with dissipation on the order of tenths of a watt. *IAM*

Ref.: Gassanov (1988), p. 81.

A **point-contact [-junction] diode** has a metal-semiconductor junction in which a tungsten or phosphor-bronze electrode is clamped against a semiconductor crystal. There are germanium, silicon, and gallium arsenide point-contact diodes. The desired voltage-current characteristic (see Fig. D38) is obtained through selection of the point of contact and the clamping force. The clamped contact results in a large variation of the junction parameters and mechanical instability, making point-contact diodes sensitive to vibration and shocks. These diodes also exhibit a high reverse current and low dielectric strength.

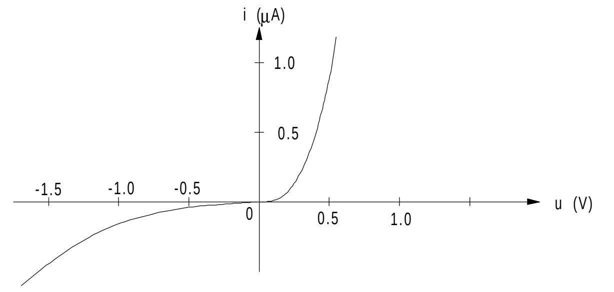


Figure D38 Voltage-current characteristic of a silicon point-contact diode (after Gassanov, Fig. 3.2, p. 74).

Point-contact diodes were used in older microwave detectors and mixers.

Related to point-contact diodes are back diodes developed with microalloy technology, which are distinguished by stable parameters and a high current sensitivity. *IAM*

Ref.: Gassanov (1988), p. 75; Fink (1975), p. 9.60.

A **positive-intrinsic-negative (PIN) diode** has strongly alloyed p- and n-regions, separated from one another by an extended i-region with a carrier concentration close to the concentration of fundamental carriers in a semiconductor. They are characterized by a small forward impedance and a large reverse impedance and a low capacitance with a weak voltage dependence.

They are used in switching diodes, attenuators, amplitude modulators, continuous and graduated microwave phase shifters, and also as powerful rectifier diodes.

With a wide i-region (0.1 to 0.5 mm), microwave PIN diodes may be used at voltages exceeding 1 kV and pulse power higher than 10 kW. *IAM*

Ref.: ITT (1975), p. 19.5; Fink (1982), p. 9.69; Andrushko (1981), p. 89; Brookner (1977), pp. 330, 384.

A **positive-negative (PN) diode** has a p-n junction, formed between two parts of a semiconductor that conduct holes and electrons. Such a diode conducts current in one direction only (see Fig. D39). The resistive loss is on the order of an ohm, and the junction capacitance depends on the voltage. Such diodes are distinguished by rather high breakdown voltages and are the diodes most commonly used in radar equipment. *IAM*

Ref.: Gorbachev (1968), p. 10; Jordan (1985), p. 18.10.

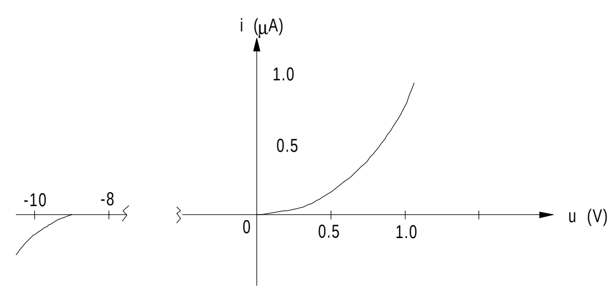


Figure D39 Typical voltage-current response for a PN diode.

A **pulse diode** is designed for use in pulse circuits, including pulse amplifiers and pulse generators. Pulse diodes can be PIN diodes, Schottky-barrier diodes, tunnel diodes, and also PN diodes of various types. *IAM*

Ref.: Erofeev (1989), p. 99.

A **Schottky(-barrier) diode** is based on a metal-semiconductor contact formed through the vacuum deposition of metal on the semiconductor (a Schottky junction). Most often, n-type gallium arsenide is used, with silicon being used more rarely. Due to the thinness of the n-layer formed by the metal junction, the slope of the voltage-current curve (see Fig. D40) and the dielectric strength are both higher than in point-contact diodes. The diode's parameters exhibit low variance and high stability. Schottky-barrier diodes have almost no injection or accumulation of minority carriers and therefore operate at higher frequencies and with less noise than PN diodes.

Schottky-barrier diodes are used as detector and mixer diodes at frequencies up to 300 GHz. The noise figure at 170 GHz is between 4.8 and 5.5 dB. These diodes are also used in frequency multipliers and converters and as low-power, fast-acting switching diodes. *IAM*

Ref.: Andrushko (1981), p. 90; Gassanov (1988), p. 75; Fink (1975), pp. 8.38, 9.62.

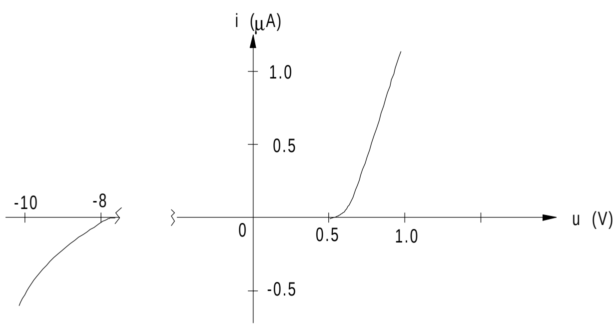


Figure D40 Voltage-current curve for a GaAs Schottky-barrier diode.

A **small-signal diode** is a diode operating in the small-signal mode as a switch, demodulator, limiter, nonlinear resistor, and so forth. *SAL*

Ref.: Fink (1975), p. 7.34.

A **switching diode** is used in microwave circuits in those cases where small size, fast switching speed, and low controlling power are required. Switching diodes may be PN, PIN or Schottky-barrier diodes. *IAM*

Ref.: Gorbachev (1968), p. 6.

TRAPATT diode (see **ATT diode**).

A **tuning diode** is a variant of the varactor diode used for electronic tuning of microwave filter circuits, amplifiers and oscillators, and also in continuous phase shifters.

The basic parameters of a tuning diode include the capacitance ratio $K_c = C_{max}/C_{min}$, its Q-factor, the capacitance for a given bias, the maximum reverse voltage, and the maximum microwave power.

The highest values for K_c (10 to 15) are obtained in diodes with an extremely sharp PN junction. The Q-factor for diodes with sharp p-n junctions can reach 300 to 400 for silicon diodes, and 500 to 800 for gallium arsenide diodes, at a frequency of 1 GHz and a reverse voltage of 4 to 6 volts. The Q-factor for diodes with extremely sharp junctions is several times smaller, limited to about 300, due to the large resistance of the low-alloy n-base. *IAM*

Ref.: Gassanov (1988), p. 83.

A **tunnel diode** exploits the tunnel effect in thin, strongly alloyed junctions. They exhibit a negative active conductivity (see Fig. D41) over a wide range of frequencies.

Tunnel diodes are made from germanium (Ge), silicon (Si), gallium arsenide (GaAs), and gallium antimonide (GaSb). The parameters of the voltage-current curve for a tunnel diode is determined by the materials used, as summarized in Table D7.

Among the parameters of a tunnel diode that are important in its application in amplifiers and oscillators are the resonant and cutoff frequencies. Above the cutoff frequency the resistance of the diode becomes positive, and amplification and signal generation become impossible. At the resonant frequency, the reactive component returns to 0; this is the most dangerous frequency in terms of exciting the amplifier. Tunnel diodes may or may not be encased. *IAM*

Ref.: Rudenko (1971), pp. 92–100; Fink (1975), pp. 8.38, 9.70.

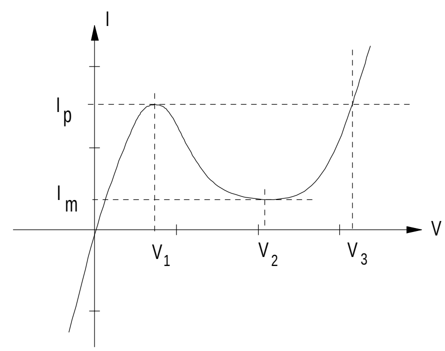


Figure D41 A typical voltage-current curve for a tunnel diode. I_p = peak current, I_m = minimum current (after Rudenko, 1971, Fig. 4.1, p. 93).

Table D7
Characteristics of Tunnel Diodes

Material	V_1, mV	V_2, mV	V_3, mV
Ge	40-70	270-350	400
Si	80-100	400-500	700
GaAs	90-120	0-600	1000
GaSb	30-50	200-250	450

A **varactor diode** (or variable-reactance diode) is a diode that uses the change in capacitance of a reverse-biased PN junction as a function of applied voltage. Typical applications of varactor diodes are harmonic generation, parametric amplification, and electronic tuning. *SAL*

Ref.: Fink (1975), p. 9.60.

A **Zener (breakdown) diode** is a semiconductor diode in which avalanche-breakdown current becomes well developed as the reverse potential is increased beyond the knee of the current-voltage curve (this occurs at the Zener voltage V_z). The typical use is in applications where a source of stable reference voltage is required. *SAL*

Ref.: Fink (1975), p. 7.34.

DIPLEX (mode) refers to the mode of operation of two transmitters alternately at two frequencies within the same frequency band, using a common antenna. One procedure is to pulse each transmitter at one-half of the desired pulse repetition frequency, 180° out-of-phase, while another procedure divides each pulse into two contiguous subpulses, generated by the two transmitters. The advantage is that higher total average power is possible because each transmitter is operating at one-half the total duty cycle of the waveform.

The return signals can be amplified in a common RF amplifier and separated into two individual channels before the mixer by means of RF filters, or after a wideband IF amplifier stage by means of IF filters matched to the bandwidth of each pulse. This mode of operation is advantageous for solid-state transmitters where peak power limitations more stressing than the average power limitations and their cost can be reduced if longer pulse duration can be tolerated. In some cases when unequal pulse lengths are employed, diplex operation allows the pulse duration to be more than doubled, reducing the peak power requirements. *SAL*

Ref.: Johnston (1979), p. 58; Skolnik (1990), p. 3.54.

DIPOLE. A dipole is a radiator of electromagnetic waves in the form of a thin conductor. It can radiate and receive RF waves, and typically its length is close to half the wavelength (in which case it is termed a *half-wave dipole*). In antenna theory, the concept of the Hertzian dipole is widely used. The Hertzian dipole is a thread in which the current meets the following requirements: (a) it is straight; (b) its length is extremely small relative to the wavelength; and (c) the amplitudes and phases of the currents are equal throughout the length of the conductor.

Practical dipoles can be of cylindrical, biconical, folded, sleeve, and other types and can be used as feeds for reflector and lens antennas or as antenna array elements. *SAL*

Ref.: Johnson (1984), Ch. 4; Sazonov (1988), p. 222.

DIRECTION FINDER, DIRECTION FINDING. Direction finding is "a procedure for determining the bearing, at a receiving point, of the source of a radio signal by observing the direction of arrival and other properties of the signal." The device used to determine direction of arrival is a direction finder (DF). DFs are typically classified as amplitude-comparison or phase-comparison types.

An amplitude-comparison DF extracts the information about the direction of signal arrival from the amplitude ratios in its antenna patterns. Typically, the modern amplitude-comparison DF, such as is used in an ESM system, includes a certain number of channels connected to directive antennas to form a monopulse network (Fig. D42).

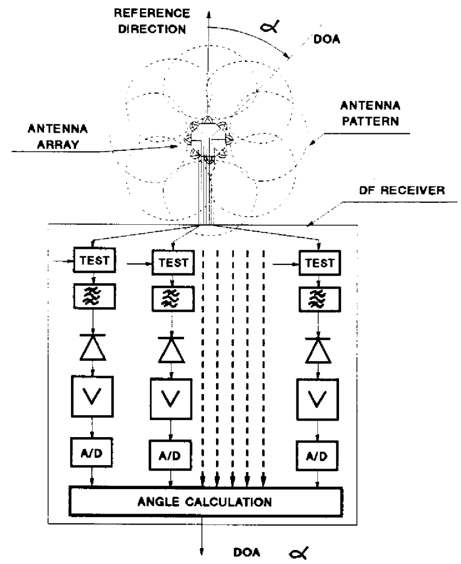


Figure D42 Amplitude-comparison direction finder from Neri, 1991, Fig. 4.20, p. 303).

The amplitude-comparison DF is less accurate than the phase-comparison DF, but it is used more extensively due to its lower complexity and cost.

A phase-comparison DF extracts the information about the direction of arrival of a signal from the phase differences in its antenna patterns (Fig. D43). In this case a phase shift ϕ carries an information about the angle-of-arrival direction α:

$$\varphi = \frac{2\pi L \sin\alpha}{\lambda}$$

where L is the base, and λ is the wavelength.

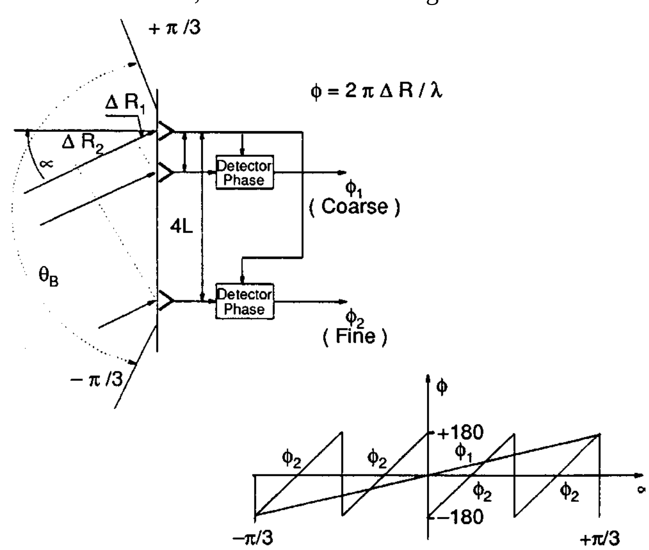


Figure D43 Phase-comparison direction finder (from Neri, 1991, Fig. 4.23, p. 306).

The main problem in application is similar to that encountered with interferometers and lies in measurement ambiguity. To remove ambiguity, a third channel is introduced, with a smaller baseline, capable of giving a coarse measurement. The greater angular width of antenna patterns of this channel makes it possible to remove the ambiguity in the more precise channel.

Sometimes the DF uses a remote station (e.g., a satellite) separated from a radar to receive jamming signals. In this way the range and azimuth of many jammers may be obtained by correlating the signals received from several stations.

The term *passive DF* is sometimes used to stress the passive nature of DF operation as an ECCM technique, or *active DF* to describe the angle measurement capabilities of an active radar antenna. *SAL*

Ref.: IEEE (1993), p. 1,059; Schleher (1986), p. 96; Neri (1991), pp. 302, 306; Jenkins (1991); Gething (1978).

DISCHARGER (see **CROWBAR**).

DISCRIMINATION, DISCRIMINATOR, of signals. In radar applications a discriminator is "a circuit in which the output is dependent upon how an input signal differs in some aspect from a standard or from another signal." Discriminators are categorized as amplitude, frequency, phase, angular, or time discriminators and their combinations (e.g., time-frequency discriminators), and as analog or digital. The first three are actually detectors of the corresponding signal properties (see **DETECTOR**). The angular discriminator uses the difference in response of two antenna channels to determine the angular coordinates of the target, and is a major component of monopulse radar (see **MONOPULSE**). Discriminators are used in radar estimators and trackers. *SAL*

Ref.: IEEE (1993), p. 367; Dulevich (1978), pp. 323–368; Johnston (1979), p. 65; Leonov (1986), p. 12; Nathanson (1990), p. 112.

angular discriminator (see **MONOPULSE**).

frequency discriminator (see **DETECTOR, frequency**).

phase discriminator (see **DETECTOR, phase**).

A **pulse-width discriminator** is a device that inhibits return signals whose time durations do not fall into some predetermined design tolerances. It can be used for clutter and jamming rejection. This technique is especially effective against long-pulse jamming and ECM with low-frequency noise modulation, but it affords little or no discrimination against short pulses, HF noise modulation, and barrage jamming. *SAL*

Ref.: Nathanson (1990), p. 112; Johnston (1979), p. 65.

target discrimination (see **TARGET RECOGNITION AND IDENTIFICATION**).

A **time discriminator** extracts information about time delay of a pulsed waveform to determine target range. Figure D44 depicts a diagram of a time discriminator with two unbalanced channels. The received pulse signal $y(t)$ is multiplied into two reference signals $u_1(t)$ and $u_2(t)$. These signals are delayed relative to the transmitted signal by $\tau_e + \delta$ and $\tau_e - \delta$. After multiplication, the signals pass through filters and are detected and are subtracted, forming a voltage at the discriminator output.

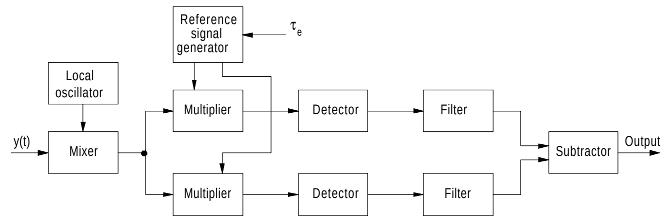


Figure D44 Time discriminator (after Tartakovskiy, 1964, Fig. 8.2, p. 447).

The most prevalent form of time discriminator is the split-gate time discriminator (Fig. D45). In this case the gate is split into the early and late gates, and when the split gate is positioned so that these gates capture equal shares of the target pulse, the difference of the two gate outputs is equal to zero. Then the gate can be said to be centered on the target. As the acceptance gate (range gate) is slaved to the split gate, the target pulse falls within the range gate and passes along for processing by the subsequent loops of radar receiver. *SAL, AIL*

Ref.: Bakut (1964), vol. 2, p. 446; Lothes (1990), p. 55.

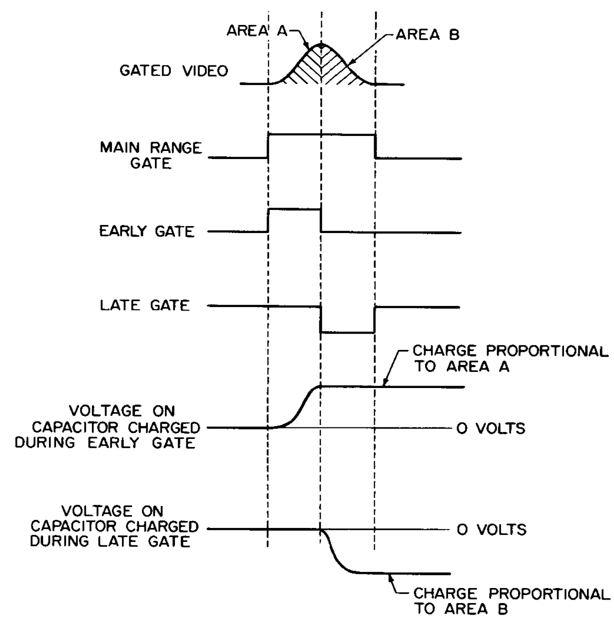


Figure D45 Split-gate time discriminator (from Skolnik, 1970, Fig. 41, p. 21.41, reprinted by permission of McGraw-Hill).

A **time-frequency discriminator** is a time discriminator for continuous-wave waveforms. Figure D46 is a block diagram of such a discriminator. It comprises time and frequency discriminators, as well as a reference voltage generator. Three reference voltages created by a common generator and carrying information on predicted values of time delay t_e and frequency f_e are applied to discriminator multipliers. Reference voltages are matched with the anticipated signal base using a frequency-modulation law. The input signal $y(t)$ is applied to

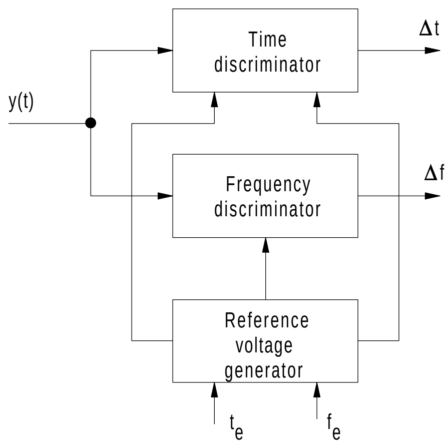


Figure D46 Time-frequency discriminator (after Shirman, 1981, Fig. 14.12, p. 214).

the time and frequency discriminators. Time-delay error is generated in the time discriminator, while frequency error is generated in the frequency discriminator. *AIL*

Ref.: Shirman (1981), p. 214.

Video discrimination is the process used to reduce the frequency band of the video amplifier stage in which it is used. *SAL*

Ref.: Johnston (1979), p. 68.

DISH (see **ANTENNA, reflector**).

DISPLAY, radar. A radar display is an electronic instrument for visual representation of radar data. Radar displays can be classified from the standpoint of their functions, the physical principles of their implementation, type of information displayed, and so forth. From the viewpoint of function, they can be detection displays, measurement displays, or special displays. From the viewpoint of number of displayed coordinates, they can be one-dimensional (1D), two-dimensional (2D), or three-dimensional (3D). An example of a 1D display is the range display (A-scope). Most widely used are 2D displays, represented by the altitude-range display (range-height indicator, or RHI), azimuth-elevation display (C-scope), azimuth-range display (B-scope), elevation-range display (E-scope), and plan-position indicator (PPI). These letter descriptions date back to World War II, and many of them are obsolete. From the viewpoint of physical implementation, active and passive displays are distinguished. The former are represented mainly by cathode-ray-tube (CRT) displays and semiconductor displays. Passive displays can be of liquid-crystal or ferroelectric types. In most radar applications CRT displays remain the best choice because of their good performance and low cost.

From the viewpoint of displayed information, displays can be classified as presenting radar signal data, alphanumerics, or combined displays. These can be driven by analog data (analog or raw-video displays) or digital data (digital or synthetic-video displays). Displays in modern radar are typically synthetic-video combined displays, often using the monitors of computer-based work stations. *SAL*

Ref.: IEEE (1993), p. 369; Poole (1966); Skolnik (1970), Ch. 6, (1980), Ch. 9; Bystrov (1985).

An **active display** is based on conversion of electrical energy into luminous energy through various physical effects (Table D8).

One of the basic parameters of active displays is the light output of the device, which defines the power consumed by the display at a normal level of brightness of 350 cd/m^2.

In contrast to passive displays, active displays have greater brightness, good multiplexing capabilities, and better quality of reproduction of color images. The CRT display is a common example of an active display.

Table D8

Types of Active Displays

Type of display	Physical effect	Max. light output lm/W
Electron beam	High voltage cathode luminescence	100
Vacuum luminescent	Low voltage cathode luminescence	5
Gas discharge	Radiation of gas discharge	5
Electro-luminescent	Pre-breakdown electroluminescence	20
Semiconductor	Injection electroluminescence	30

Semiconductor displays are active displays based on the effect of an injection luminescence that takes place when the carriers are recombined on a junction of the semiconductor crystal switched in the forward direction.

Semiconductor displays are also known as light-emitting diode displays. They are characterized by a low operating voltage, the ability to overlap with semiconductor logic circuits, small dimensions, a long service life, a high degree of pixel brightness, and a capability for multiplex addressing. Matrix displays with 6,000 to 40,000 elements have an 0.8-mm space between the elements, a brightness of 140 to 240 candles/m^2, a power consumption of 2 to 112W. *IAM*

Ref.: Bystrov (1985), p. 98; Fink (1982), pp. 11.55, 23.75.

An **alphanumeric (data) display** provides numerical and alphabetical information. Typically, character-modulating and character-synthesizing alphanumeric displays are distinguished. In the former case, a light or electronic beam is shaped into the form of a character (e.g., CRT displays with character-shaping matrices). An electronic beam passing through the lower part of the matrix “prints” the character on the screen. In character-synthesis displays the characters are formed based on the mosaic principle. In this case shaping takes place by means of an image mosaic independent of the controlled elements, each being a light-signal converter. The following displays are based on the principle of image shaping: segment character synthesizing displays, the elements of which are segments and grouped in one or several character locations; matrix character synthesizing displays, the image elements of which form an orthogonal matrix; and mnemonic character synthesizing displays: that is, displays designed to display information in the form of mnemocircuits.

Segment character-synthesizing displays only allow the display of numbers (numerical character-synthesizing displays), or numbers and letters (alphanumeric character-synthesizing displays). In a number of cases, scale and digital-analog character-synthesizing displays with an analog form of information presentation are used. Character-synthesizing displays can be constructed using gas discharge, semiconductor, and other types of displays. In particular, in CRT character-synthesizing displays the characters are synthesized from Lissajous figures. These types of displays can be used for displaying information provided by radar control and monitoring subsystems (see **SUBSYSTEM, radar**). *IAM*

Ref.: Poole (1966), Ch. 13; Bystrov (1985), p. 6; Stevens (1988), p. 4.

An **altitude-range display** shows target position in altitude (height)-range coordinates. The current of the horizontal sweep of the display is proportional to the horizontal range to the target; the current of the vertical sweep is proportional to its altitude. If a small elevation sector is displayed, time (slant range) may be used as the horizontal coordinate, and the sine of the elevation angle as the vertical coordinate.

When determining the altitude relative to the earth's surface, the current in the vertical deflection coils should have a supplemental parabolic component to compensate for the curvature of the earth and the refraction of the radio waves. In contrast to the elevation-range display, reading of the altitude becomes easier in altitude-range displays as the elevation increases. This display is also termed a *range-height indicator (RHI)*. *IAM*

Ref.: IEEE (1993), p. 1066; Druzhinin (1967), p. 415; Sauvageot (1992), p. 15.

An **amplitude display** is a display with an amplitude blip used to determine the angular coordinate. Amplitude displays are employed using the equal signal method and the method of dividing the envelope of the signal burst. In the first method the blips of the video signals of the target are observed on the CRT. These blips correspond to the extreme positions of the beam scanning in the plane determined by the angular coordinate. The operator or the automatic tracker turns the antenna until the heights (amplitudes) of the blips are aligned. In the method of dividing the envelope a temporary angular sweep is created in the amplitude display along a specific angular coordinate synchronized with the shifting of the beam in the plane of that coordinate. The operator or automatic tracker reads the coordinate according to the position of the center of the signal burst blip on the sweep line or turns the antenna until it is pointed at the target along the specified coordinate. *IAM*

Ref.: Skolnik (1962), p. 391; Barton (1964), p. 6; Rakov (1970), vol 2, p. 385.

An **amplitude-differential display** is a display with an amplitude blip for determining the angular coordinate based on the observed difference of the video signal voltages, obtained at the two extreme positions of the beam, scanning in the plane of the specified angular coordinate. The amplitude-differential display is used with the equal signal method of measurement. The operator (or automatic tracker) turns the antenna until the difference of the mean values of the video signals is zero.

In two-dimensional displays the azimuth and elevation discrepancy signals are fed to the horizontal and vertical deflecting plates of the CRT. This type of display is also called a *K-scope*. *IAM*

Ref.: Barton (1964), p. 8; Vasin (1977), p. 193.

analog display (see **raw-video display**).

An **azimuth-elevation display** is a two-dimensional display for presenting and measuring the azimuth and elevation based on the bright blip of the target. The sweep of the beam in the CRT display is carried out by two sweep generators that process voltage, the linearity of which depends on the angle of rotation in the plane of the azimuth and elevation, respectively. The azimuth-elevation display has no range resolution. Therefore, for the individual monitoring of targets that have the same angular coordinates, gating of the receiver is used for the time of the reception of the reflected pulse of the target located at the specific range. In this case the capability for simultaneous monitoring of all the targets in the scanning sector is lost. For the approximate determination of the range to the dot of the blip, dashes are added to the right and the left, the length of which is inversely proportional to the range. This type of display is also termed a *C-scope*. *IAM*

Ref.: Barton (1964), p. 7; Vasin (1977), p. 195

An **azimuth-range display** is a two-dimensional display for the recording and measurement of azimuth and range based on the bright blip of the target. In azimuth-range displays, either a panoramic sweep (plan-position indicator) or a sweep in rectangular coordinates (B-scope) is used. In the latter case, the azimuth sweep is shaped by voltage that is a function of the angular displacement of the antenna based on the azimuth. This is achieved using a transmitter in the form of a variable capacitor, a circular potentiometer, or a synchro in the sweep generator.

The advantages of the azimuth-range rectangular sweep are the even and higher resolution in the azimuth and the ease in changing the scale, which gives it preferred usage in guidance radar compared with the PPI often used in search radars. *IAM*

Ref.: Barton (1964), p. 7; Vasin (1997), p. 193.

An **A-(scope) display** is one "in which targets appear as vertical deflections from a horizontal line representing a time base (Fig. D47). Target distance [range] is indicated by the horizontal position of the deflection from one end of the time base. The amplitude of the vertical deflection is a function of the signal intensity."

Ref.: IEEE (1993), p. 18.

A **bright display** is used in applications where it is not possible or convenient to use a conventional CRT in a darkened environment. An example of such applications is the aircraft

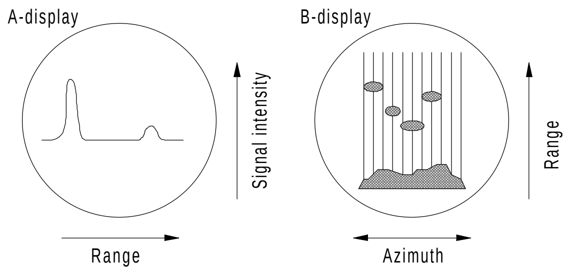


Figure D47 A- and B-displays.

cockpit or airfield control tower. In these cases a television-type display, one using a direct-view storage tube, a dark-trace display, or digital (plasma) panel may be used. *SAL*

Ref.: Skolnik (1980), p. 357.

A **B-(scope) display** is "a rectangular display in which each target appears as an intensity-modulated blip, with azimuth indicated by the horizontal coordinate and range by the vertical coordinate" (Fig. D47).

Ref.: IEEE (1993), p. 100.

A **cathode ray tube (CRT) display** is an active display based on the phenomenon of luminescence of a display screen under the effect of an electronic beam. Monochrome CRT displays are more economical and usually have a high resolution (around 0.1 mm^{-1}). They are usually constructed in the form of a CRT containing projectors in the cylindrical part to form one or several beams and a screen covered with phosphor in the wide part, as well as a beam-deflection system.

The basic characteristics of monochromatic CRT displays are a brightness of 100 to 200 candles/m^2, accelerating voltage of 12 to 17 kV, and a total length not in excess of 30 cm. The diagonal measurement of the screen is from 2.5 to 70 cm.

A color image is achieved at the expense of a loss of resolution or brightness at the same power consumption (see **color display**). More compact models of CRT displays with a flat construction are being used. As a rule, they have worse characteristics when the diagonal dimension is 26 to 69 cm. To shape an image with a diagonal dimension of 100 to 250 cm projection CRTs containing additionally an optical display system with a small luminescent target screen on a large screen are used. Vacuum luminescent displays, which are essentially distinguished by voltage level, material, and construction, also fall into the category of CRT displays (see **CATHODE-RAY TUBE**). *IAM*

Ref.: Skolnik (1962), p. 391; Poole (1966); Bystrov (1985), p. 41.

A **color display** presents information in color. The parameters that determine the quality of a color indicator are the number of colors, the brightness of the colors, and the speed with which they can be switched. A number of CRT displays, gas discharge, and liquid crystal displays have good color properties. CRT displays with digital phosphor screens possess the best characteristics based on the number and quality of colors. This design comprises the totality of the dots and bands of the three primary colors (blue, yellow, red).

In radar, the use of color displays makes it possible to visually observe a broader dynamic range, of parameters and use color as a third coordinate; for example, azimuth, range, and intensity (color gradation). *IAM*

Ref.: Poole (1966), Ch. 3; Skolnik (1980), p. 357; Bystrov (1985), p. 43; Mel'nik (1980), p. 187.

A **combined display** simultaneously presents various data (dynamic and statistical). Combined radar displays usually display coordinate (signal) data simultaneously with alphanumeric data concerning a large number of objects, and statistical data (map contours, routes, etc.). Combined displays employ special CRTs with several guns operating on a common screen and having optical windows for the projection of statistical data. One of the guns can operate in the character-printing mode. The combining of statistical and dynamic data is simplified using the conversion of a radar image into a television image by entering character data during the flyback time and by shaping synthesized radar images based on digital methods of data processing and data readout. *IAM*

Ref.: Poole (1966), p. 227; Finkel'shteyn (1983), p. 519; Stevens (1988), p. 4.

A **C-(scope) display** is "a rectangular display in which each target appears as an intensity-modulated blip with azimuth indicated by the horizontal coordinate and angle of elevation by the vertical coordinate" (Fig. D48).

Ref.: IEEE (1993), p. 168.

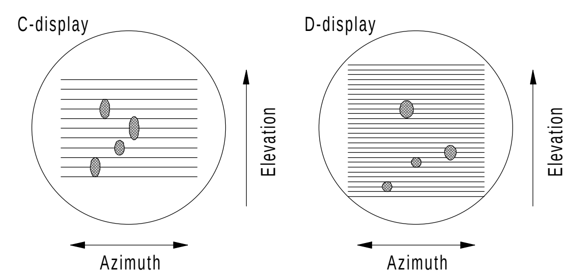


Figure D48 C- and D-displays.

A **dark display** is a (dark-trace) CRT display with data display resulting from the contrast between the sections of reflection and the absorption of light. During irradiation by an external light source, the color of the phosphor of the CRT is usually yellow-white. During recording by the electronic beam, absorption occurs in the wavelength interval in the visible range, as a result of which the recording places appear dark. Dark displays possess a long-term memory (up to several months). The data are erased by heating the screen with a built-in spiral or using an external infrared source. *IAM*

Ref.: Ridenour (1947), p. 483; Skolnik (1970), p. 6.12.

A **detection display** is designed to determine the presence of a target in the radar scanning coverage without indication of the precise coordinates. Visual or acoustic methods may be used. In the first case light signal boards and other displays are used. In acoustic "displays," detection is indicated by a sound signal; for example, by the change of the tone in the telephones.

Detection displays are usually used as auxiliary displays along with regular displays and also as independent displays in passive warning and signals intelligence radar. *IAM*

Ref.:Barton (1964), p. 6; Vasin (1977), p. 180.

A **digital display** is (a) a display driven by the digital output of a radar digital signal processor or (b) a display that presents data in digital form. In the first case it is typically a synthetic-video display that gives great flexibility in presenting radar data. In the second it may be an alphanumeric display using a digital panel or digital gas-discharge display. *SAL*

Ref.: Poole (1966), Ch. 13; Popov (1980), p. 472; Bystrov (1985), p. 6.

A **D-(scope) display** is "similar to a C-display, but composed of a series of horizontal stripes representing successive elevation angles (Fig. D48). Each stripe is a miniature B-display with compressed vertical scale. Horizontal position of a blip represents azimuth, the gross vertical scale (the stripe in which the blip appears) represents elevation, and vertical position within the stripe represents range." The term is obsolete or rare.

Ref.: IEEE (1993), p. 311.

An **electro-luminescent display** is based on the radiation of light by a body under the effect of an electrical field. In electro-luminescent displays the flashover luminescence of powder or film electrophosphors is used while the intensities of the field are close to the breakdown point. Semiconductor displays are also electro-luminescent. Structurally, the displays are constructed by depositing layers of phosphor, dielectric, and electrodes on a base layer of glass. A color display is achieved by depositing the necessary number of phosphors with transparent electrodes for each layer.

The basic characteristics of the various types of electro-luminescent displays are within the limits: resolution 0.3 to 1 mm^{-1}; brightness 2 to 100 candles/m^2; contrast (5 to 50):1; operating voltage 80 to 350V.

Electro-luminescent displays have a very small mass and thickness, they have good multiplexing capabilities, but at the same time they are characterized by a number of shortcomings: large dimensions and limited number of colors. *IAM*

Ref.: Poole (1966), Ch. 7; Fink (1982), p. 11.55; Bystrov (1985), p. 87.

An **elevation-range display** uses rectangular sweep to record the position of the target in the elevation (vertical sweep) and range (horizontal sweep) coordinates. Its construction is analogous to the azimuth-range display. To read the target altitude, lines of equal altitudes have been drawn on the transparent light filter in front of the display screen (CRT) in the form of a hyperbola.

Basic shortcomings of the display: at great distances it is difficult to distinguish the lines of equal altitude from one another, which leads to an increase in errors in reading the altitudes; a significant part of the usable space of the display is not used. To correct these shortcomings two scales are often used for range and elevation. This display is also called an *E-scope*. *IAM*

Ref.: Barton (1964), p. 7; Druzhinin (1967), p. 411.

An **E-(scope) display** is "a rectangular display in which targets appear as intensity-modulated blips with range indicated by the horizontal coordinate and elevation angle by the vertical coordinate" (Fig. D49). Sometimes the term "E-display" has been applied to a display in which height or altitude is the vertical coordinate, but this usage is deprecated because of ambiguity and the "range-height indication" term for such a display is preferred.

Ref.: IEEE (1993), p. 403.

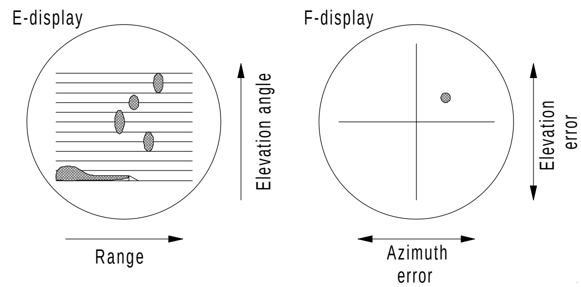


Figure D49 E- and F-displays.

A **ferroelectric display** is a passive display based on the phenomenon of double beam refraction in ferroelectric under the action of an electric field. Its special features are linearity of the effect and ability to achieve contrast in white light. Its drawbacks are low operating temperature (100 to 200K) and high voltage (3.6 kV to achieve a path-length difference of half the length of the wave).

For this reason, usually ferroelectric displays are used as light-reflecting or light-transmitting targets in a CRT (so-called light valves), which provides cathode ray commutation in a crystal matrix. Depending on the light source used, the contrast image of the moving object is sufficiently bright on a large screen (10 to 40 m^2). The dimensions of the targets are on the order of 30 to 40 cm, where the depth is 0.25 mm with a resolution of 10 mm^{-1}. *IAM*

Ref.: Poole (1966), p.113; Bystrov (1985), p. 150.

An **F-(scope) display** is "a rectangular display in which a target appears as a centralized blip when the radar antenna is aimed at it. Horizontal and vertical aiming errors are respectively indicated by horizontal and vertical displacement of the blip" (Fig. D49).

Ref.: IEEE (1993), p. 485.

A **G-(scope) display** is a modified F-display in which wings appear to grow on the blip, the width of the wings being inversely proportional to target range (Fig. D50). (Rare.)

Ref.: IEEE (1993), p. 551.

A **gas-discharge display** uses gas-discharge radiation. They operate in the glow discharge mode with a cold cathode at a gas pressure of several hundred pascals and a voltage of 100 to 200V. Monochrome and color displays with phosphor screens of the three primary colors are used. Structurally the displays are in the form of gas-discharge panels (matrix displays), forming a single unit with a great number of gas-discharge cells controlled by direct or alternating high-frequency current. In the latter case, the effect of an internal memory is

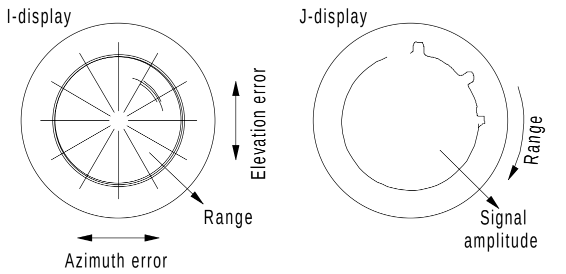


Figure D50 G- and H-displays.

achieved. The resolution of gas-discharge matrix displays reaches 25 elements/cm, brightness 40 candles/m^2, range of working temperatures 0 to 50°C.

Shortcomings of the gas-discharge displays are low stability of characteristics over the course of the service life, need for special measures to reduce the delay of the discharge glow, and spread of the luminosity from cell to cell. *IAM*

Ref.:Fink (1982), p. 23.74; Bystrov (1985), p. 72.

An **H-(scope) display** is "a B-display modified to include an indication of angle of elevation. The target appears as two closely spaced blips approximating a short bright line, the slope of which is in proportion to the tangent of the angle of target elevation" (Fig. D50). (Obsolete or rare.)

Ref.: IEEE (1993), p. 589.

An **I-(scope) display** is used "in a conical-scan radar, in which a target appears as a complete circle when the radar antenna is pointed at it and in which the radius of the circle is proportional to target range" (Fig. D51). (Rare.)

Ref.: IEEE (1993), p. 614.

A **J-(scope) display** is a modified A-display in which the time base is a circle, and targets appear as radial deflections from the time base (Fig. D51).

Ref.: IEEE (1993), p. 691.

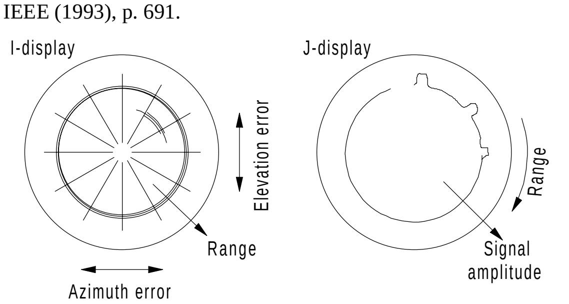


Figure D51 I- and J-displays.

A **K-(scope) display** is "a modified A-display used with a lobe-switching antenna, in which a target appears as a pair of vertical deflections. When the radar antenna is correctly pointed at the target, the deflections (blips) are of equal height, and when not so pointed, the difference in the blip height is an indication of the direction and magnitude of pointing error" (Fig. D52). (Rare.)

Ref.: IEEE (1993), p. 696.

An **L-(scope) display** is "similar to a K-display, but signals from the two lobes are placed back to back. A target appears as a pair of deflections, one on each side of a central time base representing range (Fig. D52). Both deflections are of equal amplitude when radar antenna is pointed directly at the target, any inequality representing relative pointing error. The time base (range scale) can be vertical, as in the L-display illustration, or horizontal." The L-display is also known as a bearing-deviation indicator.

Ref.: IEEE (1993), p. 707.

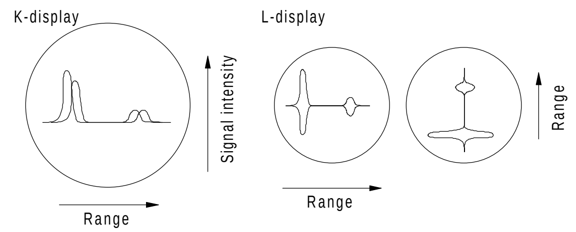


Figure D52 K- and L-displays.

A **liquid crystal display** is a passive display based on the use of the electro-optical properties of liquid crystals. At low voltages and extremely low power consumption the structure of the crystal changes. This is visually fixed as a result of the great anisotropy of the optical properties. The cell of the liquid crystal display contains two glass plates with semitransparent electrodes applied to them and a layer of liquid crystal between them. Depending on the effect employed the liquid crystal displays have the following parameters: reaction time within the limits 0.03 to 30 μs, relaxation time 0.1 to 1060 μs, operating voltage 3 to 100V, current density 1 to 10 μA/cm^2, operating temperatures –10 to +60°C.

The advantages of these displays are low power consumption, capability to operate under conditions of high levels of external background noise, capability for the manufacture of displays of extremely small thickness (a single micron), and practically any dimensions. Some of the main problems are increasing longevity and depth of colors in polychromatic displays. *IAM*

Ref.: Fink (1982), p. 23.75; Bystrov (1985), p. 108.

The **main display** is one used by operators to perform the basic task. For example, in detection radar, the main display is a plan-position indicator. In addition to the main display other displays are used (e.g., displays of control and monitoring subsystems). *IAM*

Ref.: Popov (1980), p. 269.

A **measurement display** provides precise measurement of the target coordinates. Depending on the number of simultaneously measured coordinates, measurement displays can be categorized as one-dimensional, two-dimensional, and three-dimensional, with the display in rectangular and panoramic coordinates (plan-position indicator). Measurement displays are characterized by a resolution that for two-dimensional displays is rated by the value of the dot (bright blip) in the coordinates being measured (kilometers, degrees), and by the scale of the image in the form of the ratio of the limit value of the scale to its length. To increase the accuracy of the measurements, the scale is enlarged by reducing the area being observed (see **sector display**). *IAM*

Ref.: Barton (1964), p. 7; Vasin (1977), p. 180.

A **mosaic display** is a complex display of a radar detection system that combines data from several nonsynchronized radar sets, each of which detects targets within an adjacent area. The mosaic display is based on the use of the principle of sweep conversion, which enables the operator to select any zone within the coverage of a large number of radar sets for monitoring on his own display.

Mosaic displays are used in air traffic control systems of large airports with intensive traffic. *IAM*

Ref.: Popov (1980), p. 236.

A **moving-map display** shows the current position of the moving object against a background of a map of the area projected on the display. This display system is provided by the use of an optical slide projector, which has computer-controlled selection and orientation of the slides. *IAM*

Ref.: Skolnik (1970), p. 31.19; Popov (1980), p. 152.

An **M-(scope) display** is "a type of A-display in which one target range is determined by moving an adjustable pedestal, notch, or step along the baseline until it coincides with the horizontal position of the target-signal deflection; the control that moves the pedestal is calibrated in range (Fig. D53). The use of the term M-display is uncommon. More often this display is defined as a variant of an A-display."

Ref.: IEEE (1993), p. 789.

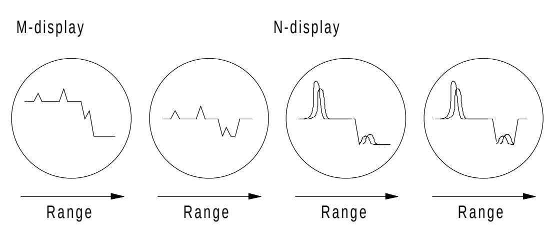


Figure D53 M- and N-displays.

An **N-(scope) display** is "a K-display having an adjustable pedestal, notch, or step, as in the M-display, for the measurement of range. This display is usually regarded as a variant of an A-display or K-display rather than a separate type" (Fig. D53).

Ref.: IEEE (1993), p. 838.

An **O-(scope) display** is "an A-display modified by the inclusion of an adjustable notch for measuring range" (Fig. D54). (Obsolete.)

Ref.: IEEE (1993), p. 873.

A **one-dimensional (1D) display** provides for measuring a single target coordinate. Usually 1D range displays are used. The amplitudinal blip is formed by feeding the signal from the output of the receiver to the deflecting plates of the CRT. The blip gives the visual information concerning the form and

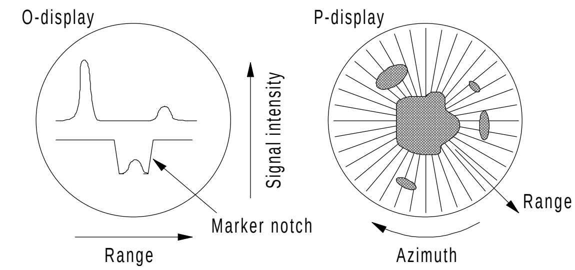


Figure D54 O- and P-displays.

intensity of the reflected signal and the character of the fluctuation. In 1D CRT displays, an electrostatic deflecting system is usually used. This system has a broader frequency band compared with magnetic deflecting systems and distorts the broadband radar signal to a lesser degree. *IAM*

Ref.: Barton (1964), p. 8; Vasin (1977), p. 180.

A **passive display** is based on the modulation of the external luminous flux as a result of the action of the electrical field or current. The electrical signal causes a change in the optical indicators of the material, the amplitude, phase, polarization plane, and direction of propagation of the light wave. Usually passive displays with intensity modulation are used. Liquid crystal (electro-optical effects in liquid crystals), electrochrome (electrochrome effect), electrophoretic (electrophoresis effect), and ferroelectric (double-beam refraction effect) displays are differentiated based on the types of light-modulating effects and the materials used.

Electrochrome and electrophoretic displays have a long reaction and relaxation time and are used to display slowly changing data. Liquid crystal displays have a relatively short switching time and an extremely low power consumption and thickness. The special features of ferroelectric displays are high control voltage and capability to switch high-power luminous flux, making it possible to obtain an image on a large screen.

In contrast to active displays, passive displays require less power but are slower and limited in their capability for multiplex addressing. *IAM*

Ref.: Fink (1982), p. 23.75; Bystrov (1985), pp. 6, 108.

A **phase-differential display** shows the angular coordinates with an indication of the average voltages of the signals characterizing the phase difference between the envelope of the video signals and the reference voltages. These displays are used when performing conical scanning and sending a large number of pulses in one scanning cycle. The operator (or automatic tracker) turns the antenna until a zero reading of the average discrepancy voltages is attained. *IAM*

Ref.: Rakov (1970), vol. 2, p. 385.

A **plan-position indicator (PPI)** is "a display in which target blips are shown on plan position, thus forming a map-like display, with radial distance from the center representing range and with the angle of the radius vector representing azimuth angle." (See Figs. D54, D55). There are various types of PPI displays implementation; the primary ones are: azimuth-stabilized PPI (the reference direction remains fixed with respect to the indicator regardless of the vehicle orientation), delayed PPI (the initiation of time base is delayed), off-center PPI (the zero position of the time base is located at a point other than the center of display, so equivalent of a larger display can be provided for a selected portion of the service area), open-center PPI (the display of the initiation of the time base precedes that of the transmitted pulse), and north-stabilized PPI (an azimuth-stabilized PPI in which the reference detection is north). This is the type of radar display normally associated with volume search or surveillance radars that allows a human operator to assess the tactical situation within the radius of coverage of the radar. The PPI presents a map-like circular presentation, usually on a CRT (CRT) of azimuth versus range. The display is refreshed at the radar azimuth scan rate. Targets and clutter remain visible from scan-to-scan due to the long-persistence properties of the phosphor coating on the CRT's inside surface. The PPI display may show a full 360° view, as in Fig. D55, or may be restricted to displaying a particular sector during each scan or may reflect operation of the radar in a more limited sector-scan mode.

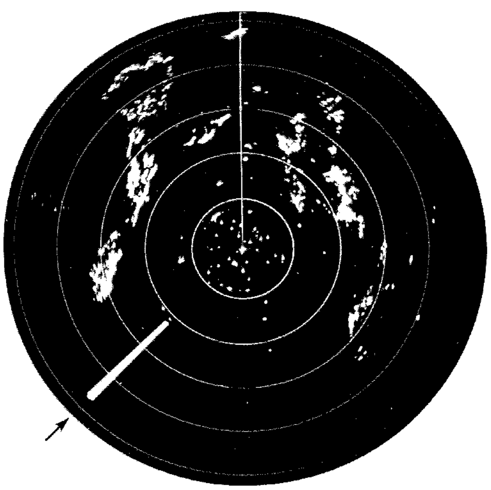


Figure D55 PPI display for maritime radar, with range rings 5 nmi apart (from Skolnik, 1970, p. 31.33, reprinted by permission of McGraw-Hill).

The PPI display shown in the figure is typical of older displays in which the analog or raw video signal out of the receiver was used to directly drive the display. The arrow shows the azimuth of a target-borne navigation beacon. Though still in use, the analog display is being replaced in many modern radars by the digital display, which is driven by the digital output of the digital signal processor. In digital displays, clutter returns can be suppressed and "true" targets indicated by synthetically generated symbols. Additional data relative to each target may also be displayed alongside the target symbol (e.g., digital readouts of target closing speed, engagement priority (for air defense radars), and target altitude if available). *PCH, SAL*

Ref: IEEE (1993), p. 960; Skolnik (1970), p. 31-33, (1990), p. 1.5.

A **P-(scope) display** is another name for the plan-position-indicator type of display commonly known as a *PPI display* (Fig. D54).

Ref.: IEEE (1993), p. 924.

Range display is a one-dimensional radar display making it possible to record and determine the range to a target. It is usually a CRT (see **CRT display**) with a static ray deflection and electronic control unit. Under the effect of the sweep voltage produced by the control unit and the signals from the receiver output, an amplitude blip is displayed on the screen against the background of a range scale (e.g., a linear scale). During the return trace of the sweep, the CRT is blanked with negative bias to remove blips of noise and false targets.

The range display is characterized by a range scale and by the value of the dot of the ray on the screen in kilometers, representing the resolution of the display.

The advantage of the range display is the capability to read signal amplitude. The inability to observe simultaneously on the screen all the targets in the radar coverage zone and to read other coordinates makes it necessary that these displays be employed in combination with other types of displays in detection and guidance radar. *IAM*

Ref.: Barton (1964), p. 7; Vasin (1977), p. 182.

A **range-height indicator (RHI)** is a display on which the signals are presented in range-height coordinates. (See **altitude-range display**. *SAL*

Ref.: IEEE (1996), p. 1,066; Sauvageot (1992), p. 16.

A **raw-video display** is one that is driven by the analog signal from the receiver (raw video). This type of display, also called an *analog display*, was used in older types of radar and in modern radar is being replaced by more flexible digital displays. *SAL*

Ref.: Skolnik (1980), p. 355.

A **rectangular coordinate display** has a rectangular screen that depicts elevation-range, altitude-range, and azimuth-range in rectangular coordinates. The rectangular coordinate display provides greater accuracy in the reading of the coordinates than the polar coordinate displays using a plan-position indicator. Examples of this type of display are the B-display, C-display, and E-display. *IAM*

Ref.: Barton (1964), p. 7; Pereverzentsev (1981), p. 364.

A **remote display** is one deployed outside the radar site and permitting monitoring and display of radar data at a remote site (e.g., at an air traffic control center). *IAM*

Ref.: Ridenour (1947), Ch. 17; Popov (1980), p. 75.

An **R-(scope) display** is "an A-display with a segment of the time base expanded near the blip for greater precision in range measurement and visibility of pulse shape. Usually regarded as an optional feature of an A-display rather than being identified by the term R-display."

Ref.: IEEE (1993), p. 1078.

A **sector display** produces an enlarged (zoomed) image of an isolated area of space. Two types of sector displays are used:

with polar or rectangular sweep. These are actually modifications of the plan-position indicator (PPI) or the rectangular-coordinate display. For example, in a PPI the center of rotation of the sweep can be offset from the center of the screen to the side opposite the enlarged sector and the scale expanded in range and angle by a factor typically two to four (an off-center PPI). *IAM*

Ref.: Ridenour (1947), p. 168; Rakov (1966), vol. 2, p. 140.

A **special display** is one used in a specific type of radar or under specific conditions of operation. Examples are bright displays operating in high-ambient-light environments, displays for sidelooking radars that must present high-resolution ground maps, those of terrain-avoidance radars that must carry the information on aircraft clearance height, and large-screen displays that present the same information to several people (e.g., in command and control centers). These last take the form of projection displays (dynamic slides, photochromic-film projection, etc., or liquid-crystal displays). *IAM*

Ref.: Skolnik (1970), p. 6.24.

A **synthetic-video display** is one in which target information is presented with standard symbols and accompanying alphanumerics. The computer is used to generate the graphics and control the CRT display, so it offers flexibility in choice of different modes of operation and provides the possibility of implementing many functions necessary, for example, in air-traffic-control or air-defence systems (e.g., blanking the areas with excessive interference, superimposing simulated targets to provide the training of radar operators,). Typically this type of display will be digital. *SAL*

Ref.: Skolnik (1980), p. 359.

A **television [-type] display** is a radar display in which the brightness of the image is increased as a result of multiple regeneration of the area of the radar image recorded within the space of one scanning cycle and the display of the regenerative images with a great repetition frequency. The operation of the display is based on the conversion of the radar image into a television image, which is usually accomplished using CRTs with storage capability. The brightness of this type of television image is 170 to 350 candles/m^2. However, at the same time, the resolution and number of gradations of brightness are reduced. *IAM*

Ref.: Pereverzentsev (1981), p. 340; Skolnik (1970), p. 6.24.

A **three-dimensional (3D) display** is one that shows the three coordinates of the radar target. To reproduce the third coordinate implicit techniques are used, making it possible to approximate its value. The two basic coordinates are displayed analogously to the two-coordinate display.

This display employs a CRT with persistence and a bright blip. The 3D display speeds up the operator's evaluation of the radar situation compared with the use of several two-dimensional displays. *IAM*

Ref.: Poole (1966), p. 177; Vasin (1977), p. 182.

A **true-motion display** is one "in vehicle-mounted radar that shows the motions of the radar and of targets tracked by that radar, relative to a fixed background; accomplished by inserting compensation for the motion of the vehicle carrying the radar." *SAL*

Ref.: IEEE (1993), p. 1417.

A **two-dimensional (2D) display** is one that presents two coordinates of a radar target. Usually CRT displays are used with a bright blip from the target in the rectangular or polar (plan-position indicator) coordinate system. Thanks to the use of a screen with persistence, the bright blips carrying the information concerning the air or ground situation are maintained for the scanning period. CRTs with electrostatic deflection and magnetic deflection are used to ensure a better focusing of the beam.

Two or more 2D indicators that have a common indexed coordinate for the correlation of the displays are used for the full display of the spatial disposition of the targets (e.g., azimuth-range and altitude-range displays). *IAM*

Ref.: Barton (1964), p. 7; Vasin (1977), p. 181.

A **volume display** presents a volume image. It is based on the stereo effect, using a pair of conventional radar displays or by using optometrical means. In the latter case tipping or rotating screens or mirrors are used on which by optical means an image is projected that is usually the bright blip of a target. The mirror is connected by means of a synchronized tracking system with the radar antenna and repeats the movement of the beam. Two other coordinates lie in the plane of the screen. *IAM*

Ref.: Poole (1966), p. 276; Druzhinin (1967), p. 425.

DISTANCE (see **RANGE**).

DISTRIBUTION. The term distribution is used in radar in two ways: (1) to describe the relative frequency of occurrence of random events (probability density and probability distribution functions) and (2) as a synonym for taper, illumination, or weighting of an antenna aperture. The probability distribution functions ware described below. The weighting of an aperture, or equivalent time or frequency function in signal processing, is described under the **WEIGHTING** entry. *SAL*

aperture distribution (see APERTURE illumination, WEIGHTING).

Bickmore-Spellmire distribution (see **WEIGHTING**).

The **binomial distribution** is the distribution describing the probability *P* that *j* outcomes will take place in *n* independent trials when the probability of each outcome is equal to *p:*

$$P(j) = \frac{n!}{j!(n-j)!}p^j(1-p)^{n-j}$$

Such a distribution describes, for example, the performance of the binary integrator operating with *n* equal pulses. Then *P*(*j*) is the probability that exactly *j* crossings of the threshold will occur in *n* trials if the probability of each threshold crossing is equal to *p*. *SAL*

Ref.: Barton (1989), p. 73.

Chebyshev distribution (see **WEIGHTING**).

The **chi-square distribution** is defined by

$$W(X) = \frac{k}{(k-1)!\bar{\bar{X}}}\left(\frac{kX}{\bar{\bar{X}}}\right)^{k-1}\exp\left(\frac{-kX}{\bar{\bar{X}}}\right)$$

where $X > 0$.

For description of target RCS statistics, this distribution was introduced by Swerling, using for k a value of 1 (case 1 or 2 target) or 2 (case 3 or 4 target). In the convectional theory of statistics the parameter $2k$ is typically called number of degrees of freedom, which must be an integer. This distribution, for example, can be used to describe the effect of noncoherent integration of fluctuating targets plus noise. *SAL*

Ref.: Currie (1989), p. 235.

clutter amplitude distribution (see **CLUTTER (amplitude) distribution**).

The **drop-size distribution (DSD)** describes the size of precipitation drops and is widely used in radar meteorology. The typical notation is $N(D)$, where D is the drop diameter. One of the most general forms of the DSD is given by the modified gamma distribution

$$N(D) = N_0 D^m \exp(-\Lambda D^q)$$

where Λ, m, and q are the distribution parameters. When m = 0 and q = 1 the DSD assumes the exponential form

$$N(D) = N_0 \exp(-\Lambda D)$$

that has been used extensively to describe rain, snow, hail, and clouds. This distribution originally was proposed by Marshall and Palmer, and is called the Marshall-Palmer distribution. *SAL*

Ref.: Meneghini (1990), p. 133; Sauvageot (1992), p. 78.

The **exponential distribution** is the distribution of random magnitude x with probability density function:

$$f_e(x) = \lambda e^{-\lambda x}$$

and probability distribution function:

$$F_e(x) = 1 - e^{-\lambda x}$$

where λ is the distribution parameter. See **CLUTTER (amplitude) distribution.**

An exponential distribution is widely used in evaluating radar reliability, attenuation of a signal during its propagation in the atmosphere, and so forth. *AIL*

Ref.: Skolnik (1962), p. 27; Tikhonov (1980), p. 44.

The **gamma distribution** is the distribution of random magnitude x with probability density function:

$$f_\Gamma(x) = \frac{1}{\beta^{b+1}\Gamma(b+1)} x^b e^{-\frac{x}{\beta}}, \; b > -1, \beta > 0$$

and probability distribution function:

$$F_\Gamma(x) = \frac{\Gamma\left(b+1; \frac{x}{\beta}\right)}{\Gamma(b+1)}, \; x > 0$$

where $\Gamma(\cdot)$ is the gamma function, and $\Gamma(\cdot\cdot)$ is the incomplete gamma function.

A gamma distribution finds use in description of noise statistics, RCS fluctuations, and in radar reliability. *AIL*

Ref.: Skolnik (1980), p. 50; Tikhonov (1980), p. 36.

The **Gaussian distribution** is the distribution of random magnitude x with probability density function:

$$f_g(x) = \frac{1}{\sigma_x\sqrt{2\pi}} \exp\left[-\frac{(x-m_x)^2}{2\sigma_x^2}\right]$$

and probability distribution function:

$$F_g(x) = \Phi\left(\frac{x-m_x}{\sigma_x}\right)$$

where $\Phi(\cdot)$ is the Laplace function, m_x = mean value of x, and σ^2 = variance of x.

A Gaussian distribution is widely used for describing noise, interference, radar coordinate measurement errors, and other statistical parameters. It is also called the *normal distribution*. (See **CLUTTER (amplitude) distribution.**) *AIL*

Ref.: Barton (1969), p. 210.

The ***K*-distribution** is defined by

$$f_k(u) = \frac{b^{\nu+1}u^\nu}{2^{\nu-1}\Gamma(\nu)} K_{\nu-1}(bu) \quad u \ge 0, \nu \ge 0, b > 0$$

where ν and b are a shape and a scale parameter, respectively, related to the common variance of the quadrature components by $\sigma^2 = 2\nu/b^2$, and K is the modified Bessel function. *DKB*

Ref.: Schleher (1991), p. 33.

The **log-normal distribution** is the distribution of random magnitude x with probability density function:

$$f_l(x) = \frac{1}{x\sigma_x\sqrt{2\pi}} \exp\left[-\frac{(\log x - m_x)^2}{2\sigma_x^2}\right]$$

and probability distribution function:

$$F_l(x) = \Phi\left(\frac{\log x - m_x}{2\sigma_x^2}\right), \; x > 0$$

where $m_x = M$ (log x), σ^2 = standard deviation of log x, and $\Phi(\cdot)$ is the Laplace function.

A log-normal distribution finds use in description of radar clutter and RCS fluctuations. See **CLUTTER (amplitude) distribution.** *AIL*

Ref.: Tikhonov (1986), p. 36; Schleher (1991), p. 33.

Marshall-Palmer distribution (see **drop-size distribution**).

The **Nakagami distribution** is the distribution of random magnitude x with probability density function:

$$f_N(x) = \frac{2}{\Gamma(m)}\left(\frac{m}{\sigma^2}\right)^m x^{2m-1} \exp\left[-\left(\frac{mx^2}{\sigma^2}\right)\right], \; m \ge \frac{1}{2}$$

and probability distribution function:

$$F_N(x) = \frac{\Gamma\left(m;\dfrac{mx^2}{\sigma^2}\right)}{\Gamma(m)}, \, x > 0$$

where $\Gamma(\cdot)$ is the gamma function, and $\Gamma(\cdot\cdot)$ is the incomplete gamma function.

A Nakagami distribution is used in radar detection theory. *AIL*

Ref.: Bakut (1984), p. 341.

normal distribution (see **Gaussian distribution**).

The **Poisson distribution** is the distribution of discrete random magnitudes. The random magnitude x is considered to be distributed according to Poisson's law if the probability that it will take on specific value k is expressed as

$$P_k(x) = \frac{m^k}{k!}e^{-m}$$

where m = mean value of random magnitude x.

A Poisson distribution is used in radar signal detection theory and when calculating reliability and the required number of spares for radar maintenance. *AIL*

Ref.: Korn (1961), p. 18.8.1; Tikhonov (1980), p. 30.

The **Rayleigh distribution** is the distribution of random magnitude x with probability density function:

$$f_R(x) = \frac{x}{\sigma^2}\exp\left(-\frac{x^2}{2\sigma^2}\right), \, x \ge 0$$

and probability distribution function:

$$F_R(x) = 1 - \exp\left(-\frac{x^2}{2\sigma^2}\right), \, x \ge 0$$

where σ is the distribution parameter.

A Rayleigh distribution is used in a statistical description of radar receiver noise and of the echo area of certain types of targets, as well as in signal detection theory. See **CLUTTER (amplitude) distribution.** *AIL*

Ref.: Lawson (1950), p. 53; Bakut (1984), p. 341.

The **Rician distribution** is distribution of random magnitude x with probability density function:

$$f_0(x) = \frac{x}{\sigma^2}\exp\left(-\frac{x^2+a^2}{2\sigma^2}\right)I_0\left(\frac{ax}{\sigma^2}\right), \, x \ge 0$$

and probability distribution function:

$$F_0(x) = \exp\left(-\frac{a^2}{2\sigma^2}\right)\sum_{k=0}^{\infty}\frac{x}{(k!)^2}\left(\frac{a^2}{2\sigma^2}\right)\Gamma\left(a+1;\frac{x^2}{2\sigma^2}\right), \, x \ge 0$$

where $I_0(\cdot)$ is the Bessel function of the zero order, $\Gamma(\cdot;\cdot)$ is the incomplete gamma function, and a and c are distribution parameters.

A Rician distribution sometimes is referred to as a *generalized Rayleigh distribution* and finds use in a statistical description of receiver noise, in signal detection theory, radar clutter, and so forth. When $a = 0$, the Rician distribution evolves into the Rayleigh distribution. See **CLUTTER (amplitude) distribution.** *AIL*

Ref.: Skolnik (1980), p. 50; Tikhonov (1980), p. 40.

The **Student distribution** is the distribution of random magnitude x with probability density function:

$$f_S(x) = \frac{\Gamma(n/2)}{\sqrt{\pi(n-1)}\Gamma[(n-1)/2]}\left(1+\frac{x^2}{n-1}\right)^{-n/2}$$

and probability distribution function:

$$F_S(x) = \frac{\Gamma(n/2)}{\sqrt{\pi(n-1)}\Gamma[(n-1)/2]}\int_{-\infty}^{x}\left(1+\frac{t^2}{n-1}\right)^{-n/2}dt$$

where $\Gamma(\cdot)$ is the gamma function, and n is the distribution parameter.

A Student distribution finds use when matching results of an evaluation of radar parameters in radar tests. *AIL*

The **uniform distribution** in the interval (a, b) is the distribution of a random magnitude x with probability density function:

$$\begin{aligned} f(x) &= \frac{1}{b-a}, \, a \le x \le b \\ &= 0, \text{ otherwise} \end{aligned}$$

that is, it is constant within the specified interval. This distribution is often used for statistical evaluation of random phase, which is considered to have uniform distribution within the interval $(0, 2\pi)$. *SAL*

The **Weibull distribution** is the distribution of random magnitude x with probability density function:

$$f_W(x) = abx^{b-1}\exp(-ax^b), \, a > 0, \, b > 0$$

and probability distribution function:

$$F_W(x) = 1 - \exp(-ax^b), \, x > 0$$

where a and b are the distribution functions.

A Weibull distribution is used in statistical description of receiver noise, clutter, and reliability (malfunctions in radar assemblies and units). See **CLUTTER (amplitude) distribution.** *AIL*

Ref.: Bakut (1984), p. 341; Schleher (1991), p. 34; Currie (1992), p. 22.

The **Weinstock distribution** is a chi-square distribution for positive k values less than unity. Weinstock was the first to observe that fluctuations of some simple targets (e.g., cylinders) can be described with values of k between 0.3 and 2, as in previous radar practice the target RCS fluctuations were described by Swerling's models, using k values of one or greater. *SAL*

Ref.: Currie (1989), p. 236.

divergence factor (see **REFLECTION from the spherical earth**).

DIVERSITY is the mode of radar operation when the basic radar parameters are changed in time, space, frequency, or polarization to improve radar performance. *SAL*

Combined diversity is the mode in which both time and frequency diversity are used to increase the number of samples to reduce the fluctuation loss in the detection of fluctuating targets. The total number of samples can be expressed as

$$n_e = \left(1 + \frac{t_0}{t_c}\right)\left(1 + \frac{\Delta f}{f_c}\right)$$

where t_o is the integration time, t_c is the correlation time, Δf is the frequency bandwidth, and f_c is the correlation frequency of the target. *DKB, SAL*

Ref.: Barton (1989), p. 86, (1991), p. 4.21.

Frequency diversity is the mode of radar operation in which parallel channels or burst-to-burst frequency changes are used to obtain independent samples to reduce the fluctuation loss in detection of fluctuating targets. It offers a solution to the problem of large fluctuation loss, without compromising MTI or doppler performance. A number of channels (often two) may be operated in parallel, spaced by at least the correlation frequency f_c within the radar band, or the frequencies may be transmitted by changing from one burst to the next. Pulses are received in each channel to increase the integration gain and to decrease the fluctuation loss. *DKB, SAL*

Ref.: Barton (1989), p. 8, (1991), p. 4.21.

Diversity gain is "the reduction in predetection signal-to-interference ratio required to achieve a given level of performance, relative to that of nondiversity radar, resulting from the use of polarization, space, time, frequency, or other characteristics." The usual notation is $G_d(n_e)$, where n_e is the number of independent samples provided by diversity. When the task is to estimate the reduction in fluctuation loss for radar detecting a fluctuating target due to taking samples over intervals in time and frequency, the diversity gain can be defined as

$$G_d(n_e) \equiv \frac{L_f(1)}{L_f(n_e)} = L_f(1)^{1-1/n_e}$$

where $L_f(1)$ and $L_f(n_e)$ are the target fluctuation losses for one and n_e number of samples correspondingly. In dB notation, the expression for the diversity gain can be written as

$$G_d(n_e)_{dB} = \left(1 - \frac{1}{n_e}\right)L_f(1)_{dB}$$

DKB, SAL

Ref.: IEEE (1993), p. 379; Barton (1988), p. 85.

Polarization diversity is the mode of radar operation in which multiple polarizations are used to avoid interference, to increase signal detectability, or to identify targets. *SAL*

Ref.: Brookner (1977), p. 55.

Space diversity is the mode of radar operation in which the transmitting and/or receiving antennas are displaced in space. (See **RADAR, multistatic.**) *SAL*

Ref.: Barton (1964), p. 501.

Time diversity is the mode of radar operation in which independent samples are received at intervals equal to the correlation time of the target. It requires that the integration time t_o exceed the correlation time t_c of the target. The number of samples available is

$$n_e = 1 + \frac{t_0}{t_c}$$

For rigid targets, the correlation time may be estimated as

$$t_c = \frac{\lambda}{2\omega_a L_x}$$

where λ is radar wavelength, ω_α is the rate of rotation of the target about the line of sight, and L_x is the target dimension normal to the line of sight and the axis of rotation. *DKB, SAL*

Ref.: Barton (1989), p. 86.

DIVIDER: A **power divider** is "a device for producing a desired distribution of power at a branch point." Typically, it is an assembly of transmission lines for the distribution of the power fed to the input between several output channels. As a rule, power dividers are reciprocal units (i.e., they are capable of dividing and summing power). Depending on the their function, dividers perform even and uneven division into two or a greater number of channels.

The quality of operation of a power divider in a frequency band is evaluated by the voltage standing-wave ratio of each input and the transmission coefficient. Sequential and parallel (in the form of a star) power dividers are used. A power divider with two outputs is a T-junction with absorbing elements for the matching and isolation of the outputs. Bridges and directional couplers are widely used in divider networks. *IAM*

Ref.: IEEE (1993), p. 986; Veselov (1988), p. 68; Sazonov (1988), p. 113.

A **voltage divider** is "a network consisting of impedance elements connected in series, to which voltage is applied, and from which one or more voltages can be obtained across any portion of the network." Typically, the output voltage is reduced in relation to the input voltage by a specific factor. The most simple and widely used voltage divider is constructed on resistors. For the division of alternating voltage, dividers with reactive resistors are often used; for example, capacitance dividers, consisting of two or several capacitors connected in series. *IAM*

Ref.: IEEE (1993), p. 1,467; Popov (1980), p. 104.

Dolph-Chebyshev weighting (see **WEIGHTING**).

DOPPLER EFFECT. The doppler effect is "the effective change of frequency of a received signal due to the relative velocity of transmitter with respect to receiver." In radar it primarily manifests itself in the effect that the carrier frequency f_r of received signal differs from the carrier frequency f_0 of the transmitted signal when reflected (or retransmitted) by the moving target.

For a case when the target is moving and the radar is stationary, the relation between f_0 and f_r can be written

$$f_r = f_0\left(\frac{1 - v_r/c}{1 + v_r/c}\right)$$

where v_r is the target radial velocity (along the line connecting the radar and target) and c is the velocity of light. If we denote the difference as the doppler frequency, $f_d = f_r - f_0$, we obtain

$$f_d = f_0\left(\frac{1 - v_r/c}{1 + v_r/c} - 1\right)$$

and $f_d < 0$ for $v_r > 0$ (increasing range). Typical target velocities are much less than the velocity of light, $v_r << c$, so this equation can be expended into a series:

$$f_d \approx f_0\left(1 - \frac{2v_r}{c} + \ldots - 1\right) \approx -f_0\frac{2v_r}{c} = -\frac{2v_r}{\lambda}$$

where λ is the wavelength of the transmission. This final form is used in most radar applications.

The formula is valid for active radar, where the doppler shift results from change in the two-way path as the target moves. For a passive radar:

$$f_d \approx \frac{v_r}{\lambda}$$

because the shift applies only for the one-way path.

For pulsed radar, the doppler effect also produces a change in pulse width and pulse repetition frequency, but because the doppler shift is proportional to the frequency f_0, it is only significant for the RF carrier.

The doppler effect is of fundamental importance in radar applications, where it is used to determine target radial velocity and to discriminate between moving targets and clutter. *SAL*

Ref.: IEEE (1993), p. 381; Shirman (1970), p. 51; Dulevich (1978), pp. 236–239; Skolnik (1980), p. 68.

Doppler ambiguity (see **AMBIGUITY, doppler**).

Doppler beam-sharpening is "a special form of synthetic aperture radar processing that uses a constant frequency reference." It is a coherent, airborne radar technique associated with ground mapping and the detection of fixed land targets that uses doppler filtering to synthetically reduce the radar's azimuth beamwidth.

In a conventional (nondoppler) radar, airborne detection of a small, fixed target (such as a automobile, truck, or tank) in land clutter is difficult due to the large clutter-to-signal ratio attendant to the radar mainbeam illumination of the surface, and the clutter cell dimensions are defined by the effective transmitted pulsewidth, range to the target, radar antenna azimuth beamwidth, and relative geometry. For example, even with a narrowbeam antenna, e.g., 1°, a conventional airborne radar with an effective pulsewidth on the order of a few tenths of a microsecond would be unable to detect a 5 m^2 target at a range of 20 km in moderate land clutter.

Figure D56 depicts the geometry for a doppler-beam-sharpening approach to the problem just described. The pre-doppler-filtering clutter cell dimensions are

$$R = \frac{\tau_{\text{eff}}c}{2}\sec\psi$$

in range, where ψ is the grazing angle from the radar to the clutter cell, and

$$\Delta Az = R_c\theta_{AZ}$$

in cross range, where R_c is the range to the clutter cell. The doppler shift to any point n within the range cell is given by

$$f_{dn} = \frac{2V_m}{\lambda}\cos\alpha_n$$

where V_m is the radar platform velocity, and α_n is the angle between the point n and the radar platform velocity vector.

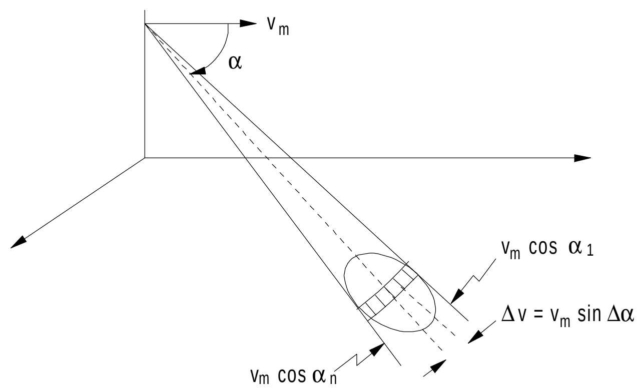


Figure D56 Doppler-beam-sharpening geometry, airborne radar.

If the azimuth dimension of the clutter cell is further divided into a number of smaller cells of equal size, the doppler frequency difference across each cell will be shown that

$$\Delta f_n = \left(\frac{2V_m}{\lambda}\right)(\cos\alpha_{n1} - \cos\alpha_{n2})$$

If the doppler filter bandwidth B_D is set equal to the doppler spread Δf_n, the angular resolution due to beam sharpening will be

$$\Delta\theta = \frac{\lambda B_D}{2V_m \sin\alpha_m}$$

PCH

Ref.: IEEE (1990), p. 15; Schleher (1991), pp. 469–472.

The **doppler bin** is the filter channel of a CW or pulsed doppler radar. In a tracking radar, a single doppler bin may be used that tracks the target echo. In a search or acquisition processor, multiple bins are formed with a filter bank, implemented by parallel doppler filters or a digital signal processor using a fast Fourier transform (FFT). *DKB*

doppler filter (see **FILTER, doppler**).

doppler (frequency) shift (see **DOPPLER EFFECT**).

A **doppler navigator** is "a self-contained dead reckoning navigation aid transmitting two or more beams of electromagnetic or acoustic energy outward and downward from the vehicle and using the doppler effect of the reflected energy, a reference direction, and the relationship of the beams to the vehicle to determine speed and direction of motion over the reflecting surface."

Ref.: IEEE (1993), p. 382; Skolnik (1962), p. 103.

doppler spread (see **CLUTTER spectrum**).

Doppler velocity is the radial velocity of target derived from the measurement of doppler frequency, f_d. For active radar:

$$v_2 \approx \frac{\lambda f_d}{2}$$

and for passive radar:

$$v_1 = \lambda f_d$$

SAL

Ref.: Skolnik (1980), p. 69.

DRIVE, antenna. An antenna drive is a device serving for rotation of an antenna in one or two planes for the spatial displacement of the beam. An antenna drive comprises one or several electrical motors, control equipment, and a system of mechanical or electrical transmission of motion (rotation). A power servo often is used in radars. *AIL*

Ref.: James (1947); Popov (1980), p. 39.

DUCT, DUCTING. Ducting is "confinement of electromagnetic wave propagation to a restricted atmospheric layer by steep gradients in the index of refraction with altitude." Ducting occurs when the gradient of the index of refraction *dn/dh* is less than -1.57×10^{-7} m^{-1} and results in anomalous propagation that can extend radar range considerably. Ducting is usually a fine-weather phenomenon (with the exception of thunderstorms), occurring when the upper air is exceptionally warm and dry in comparison with that at the surface. Although radar range increases within the duct, the coverage in other directions can be reduced and so-called radar holes can appear. A duct (or atmospheric duct) is a natural "waveguide" formed in atmosphere inside which the ducted propagation takes place. Elevated ducts, evaporation ducts, and ground-based (or surface) ducts are distinguished. *SAL*

Ref.: IEEE (1993), p. 392; Barton (1991), pp. 5.16, 8.17; Morchin (1993), p. 312.

The **ducting effec**t is the effect of forming of an atmospheric duct in particular atmospheric conditions. *SAL*

Ref.: Neri (1991), p. 87.

An **elevated duct** is one that lies above the ground surface. An example of propagation in elevated ducts is found in the "tradewind region" between the midocean, high-pressure cells and the equatorial doldrums. The best known tradewind areas lie between Brazil and the Ascension Islands, and between Southern California and Hawaii. The propagation in the elevated ducts is seasonal. *SAL*

Ref.: Skolnik (1980), p. 453.

An **evaporation duct** is one that lies above the surface of the sea and results from the water vapor evaporated from the sea. This duct exists practically all the time over the ocean. The height of the duct typically lies between 6 and 30m, and it varies with the geographic location, season, time of day, and wind speed. The water surface temperature, air temperature, relative humidity, and wind speed are sufficient measurement parameters to describe ducting conditions. *SAL*

Ref.: Skolnik (1980), p. 453.

A **ground-based [surface] duct** is one that lies close to the surface of the ground, typically at a height of 10 or 20m (never more than 150 to 200m). Ground-based ducts are more usual than elevated ducts. Sometimes this duct is called a *surface duct*. *SAL*

Ref.: Skolnik (1980), p. 451.

Ionospheric ducting is the long-distance propagation of electromagnetic waves by ducting along field-aligned ionization in the ionosphere. *SAL*

Ref.: Fink (1975), p. 18.116; Kolosov (1987), p. 31.

DUPLEXER. A duplexer is "a device that utilizes the finite delay between the transmission of a pulse and the echo thereof so as to permit the connection of the transmitter and receiver to a common antenna." On transmission it protects the receiver from the damage by the high power of the transmitter, and on reception it channels the echo signal to receiver. Typically, duplexers are divided into devices using gas-discharge tubes (gas-tube or TR-tube duplexers), solid-state devices (diode duplexers), and ferrite circulators (ferrite circulator duplexers). The most common configurations are balanced duplexer and branch-type duplexer. In a typical duplexer application, the transmitter peak power might be a megawatt or more. Ensuring a safe power level for the receiver, typically less than a watt, more than 60 dB of isolation may be required. Usually a receiver protector may be inserted between the duplexer and the receiver for added protection. *SAL*

Ref.: IEEE (1993), p. 393; Fink (1975), p. 25.70; Skolnik (1980), p. 4.4; Skolnik (1990), pp. 359–368.

A **balanced duplexer** is based on the short-slot hybrid junction that consists of two sections of waveguides joined along one of their narrow walls (Fig. D57). Such a hybrid may be considered a broadband directional coupler (coupling ratio is 3 dB). The slot is cut in the common narrow wall of the junction to provide the coupling. In the transmit condition the TR tubes fire and reflect the incident power out the antenna arm, while on reception the TR tubes are unfired and the echo signals pass through the duplexer into the receiver. The hybrid junction provides an additional 20 to 30 dB of isolation in addition to attenuation provided by the TR tubes.

There is also another configuration of balanced duplexer, called ATR type, that can be employed. This form uses two hybrid junctions and four ATR tubes (Fig. D58). Dashed lines show the flow of power. This type of balanced duplexer has higher power-handling capability than the first one, but its bandwidth is less. *SAL*

Ref.: Skolnik (1980), p. 360.

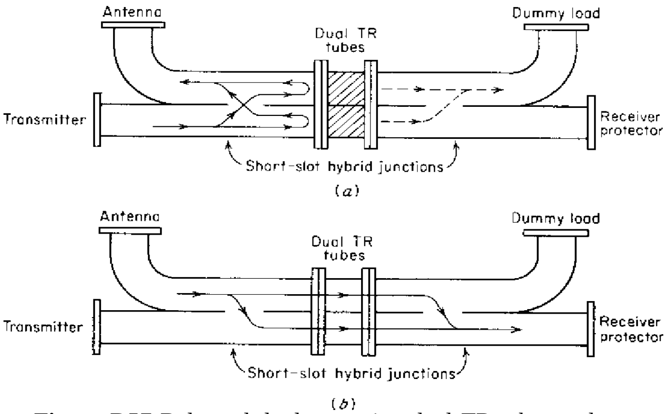


Figure D57 Balanced duplexer using dual TR tubes and two short-slot hybrid junctions: (a) transmit condition (b) receive condition (from Skolnik, 1980, Fig. 9.6, p. 361, reprinted by permission of McGraw-Hill).

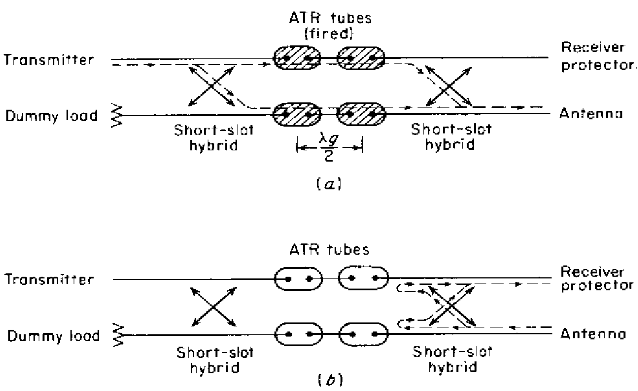


Figure D58 Balanced duplexer using ATR tubes: (a) transmit condition; (b) receive condition (from Skolnik, 1980, Fig. 9.7, p. 362, reprinted by permission of McGraw-Hill).

A **branch-type duplexer** consists of a TR switch and an ATR switch based on gas-discharge tubes (Fig. D59). In the trans-

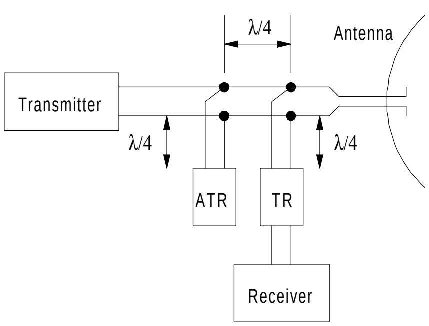


Figure D59 A branch-type duplexer, parallel configuration (after Skolnik, 1980, Fig. 9.5, p. 360).

mit condition the TR and ATR tubes are fired (ionized); in reception the transmitter is off, and the tubes are not ionized. In the first case the TR acts as a short circuit to prevent transmitter power from entering the receiver; in the second case the open circuit of the ATR, being a quarter-wave from the transmission line, appears as a short circuit across the line. This disconnects the transmitter from the line because the short circuit is a quarter wave from the receiver branch-line, and so the echo signal is directed to the receiver.

This type of duplexer is one of the earliest duplexers used in radar. It has limited power-handling capability and bandwidth, and in modern radars has generally been replaced by balanced duplexers and other protection devices. *SAL*

Ref.: Skolnik (1980), p. 360.

A **diode duplexer** uses diode switches in place of TR tubes. Typically, PN and PIN diodes operating unbiased or with a dc forward bias current are used. Unbiased (passive) operation is used for low-power application, while biased operation (active) is capable of switching high power. If the single diode cannot withstand the required voltage, multiple diodes can be used. The typical example is a balanced duplexer that uses 32 PIN diodes instead of TR tubes and handles 150 kW peak power and 10 kW average power, with a pulse width of 200 μs. *SAL*

Ref.: Fink (1975), p. 25.71; Skolnik (1980), p. 363.

A **ferrite duplexer** uses a ferrite circulator instead of TR tubes. The circulator does not provide sufficient protection by itself and requires a receiver protector (Fig. D60). This type of duplexer is attractive for radar applications because of its long life, wide bandwidth, and compact design. *SAL*

Ref.: Skolnik (1980), p. 365; Skolnik (1990), p. 4.4.

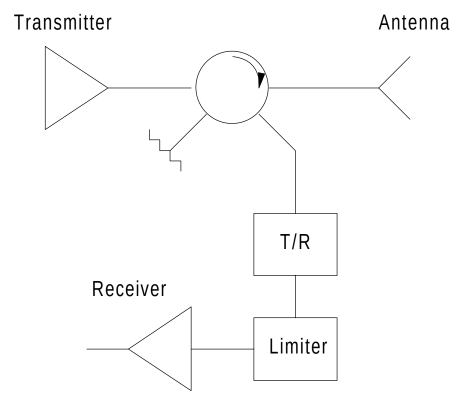


Figure D60 Ferrite circulator duplexer (after Skolnik, 1990, Fig. 4.2b, p. 4.4).

A **gas-tube duplexer** uses a gas-discharge tube as the switching element. Typically it is a gas-filled TR tube that fires in the presence of high power produced by the transmitter signal. The typical configuration of a gas-tube duplexer is given in Fig. D61.

A **TR-tube duplexer** uses a TR tube as the switching element. *SAL*

DUTY FACTOR [CYCLE, RATIO] is "the ratio of the active or ON time within a specified period to the duration of the specified period." In pulsed radar, it is the ratio of pulse-width τ to the pulse repetition interval t_r: $D_u = \tau/t_r$. Since

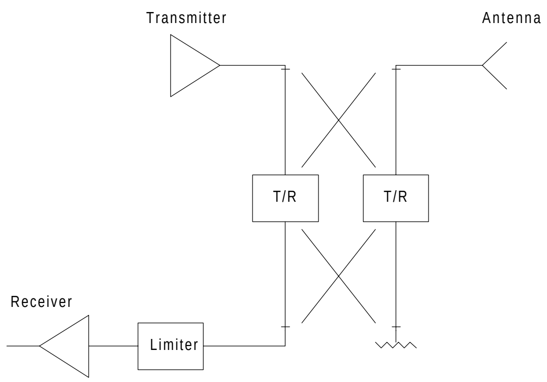


Figure D61 Gas-tube duplexer (after Skolnik, 1990, Fig. 4.2a, p. 4.4)

average power $P_{av} = P_t \cdot \tau/t_r = P_t \cdot D_u$, the duty factor is the ratio of P_{av} to peak power P_t: $D_u = P_{av}/P_t$. *SAL*

Ref.: IEEE (1993), p. 395.

DYNAMIC RANGE. Dynamic range is "the difference, in decibels, between the overload level and the minimum acceptable signal level in a system or transducer." In radar applications, this concept is typically applied to receiver dynamic range. *SAL*

Ref.: IEEE (1993), p. 397.

Dynamic range compression is the reduction in dynamic range of the input signal to match it with the dynamic range of signals processors and filters, as their dynamic ranges are typically less than that of the input. Theoretically, compression of the dynamic range of signal amplitudes is optimum at the output of a filter. However in practice it is advisable to perform compression at the input of the filter. This makes it possible to obtain greater compression, to avoid additional equipment errors when multichannel filters are used, and in a number of other cases to normalize interference intensity simultaneously.

Dynamic range compression is done by a nonlinear circuit with a logarithmic amplitude characteristic or by one that places a strict restriction of the amplitude of the input signal (e.g., a limiter). Compensation of the spread of levels of the input signal due to interference occurs as a result of normalization of their intensity. One of the basic methods of normalizing intensity of interference is to change the threshold level or control the gain of the processing channel with a fixed threshold value. (See **GAIN, automatic gain control**.) *IAM*

Ref.: Barton (1964), p. 196; Sloka (1970), p. 125.

Instantaneous dynamic range is the range between the noise level and the saturation level in a receiver when the gain of the receiver stages is constant. Typically, it is less than the total receiver dynamic range because the gain of receiver stages may be varied. *SAL*

Ref.: Currie (1987), p. 495.

E

ECHO, radar. A radar echo is the portion of energy of the transmitted pulse that is reflected to a receiver. *SAL*

Ref.: IEEE (1990), p. 15.

angel echo (see **ANGEL**).

An **echo box** is "a high-Q resonant cavity that stores part of the transmitted pulse power and feeds the resulting exponentially decaying power into the receiver after completion of the pulse transmission." As a result of their simplicity and reliability, echo boxes were widely used as built-in test equipment during the 1940s and 1950s but have subsequently been replaced with electronic test equipment. *SAL*

Ref.: IEEE (1990), p. 15.

An **echo intensifier** is a device typically mounted on the target and used to increase abnormally the amplitude of the reflected signal. *SAL*

Ref.: Johnston (1979), p. 58.

echo reduction (see **RADAR CROSS SECTION reduction**).

A **false echo device** is a device that produces an echo that differs from the one normally observed in some essential parameter (in character, time, etc.). It is used to prevent accurate target recognition. *SAL*

Ref.: Johnston (1979), p. 60.

ground echo (see **CLUTTER, land**).

A **mantle-shaped echo** is an angel echo having a U- or V-shape and is associated with meteorological effects in the atmosphere. *SAL*

Ref.: Skolnik (1980), p. 511.

meteorological echo (see **CLUTTER**).

A **multiple-[*n*th] time-around echo** is one that arrives at the radar with delay greater than the pulse repetition interval (PRI), leading to a possible ambiguous range measurement when a constant-PRI is used. (See **AMBIGUITY, range.**) When the echo arrives with a delay from one to two PRIs, it is called a *second-time-around echo*. *SAL*

Ref.: Skolnik (1980), p. 3.

A **ring echo** is an echo that starts as a point on radar display, and then forms a rapidly expanding ring. Typically, it is an angel echo associated with birds flying away from roosting areas. *SAL*

Ref.: Skolnik (1980), p. 512.

sea echo (see **CLUTTER, sea**).

ECLIPSING is "the loss of information on radar echoes during intervals when the receiver is blanked because of the occurrence of a transmitter pulse. Numerous such blanking can occur in radars having high-pulse-repetition frequencies." The eclipsing effect applies to pulsed radars (Fig. E1), as in CW radars reception can be accomplished all the time and no eclipsing of radar returns will occur. Sometimes eclipsing is also termed an eclipsing effect or an eclipsing loss.

Ref.: IEEE (1993), p. 403; Hovanessian (1984), pp.147–150.

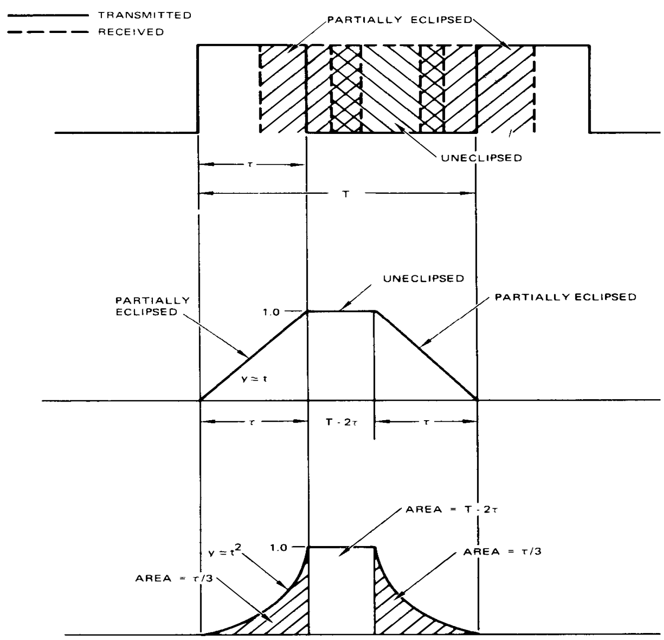


Figure E1 Eclipsing of radar return (from Hovanessian, 1984, p. 149).

Average eclipsing is an average power loss due to the eclipsing effect. Practically it can be represented by the area under the curve of Fig. E1(c).

Ref.: Hovanessian (1984), p. 148.

Deterministic eclipsing is the amount of power loss due to eclipsing that can be calculated in a deterministic fashion. The simplified series of eclipsing diagrams is cited on Fig. E2. Typically, using the deterministic eclipsing calculation

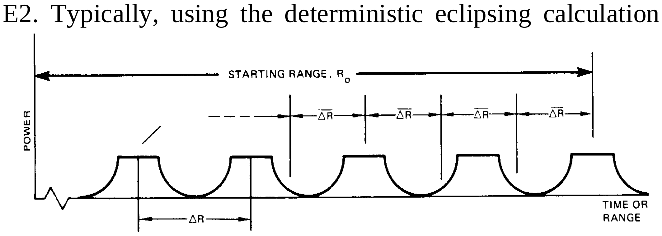


Figure E2 Deterministic eclipsing calculation (from Hovanessian, 1984, p. 151).

method, the amount of power loss due to eclipsing can be determined by dividing the radar-target range during each target illumination by $\Delta R = c\tau/2$, where ΔR is the distance between successive uneclipsed periods, and calculating the signal power loss from the decimal fraction portion of this ratio. *SAL*

Ref.: Hovanessian (1984), p. 150.

eclipsing loss (see **LOSS, eclipsing**).

EFFECTIVE ECHOING AREA is the same as *radar cross section*.

Ref.: IEEE (1990), p. 15.

EFFICIENCY

antenna aperture efficiency (see **APERTURE**).

integration efficiency (see **LOSS, integration**).

ELECTROMAGNETIC COMPATIBILITY (EMC) is “the capability of electronic equipment or systems to be operated in the intended operation electromagnetic environment at designed levels of efficiency.” While operating a radar, it must not have a harmful effect on the operation of other electronic equipment and it must be protected from the harmful influence of this equipment. Typically, the electromagnetic environment in which a radar operates is characterized by radiation of neighboring radars, other electronic equipment, and natural and industrial interference sources.

The main measures to control mutual interference are using different frequencies, choosing appropriate antenna locations for interfering equipment, reducing spurious radiation levels of transmitters, and using waveform coding, pulse-repetition-frequency agility, polarization selection, and other special measures. Usually, none of these measures is able to ensure the required level of electromagnetic compatibility by itself. The combination of technical and organizational measures must be undertaken. *AIL, SAL*

Ref.: IEEE (1978), p. 222; Maksimov (1976), p. 96-108; White (1971, 1973); Finkel’shteyn (1983), pp. 351–354.

ELECTRONIC COUNTER-COUNTERMEASURES (ECCM) are “actions taken to ensure friendly use of the electromagnetic spectrum against electronic warfare.” Their main objective is to eliminate or reduce the efficiency of the enemy’s ECM. They fall into two broad classes: electronic ECCM and operational ECCM. Electronic techniques are included in the main radar subsystems and are typically described, following the usage of Johnston (1979), as antenna-related ECCM, transmitter-related ECCM, and receiver- and signal-processing ECCM. From the point of view of radar types where these techniques are implemented, search (surveillance) radar ECCM and tracking radar ECCM are distinguished. *SAL*

Ref.: Johnston (1979), p. 58; Schleher (1986), pp. 199–301; Barton (1991), p. 12.11; Skolnik (1990), pp. 9.7–9.35; Neri (1991), pp. 417–453; Farina (1992).

Antenna-related ECCM are ECCM techniques based on the properties of antenna systems to reduce the effectiveness of ECM. Space selection based on antenna directivity and polarization selection based on the polarization properties of electromagnetic waves are the main ECCM strategies of discrimination the useful signals and interference.

The main techniques for antenna-related ECCM based on spatial selection are coverage and scan control, reduction of main-beam width, reduction of sidelobe level, and employing of adaptive antennas. The first group of methods based on antenna pattern and scan control may include blanking or turning of the receiver while the radar is observing the part of space containing a jammer, using multiple-beam configuration to detect a target by a beam not afflicted by jammers, random scanning to prevent deception jammers from synchronizing with the antenna scan rate, and other relevant measures. Reduction of beamwidth increases the angular resolution and is a valuable feature of any radar operating in a dense ECM environment. It should be kept in mind, however, that the reduction of beamwidth, for a given aperture size, results in increasing the sidelobe level that worsens a radar's antijamming capability, so the reasonable compromise should be found in specifying the antenna radiation pattern. The main sidelobe-related techniques are usage of low- and ultra-low sidelobe antennas, sidelobe blanking (SLB) and sidelobe cancellation (SLC). The generalization of SLC techniques in combination with adaptive processing technique is the concept of adaptive arrays. This technique is very promising as it is based on advanced methods of digital beam-forming and digital signal processing and permits superresolution capabilities that can be very useful for ECCM.

Polarization selection takes advantage of electromagnetic wave polarization features for discrimination of useful signals at the interference background. In this case, if only signals with copolarization are useful, the cross-polarized response should be kept as low as possible to protect against cross-polarized jamming. In more complex cases the two orthogonally polarized components may be used by the antenna to discriminate the useful signal from those received from jammers and chaff. *SAL*

Ref.: Skolnik (1990), pp. 9.7–9.16; Farina (1992), p. 6.

The **ECCM improvement factor** is "the power ratio of the ECM signal level required to produce a given output signal-to-interference ratio from a receiver using an ECCM technique to the ECM signal level producing the same output signal-to-interference ratio from the same receiver without the ECCM technique." *SAL*

Ref.: IEEE (1990), p. 15; Johnston (1979), p. 58; Skolnik (1990), p. 9.29.

Electro-optic counter-countermeasures are "the actions taken to ensure the effective friendly use of the electro-optic spectrum despite the enemy's use of countermeasures in this spectrum." *SAL*

Ref.: Johnston (1979), p. 59.

Human-factors ECCM is the "ECCM technique that covers the ability of an air-defence officer, a radar operator, a commanding officer, and/or any other air defence associated personnel to recognize the various kinds of ECM, to analyze the effect of ECM, to decide what the appropriate ECCM should be and/or to take the necessary ECCM actions within the framework of the person's command structure." *SAL*

Ref.: Skolnik (1990), p. 9.22.

Operational ECCM is the combination of operational modes of the radar in response to a specific ECM threat. This group of techniques can be subdivided into those involving methods of operation, radar deployment tactics, work of radar operator, and friendly electronic support measures in aid of ECCM. *SAL*

Ref.: Farina (1992), p. 6; Skolnik (1990), p. 9.21-9.22.

Range gate pull-off ECCM applies to measures taken, either with radar operator support or automatically through ECM sensing and response circuits, to defeat the range gate pull-off (RGPO) type of jamming employed by deceptive ECM (DECM) repeater jammer systems. Nondoppler measuring radars may use a human operator to identify the occurrence of RGPO, in which case the operator can manually maintain the radar tracking cursor over the correct target. If a victim of RGPO jamming is netted to another tracking radar, true target range may be derived through triangulation. Doppler radars can compare target doppler with range rate derived from range data differentiation and thus determine which is the correct target. This ECCM tactic may be frustrated if the jammer also employs velocity gate pull-off (VGPO) techniques, but other ECCM procedures can be invoked to deny the effectiveness of VGPO (see velocity gate pull-off ECCM). RGPO is seldom effective against modern radars when used alone, and to be effective, a radar ECCM strategy must be devised that deals with the several combined forms of range, angle, and doppler jamming techniques that the radar may encounter. *PCH*

Ref.: Schleher (1986), pp. 115, 124–125.

Receiver- and signal-processing-related ECCM is ECCM based on the properties of the receiver and signal processing to reduce the influence of ECM on radar performance. The main technique associated with receivers is using a wide dynamic range to avoid saturation of signal-processing chains by a jamming signal. Logarithmic and linear-logarithmic receivers are normally used to avoid the saturation. Other techniques often used in radar receivers (e.g., constant-false-alarm-rate circuits, automatic gain control, fast-time- constant circuits, etc.) may be useful to prevent saturation, although normally they are not referred as to ECCM techniques. Dicke-Fix devices and various kinds of limiters also can be used to counter jamming. The main type of signal processing that can significantly reduce the effectiveness of ECM is coherent signal processing. In modern radars it is implemented in the form of coherent digital signal processing, including digital MTI. *SAL*

Ref.: Skolnik (1990), p. 9.18; Farina (1992), p. 9.

Search radar ECCM is employed in a radar whose purpose is to search a large volume of space and locate the position of detected targets. The major ECM techniques that are threats for such radars are noise jamming, deception jamming, decoys, chaff, and antiradiation missiles. The main techniques that can be employed against noise jamming are based on the concepts of maximizing radar energy delivered to the target of interest (burnthrough mode) or/and minimizing the amount of jamming energy received by the radar (ultra- and low-sidelobe antennas, sidelobe blanking and cancellation, frequency agility and frequency diversity, use of coded waveforms). Techniques employed against the deception jamming can use PRF stagger and jitter or measurement and analysis of angular, doppler or polarization characteristics of radar echoes to discriminate between real and false-target returns. The main ECCM technique against chaff is discrimination of real and false returns on the basis of doppler processing (e.g., MTI). Antiradiation missiles are threats to search radars. One of the best methods of defence is to use a blinking network of radars. *SAL*

Ref.: Skolnik (1990), p. 9.25; Neri (1991), pp. 418–436; Chrzanowski (1990), pp. 81, 99.

Tracking radar ECCM is employed in radars whose purpose is to provide good resolution and precise measurement of targets parameters. It is based on antenna-, transmitter-, receiver- and signal-processing-related ECCM techniques. Practically all measures referred to as search radar ECCM, including maximizing the radiated energy delivered to the target and minimizing the jammer signal entering signal-processing circuits, are helpful to counter noise jamming in tracking radars. A more threatening ECM technique against tracking radars is deception jamming. Typically, it is deception in angle (angle-gate stealing), range (range-gate stealing or range-gate pull-off) and velocity (speedgate stealing or velocity-gate pull-off). Angle-gate stealing is especially effective against radars employing conical scanning and sequential lobing, so monopulse tracking, inherently insensitive to angle deception jamming from a single point, must be used in radars where good ECCM efficacy against jamming in angle is required. Using of leading-edge tracker, tracking of both real and false targets in both range and doppler, using PRF jitter, the mode of multiple PRF operation (low, high and medium PRF) and frequency agility are the usual measures against deception jammers in range and doppler domains. *SAL*

Ref.: Skolnik (1990), p. 9.25; Neri (1991), pp. 436–438; Chrzanowski (1990), p. 127.

Transmitter-related ECCM is based on the property of a radar transmitter to control the frequency, power, and waveform of the radiated signal to reduce the effectiveness of ECM. Increasing of the transmitter frequency is analogous to increasing the antenna aperture in a fixed-frequency system, narrowing the antenna beamwidth and in turn improving the spatial resolution (see **antenna-related ECCM**). Control of frequency involves primarily frequency agility and frequency diversity. The objective of frequency agility and frequency diversity is to force the jammer to spread its energy over the wider bandwidth and to reduce the power density of jamming at the radar input. The methods of power control are based mainly on the brute-force approach of increasing radar transmitter power or more flexible approaches based on management of power in time and space. The control of the waveform includes using waveform coding based on intrapulse modulation that enables increased signal bandwidth, shaping of the transmitted radar pulse, and coding of various parameters of radar signal (e.g. PRF jitter and PRF stagger). *SAL*

Ref.: Skolnik (1990), p. 9.16; Farina (1992), p. 8.

ELECTRONIC COUNTERMEASURES (ECM) are "actions taken to prevent or reduce the enemy's effective use of the electromagnetic spectrum." It is one of the three components of electronic warfare. In radar applications its main objectives are to deny or to falsify information (detection, measurement, discrimination, and classification data) that the radar tries to obtain. There are a number of ways to classify ECM tactics and techniques. From the point of view whether electromagnetic energy is radiated or not, ECM is divided into active and passive. From the standpoint of what main parameters of radar information are influenced, ECM is divided into angle-measurement ECM, range-measurement ECM, and velocity-measurement ECM. From the viewpoint of the types of jammed radars, it is primarily divided into ECM versus search (surveillance) radars and ECM versus tracking radars. From the viewpoint of the ECM systems location, onboard and offboard ECM are distinguished. As to tactics of ECM combat employment, typically five classes can be identified: escort ECM, cooperative (mutual-support) ECM, self-protection (self-screening) ECM, stand-forward ECM, and stand-off ECM. Some authors include defence suppression as an ECM, although others consider it a separate ingredient of electronic warfare.

ECM is primarily based on jamming: both noise jamming and deception jamming. Modern ECM systems are designed to cope with different types of radars, and they have to operate in a dense threat environment that requires computer control of the system. Typical parameters of an advanced, self-protection active ECM system designed for fighter-type aircraft are given in Table E1. *SAL*

Ref.: Schleher (1986), pp. 109–198; Barton (1991), pp. 12.2–12.9; Skolnik (1990), p. 9.4; Wiegand (1991); Chrzanowski (1990); Neri (1991), pp. 337–416; VanBrunt (1978), (1982), (1995).

Table E1
Typical Advanced Self-Protection Jammer Parameters

Frequency coverage	0.7 to 18 GHz
System response time	0.1 to 0.25 sec
Receiver-processor:	
(a) Dynamic range	50 dB
(b) Sensitivity	–71 dBm
(c) Resolution	5 MHz
(d) Instantaneous bandwidth	1.44 GHz
(e) Pulsewidth	0.1 µs (min)
(f) Input pulse amplifier	20 dBm (max)
(g) False alarm rate	5/h (max)
(h) Signal detection capability	Pulsed, CW, PD, agile
Jammer:	
(a) Peak power	58 to 63 dBm
(b) Pulse-up capability	5 to 7 dB
(c) Set-on accuracy	±0.5 to ±20 MHz
(d) Duty cycle	5 to 10%
(e) Jamming capability	16 to 32 signals
(f) Modes	Noise, deception
(from Schleher, 1986, p. 169).	

Active ECM incorporates devices and methods based on deliberate radiation of electromagnetic energy to disrupt the operation of the victim radar. Active ECM typically means active jamming. *SAL*

Ref.: Barton (1991), p. 12-5; Chrzanowski (1990), p. 8.

Angle-measurement ECM is designed to cause disruption of correct determining of the target's angular direction both by search and tracking radars. The main methods to counter a search radar are to jam it through antenna sidelobes, based on the fact that, if a signal is received and detected, it is considered to be in the antenna main beam. So, if the energy of the signal radiated by a jammer and received by radar through antenna sidelobes is sufficient to exceed the detection threshold, the radar concludes that the signal was in the main beam and, therefore, considerable error is introduced in angular measurements. It is obvious that extremely large amounts of energy are required to overcome the main-to-sidelobe ratio of the victim radar. High-power noise jamming can be employed to implement this task, and special vehicles are being used in escort or stand-off ECM modes to carry the jammers required to protect the penetrating vehicles.

The main type of modern tracking radar is monopulse radar. Typically, jamming techniques against this type of radar fall into two categories. The first uses imperfection in the monopulse design and hardware implementation to reverse the sense of the angle-error signal. These involve cross-polarization jamming, image-frequency jamming, and skirt-frequency jamming. The second class uses the effect of multipoint jamming to distort true angle-of-arrival of real signal. These technique include blinking jamming, cross-eye jamming, formation jamming, and ground-bounce jamming. Down-link jamming may be a useful technique to reduce angular measurement capability of track-while-scan radars. *SAL*

Ref.: Leonov (1970), Ch. 8; Schleher (1986), p. 152; Chrzanowski (1990), pp. 91–107, 133–164.

Cooperative ECM is that "involving the coordinated conduct of electronic countermeasures by combat elements against hostile acquisition and weapon-control radars." Its advantages are coordinated tactics and greater effective radiated power that can be achieved due to jamming from various platforms. An example of this tactic is blinking jamming. This type of ECM is also termed *mutual-support ECM*. *SAL*

Ref.: Schleher (1986), p. 13; Skolnik (1990), p. 9.6.

Cross-eye ECM is a deceptive, self-protection technique designed to frustrate target-tracking radars or radar seekers by causing the radar to track a jamming signal whose phase front differs significantly from that of the true target return signal.

Cross-eye is one of several multisource jamming concepts (others include cross-polarization and terrain-bounce) designed to attack monopulse tracking radars, which are invulnerable to angle-jamming techniques such as inverse gain jamming effectively used against sequential-lobing and conical-scan tracking radars. Figure E3 shows a generic implementation of the cross-eye jamming technique, consisting of two repeaters connected to two separate antenna systems.

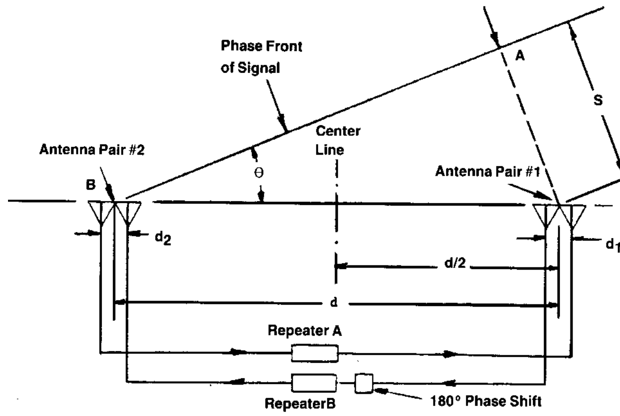


Figure E3 Block diagram and wavefronts in cross-eye ECM (from Schleher, 1986, Fig. 3-7, p. 159).

The transmit and receive antennas for each repeater are separated by the baseline distance *d*. One of the repeater paths includes a 180° phase shifter, and one path contains controls to ensure proper adjustment of relative phase and amplitude between the two repeater paths. With this configuration, the coherent jamming signals arrive at the victim radar matched in amplitude, but 180° out of phase. This creates an interferometric null between the two jamming signals in the direction of the victim radar antenna. Jammer effectiveness relies on the presence of true target-generated angle noise within the victim radar's angle tracking loop to drive the radar antenna off the jamming signal null so that a sufficiently high jammer-to-signal (*J*/*S*) ratio is developed.

The cross-eye concept has two inherent vulnerabilities: (1) the delay time introduced by the baseline distance between jammer antenna sets provides a potential opportunity for the victim radar to detect and track the leading edge of the true target return before the repeated jamming signal arrives, and (2) a relatively high J/S ratio, on the order of 20 dB or greater, is required for the technique to work effectively. *PCH*

Ref.: Schleher (1986), pp. 158–160.

Cross-polarization ECM is a deceptive ECM (DECM) technique, designed for use against monopulse tracking radars, that attempts to exploit the victim radar's response to a repeater jamming signal whose polarization is orthogonal to that of the radar receiving antenna. The response of the victim radar's angle error discriminator can be significantly different for cross-polarized signals, leading to large angle tracking errors and eventual break-lock.

Reflector antennas, particularly parabolic-dish designs, are most susceptible to cross-polarization jamming in that their response to orthogonally polarized signals may be reduced only 15 to 30 dB relative to their copolarized response. Flat-plate planar arrays are inherently less susceptible to cross-polarization jamming in that their response is typically reduced by 40 to 50 dB relative to copolarized signals. However, this apparent immunity can be largely negated by the presence of a radome, which may create significant levels of cross-polarized components, depending on the radome design.

Counter-countermeasures to cross-polarization jamming include the use of polarization screens to block signals with undesired polarizations. A more expensive and complex option entails the use of antennas capable of receiving multiple polarizations, perhaps in conjunction with one or more separate receiver and signal-processing channel.

To be effective, even against single-polarization radars, cross-polarization jammers must realize high *J/S* ratios while accurately maintaining jammer-to-radar signal polarization orthogonality (typically to within plus-or-minus 5°). If these conditions are not met, the jammer signal can become the equivalent of a target-borne radar beacon, making the jammer platform highly susceptible to engagement by radar homing missiles.

A block diagram of a cross-pol jammer is shown in Fig. E4. Signal components received by the vertically polarized antenna are amplified and retransmitted through the horizontally polarized antenna, and vice versa. As a result, any arbitrary elliptically polarized signal will be returned with the orthogonal polarization. The avoids the necessity of analyzing each received signal and adjusting the repeated signal polarization to be orthogonal to it. *PCH*

Ref.: Schleher (1986), pp. 153–154.

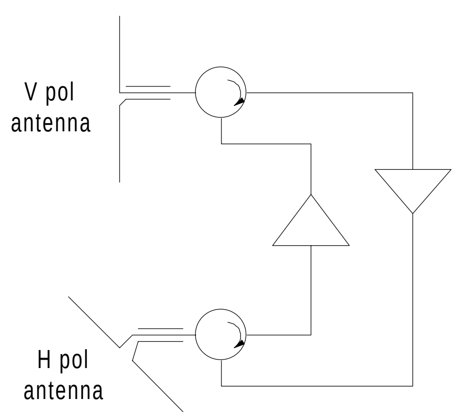


Figure E4 Block diagram of cross-pol ECM.

Deception [deceptive] ECM employs "the intentional and deliberate transmission or retransmission of amplitude, frequency, phase, or otherwise modulated intermittent or CW signals or the purpose of misleading in the interpretation or use of information by electronic system." The common acronym is DECM. In respect to radar, its main objective is to mask the real radar signal by injecting suitably modified replicas of target return into the victim radar. Typically, it is performed by deception jamming.

Angle deception is any ECM technique designed to frustrate a radar's ability to determine the relative angular position of the true target. When angle deception is employed against surveillance radars, the objective of the jammer is usually to insert multiple false targets into the radar's processor by injecting high-power signals through the victim radar's antenna sidelobes, thus making discrimination of the true target more difficult or impossible. Angle deception can also be directed to attack the angle-tracking capability of a tracking radar or radar homing missile seeker. In this role angle deception jamming is usually an integral component of a general class of DECM techniques used by self-screening jammers (SSJ) to protect the host platform (e.g., an aircraft, ship, or missile) from engagement by an air-defense system. Most angle deception techniques are implemented via a repeater jammer, which retransmits an amplified version of the victim radar's signal that has been modulated such that the angle data recovered by the victim radar no longer represents the true target's angular position relative to the radar.

To be effective, an angle deception jamming technique must be tailored to the particular way in which the victim radar derives target angle information. Track-while-scan (TWS) radars, sequential lobing, and conical-scan trackers derive target angle by sequential measurements of target energy as the radar antenna is pointed at different positions about the target location. Angle deception against these radar types is readily effected through a technique known as *inverse gain jamming*, in which an amplified replica of the radar signal is transmitted, but with amplitude modulation 180° out of phase from the original, which has the net result of canceling the amplitude modulation needed to derive target angle information.

In monopulse radars, all of the target angle information is available on a single pulse, rendering this type of tracking radar more difficult to jam. Deceptive techniques other than inverse gain jamming are required and these fall into two general classes:

(1) Those that exploit a particular mechanization or imperfection in the monopulse design.

(2) Those that exploit multiple signal source effects to distort the angle-of-arrival of the radar return signal. Techniques in class (2) are more "robust" in that they do not rely on detailed knowledge of a particular radar or missile seeker design for their effect. Angle deception techniques in this class include:

(1) Cross-polarization jamming, in which the repeated signal, with a high jammer-to-signal (*J/S*) ratio, has a polarization orthogonal to that of the original signal.

(2) Cross-eye, a technique that utilizes two separate repeaters, whose antennas are separated in space (e.g., one at each wingtip of an aircraft) to develop a combined signal whose phase front differs significantly from that of the true target return signal.

(3) Blinking jamming, in which two repeater jammers separated in space "blink," or transmit alternately in time at a rate designed to defeat the angle-tracking servo dynamics of the radar.

(4) Terrain bounce, which uses the earth's surface to reflect an airborne jamming signal and thus create a multipath situation in the elevation plane for the victim radar or missile seeker. If the victim is, for example, a semi-active homing missile, the objective of the terrain bounce jammer is to cause

the seeker to home on the jammer-illuminated patch on the earth's surface rather than on the jammer itself.

Both blinking and terrain bounce techniques can be implemented in either repeater jammer or noise jammer configurations. (See also **JAMMING, deception**.) *PCH, SAL*

Ref.: Schleher (1986), pp. 138, 145–160; Skolnik (1990), p. 9.5.

Downlink ECM is used against satellites, aircraft, and missile-borne radars that have separate, onboard systems to "downlink" target information or other radar sensor data, such as platform position and radar mapping data, from the radar acquisition platform to a ground radar or data processing center. Downlink ECM is directed at disrupting this line of communication.

In the interest of efficiency as well as security, most of the waveforms used by downlink transmitters are modulated in either time, frequency, phase, amplitude, or combinations thereof. Encryption algorithms may be superimposed for added security. In the absence of detailed knowledge of the downlinked signal coding, attempts to interfere with the downlink are mostly restricted to the use of stand-off noise jamming.

Downlink jamming is made more difficult still, due to the one-way transmission path of the downlink, and the usually high power transmitters employed by downlink systems. Downlink transmission systems generally employ fairly wide beamwidth antennas, but even coarse directivity can pose significant problems for a noise jammer, forcing it, in many circumstances (e.g., a missile downlink) to jam through the downlink antenna sidelobes. Yet another complication for the missile downlink jammer occurs when the downlink is used only infrequently, i.e., has a low duty cycle and so is vulnerable only for a small portion of the total missile flight time. (See also **JAMMING, downlink**.) *PCH*

ECM (range) equation (see **RANGE EQUATION**).

Escort ECM is the "ECM tactic in which the jamming platform accompanies the strike vehicle and jams radars to protect the strike vehicles," This tactic is applied to aircraft combat operations. It is generally used when the strike aircraft has no enough available power or payload for self-protection. Its effectiveness usually is greater than for stand-off ECM due to its closer proximity to the victim system. *SAL*

Ref.: Schleher (1986), p. 13; Skolnik (1990), p. 9.6.

Expendable ECM is "deployed once for a limited time offboard the platform which they are designed to protect." This technique is typically divided into active and passive expendable ECM systems. First usually are active decoys (miniature jammers), and the second type is primarily represented by passive decoys and chaff. To be cost effective, expendable ECM systems must be relatively cheap. Active expendable systems are usually more expensive than passive ones, and they tend to be used where passive devices are not effective. *SAL*

Ref.: Schleher (1986), p. 178.

Offboard ECM systems are designed to be used remotely from the defended platform (e.g., airborne or naval platforms). The main representatives of this systems are decoys (active and passive) and chaff. *SAL*

Ref.: Neri (1991), pp. 399–411.

Onboard ECM systems are ECM devices within the defended platform. Typically, they are active systems: noise and deception jammers. Sometimes various means for RCS reduction are classified as passive ingredients of onboard ECM systems. *SAL*

Ref.: Neri (1991), pp. 338–364.

Passive ECM refers to electronic countermeasures involving devices reflecting electromagnetic energy in such a manner that the reradiated signal competes with true target return to conceal real reflection. A typical example of passive ECM is chaff. *SAL*

Ref.: Barton (1991), p. 12-8; Chrzanowski (1990), p. 8.

Range-measurement ECM provides interfering signals, the main goal of which is disrupting the measurement of time of arrival to determine target range. Typically, it may be implemented in two ways. First is to cover or suppress echo return before it can be detected in radar receiver. Noise jamming is an appropriate technique for this purpose; only the high radiated power is required, especially if frequency agility is employed in victim radar. The second is to confuse a radar as to the true location of the target. This can be implemented by use of deception jamming. In this case false targets (passive decoys) presenting many radar blips with different ranges and range-gate pull-off are the effective measures.

Range gate pull-off (RGPO) is an range-measurement ECM technique in which a repeater captures the tracking range gate and introduces delay or advance to move the gate away from the target echo. A typical RGPO cycle is shown in Fig. E5. In the first phase of the cycle, the repeater pulse is

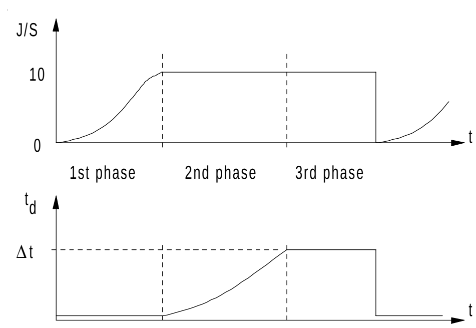


Figure E5 Range gate pull-off operation.

introduced and raised smoothly to a level sufficient to suppress the target echo pulse (without creating transients that might disclose the presence of the ECM). In the second phase the repeater pulse is delayed or advanced to pull the gate off the target echo. Only delay is possible in cases of PRF stagger or carrier frequency agile radars, except that a linear FM waveform permits a frequency shift to produce an early output from the pulse compression filter).

Against sophisticated trackers the apparent target acceleration of the repeater pulse must be maintained within the limits that can represent an intended target. In the third phase, when the target echo is not contributing to the output of the gate, angle deception modulations may be applied to the repeater pulse in an attempt to introduce an error in angle tracking rate. In the fourth phase, the repeater pulse is turned off, breaking the tracking loops. If sufficient angle rate has been introduced into the tracking loops, the beam may drift off the target before the receiver gain is increased and reacquisition of the target is attempted. This will force the radar to reinitiate its complete acquisition scan process, during which time the target is not under track. In the absence of a significant angle rate, the range gate can be swept to reacquire the target within fractions of a second. *SAL, DKB*

Ref.: Schleher (1986), pp. 143–145; Chrzanowski (1990), pp. 71–86, 114–117; Lothes (1990), pp. 114–119.

ECM versus search radar is used to prevent detection and acquisition of the target. To achieve this goal all main methods of active and passive jamming may be employed, including noise jamming, deception jamming, and use of chaff and decoys. The main types of ECM and relevant ECCM techniques are cited in Table E2. *SAL*

Ref.: Barton (1991), p. 12.15; Chrzanowski (1990), pp. 51–109.

Self-screening ECM is "conducted by individual combat elements to deny acquisition, tracking, or fire-control data to hostile weapon system." Typically, it requires that the single aircraft penetrating through an air defence system rely upon its own ECM capabilities. *SAL*

Ref.: Schleher (1986), p. 13; Skolnik (1990), p. 9.6

Stand-forward ECM is the ECM tactic "in which the jamming platform is located between the weapon systems and the strike vehicles and jams the radar to protect the strike vehicles." In this case only remotely piloted vehicles can meet the safety requirements as the platform with the jammer is usually located within the lethal range of defensive weapon system for a long period of time. *SAL*

Ref.: Skolnik (1990), p. 9.6.

Stand-off ECM is the ECM tactic in which missions are "conducted outside the lethal zones of hostile weapon-control systems to provide ECM support for friendly forces subject to hostile fire." Typically, the stand-off ECM system must provide high-power noise jamming able to jam the victim radar through the sidelobes at long ranges. *SAL*

Ref.: Schleher (1986), p. 12; Skolnik (1990), p. 9.6.

ECM versus tracking radar is used to prevent or delay acquisition of the target and to disrupt the tracking function of the radar (or at least introduce intolerable measurement errors). The primary types of ECM against tracking radar and relevant ECCM techniques are cited in Table E3. *SAL*

Ref.: Barton (1991), p. 12-17; Chrzanowski (1990), p. 109-164.

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Table E2
ECM against Search Radar

ECM type	Effects of ECM on radars	ECCM techniques
Barrage noise:		
• SOJ	Prevent target detection and acquisition	Low-sidelobe antennas, sidelobe cancellation, coherent integration
• SSJ and ESJ	Prevent target detection and acquisition	Jammer strobe processing, coherent integration
Spot noise:		
• SOJ	Prevent target detection and acquisition	Same as barrage noise SOJ plus frequency agility
• SSJ and ESJ	As above, plus receiver saturation	Same as barrage noise SSJ, plus frequency agility, wide dynamic range receivers
Swept noise	Same as barrage noise, plus false target generation	Guardband controlled blanking
Repeater (SSJ and ESJ)	False target generation, track file saturation[1]	Multiple track file maintenance[1], sidelobe blanking (ESJ)
Decoys: • Active repeater • Passive	False target generation, track file saturation[1]	Multiple track file maintenance[1], target analysis
Chaff	Prevent target detection and acquisition, false target generation	Doppler filtering and MTI

[1] Assumes track-while-scan (TWS) operation

from Barton (1991), p. 12.16.

Velocity-measurement ECM is used to disrupt the measurement of velocity based on doppler shift measurement with a doppler filter bank. In this case, deception jamming has a preference over noise jamming, as the latter has to spread its energy over the expected range of frequencies of possible radar returns (that may be tens of percent of the carrier frequency), resulting in very high requirements of effective radiated power. The main type of deception jamming against doppler radar is velocity-gate pull-off (VGPO), which is a deceptive jamming technique, often used by self-protection airborne jammers against coherent tracking radars or missile seekers, to produce erroneous target doppler measurement and ultimately to cause break-lock of the doppler-tracking loop. The jammer is either a true repeater or a transponder repeater that captures the victim radar or missile seeker velocity-tracking gate by reradiating the target signal at a high jammer-to-signal ratio and then pulls the captured gate off the true target doppler signal. A primary tactic with VGPO is to pull the doppler gate off and then momentarily shut off the jammer, forcing reacquisition of the target by the radar or seeker. The doppler pull-off rate can be calculated to avoid early recognition and counter by the victim, but spectral analysis and audio indications can reveal the presence of VGPO and, in the case of tracking radars and given enough time, manual tracking can defeat the VGPO. Modern doppler radars and missile seekers can be expected to contain effective and automatic countermeasures to simple VGPO; for this reason, the technique is seldom used alone but is combined in a repertoire of other deceptive jamming techniques. *PCH, SAL*

Ref.: Chrzanowski (1990), p. 123; Schleher (1986), p. 160.

Table E3
ECM against Tracking Radar

ECM type	Effects of ECM on radars	ECCM techniques
Barrage noise:		
• SOJ	Prevent target acquisition	Low-sidelobe antennas
• SSJ and ESJ	Deny target range	Sidelobe cancelation, coherent integration, jammer strobe processing (track on jamming), mainlobe canceler (ESJ)
Spot noise: SSJ and ESJ	Deny target range, receiver saturation	As above, plus wide dynamic range receivers
Repeater (SSJ and ESJ):		
• Simple type	Generate false targets (delay acquisition), track file saturation	Multiple track file maintenance, target signature analysis
• Range-gate pull-off (RGPO)	Break track	As above, plus dynamic track filtering
• Velocity-gate pull-off	Break track	As above, plus dynamic track filtering
• Cross-polarization jammer	Generate large tracking errors or break track	Planar antennas and radomes, polarization diversity
• Cross-eye jammer	Generate large tracking errors or break track	Multiple-phase-center antennas, separate transmit and receive antennas
• Surface-bounce	Generate large tracking errors or break track	Narrow-beam antennas, target signature analysis

Table E3
ECM against Tracking Radar

ECM type	Effects of ECM on radars	ECCM techniques
Special jammers:		
• Blinking (noise) jammer	Upset AGC circuits, break track (multiple SSJs)	Fast AGC, wide dynamic range receivers
• Skirt frequency jammer	Saturate receiver	Multiple track file maintenance
• Image frequency jammer	Generate large tracking errors or break track	Improved AFC, Σ and Δ channel phase matching
• Two-frequency jammer	Generate large tracking errors or break track	RF preselection filter
Chaff	Prevent target detection, false target generation	Doppler filtering and MTI, multiple track file maintenance
Passive decoys	Generate false targets, track file saturation	Multiple track file maintenance
from Barton (1991), pp. 12.18–19.		

ELECTRONIC INTELLIGENCE (ELINT) is “information that is the product of activities in the collection and processing, for subsequent intelligence purposes, of potentially hostile, non-communications electromagnetic radiations which emanate from other than nuclear detonations and radioactive sources.” The primary ELINT activities collect data on all types of radar, but navigation systems, command and control links, data links, and test range instrumentation (telemetry) may be included as well.

ELINT data may be used in several different ways:
(1) Operational capability: to evaluate the operational capability of radar and associated equipment as revealed by an analysis of key observable radar characteristics and parameters, including functions performed, modes, and performance.
(2) State-of-the-art assessment: by observing existing capabilities and features and comparing these with past records, an assessment of the technical progress made within a given time period may be made. Also, a review of the key features and performance may imply future trends and the ability to pursue certain technological radar solutions.
(3) Electronic order-of-battle: wide-area electronic surveillance can result in an accurate, timely geographically based inventory of a potential enemy’s radar deployment. Such an inventory may reveal potential strategic options available to the potential enemy, or indicate changes in strategic or tactical intent.
(4) Electronic countermeasures (ECM): all of the uses indicated above may serve to materially assist the ELINT host party in the development of new ECM techniques and equipment that may be required to respond effectively to observed

radar modifications and new radar capability. In addition, the electronic order-of-battle, if kept current, can be used to ensure that the appropriate sets of ECM equipment are available and deployed as the situation requires. Such knowledge can often mean the difference between success and failure when air attacks against the radar air-defense systems are contemplated. *PCH*

Ref.: Schleher (1986); Wiley (1982).

ELECTRONIC INTERFERENCE (see **INTERFERENCE**).

ELECTRONIC JAMMERS (see **JAMMER**).

ELECTRONIC JAMMING (see **JAMMING**).

ELECTRONIC RECONNAISSANCE (see **ELECTRONIC SUPPORT MEASURES**).

ELECTRONIC SIGHT (see **SIGHT**).

ELECTRONIC (WARFARE) SUPPORT MEASURES (ESM) are “actions taken to search for, intercept, locate, record, and analyze radiated electromagnetic energy for the purpose of exploiting such radiations in support of military operations.” The main difference between ESM and signal intelligence is that the first requires immediate actions while the latter gathers the intelligence information which can be processed later. The major constituent parts of ESM systems are radar warning receivers and appropriate antennas. Typically, an ESM system is completely passive and it locates the sources of electromagnetic energy (e.g., radars) only in angle by measuring the angle of arrival. It is an important advantage, as it ensures the security of operation and permits to detect the sources of radiation at ranges greater than even the range of intercepted sensors (radars, lasers, etc.). The disadvantage of the passive mode is that the range to the source of radiation must be generally obtained using multiple ESM fixes on the target (for example, by means of triangulation). To accomplish the comprehensive analysis of the intercepted sources including detecting, locating, classifying, and identifying the sensors, ESM system must have a threat library representing the main features of the enemy’s electronic order-of-battle stored in its microprocessor. The main ways to reduce the efficiency of ESM is to decrease the time when the active sensor that may be intercepted by this system is turned on to minimize warning time, and to use when possible completely passive weapons, such as antiradiation missiles and heat-seeking missiles. Typically, the ESM subsystem and ECM subsystem are the integrated parts of the same EW system, and the link between them is often automatic. *SAL*

Ref.: IEEE (1990), p. 16; Barton (1991), p. 12-2; Skolnik (1990), p. 9.2; Schleher (1986), p. 6; Neri (1991), p. 279.

ELECTRONIC WARFARE (EW) is “a military action involving the use of electromagnetic energy to determine, exploit, reduce, or prevent hostile use of the electromagnetic spectrum and action while retains friendly use of the electromagnetic spectrum.” EW is divided into three classic categories: electronic (warfare) support measures (ESM), electronic countermeasures (ECM), and electronic counter-countermeasures (ECCM). The structural tree depicting basic EW areas is shown in Fig. E6.

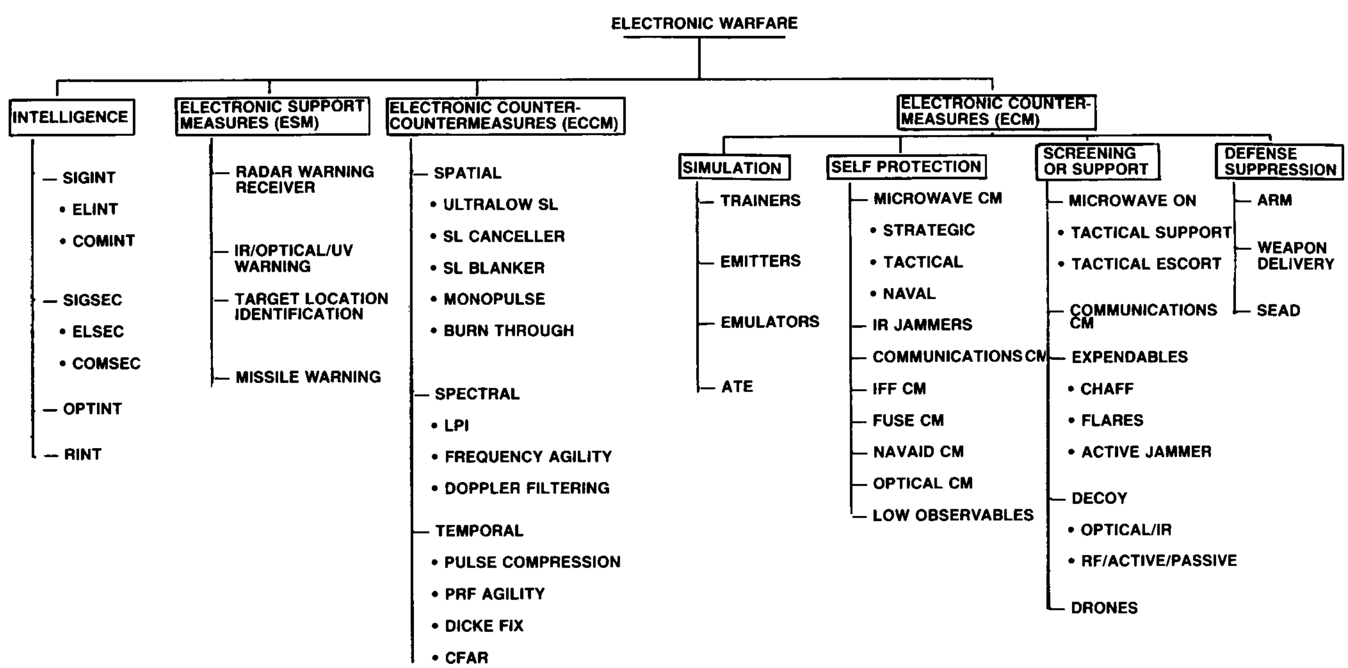


Figure E6 Electronic warfare areas (from Schleher, 1986, Fig. 1.1, p. 7).

In a broad sense EW typically incorporates the following constituent parts: electronic (warfare) support measures, signal intelligence, signal security, electronic countermeasures, electronic counter-countermeasures, defence suppression, and electronic warfare simulators. *SAL*

Ref.: Schleher (1986).

EW intelligence refers primarily to measures used to gain information on potential targets for ECM activities. Security measures such as radio or radar silence and encryption, intended to deny a potential enemy the opportunity to intercept one's own signals, are sometimes included in the definition of EW intelligence.

EW intelligence can be either tactical or strategic in nature. Tactical use of EW intelligence generally comes under the heading of electronic support measures (ESM), and refers to real-time collection and analysis of electronic intelligence (ELINT) for the purpose of updating an electronic order-of-battle (EOB) and for selecting ECM responses to a battlefield deployment of air-defense systems. Such information may be augmented by communications intelligence (COMINT) data, derived from intercepts of radio transmissions that may reveal tactical intent on the part of an enemy and battlefield deployments. Tactical combat aircraft are generally equipped with passive ESM equipment interfaced with active ECM systems for automatic response to threatening air-defense systems. ELINT and COMINT are subsets of the more general category of signals intelligence (SIGINT).

SIGINT can be a major sources of strategic intelligence that provides not only an up-to-date EOB, but technical data that may reveal important information about specific radar and radar weapon system capabilities.

Other categories of EW intelligence include operations intelligence (OPINT), which are nonelectronic means of collecting information that may reveal operational procedures, or tactical movement of enemy forces on the battlefield, and radiation intelligence (RINT), which is different than ELINT in that RINT relies on the interception of spurious and unintended radiation to derive information relative to the presence and characteristics of hostile electronic systems, including radar and EW systems.

Measures taken to deny an enemy knowledge of one's own presence and electronic systems capabilities include emission control (EMCON), an operational state in which no radiation is permitted (radar and radio silence), use of data encryption, and employment of low probability of intercept (LPI) radar or radio systems. *PCH*

Ref.: Schleher (1986), pp. 6–9.

Electronic warfare simulators are the simulators used for training of EW operators in a realistic combat environment, test and evaluation. The main purpose of EW simulators is to present EW data as would be encountered in an operating situation and simulate realistic threat scenarios common to modern EW environment. The basic types of these simulators involve the usage of pure hardware signal radiators, pure computer simulation, and combination of hardware and software usually referred to as hybrid simulators. *SAL*

Ref.: Schleher (1986), p. 19.

EW support measures (see **ELECTRONIC SUPPORT MEASURES**).

ENCODER, **shaft angle** (see **CONVERTER, angle- to-code**).

ELEVATION (see **ANGLE, elevation**).

ENVELOPE (see **DETECTOR, amplitude**).

EQUALIZATION is the process of removing amplitude and phase distortions in radar channels (amplitude equalization and phase equalization correspondingly). Typically, in hardware, the process of equalization is implemented by means of equalization filters which are designed by synthesizing an auxiliary transfer functions to produce flat amplitude and linear phase transfer characteristics. The transversal filter is widely used in the equalization of amplitude-phase distortions. An example of equalizing transversal filter based on wideband, dispersion-free, IF, tapped-delay-line used in pulse compression radar is shown in Fig.E7.

The necessity of equalization is especially acute in wideband radars. *SAL*

Ref.: Skolnik (1990), p. 10.36; Wehner (1987), p. 59.

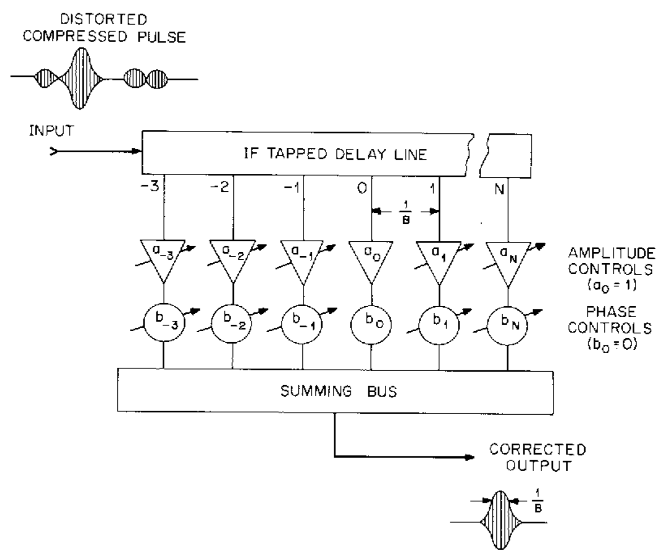


Figure E7 Transversal equalizer for reduction of time sidelobes (from Skolnik, 1990, p. 10.36, reprinted by permission of McGraw-Hill).

EQUIVALENCE PRINCIPLE. The equivalence principle is the principle based on the assumption that radiation of antenna computed from the electric and magnetic current sheets on the closed surface so that the fields inside a closed surface are zero and supposing that a given distribution of electric and magnetic fields was substituted by appropriate distribution of electric and magnetic current sheets is identical (except for a difference in sign) to the radiation that would have been produced by the original sources inside the closed surface drawn about the antenna structure. The principle was given by Schelkunoff. It is widely used in antenna theory to determine radiation from different types of aperture antennae: horn antennas, reflector antennas, and so forth. *SAL*

Ref.: Johnson (1984), p. 2-8.

ERROR, measurement. An error is a discrepancy between the true value of a target parameter and its measured value (the estimate at the output of a radar estimator). The main measured parameters in radar applications are signal time delay (target range), signal angle of arrival (target angular coordinates), signal doppler frequency (target radial velocity), and signal amplitude (target RCS). The fundamental source of measurement error is the presence of random noise in the receiver that introduces stochastic uncertainty in the process of radar measurement. In the general case the measurement error can be represented as a random function of time (see **error model**) and of target coordinates, the latter representation being much more complicated and applying to measurement in phased-array radars. Based on the representation of the target-radar measurement channel, the errors may be classified as target-dependent, propagation, radar-dependent, and, if the radar is located on a moving platform (ship, aircraft, spacecraft, etc.), platform-dependent errors. From the viewpoint of the measured parameter, angle, doppler, range, and RCS errors are distinguished. From the viewpoint of time variation (correlation function), they are classified as systematic (bias) and varying (slow or rapid) errors. The rapidly varying errors caused by receiver noise are called noise errors, and as they are uncorrelated from pulse to pulse they can be reduced by data smoothing. As a result, the final accuracy of the data is usually dominated by bias and slowly varying errors. Bias error can be reduced by proper radar pointing or calibration. *SAL*

Ref.: Barton (1964), (1969), (1988); Leonov (1990).

Acceleration error refers to the dynamic lag error in a tracking radar or track-while-scan loop, caused by target acceleration. This component of error can be found as the ratio of target acceleration a_t to the acceleration error constant K_a of the loop:

$$\varepsilon_a = \frac{a_t}{K_a}$$

When the acceleration is expressed in a linear coordinate, as m/s^2, the error will be in meters in that coordinate, while if angular acceleration is used, r/s^2, the error will be in radians in that coordinate.

The acceleration error constant can be expressed in terms of the closed-loop (noise) bandwidth, β_n, in hertz:

$$K_a = 2.5\beta_n^2$$

For a track-while-scan loop of the α-β filter type, K_a may be found as (Blackman, 1986)

$$K_a = \frac{\beta}{(1-\alpha)T^2}$$

where T is the data sample interval. *DKB*

Ref.: Barton (1988), pp. 463–466; Blackman (1986), p. 46.

Angular (measurement) error is the error in measurement of radar angular coordinates (e.g., azimuth and elevation in a land-based radar, yaw and pitch in an airborne radar or missile seeker, or other angular coordinates depending on the nature of the radar platform and application). The many sources of angular error may be divided into tracking errors, which cause the radar antenna axis to depart from the target angles; translation errors, which cause the axis angles to be reported incorrectly; and propagation errors. The several components of each type of error may further be divided into radar-dependent, target-dependent, and platform-dependent classes, and into bias and noise error components, as shown for a tracking radar of the monopulse type, in Table E4. *DKB*

Ref.: Barton (1988), pp. 533–548.

Table E4
Angular Error Components

Error class	Bias components	Noise components
Radar-dependent tracking errors	Boresight axis setting and drift; torque caused by wind and gravity; servo unbalance and drift	Thermal noise; multipath; clutter; jamming; torque caused by wind gusts; servo noise; deflection of antenna caused by acceleration
Radar-dependent translation errors	Pedestal leveling; azimuth alignment; orthogonality of axes; pedestal flexure caused by gravity and solar heating	Bearing wobble; data gear nonlinearity and backlash; data takeoff nonlinearity and granularity; pedestal deflection caused by acceleration; phase shifter error
Target-dependent tracking errors	Dynamic lag	Glint; dynamic lag variation; scintillation or beacon modulation
Propagation errors	Average refraction of troposphere and ionosphere	Irregularities in refraction of troposphere and ionosphere

An **anomalous error** is an error in evaluation of parameter that exceeds some specified value. It can arise because the conditions under which measurements are made deviate significantly from the standard case (e.g., because of signal-to-noise ratio decreases caused by jamming, clutter, or other adverse factors). *AIL*

Ref.: Kulikov (1978), p. 29.

Beam-steering error is the error between the commanded (and indicated) position of a phased-array beam and the actual position. For a linear array or rectangular array with common row or column phase-shift settings, the error due to granularity of an m-bit phase shifter is

$$\varepsilon_\theta = \frac{\theta_3}{2^m}\sqrt{\frac{G_d(\theta)}{G_0}}$$

where θ_3 is the half-power beamwidth, $G_d(\theta)$ is the normalized pattern derivative, or monopulse pattern for the scan angle, and G_0 is the gain for uniform illumination of the antenna aperture.

The average angular error for phase-shift errors σ_ϕ radian, which are independent over T elements in an array, is

$$\sigma_\theta = \frac{\theta_3 \sigma_\phi}{1.61\sqrt{T}}$$

DKB

Ref.: Barton (1969), pp. 196–197; Barton (1988), p. 171.

Bias error is the systematic component of error, which is constant in time or which varies slowly, preventing its reduction by averaging of sequential readings (smoothing) of the radar output. *DKB*

Boresight error is the displacement of the antenna electrical axis, or monopulse null axis, from the mechanical axis to which the output data are referred. *DKB*

Clutter error refers to the radar tracking error caused by presence of uncanceled clutter at the output of the receiver and signal processor. When random clutter sources extend across the radar beam, producing uncanceled clutter of power C_Δ in a monopulse difference channel, and C in the sum channel, the resulting angular clutter error component is

$$\sigma_\theta = \frac{\theta_3}{k_m}\sqrt{\left(\frac{C_\Delta}{2Sn_c}\right)\left(1 + \frac{C}{S}\right)}$$

where θ_3 is the half-power beamwidth, k_m is the monopulse difference slope, S is the signal power in the sum channel, and n_c is the number of independent samples of S/C_Δ averaged during the tracking time constant. If clutter is concentrated at a particular region in the difference pattern of the antenna, producing correlation between a component C_c in the sum channel and $C_{\Delta c}$ in the difference channel, an additional bias error may appear:

$$\sigma_{\theta 2} = \frac{\theta_3}{k_m}\sqrt{\frac{C_{\Delta c} \cdot C_c}{2S^2}}$$

This error is not reducing by averaging over successive pulses. *DKB*

Ref.: Barton (1988), pp. 531–533.

Collimation error refers to the boresight error component resulting from failure to perform initial alignment of the electrical and mechanical axes of the antenna. *DKB*

Cross-polarization error is the boresight error component resulting from reception of a signal component whose polarization is orthogonal to that of the intended antenna pattern. The antenna is collimated on a signal having the intended polarization, with response Δ in the difference channel going to zero at the axis. A component received with the orthogonal polarization is passed to the receiver as a result of undesired antenna response Δ_c in the antenna difference pattern. The resulting cross-polarization error is

$$\sigma_\theta = \left(\frac{\theta_3}{k_m}\right)\left(\frac{\Delta_c}{\Sigma}\right)\sqrt{\frac{\sigma_c}{2\sigma}}$$

where θ_3 is the half-power beamwidth, k_m is the monopulse difference slope, Σ is the on-axis sum pattern voltage response for the intended polarization, σ is the target RCS in the intended polarization, and σ_c that in the orthogonal polarization.

While the ratio σ_c/σ is usually < 1 for reflective targets, the presence of a jammer may greatly increase this ratio, causing large errors or breaking the track. *DKB*

Ref.: Barton (1988), pp. 415–416.

Diffraction error is a component of multipath error that is the result of signal components arriving at the radar antenna via diffraction from an obstacle underlying the target-radar path. *DKB*

Doppler (measurement) error is the error in radar measurement of target doppler (frequency) shift. The many sources of doppler error may be divided into tracking errors, which cause the radar tracking filter (velocity gate) to depart from the centroid of the target echo spectrum, translation errors, which cause the frequency of the filter or tracking oscillator to be reported incorrectly, and propagation errors. The several components of each type of error may further be divided into radar-dependent and target-dependent classes, and into bias and noise components, as shown for a tracking radar in Table E5. *DKB*

Table E5
Sources of Doppler Error

Error Class	Bias Components	Noise Components
Radar-dependent tracking errors	Discriminator zero setting and drift; gradient of receiver delay	Thermal noise; multipath; clutter; jamming; variation in receiver delay
Radar-dependent translation errors	Transmitting oscillator (velocity of light)	Range-doppler coupling; VCO frequency measurement; radar frequency stability
Target-dependent tracking errors	Dynamic lag	Glint (target rotation); target modulation
Propagation errors	Gradient of refraction of troposphere and ionosphere	Irregularities in refraction of troposphere and ionosphere

Ref.: Barton (1988), pp. 551–552.

dynamic error (see **lag error**).

fluctuation error (see **scintillation error**).

Geodetic error refers to the error in target position resulting from discrepancies between the model of the geoid (the average earth's surface) and the actual average surface, or between models used at the radar and in systems using radar data. It is usually a component of the platform-dependent error. *DKB*

Glint error refers to the random tracking error component of a radar operating against a complex target, resulting from interference between reflections form different scattering centers of the target. The glint error may exceed the physical extent of the target, as a result of ripples in the phase front received at the radar antenna. *DKB*

Ref.: Skolnik (1970), Ch. 28; Barton (1969), pp. 167–171.

Indicator (measurement error) is an error in parameter estimation due to indicator (usually display) equipment errors. Indicator errors arise for the following reasons. Errors caused by synchronization inaccuracy are determined by the error time between start of the indicator scan and radar transmitter signal, those arising in parameter estimation, and those due to scale and method of calculating the measured parameter. In modern radar, automatic coordinate measurement is used, and as a result indicator error does not affect parameter evaluation. *AIL*

Ref.: Dymova (1975), p. 129.

Instrumentation error is the error in evaluating radar accuracy that actually results from optical or other test instrumentation used as a tracking reference, rather than from the radar itself. When a boresight telescope is used as a reference, it is subject to parallax error relative to the radar tracking axis, as well as to errors in stability of the optics and reading of the film. Similar errors may arise in use of external reference instrumentation. *DKB*

Ref.: Barton (1964), p. 325.

Lag error refers to the failure of a tracking system to keep up with target velocity, acceleration, or higher derivatives of motion in radar coordinates. The conventional servo error analysis expresses total lag error in terms of a Taylor expansion of the target trajectory, each term being reduced by an appropriate error constant:

$$\varepsilon = \frac{x}{K_p} + \frac{\dot{x}}{K_v} + \frac{\ddot{x}}{K_a} + \ldots$$

where K_p is the position error constant (normally infinite), K_v is the velocity error constant, K_a is the acceleration error constant, and higher order terms are generally negligible. The velocity error constant can be made arbitrarily high, although transient effects limit its practical value. This error is also called *dynamic error*. See also **acceleration error**. *DKB*

Ref.: Barton (1988), pp. 463–466.

An **error model** is a mathematical description representing radar measurement error as a function of the radar and environmental parameters causing the error. The error is described as a random process (i.e., as a random function of time):

$$\xi = \xi(\vec{p}, t)$$

where $\vec{p}$ is the vector of parameters causing the specified error type (see **angular error**, **doppler error**, **range error**). Measurement errors usually include both systematic (bias) and random (noise) components (Fig. E8).

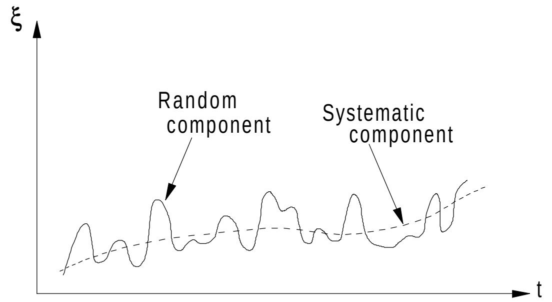


Figure E8 Radar measurement error.

The complete description of $\xi(t)$ requires knowledge of the nth-order probability distribution function, $F_n(\vec{x})$ from which one can determine the basic parameters of the error: its mathematical expectation (mean value), m_x; variance, D_x (and rms value, $\sigma_x = \sqrt{D_x}$); correlation function, $K_x(t_1,t_2)$; and power spectrum, $G_x(\omega)$. In practice, the nth-order pdf is seldom available, and these parameters are estimated from the samples, $z(tn)$ of the function $\xi(t)$ at N moments of time $n = 0, ..., N$. The function $\xi(t)$ is usually assumed to be stationary and ergodic. In that case, convenient expressions to estimate the basic error parameters are

$$D_x = \frac{1}{N+1}\sum_{n=0}^{N}[\xi(n\cdot\Delta t)]^2 - \frac{1}{(N+1)^2}\left[\sum_{n=0}^{N}\xi(n\cdot\Delta t)\right]^2$$

$$K_x(m\cdot\Delta t) = \frac{1}{N+1-m}\sum_{n=0}^{N-m}\xi(n\cdot\Delta t)\cdot\xi[(n+m)\cdot\Delta t] - \frac{1}{(N+1-m)^2}\sum_{n=0}^{N-m}\xi(n\cdot\Delta t)\sum_{n}\xi(n\cdot\Delta t)$$

$$G_x(\omega) = \frac{K_x(0)\cdot\Delta t}{2\pi} + \frac{1}{\pi}\sum_{m=1}^{N-1}K_x(m\cdot\Delta t)\cos\omega(m\cdot\Delta t\omega)\left(1-\frac{m}{N}\right)\Delta t$$

$$m_x = \frac{1}{N+1}\sum_{n=0}^{N}\xi(n\cdot\Delta t)$$

where Δt is the sampling interval. When these parameters are determined, the measurement error can be described in terms of its magnitude (mean value m_x and rms value σ_x) and its temporal behavior (i.e., whether it is a slow or fast function of time, as determined by the correlation interval, t_c, derived from $K_x(\tau)$). The final model of error can then be written as

$$\xi(t) = \sum_{i=1}^{3}[m_{x_i}(\vec{p}, t) + \sigma_{x_i}[\vec{p}, t]\cdot\eta_i(t)]$$

where m_{xi}, and σ_{xi} are mean values and rms errors for the systematic component (i = 1), the slowly fluctuating component (i = 2), and the rapidly fluctuating component (i = 3), $\eta_i(t)$ is a reference random function with zero mean value, unity variance, and specified correlation function $K_{xi}(\tau)$. Typically, $K_{x1}(\tau) = 1$, $K_{x2}(\tau)$ is defined by the slow fluctuations, and $K_{x3}(\tau) = \delta(\tau)$ (i.e., this portion of the model describes delta-correlated noise error (uncorrelated from sample to sample)).

If there are M independent components that originate each ith error constituent part,

$$m_{x_i} = \sum_{m=1}^{M} m_{x_{im}} = m_{i,j}$$

$$\sigma_{x_i} = \sqrt{\sum_{m=1}^{M} \sigma^2_{x_{im}}} = \sigma_i$$

and a convenient expression to simulate the error at the time $n\Delta t$, $n = 0, ..., N$, is

$$\xi(n \cdot \Delta t) = \sum_{i=1}^{3} [m_i + \sigma_i \cdot \eta_i(n \cdot \Delta t)]$$

as a discrete stationary random process. *SAL*

Ref.: Leonov (1990), p. 198; Barton (1988), pp. 505–512.

Monopulse (implementation) error results from imperfect balance in the RF and IF portions of the antenna feed and the sum and difference channels of a monopulse tracking radar. In Fig. E9, two input horns of a reflector or lens antenna are coupled through a hybrid network to sum and difference receiver channels. *DKB*

Ref.: Barton (1969), pp. 208–210.

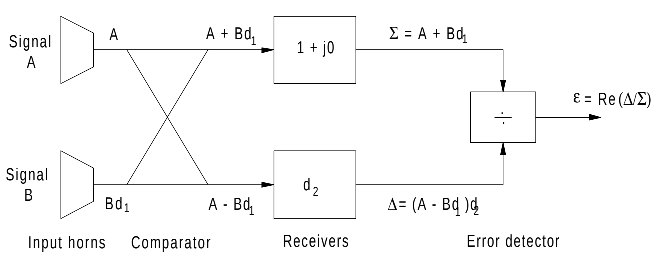


Figure E9 Monopulse implementation errors (after Barton, 1969, p. 208).

Multipath error occurs in radar tracking and measurement as a result of reflection of the target signal from the surface underlying the direct path. The most serious error appears in the elevation coordinate, where the specularly reflected image of the target, possibly replaced or surrounded by diffuse reflections, produces a broad signal distribution extending from the horizon (near zero elevation) to angles below the horizon in excess of the actual target elevation angle (Fig. E10).

The distribution of reflected power in the elevation coordinate may be calculated and plotted as in Fig. E11, with the

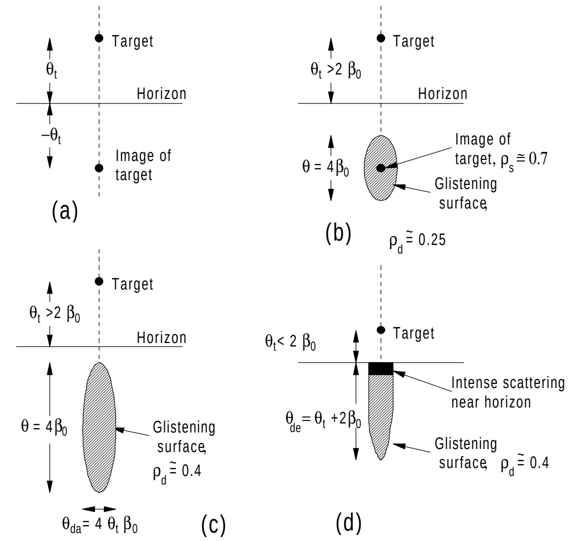


Figure E10 Multipath reflections for (a) smooth surface, (b) slightly rough surface, (c) rough surface, and (d) rough surface with low-elevation target (after Barton, 1969, p. 150).

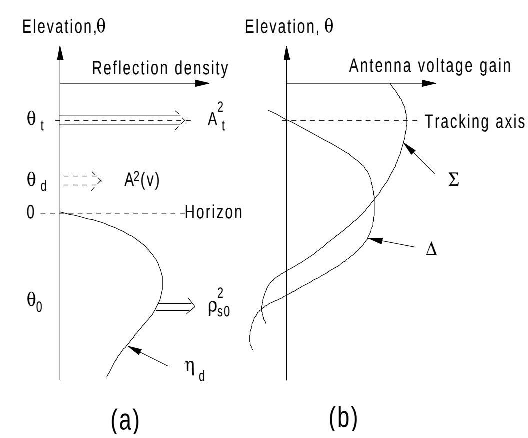


Figure E11 Multipath reflection density and antenna pattern weighting functions in elevation coordinate. (a) Target and reflection components, (b) sum and difference antenna gains (after Barton, 1988, p. 520).

voltage gain patterns of the tracking antenna. The power ratio of multipath reflection to signal is then found by integration of the pattern-weighted density over elevation:

$$\frac{I_\Delta}{S} = \frac{1}{A_t^2}\left[\int_\theta A_r^2(\theta)\rho_0^2(\theta)\eta_d\Delta(\theta)d\theta + A_r^2(\theta_0)\rho_0^2(\theta_0)\Delta^2(\theta_0)A_r^2(\theta_d)A^2(v)(\theta_d)\right]$$

The multipath interference causes a tracking error given by

$$\sigma_\theta = \frac{\theta_e}{k_m\sqrt{2\left(\frac{S}{I_\Delta}\right)n_e}}$$

where θ_e is the half-power beamwidth, $k_m \approx 1.5$ is the monopulse difference slope, and $n_e \approx 1$ is the number of independent multipath samples averaged in the tracker time constant. Typical variation in elevation multipath error as a function of target elevation is shown in Fig. E12, calculated for a radar with 40-mrad beamwidth but normalized to beamwidth for use with other values. It can be seen that the error varies between $0.05\theta_e$ and $0.1\theta_e$ for targets below one beamwidth, depending on surface conditions.

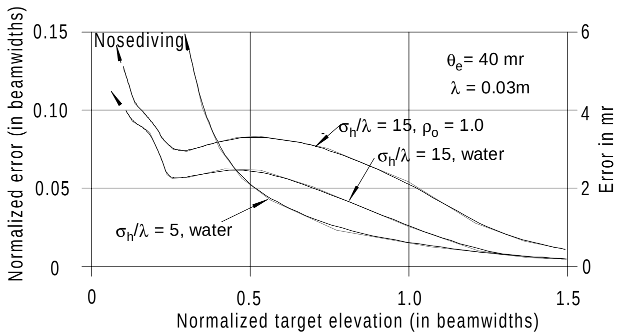


Figure E12 Typical elevation multipath error for radar tracking over a sea surface (from Barton, 1988, p. 526).

When the specular reflection coefficient is large ($\rho_0\ \rho_s > 0.7$), a "nosediving" phenomenon may occur for radar tracking at elevation angles below about $0.7\theta_e$. The tracking point lies at or near the horizon except when the reflected signal is 180° out of phase with the direct signal, in which case it rises abruptly to an elevation near $0.7\theta_e$. This is shown in Fig. E13, for an antenna beamwidth θ_e = 20 mrad, over rough, medium, and smooth surfaces. The large positive excursions occur when the signal fades into a reflection null.

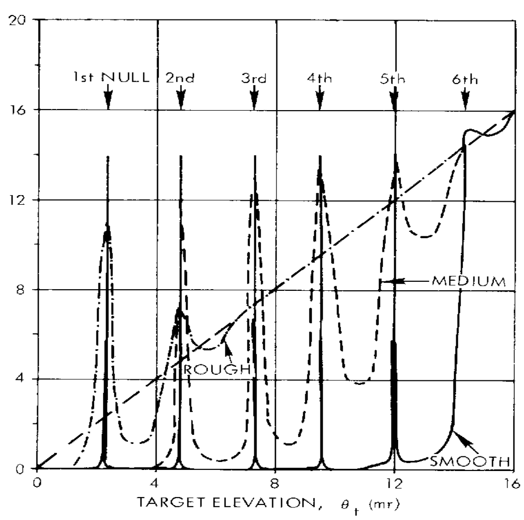


Figure E13 Tracking error with strong specular reflection (from Barton, 1988, p. 524).

Errors in azimuth and range coordinates will also be produced, but these are usually of lesser importance than the elevation error. *DKB*

Ref.: Barton (1988), pp. 512–531.

Noise error is an error caused by noise (generally of thermal origin) entering the radar receiver. This error component defines the fundamental accuracy of radar measurement (see **ACCURACY**); that is, the error below which no real system can measure. In most cases, there will be other significant errors (e.g., target-dependent, platform-dependent, and propagation errors). Detailed models for noise errors in all radar coordinates are given in Barton and Ward (1969). The general expression for noise error in a radar coordinate z_i (where z_1 = range, z_2 = angle, z_3 = radial velocity) can be given as

$$\sigma_{z_i} = \frac{K_i}{\sqrt{2(S/N)}}$$

where K_i is a coefficient proportional to the resolution of the radar in the coordinate concerned and S/N is the signal-to-noise ratio. The rule of thumb to estimate K_i is

$K_1 = c/(2B)$ for range measurement, where B is signal bandwidth and c is the velocity of light.

$K_2 = \lambda/(L_{ef})$ for angular measurement, where L_{ef} is effective aperture width and λ is wavelength.

$K_3 = \lambda/(2\tau_{ef})$ for velocity measurement, where τ_{ef} is signal duration.

Thus, for a given signal-to-noise ratio, range measurement is more accurate for waveforms of greater bandwidth, angular measurement for larger apertures, and velocity measurement for longer waveforms.

For angular measurements, the rms value of noise error in a monopulse system is given by

$$\sigma_\theta = \frac{\theta_3}{k_m\sqrt{2\left(\frac{S}{N}\right)n}}$$

where θ_3 is the half-power beamwidth, k_m is the monopulse difference slope, S/N is the sum-channel signal-to-noise ratio, and n is the number of independent noise samples averaged in the tracker time constant. For conical-scan radars, a slope k_s replaces k_m and the factor of two is absent from the denominator. In the more general case of interference, $(S/N)n$ in the denominator are replaced by $(S/I_\Delta)n_e$, where I_Δ is the interference entering the difference channel and n_e is the number of independent interference samples averaged during the tracker time constant.

When the parameters of the radar and waveform are fixed, higher accuracies are achieved with higher signal-to-noise ratios, following the relationship $\sigma \sim 1/(\sqrt{S/N})$. The noise error can be reduced by smoothing (time averaging) over a series of measurements, leaving the error close to the level set by equipment tolerances, propagation effects, and so forth. *DKB, SAL*

Ref.: Barton (1969), (1988), pp. 401–404; Leonov (1988), p. 27.

Peak error is the magnitude of the largest error in radar data, generally two to three times the rms error.

Peak-to-peak error is the difference between the largest positive and negative errors in radar data, generally four to six times the rms error.

Platform-dependent error is the error introduced by a movable platform (e.g., a ship, an aircraft, a spacecraft) in the measurement of the target coordinates. These errors are caused by a variety of factors: the errors in evaluation of the platform location in three-dimensional space by the navigation systems (autonomous or external), the errors in evaluation of the platform orientation (yaw, pitch, and roll), the spatial oscillations of the platform relative to its center of gravity, the dynamic deformations of the platform itself and radar antennae because of these oscillations, and so on. All of the errors caused by the variety of the different external factors depending on the type of movable platform can be classified as the errors of nonradar character arising from the linear and angular platform motion in 3D space (including oscillating motion relative to its center of gravity) and the attendant factors (e.g., deformations of the platform), and the errors of the radar-performance character arising from the distortion of radar characteristics due to the platform motion (e.g., distortion of angle-sensing response due to dynamic antenna deformation).

Table E6
Platform-Dependent Errors

Coordinate	Evaluation formula	Source of error
Azimuth, ϕ	$\Delta\phi_i = \begin{cases} \operatorname{acos}(\cos\phi + \delta\phi_i) - \phi\ ,\ 0 \le \phi \le \pi \\ 2\pi - \operatorname{acos}(\cos\phi + \delta\phi_i) - \phi\ ,\ \pi \le \phi \le 2\pi \end{cases} \quad i = 1, 2$ $\delta\phi_1 = \sin\phi \cdot \Delta\alpha_1 + \cot\theta[(\sin\phi)^2 \cdot \Delta\alpha_2 - 0.5\sin 2\phi \cdot \Delta\alpha_3]$ $\delta\phi_2 = \varepsilon\csc\theta\sin\eta_3(\cos\phi\cos\beta - \cos\eta_2)\ ,\theta \ne 0, \pi$	For $\Delta\phi_1$, $\Delta\theta_1$: errors in orientation of reference frame (Fig. E14), e.g., errors in measurement of yaw ($\Delta\alpha_1$), pitch ($\Delta\alpha_2$), roll ($\Delta\alpha_3$); deformations causing angular displacement ($\Delta\alpha_1$, $\Delta\alpha_2$, $\Delta\alpha_3$) etc.
Elevation, θ	$\Delta\theta_i = \operatorname{acos}(\cos\theta + \delta\theta_i) - \theta\ ,\ 0 \le \theta \le \pi$ $\delta\theta_1 = -\sin\theta(\cos\phi \cdot \Delta\alpha_2 + \sin\theta \cdot \Delta\alpha_3)$ $\varepsilon = \eta_1/R\ ;\ \beta = \phi - \eta_2$ $\delta\theta_2 = \varepsilon[0.5\sin(2\theta)\sin\eta_3\cos\beta - (\sin\theta)^2\cos\eta_3]$	For $\Delta\phi_2$, $\Delta\theta_2$, ΔR: errors in origin of the location of the reference frame (Fig. E15); e.g., deformations causing linear displacements; errors in platform location measurement, and so forth.
Range, R	$\Delta R = -\eta_1(\sin\theta\cos\beta\sin\eta_3 + \cos\theta\cos\eta_3)$	

Figure E14

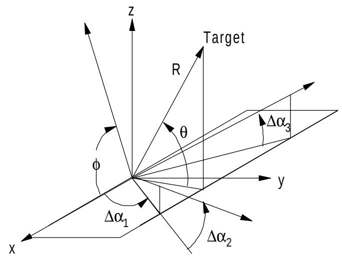
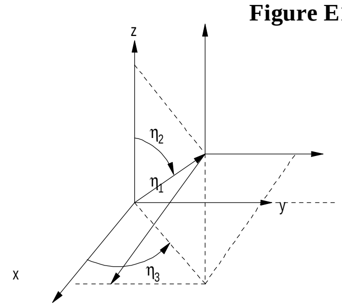


Figure E15



The factors and errors of nonradar character can be described through the introduction of different reference frames corresponding to the appropriate distortion factor and evaluation of target coordinate measurement errors as the function of ambiguity in evaluation of origin and orientation of corresponding frames (Table E6) The errors of radar-performance character can be evaluated applying the general approaches of the radar and antenna theory. The resultant errors of the target coordinates measurement for the radars located on the movable platforms (shipborne, airborne, and spaceborne radars) are the statistical sum of radar-dependent, target-dependent, platform-dependent, and propagation errors. Platform-dependent errors can be prevalent for precision movable radars operating in the lower portion of wavelengths (e.g., millimeter wave airborne and spaceborne radars). *SAL*

Ref.: Leonov (1988), p. 29; Leonov (1990), pp. 174–203

polarization error (see **cross-polarization error**).

Probable error is the value exceeded 50% of the time. For an error with normal distribution and standard deviation σ_x, the probable error is $x_{50} = 0.6745\sigma_x$. For a two-dimensional error with normal distribution and standard deviation σ_x in each coordinate, the circular probable error is $r_{50} = 1.177\sigma_x$. *DKB*

Ref.: Barton (1964), pp. 330–333.

Propagation error results from atmospheric refraction along the radar-target path. The error is evident primarily in elevation and range coordinates. The error is proportional to the refractivity N_s, measured at the radar site. The refractive bias errors in elevation and range, for a radar at sea level operating through the exponential reference atmosphere, are shown in Figures E16 and E17.

There are also small fluctuating errors in all three radar coordinates, resulting from variations in refractive index in the atmosphere. The tropospheric components of these errors are usually a few hundredths of a milliradian.

The ionospheric errors are strongly dependent on frequency. Bias errors in elevation and range are shown in Figures E18 and E19, for daytime conditions at zero elevation angle. *DKB*

Ref.: Barton (1969), pp. 366–393.

Quantization error results from the granularity of digital processes as the result of data quantization, typically in steering the antenna beam (see **beam-steering error**), reading the angles of a mechanically steered antenna, or converting signal voltages to digital form in an analog-to-digital converter. When a shaft angle encoder having m bits and a minimum angle quantum $\Delta = 360°/2^m$ is used, the peak-to-peak error is Δ and the rms error is.

$$\sigma = \frac{\Delta}{\sqrt{12}} = \frac{180°}{2^m\sqrt{3}}$$

A similar error appears when voltage is quantized with m bits representing a maximum amplitude A, in which case there will be an rms noise introduced:

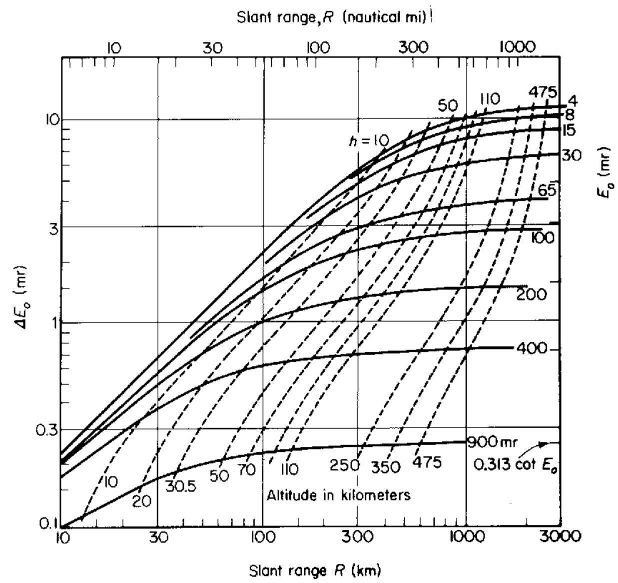


Figure E16 Elevation bias error vs. range for exponential reference atmosphere, $N_s = 313$ (from Barton, 1988, p. 306).

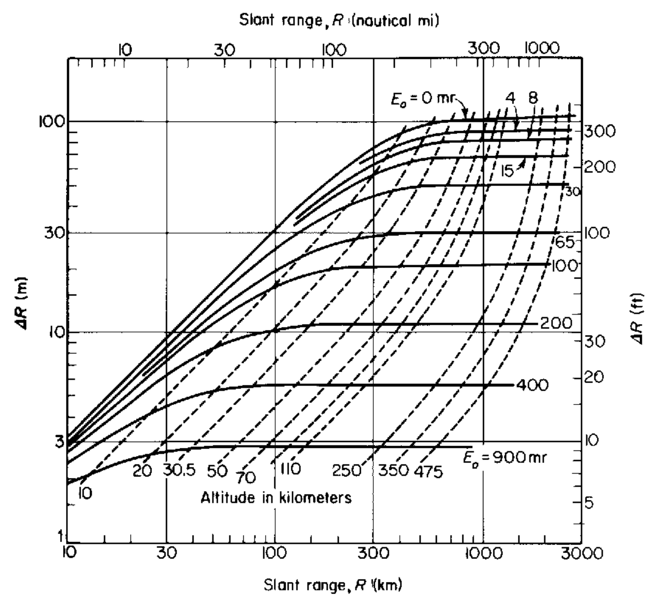


Figure E17 Range bias error vs. range for exponential reference atmosphere, $N_s = 313$ (from Barton, 1988, p. 307).

$$\sigma_a = \frac{A}{2^m\sqrt{12}}$$

Note that m bits plus a sign bit are needed to obtain this noise on a sine wave of amplitude A. *DKB*

Ref.: Barton (1969), pp. 187–193.

Radar-dependent error is the portion of error dependent primarily on design of the radar, independent of target-induced errors such as glint and dynamic lag, and of propagation errors (see **angular error, doppler error, range error**). *DKB*

random error (see **error model**).

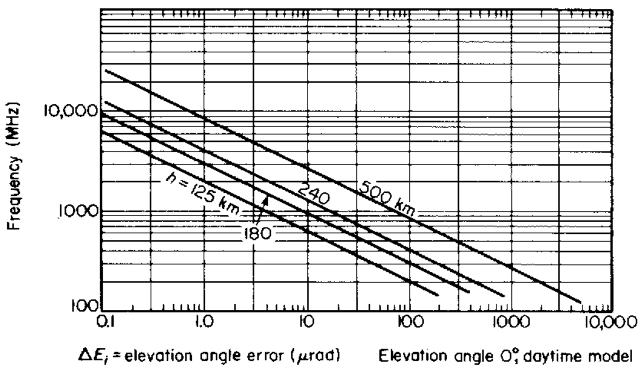


Figure E18 Ionospheric elevation error vs. frequency (from Barton, 1988, p. 312).

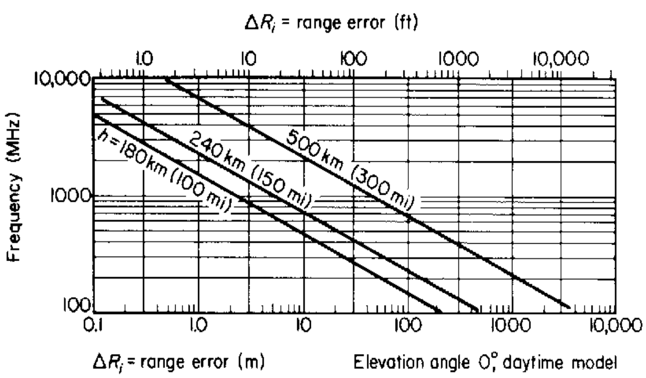


Figure E19 Ionospheric range error vs. frequency (from Barton, 1988, p. 312).

Range error is the error in measuring slant range from the radar. The many sources of range error may be divided into tracking errors, which cause the radar range gate or target strobe to depart from the centroid of the target echo pulse (or equivalent for modulated CW radar); translation errors, which cause the delay of the gate or strobe to be reported incorrectly; and propagation errors. The several components of each type of error may be further divided into radar-dependent, target-dependent, and platform-dependent classes, and into bias and noise components, as shown for a typical tracking radar in Table E7. *DKB*

Ref.: Barton (1988), pp. 533–548.

RCS measurement error is the error in measurement of radar cross section of the target, based on measurement of signal amplitude. When the amplitude of a signal received from a target at known range and angular coordinates is measured, the RCS can be determined from the relative power of that signal and known parameters of the radar. The difference between RCS measurement and target coordinate measurement is that the measured parameter itself determines the signal-to-noise ratio S/N so theoretical expressions for error are more complicated than for target coordinates. For the case of RCS measurement in a background of white noise with Gaussian distribution, the noise error in measured amplitude is

$$\sigma_A = \frac{1}{\sqrt{2(S/N)}}$$

SAL

Ref.: Skolnik (1962), p. 463; Shirman (1970), p. 177.

Table E7
Range Error Components

Error class	Bias components	Noise components
Radar-dependent tracking errors	Zero range setting; range discriminator shift; receiver delay	Thermal noise; multipath; clutter; jamming; variation in receiver delay
Radar-dependent translation errors	Range oscillator (velocity of light)	Range-doppler coupling; internal jitter; data encoding; range oscillator stability
Target-dependent tracking errors	Dynamic lag; beacon delay	Glint; dynamic lag variation; scintillation; or beacon jitter
Propagation errors	Average refraction of troposphere and ionosphere	Irregularities in refraction of troposphere and ionosphere

refraction error (see **propagation error**).

Root-mean-square (rms) error is the square root of the variance of the corresponding error represented as a random function of time. When the individual components of error are independent, it can be represented as the square root of the sum of the squares of all individual error components, including the bias errors. For a series of n data points having individual errors x_i, $i = 1, 2, \ldots, n$, the rms error is

$$x_{rms} = \sqrt{\frac{1}{n}\sum_{i=1}^{n} x_i^2} = \sqrt{\bar{x}^2 + \frac{1}{n}\sum_{i=1}^{n}(x_i - \bar{x})^2}$$

where $\bar{x}$ is the total bias error and $x_i - \bar{x}$ is the random error for the ith point. *DKB*

Ref.: Barton (1988), p. 507.

Sampling error is the result of taking discrete samples of a continuous quantity at time intervals (sampling intervals) δt. The result may be to cause aliasing of the output data, in which the frequency components above $1/2\delta t$ are folded into the spectrum below this value. *DKB*

Ref.: Barton (1969), pp. 183–187.

Scintillation error is the result of rapid target amplitude fluctuations, which interact with the measurement system. The error is a significant limitation in sequential lobing and conical scanning radars, in which the target fluctuation distorts the modulation expected as a result of antenna scan. In conical scanning at a frequency f_s, the effective signal-to-interference ratio used in the denominator of the noise error expression is

$$\left(\frac{S}{I_\Delta}\right)_{n_e} = \frac{1}{2\beta_n W(f_s)}$$

where β_n is the tracking bandwidth and $W(f_s)$ is the power spectral density of the target fluctuation at the scan frequency. The corresponding scintillation error is

$$\sigma_s = \frac{\theta_3}{k_s}\sqrt{\beta_n W(f_s)}$$

For example, if the fluctuation power density is $W(f_s) = 0.01$ per hertz at the scan frequency, with $k_s = 1.4$ and $\beta_n = 1$ Hz, the rms error is $\sigma_s = 0.07\theta_3$.

A similar error occurs in sector-scanning radar, including search radars rotating through 360°. For the usual Rayleigh fluctuating target (Swerling case 1 or 2) with correlation time t_c, the error depends on the number of independent amplitude-difference samples received during the observation time, or time-on-target t_o, according to Fig. E20. This is sometimes called *fluctuation error. DKB*

Ref.: Barton (1969), pp. 171–182.

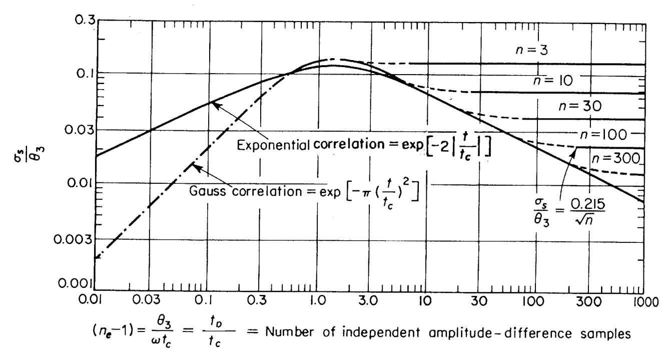


Figure E20 Scintillation error in sector-scanning radar (from Barton, 1969, p. 180).

Survey error is the error resulting from failure to locate accurately the position of the radar antenna with respect to the geodetic coordinate system to which the data are referred. In modern systems, accurate location using such techniques as GPS reduces this error to a negligible level, even for transportable and shipborne radars. *DKB*

Systematic error (see **bias error**).

Target-dependent error refers to error components that depend on target parameters and trajectory. See **angular error**, **doppler error**, **range error**. *DKB*

Tracking error is the component of error resulting from inability of the radar tracker loops to follow accurately the centroid of the target echo in angle, range, and doppler. See **angular error**, **doppler error**, **range error.** *DKB*

Translation error is the component of error resulting from inability of the radar data system to report accurately the indications of the tracking loops in angle and range. See **angular error**, **doppler error**, **range error.** *DKB*

Tuning error can refer to the error in frequency setting of the radar receiver circuits, relative to the received echo signal, or to errors in angle or range that result from the receiver tuning. The latter definition includes the effects of phase and time delay change in the receivers, resulting from a signal not centered in the bandpass. *DKB*

Weighting error refers to the failure of a receiving array network to apply accurately to the array elements the weighting function desired for low-sidelobe performance. The rms errors can be expressed as an amplitude component σ_a (normalized to the intended element amplitude at the center of the array), and a phase component σ_ϕ, in radians. For random errors, the resulting mean sidelobe power level, relative to isotropic gain, is

$$G_s = G_e(\sigma_a^2 + \sigma_\phi^2)$$

where $G_e \approx \pi$ is the array element gain. *DKB*

Ref.: Nitzberg (1992), p. 38; Mailloux (1994), pp. 395–399.

Wind-torque error is the result of wind forces on a mechanically pointed antenna. The wind may cause both mechanical deflection of the antenna surface and servo tracking error, in which the target error signal must generate sufficient torque at the servo output to balance the wind torque. *DKB*

Ref.: Barton (1964), pp. 331–335.

velocity (measurement) error (see **doppler error**).

ESTIMATOR (see **TRACKER**).

EXCITER. A radar exciter is a unit that generates waveforms used in the transmitted signal. It consists of two parts: a waveform generator that creates and shapes the waveform modulation, and an upconverter that translates this waveform to the transmitted carrier frequency. An important requirement on the exciter, especially in a coherent radar, is its frequency stability. High stability is usually achieved using crystal oscillators driving frequency multipliers in stable local oscillator chains. Exceptionally high stability for certain classes of pulsed doppler and continuous-wave radars is achieved using cavity-stabilized microwave oscillators (usually klystrons) at or near the carrier frequency. The crystal oscillator provides better long-term stability, at the expense of greater phase instability, because multiplication increases the phase instability of the basic crystal oscillator by the square of the multiplication ratio (phase noise sidebands increase in power by 20log(frequency ratio)). A typical crystal oscillator operates at 90 MHz, and is multiplied by a factor of 32 to obtain an S-band carrier output. Tunability of the crystal oscillator is very limited (e.g., 0.25%), and frequency change is usually obtained by changing crystals. The configuration of a typical exciter for coherent radar is shown in Fig. E21. The exciter can be included in the transmitter or the receiver subsystem of the radar. *SAL*

Ref.: Fink (1982), p. 25.70

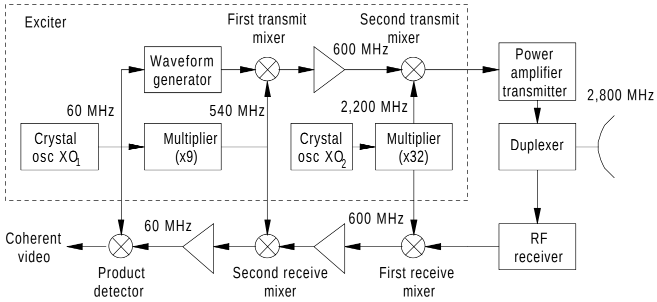


Figure E21 Coherent-radar exciter (after Fink, 1982, Fig. 25.82, p. 25.72).

F

FADING is "the variation of radio field intensity caused by changes in the transmission medium, and transmission path with time." Typically random variations appear in the signal received via various transmission media, especially for propagation via the troposphere or ionosphere at frequencies above about 100 kHz. Usually two types of such variations are distinguished: power fading (or attenuation) that is associated with comparatively large-scale changes in the medium, and variable-multipath or phase-interference fading. In radar, severe fading has a strong effect on the target detection performance. The readers who are interested in details can find many graphical results in Cantafio (1989), where the influence of various level of scintillation severity from no scintillation to worst-case Rayleigh fading on probability of detection are cited. *SAL*

Ref.: IEEE (1993), p. 476; Fink (1975), p. 18.70; Cantafio (1989), pp. 97–117.

FAILURE. A failure is an event involving disruption of the operability of a radar. The term failure must be understood to mean not only complete loss of radar operability, but also its deterioration due to excursion of the value of parameters beyond established tolerances. Failures are categorized as follows:

1. Sudden or gradual failures, based on the nature of the parameter change. A sudden failure is a failure characterized by a spasmodic change in one or several parameters. It arises randomly and unexpectedly, and cannot be corrected during radar maintenance. A gradual failure is a failure characterized by a gradual change in the values of one or several parameters. Hence, gradual failures can be forecast and averted through preventive maintenance.

2. Dependent or independent failures, based on their interconnection. A dependent failure is a failure of an element caused by the failure of another element. An independent failure is a failure of an element not caused by failures of other elements but arising for other reasons.

3. Self-correcting or intermittent failures, based on the nature of the time of operability disruption. A self-correcting failure is a failure leading to a brief disruption of operability. This failure sometimes is referred to as a malfunction. An intermittent failure is a frequently occurring, self-correcting failure of the identical nature.

4. Explicit or implicit failures, based on presence of external signs. An explicit failure is a failure that is detected immediately when it occurs without use of meters. An implicit failure is a hidden failure lacking external manifestations and that can be detected only by means of the corresponding measurements.

The number of failures per unit of time is referred to as failure rate, while the average value of the time of operation between failures is called mean-time-between-failures. Since the process of failures in a radar is random and depends on many factors, failure-free performance time is also random and a series of distributions, the most widely used being the Weibull, the exponential, and the Rayleigh distributions, are used to describe it. *AIL*

Ref.: Fink (1982), Ch. 28; Leonov (1991), p. 10.

Mean-time-between-failures (MTBF) is an expected operating time between failures. It is an indicator of failure-free performance of repaired radar elements, units, assemblies, and the station as a whole, usually expressed in hours. *AIL*

Ref.: Fink (1982), p. 28.4; Leonov (1991), p. 31.

Mean-time-to-first-failure expected value of operating time until the first failure. Time-to-first-failure is an indicator of failure-free performance of unrepaired radar elements and units and is expressed in hours. *AIL*

Ref.: Leonov (1991), p. 27.

FALSE ALARM. A false alarm is "an erroneous radar target detection decision caused by noise or other interfering signals exceeding the detection threshold." (See also **DETECTION.**) In general, it is an indication of the presence of a radar target when there is no valid target. False alarms are generated when thermal noise exceeds a preset threshold level, by the presence of spurious signals (either internal to the radar receiver or from sources external to the radar), or by equipment malfunction. A false alarm may be manifested as a momentary blip on a cathode ray tube (CRT) display, a digital signal processor output, an audio signal, or by all of these means. The task of the radar designer is to establish the threshold such that the radar target detection goals (expressed in terms of a specified detection range against a target of a specified radar cross section, with a specified probability) can be met with a signal-to-noise ratio (SNR) that is consistent with the capabilities of the radar design. If the detection threshold is set too high, there will be very few false alarms, but the SNR required will inhibit detection of valid targets. If the threshold is set too low, the large number of false alarms will mask detection of valid targets, and scarce radar resources (e.g., time) will be expended in investigating the false alarms. Solutions to the false-alarm problem involve implementation of constant false-alarm rate (CFAR) schemes that vary the detection threshold as a function of the sensed environment. See **false-alarm probability**; **CONSTANT FALSE-ALARM RATE (CFAR)**. *PCH*

Ref.: IEEE (1993), p. 480; DiFranco (1968), p. 270.

constant false-alarm rate (see **CFAR**).

False-alarm control refers to any technique applied to the output of a radar detector or detection scheme that attempts to maintain a constant false-alarm rate under varying environmental conditions. (See **CFAR**). A fixed detection threshold, if set sufficiently high, will result in a uniform false-alarm rate, but at the cost of reduced sensitivity to true target returns. Such use of a fixed threshold is generally unacceptable, and three different forms of false-alarm control have been devised in an attempt to solve the problem:

(1) Adaptive thresholding.
(2) Nonparametric detectors.
(3) Clutter maps.

The adaptive thresholding and nonparametric detector approaches both make the assumption that the samples in the reference cells, or the range cells surrounding the test cell, are independent, both in space and time, and have the same statistical distribution. The detectors then compare the return from the test cell with the statistical average from the reference cells and make a detection decision if the test cell return is sufficiently larger.

Clutter maps are most effective in fixed-frequency, ground-based radars in a stationary land clutter environment. A clutter map is made by storing an average background clutter level for each range-azimuth cell, and a target detection is declared in a particular cell if the return exceeds the average for that cell by a preestablished amount. The threshold is determined based on the average return in that cell over previous scans (e.g., 5 to 10 scans). An advantage of clutter maps is that they give the radar a useful interclutter visibility capability; they can see targets that lie between large clutter returns. (See **CFAR**.) *PCH*

Ref.: Skolnik (1990), pp. 8.12–8.21.

The **false-alarm number** is the reciprocal of the false-alarm probability $1/P_{fa}$ where P_{fa} is the probability that the interference voltage envelope will exceed some threshold voltage V_T. (See **false-alarm probability**.) *PCH*

Ref.: Skolnik (1980), p. 24.

False-alarm probability. If a threshold device is used to make a decision as to the presence or absence of a signal in a background of noise, the performance of this device can be expressed in terms of two probabilities: the probability of detection P_d and the probability of false alarm P_{fa}. The threshold can be considered the value of a receiver output voltage V_T, that when exceeded, indicates a detection. Due to the presence of thermal noise in the receiver, there is always a nonzero probability that the threshold will be exceeded, even in the absence of a target signal. The probability that the threshold value V_T is exceeded when no signal is present is the false-alarm probability. For Gaussian noise passed through a narrowband IF filter, the noise envelope has a Rayleigh distribution, and in this case we can write

$$P_{fa} = \exp\left(\frac{-V_T^2}{2\psi_0}\right)$$

where ψ_0 is the mean square value of the noise voltage in a narrowband (compared to midfrequency) IF filter.

In Fig. F1, T_k is the time between crossing of the threshold V_T by the noise envelope, when the slope of crossing noise voltage is positive, and t_k, t_{k+1}, t_{k+n} are the times during

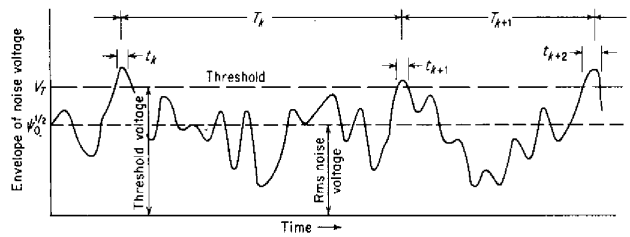


Figure F1 Envelope of receiver output showing false alarms due to noise (from Skolnik, 1980, Fig. 2.4, p. 25, reprinted by permission of McGraw-Hill).

which the noise voltage lies above the threshold. The average time interval between crossings of the threshold by noise alone is the false alarm time T_{fa} where

$$t_{fa} = \lim_{N \to \infty} \frac{1}{N} \sum_{k=1}^{N} T_k$$

The false-alarm probability may also be defined as the ratio of the duration of time the noise voltage envelope is actually above the threshold, to the total time it could have been above the threshold. The average duration of a noise pulse is approximately equal to the reciprocal of the bandwidth B, so that

$$P_{fa} = \frac{\langle t_k \rangle_{av}}{\langle T_k \rangle_{av}} = \frac{1}{t_{fa}B}$$

The false-alarm time can be seen also to equivalent to

$$t_{fa} = \frac{1}{B}\exp\left(\frac{V_T^2}{2\psi_0}\right)$$

PCH

Ref.: Skolnik (1980), pp. 23–25.

false-alarm time (see **false-alarm probability**).

FANTASTRON. A fantastron is a single-stage relaxation vacuum-tube oscillator of linearly dropping voltage operating in the free-running or delay-line mode. A fantastron processes sawtooth and rectangular voltage pulses, the duration of which can be regulated within broad limits by changing the control voltage.

Fantastrons were used in older types of radars to obtain precise adjustable pulse delays, to determine the time interval between pulses, and to generate sawtooth voltage. *IAM*

Ref.: Popov (1980), p. 454.

FARADAY

The **Faraday constant** is the angle of rotation of the polarization plane per unit path length, under the conditions of linear polarized wave propagation in a magnetized ferrite. The typical notation is

$$R_f = \frac{\omega\sqrt{\varepsilon}}{2}(\sqrt{m+\alpha}-\sqrt{m-\alpha})$$

where $\omega = 2\pi f$, f = frequency, ε = dielectric constant, m = magnetic permeability, and α = the component of the magnetic permeability tensor. *AIL*

Ref.: Nikol'skiy (1964), p. 184.

The **Faraday rotation effect** is the phenomenon of the rotation of the plane of polarization of a linearly polarized wave when it is longitudinally (parallel to the external magnetic field) propagated in a gyrotropic medium. This is explained by the difference in the phase speeds of normal waves with circular polarization when a linearly polarized wave is placed on them.

The Faraday effect occurs when radio waves with frequencies less than 3 GHz propagate in the ionosphere. It has widespread usage in ferrite microwave units with longitudinally magnetized ferrites (isolators, circulators, and phase shifters). *IAM*

Ref.: Nikol'skiy (1964), p. 186; Berkowitz (1965), p. 360; Kravtsov (1983), p. 81; Sazonov (1981), p. 264.

FEED, antenna. The antenna feed is "that portion of the antenna coupled to the terminals which functions to produce the aperture illumination." It is the means of connecting the transmitter output or receiver input to the radiating aperture. In reflector or lens antennas, including space-fed arrays, the feed is typically a horn or small end-fire antenna near the focal point, with a pattern extending over the aperture. In a constrained-feed array, the feed is a network of couplers that divide the transmitter power among the radiating elements and combine the received power from these elements. *SAL*

Ref.: IEEE (1993), p. 485; Skolnik (1980), p. 306.

Array feed networks distribute transmitter power to the array radiators and collect received power in the receiver input. There are two basic methods of feeding phased-array antennas: optical (or space) feed and constrained (or corporate) feed. There are two variants of space-fed arrays: space-fed transmission arrays and space-fed reflectarrays (see **space feed**). Constrained feeds also have two variants: series and parallel feeds (see **constrained feed**). *SAL*

A **beam-forming (feed) network** is used in a constrained-feed array to generate one or more beams, coupling each array element to the beam port(s). Networks forming multiple beams are used in monopulse tracking radars and stacked-beam surveillance radars, and in electronically scanned arrays to select beam positions without using controllable phase shifters or frequency scanning. Analog configurations include the Rotman lens (see **ARRAY, Rotman**), while discrete networks take the form of the Blass matrix (Fig. F2) and Butler matrix (Fig. F3). These networks are also called *beam-forming matrices*.

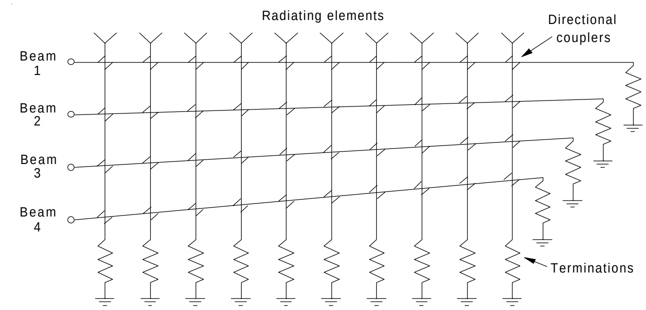


Figure F2 Blass multiple-beam-forming matrix.

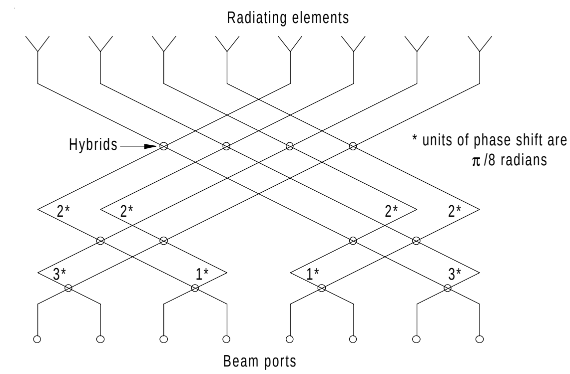


Figure F3 Butler multiple-beam-forming matrix.

In the Blass matrix, several beam-port lines are coupled to radiating-element lines through directional couplers. The illumination function is controlled by the coupling coefficients, and the direction of the beam is controlled by the relative pathlengths between couplers to adjacent element lines (the beamport lines are oriented at an angle with respect to the radiating elements to introduce the appropriate phase shifts). Adjacent beams are independent if separated in angle by more than the "standard beamwidth" (that of the array when uniformly illuminated). Wideband operation of the Blass type of matrix can be obtained using the equal-pathlength configuration of Fig. F4.

The Butler matrix is simpler, having a minimum number of couplers for a given number of elements, but the number of beam ports is equal to the number of elements, and the beams are equally spaced (in sine space) over the forward hemisphere. The weighting across the array is uniform, leading to high sidelobes that can be controlled by coupling among two or more adjacent beam ports.

Beam-forming matrices can also be implemented, for receiving only, in IF or baseband portions of the receiver (Fig. F5). In the system shown, the amplitude and phase weighting for each element is determined by a resistive network that adds baseband phasor components. In a digital beam-forming system (Fig. F6), analog-to-digital conversion is done for I and Q channels at baseband on receivers fed from each array element, and complete flexibility is provided in digital summing of these components. Arbitrary beam shapes and spacings are obtained with no concern for dissipative loss that

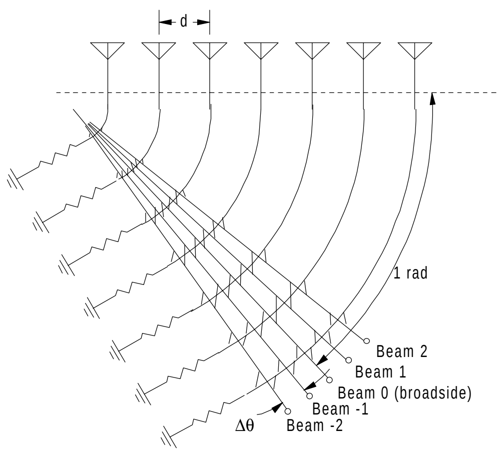


Figure F4 Time-delayed multiple-beam matrix (after Johnson, 1984, Fig. 20.46, p. 20.58).

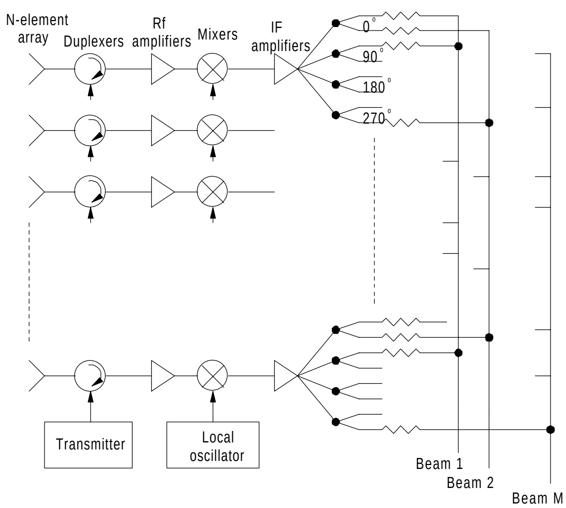


Figure F5 IF beam-forming matrix (after Barton, 1988, Fig. 4.3.11, p. 183).

affects microwave networks, and weighting tolerances that affect the analog IF beam-former. The main disadvantage is the high cost of multiple receivers and A/D converters, and the high digital throughput rate resulting from formation of *M* beams as weighed complex sums of *N* element outputs, that must be obtained at a rate at least equal to the signal bandwidth. *DKB, SAL*

Ref.: Johnson (1984), pp. 20.56–20.60; Barton (1988), pp. 178–185; Farina (1992), p. 34.

Feed blockage is the effect of shadowing of the main mirror in reflector antennas by a feed and feed-support structures. Feed blockage introduces undesirable distortions in the antenna pattern that can cause, for example, the broadening of the main beam and rise in the sidelobe level. Typical effects of reflector blockage are shown in Fig. F7. Feed and feed-support blockage can be reduced by offsetting the reflector, changing the metal support structures to dielectric ones, or other special techniques. *SAL*

Ref.: Barton, (1988), pp. 158–160; Skolnik (1990), p. 6.35.

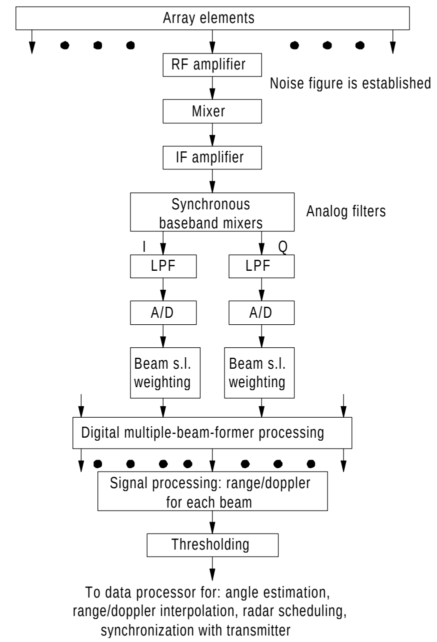


Figure F6 Digital beam-forming system (after Barton, 1988, Fig. 4.3.12, p. 184).

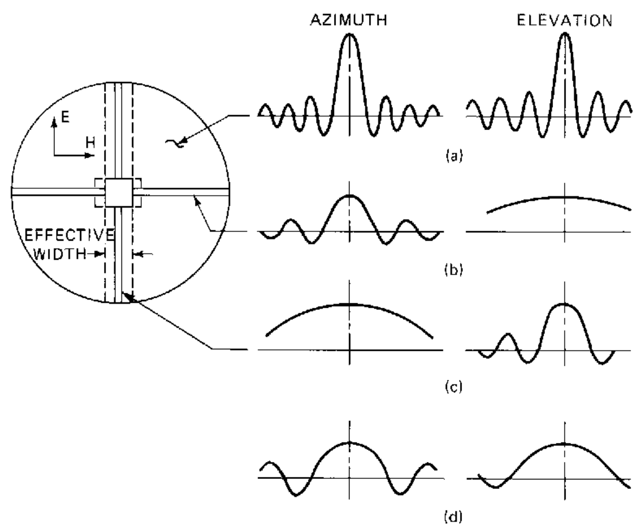


Figure F7 The typical effects of reflector blockage (from Skolnik, 1990, p. 6.35, reprinted by permission of McGraw-Hill).

A **constrained feed** is a feed concept in which the signal is confined to waveguide, stripline, or coaxial structures with discrete branches connecting to the elements. Constrained feeds are classified as either series or parallel networks.

A parallel feed is a type of constrained feed network in which paths to the individual elements proceed in parallel from couplers or power splitters. The most common technique for realizing a parallel feed structure is the matched corporate feed structure (Fig. F8), in which matched hybrids or used as power-dividing elements.

A series feed is a type of constrained feed in which radiating elements are fed from couplers or power dividers arranged in series. The basic types of this configuration are the end-fed serial feed system, in which there is a transmis-

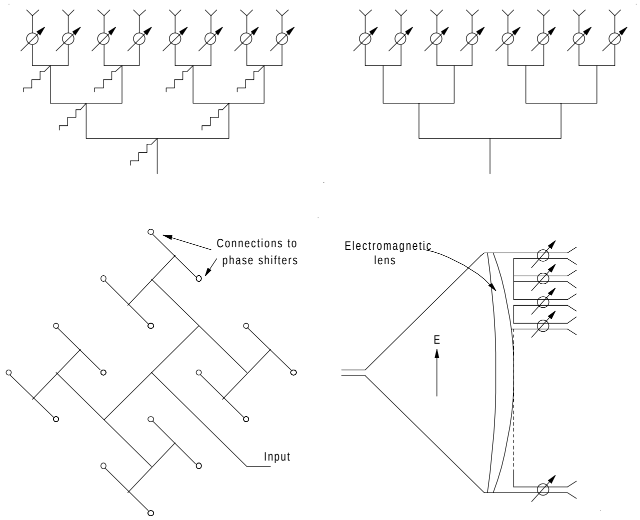


Figure F8 Parallel feed networks (after Skolnik, 1990, p. 7.61).

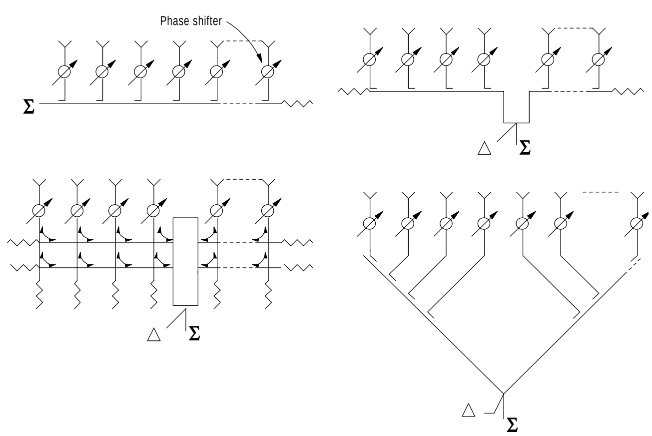


Figure F9 Series feed networks (after Skolnik, 1990, p. 7.60).

sion line from which energy is tapped through loosely coupled junctions to feed the radiating elements (Fig. F9a), and the center-fed serial feed system, which can be used to form monopulse sum-and-difference beams (Fig. F9b).

When a mixer and power amplifier are used at each element to produce lower-power-level beam steering, the feed configuration sometimes is called a *mixer-matrix feed*. A feed often used as a series feed for economy and ease of fabrication is a slotted waveguide.

The main drawback of the constrained feed is complexity, especially when a network for multibeam antenna has to be implemented. *SAL*

Ref.: Barton (1988), p. 175; Skolnik (1980), p. 307, (1990), p. 7.58; Johnson (1984), pp. 20.51–20.54.

A **dipole feed** is a dipole placed at the focal point of a lens or parabolic reflector to illuminate the aperture. The dipole feed will normally use a small reflector to restrict its radiation to the direction of the aperture.

Ref.: Sazonov (1988), p. 380; Silver (1949), p. 448.

A **dual-ladder [Lopez] feed** is a parallel, constrained feed for monopulse phased arrays using the primary line with couplers to produce the sum illumination function and a secondary line that is coupled only to the difference channel (Fig. F10). The discontinuity introduced by the center hybrid of the primary line in the difference illumination is canceled by an opposite discontinuity in the secondary line, so the desired smooth functions can be obtained on both channels. To hold the electrical tolerances over the operating band of the radar, each element has to employ two precision couplers, which makes it relatively costly to obtain low sidelobes as compared with single-line feed. *SAL*

Ref.: Barton (1988), p. 203.

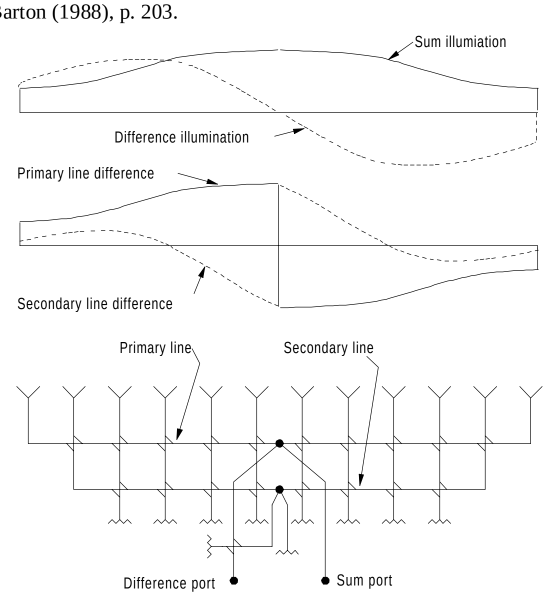


Figure F10 A dual-ladder network for separate optimization of sum- and difference-excitation of a linear array (after Barton, 1988, p. 206).

A **dummy-snake feed** is a technique for implementing a frequency-scanned monopulse antenna (Fig. F11). It has two halves of identical design to scan synchronously. Disadvantages that make it difficult to construct are difficulty in matching the two halves and inability to obtain independent control of sum and difference patterns. *SAL*

Ref.: Johnson (1984), p. 19.12.

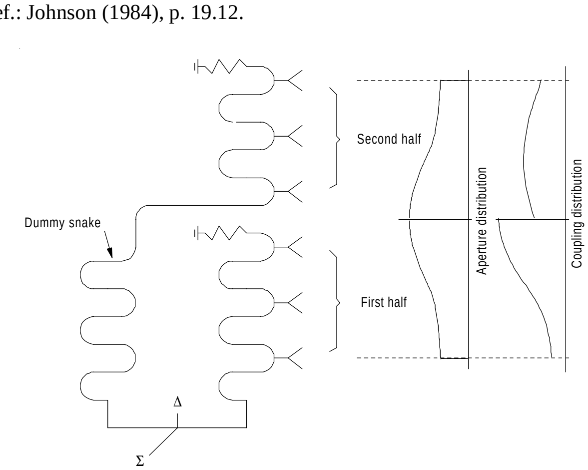


Figure F11 Dummy-snake frequency-scan monopulse antenna (after Johnson, 1984, Fig. 19.9, p. 19.12).

A **four-horn feed** is a classical monopulse feed assembly and network used in early monopulse radars. This feed is a cluster of four horn feeds displaced about the focal point of a reflector or lens antenna. Each horn can be associated with a pencil beam in space, squinted about the axis of the antenna (Fig. F12). A basic problem in such antennas is the compromise between sidelobe levels and efficiency in the sum and difference patterns. More advanced feed systems use a twelve-horn feed cluster or a multimode feed. *SAL*

Ref.: Barton (1988), p. 198.

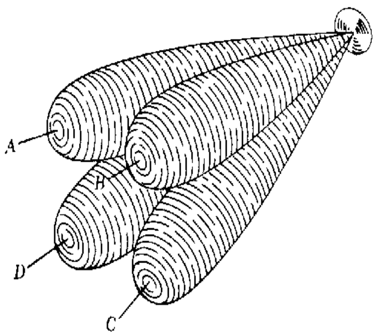


Figure F12 Beams formed with four-horn feed (from Rhodes, 1980, p. 70).

A **horn feed** is a horn antenna used to illuminate a reflector or lens.

Ref.: Silver (1949), p. 448; Sazonov (1988), p. 382.

A **linear feed** is a linear array of dipoles, horns, or waveguide slots used to feed a parabolic cylinder reflector antenna.

Ref.: Silver (1949), pp. 13, 457; Sazonov (1988) p. 383.

Lopez feed (see **dual-ladder feed**).

A **monopulse feed** is a feed network for a monopulse antenna. Typically, in a reflector or lens antenna system, horn feeds are used with a network, which forms sum and difference channels. The main configurations in this case are two-horn (dual-mode and triple-mode), four-horn (simple and triple-mode), and twelve-horn feeds. The shapes and typical performance of monopulse feeds are given in Table F1.

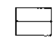
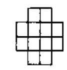
In phased-array antennas using constrained feeds, the monopulse feed network consists of a set of couplers that produce illumination functions having even and odd symmetry about the center of the aperture; the former corresponding to the sum channel and the latter the difference channels. *SAL*

Ref.: Barton (1988), pp. 198-206; Skolnik (1990), p. 6.21.

A **multiple-beam feed** is used in a multiple-beam antenna to generate multiple simultaneous beams covering, for example, the elevation sector of a 3D surveillance radar. The main types of multiple-beam feeds are monopulse feeds and multiple-beam-forming networks. (See **Blass feed**, **Butler feed**, **monopulse feed**.) *SAL*

Ref.: Johnson (1984), p. 20.56.

Table F1
Monopulse Horn Feed Performance

Type of Horn	η_a	*H-plane* $K_r\sqrt{\eta_y}$	k_m	*E-plane* $K_r\sqrt{\eta_x}$	k_m	G_{sr} (db)	G_{se} (db)	*Feed Shape*
Simple four-horn	0.58	0.52	1.2	0.48	1.2	19	10	
Two-horn dual-mode	0.75	0.68	1.6	0.55	1.2	19	10	
Two-horn triple-mode	0.75	0.81	1.6	0.55	1.2	19	10	
Twelve-horn	0.56	0.71	1.7	0.67	1.6	19	19	
Four-horn triple-mode	0.75	0.81	1.6	0.75	1.6	19	19	

(from Barton and Ward, 1969, p. 31).

A **nutating feed** is an offset feed that can rotate about the axis of the dish in a reflector antenna and maintain the plane of polarization fixed as it rotates. It is used in the technique of conical scanning. *SAL*

Ref.: IEEE (1990), p. 22; Skolnik (1980), p. 155.

optical feed (see **space feed**).

An **optical-fiber beam-forming circuit** is a circuit that uses fiber-optic delay lines as the phase-shifting components. Radio signals from the output of each of the radiating elements of the antenna array modulate the intensity of the sources of optical radiation (lasers), and then the radiation is sent to the fiber-optic delay line. The length of the fibers forming the delay line, and thus the delay in each channel, is selected so as to compensate for the output spatial delays of excitation of the radiating elements by the plane wave. Optical signals from the output of each fiber are delivered in parallel, without interaction, to the phase-sensitive area of the photodetector, adding the powers of the channels.

Fiber-optic diagram-forming circuits are marked by their low size and mass parameters and energy consumption, and they have a high wideband capacity. *IAM*

Ref.: Bratchikov, *Radioelektronika* ***32***, no. 2, 1989, p. 23; Zmuda (1994), p. 299.

parallel feed (see **constrained feed**).

A **parallel-plate feed** is a reactive feed system using a microwave structure to provide power division. Such feeds are basically used with a linear array and have to be stacked to feed a planar array. The parallel-plate horn and pillbox feed are examples. *SAL*

Ref.: Skolnik (1980), p. 309.

A **pillbox feed** is a pillbox antenna used as a feed; for example, to illuminate a linear array (Fig. F13). If the array of pickup elements is matched to the pillbox then no reflections occur, and if the back surface of the pillbox is made spherical instead of parabolic, the feed can provide many multiple-simultaneous beams with the use of multiple feed horns. *SAL*

Ref.: Johnson (1984), p. 20.56.

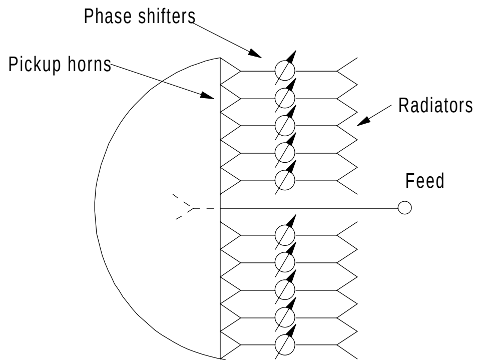


Figure F13 Pillbox feed (after Johnson, 1984, p. 20.55).

A **rotating feed** is an offset rear feed rotated about the axis of a reflector antenna to produce a conically scanned beam and causing the polarization to rotate (unlike a nutating feed). *SAL*
Ref.: Skolnik (1980), p. 155.

Rotman lens feed (see **ARRAY, Rotman**).

A **serpentine feed** is a folded electrical feed formed to bring the elements closer together physically while keeping them apart electrically (Fig. F14). Sometimes this feed is called a *sinuous feed*. Such type of a feed is used in frequency scanning radars. *SAL*
Ref.: Billetter (1989), p. 9.

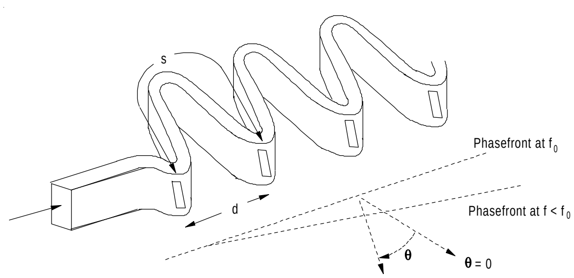


Figure F14 Serpentine feed for frequency-scanning radar (from Barton, 1988, Fig. 4.3.2, p. 169).

A **sinuous feed** is a structure used to illuminate frequency-scanned arrays. At S-band and higher frequencies, waveguides are used, and at lower frequencies, coaxial lines are the appropriate choice. A typical configuration for a sinuous feed for a linear array is shown in Fig. F14. For monopulse antennas, a tandem sinuous feed can be used, generating two beams in a Blass matrix with two parallel sinuous feeds (Fig. F15). *SAL*
Ref.: Johnson (1984), pp. 19.11, 19.16.

slot feed (see **RADIATOR**).

The **space (optical) feed** is used in phased arrays when the illumination of the array is accomplished by optically distributing the source signal through space, illuminating an array of pickup elements that are connected, through element phase shifters, to the radiating elements. These feeds are divided into the transmission type, in which the array elements are the

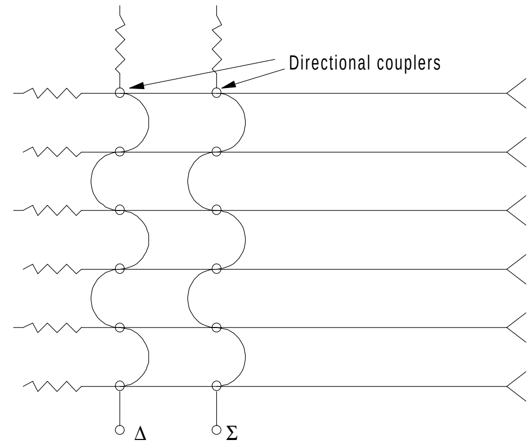


Figure F15 Tandem-fed frequency-scan monopulse antenna (after Johnson, 1984, Fig. 19.7, p. 19.11).

radiating elements of a feed-through lens (Fig. F16a) and reflection type, in which the array aperture is used as a reflector (Fig. F16b).

The main advantage of an optical feed over a constrained feed is its simplicity and correspondingly lower cost, while the main disadvantages are the volume of space required to accommodate the feed system and limited ability to control the amplitude taper. A typical space-fed lens antenna is shown in Fig. F17.

The advantages of both types of space feeds include relative simplicity, a convenient control of the amplitude distribution across the aperture by selecting the illumination pattern of the primary feed, and the ability to use complicated monopulse feedhorns. A common disadvantage of optical feeds is their increased depth as compared with constrained feeds. Furthermore, in optical feeds a part of the illumination power in not captured by the primary array; this reduces the antenna efficiency and may produce undesirable sidelobes. From a construction point of view, the reflectarray has an advantage as compared with the transmission type: easy access to all phase shifters. In addition, the phase shifters for the reflectarray are simpler than for transmission arrays. The advantages of the space-fed transmission array include the

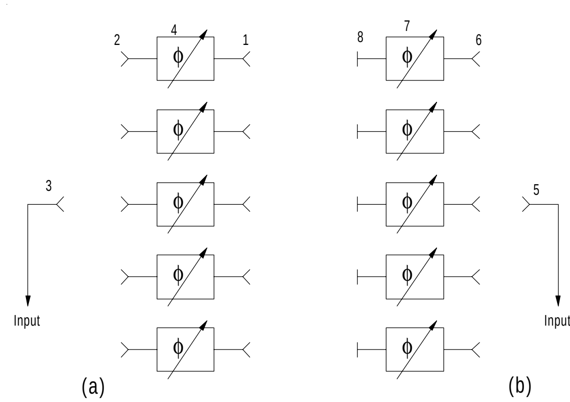


Figure F16 Space-fed arrays: (a) transmission feeds; (b) reflection feeds.

Figure F17 The SA-12 *Grill Pan* fire control radar, an example of a space-fed lens array. In the background is the *Billboard* three-dimensional surveillance radar.

absence of both aperture blockage by the feed and of reflection from the array back to the feed. *SAL*

Ref.: Barton (1988), pp. 175–179; Johnson (1984), p. 20.49.

Feed spillover is the effect of the energy losses in reflector antennas due to illumination that extends beyond the edge of the dish, resulting primarily in formation of spillover lobes (Fig. F18). Typically, spillover reduction is accomplished by shaping the feed pattern to cut off sharply at the edge of the dish and by maintaining low sidelobes in the feed pattern. *SAL*

Ref.: Skolnik (1990), p. 6.37.

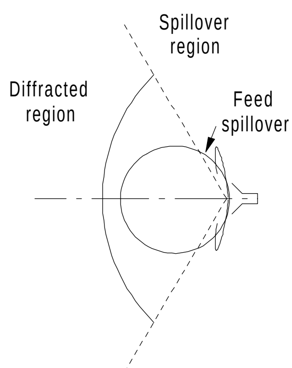


Figure F18 Spillover lobes (after Skolnik, 1990, p. 6.39).

A **twelve-horn feed** is a monopulse horn feed using a twelve-horn cluster (Fig. F19) that permits separate optimization of sum and difference illuminations. *SAL*

Ref.: Barton (1988), p. 199.

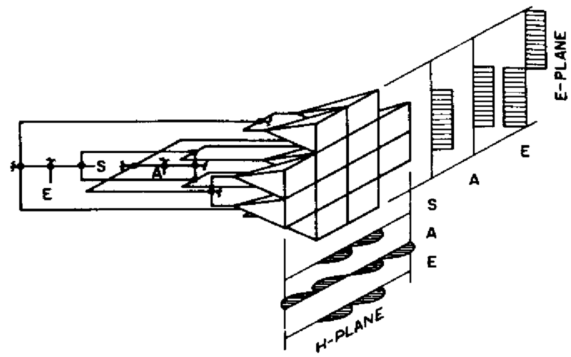


Figure F19 Twelve-horn feed (from Barton, 1988, Fig. 4.5.5, p. 201).

waveguide feed (see **linear feed**, **sinuous feed**).

FEEDBACK is "the returning of a fraction of the output to the input." A fundamental concept in automatic control systems, feedback derives its name from a signal "fed back" from a controlled process to measure its response to an input command. Feedback is based on comparing the actual response of a controlled output to the desired response. Any difference between the two values is used to adjust the amplitude of the input control signal. When the feedback signal results in a decrease in the input signal, it is termed *negative feedback*; when the input signal is increased (amplified) through feedback, it is known as *positive feedback*. Systems using feedback are referred to as *closed-loop systems*.

The feedback concept is utilized in many areas of radar systems engineering, among which are the electromechanical servomechanisms used for antenna control, and electronic systems such as radar frequency control, constant false-alarm rate (CFAR) systems, and automatic gain control (AGC). *PCH, SAL*

Ref.: IEEE (1993), p. 485; Bode (1945), p. 31; Popov (1980), p. 255.

FEEDER [FEED LINE]. A feeder is a conducting transmission line for radio-frequency electromagnetic energy. Depending on the frequency range, either open symmetrical lines of parallel wires or RF cables are used. The cables are symmetrical unshielded or shielded, or coaxial cables (see **transmission line**, **wire** and **coaxial**). Feeders are used to channel energy from the transmitter to the antenna and from the antenna to the receiver, and also to join other high-frequency assemblies of radar up to the long-wave part of the centimeter-wave range. Ultrawideband video pulses are also transmitted over high-frequency feeders (usually coaxial). *IAM*

Ref.: Popov (1980), p. 455.

FENCE. A fence is (1) a "line of networks of early warning radars; (2) The locus of the positions of a surveillance radar beam that describes the search area covered by space-based radar; (3) Concentric conducting barrier erected around a ground-based radar to serve as an artificial horizon and suppress ground clutter" (the latter is more often termed a *clutter fence*).

Ref.: IEEE (1990), p. 16; Skolnik (1970), p. 32.8.

A **clutter fence** is a "concentric conducting barrier erected around a ground-based radar to serve as an artificial horizon and suppress ground clutter." The main task of such a fence is to reduce the reflections from large clutter sources; for example, nearby mountains, which cannot be eliminated by radar clutter-suppression techniques such as MTI. Such a fence typically is made from an electromagnetically opaque material preventing the radar from viewing the clutter directly but also limiting the ability of the radar to detect low-altitude targets. The typical two-way isolation provided by such a fence is about 40 dB. *SAL*

Ref.: IEEE (1990), p. 16; Barton (1975), p. 363; Skolnik (1980), p. 497.

FERRITE. A ferrite is a ceramic material prepared from a mixture of powdered iron oxide and oxides of other metals such as copper or nickel. The powder is mixed with a binder and annealed, resulting in a hard material. Ferrites are characterized by their permeability and loss. The main types of ferrites are spinel, magnetoplambite, and garnet. They are widely used in RF components for radar, especially in circulators and phase shifters. There main limitations are fragility, difficulties in mechanical processing because of hardness, temperature dependence, and aging. *SAL*

Ref.: Fink (1982), pp. 6.77, 9.20; Zherebtsov (1989), p. 170.

FIELD, electromagnetic. The electromagnetic field consists of electric and magnetic fields completely described by four vectors:

$\vec{E}$ = electric field strength (intensity).
$\vec{D}$ = electric field induction.
$\vec{H}$ = magnetic field strength (intensity).
$\vec{B}$ = magnetic field induction.

The mutual relationships among these vectors are determined by the properties of the medium in which the field exists, and for a vacuum the following apply:

$$\vec{D} = \varepsilon_0 \cdot \vec{E}\ ;\ \vec{B} = \mu_0 \cdot \vec{H}$$

where ε_0 is the permittivity and μ_0 the permeability of free space. For a description of the resulting propagating wave, see **WAVE**. *SAL*

Ref.: Fink (1982), p. 1.38; Nikol'skiy (1964).

The **antenna field** is the electromagnetic field produced by an antenna during operation. This is divided into three regions: (1) the reactive near-field region, (2) the radiating near-field (or Fresnel) region, and (3) the far-field (or Fraunhofer) region. The boundaries are not exactly defined, but the commonly accepted boundary between the first and second regions is $\lambda/2\pi$, and between the second and third is $2D^2/\lambda$, where λ is wavelength and D is the largest dimension across the aperture. Radar targets lie in the far-field region, except for very short-range applications.

In the far field, the electrical field strength is given by

$$\vec{E}(\phi, \theta, R_0) = I_0 \cdot \frac{e^{-jkR_0}}{R_0} \cdot f(\phi, \theta) \cdot e^{j\omega t}$$

where ϕ, θ, R_0 are the spherical coordinates of point P with respect to the center of the antenna and I_0 is the complex amplitude of the current a radiating element at point O (Fig. 20). This equation shows the structure of the antenna field, which consists of four multiplicative factors:

(1) I_0 describes the intensity of a field at the antenna aperture.

(2) e^{-jkR_0}/R_0, where $k = 2\pi/\lambda$ is the wave number, describes the dependence of the field on the distance from the antenna.

(3) $f(\phi,\theta)$ describes the dependence on the angular coordinates (spatial distribution) and is virtually the antenna pattern.

(4) $e^{j\omega t}$, where ω is the angular frequency of the radiated wave, describes the temporal structure of the field (see also **ANTENNA radiation regions**). *SAL*

Ref.: Johnson (1984), p. 1.9; Fradin (1977), p. 19.

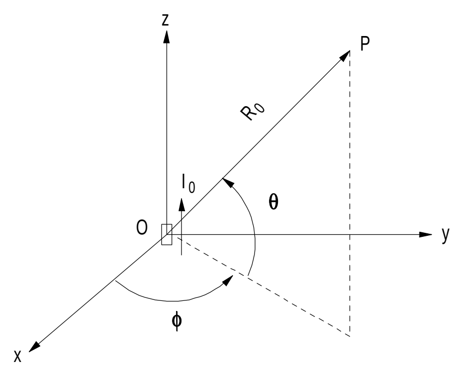


Figure F20 Geometry of the antenna field.

Electromagnetic field strength [intensity] is the field strength of the electric component of an electromagnetic field at any point, defined as the ratio of the force that a small test-body charge would experience at that point, to the force as the charge on the body approaches zero. The electric field strength is expressed by the vector equation

$$\vec{E} = \lim_{\Delta Q \to 0} \frac{\Delta \vec{F}}{dQ} = \frac{d\vec{F}}{dQ}$$

where $\overrightarrow{\Delta F}$ is an increment of force exerted on an increment of charge ΔQ at a point in the electric field. Electric field strength is measured in newtons/coulomb (SI units), or volts/meter.

The impedance of free space is 120π, and the field intensity produced by a transmitter with power P_t at range R is

$$E = \sqrt{120\pi \cdot I} = \frac{\sqrt{30 P_t}}{R} \quad \text{[V/m]}$$

where $I = P_t/4\pi R^2$ is the power density. *PCH, SAL*

Ref.: Silver (1951); Fink (1982), p. 18.56.

The **magnetic field strength** measures the magnetic effects caused by a flowing electrical current. It is a field vector $\vec{H}$ measuring the amount of force $\vec{F}$ produced per unit of magnetic flux set up by a conductor consisting of N closely spaced turns each carrying the same current I through a conductor of closed-path length l, in a medium of permeability μ. Magnetic field strength can be expressed by the vector equation

$$\vec{H} = \frac{\vec{B}}{\mu}$$

where $\vec{B}$ is the magnetic flux density. The magnetic field strength is measured in ampere-turns/meter. *PCH*

Ref.: Fink (1982), p. 1-22.

FILTER, FILTERING. In radar applications the terms filter and filtering are used in the following three senses:

(1) One type of filter is a frequency-selective circuit that separates the radar signal from its background based on the signal frequency spectrum. Filters of this type are referred to as frequency(-selective) filters, classified as low-pass, high-pass, bandpass, or band-stop filters. Depending on the center frequency, tuned filters are divided into radio-frequency (RF, or microwave) filters, intermediate-frequency (IF) filters, or video filters, and, depending on bandwidth, into narrowband or wideband filters. These are used primarily in the radar receiver.

(2) A second type of filter is a two-port device providing a required output based on its input signal. In this sense it is a major component of the radar signal processor. An optimum filter (which is termed a matched filter for a background of white noise) gives the highest signal-to-noise ratio at its output, for given input signal energy. Signal processing filters are divided into analog, discrete, and digital filters (Fig. F21), based on the representation of input and output signals, and into linear and nonlinear filters based on the equations linking the input and output signals.

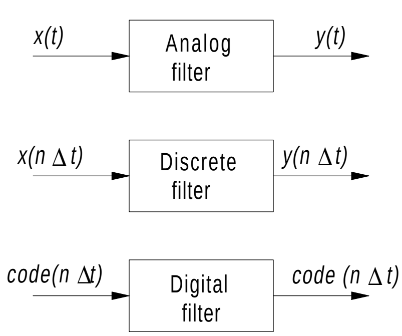


Figure F21 Types of signal-processing filters.

The basic equation describing the operation of a modern radar filter is the discrete filter equation (see **linear filter**). Digital filters have considerable advantage over analog filters, and are most commonly used in modern radar signal processors. They are implemented as tapped-delay-line filters and are classified as recursive (infinite impulse response, or IIR) or nonrecursive (finite impulse response, FIR, or transversal) filters. The latter type is most widely used.

(3) This type of filter is an algorithm processing a set of individual estimates (e.g., radar measurements obtained over successive time periods) with the objective of obtaining a new estimate for time t_n. If t_n is equal to or less than the current time t_j when the last measurement is made, the operation is termed smoothing (or interpolation); if $t_n > t_j$, it is termed prediction or extrapolation. The two basic types of filter performing these operations are the α-β(-γ) filter and the Kalman filter.

In Russian literature the operations performed by the filters described in (2) are termed "primary signal processing," and those in (3) are "secondary signal processing". *SAL*

An α–β(–γ) **filter** is a recursive filter used in track-while-scan radar and other sampled-data systems to smooth and extrapolate past data. At the time of the nth data point in a sequence, an estimate of the smoothed position x_{ns} in coordinate x is made using

$$x_{ns} = x_{pn} + \alpha(x_n - x_{pn})$$

where x_{pn} is the value predicted from past data, x_n is the measured value for the nth point, and α is the position smoothing parameter. At the same time, an estimate of smoothed velocity $\dot{x}_{ns}$ is made, using

$$\dot{x}_{ns} = \dot{x}_{n-1} + \frac{\beta}{T}(x_n - x_{pn})$$

where β is the velocity-smoothing parameter, and T is the data interval. It is apparent that $\alpha = \beta = 0$ rejects new data, producing a track that follows previous predictions, while $\alpha = \beta = 1$ rejects past data, producing a track following each new point as it is measured.

Where the quality of the data warrants, a third smoothing parameter γ may be included in an estimate of target acceleration:.

$$\ddot{x}_{ns} = \ddot{x}_{n-1} + \frac{\gamma}{T^2}(x_n - x_{pn})$$

In this case the filter is called an α-β-γ filter. As opposed to the Kalman filter, this filter uses fixed parameters as filter gains, considerably simplifying the implementation at the expense of greater error in tracking. *DKB*

Ref.: Skolnik (1980), pp. 184–186; Blackman (1986), pp. 21–25.

An **acousto-optical filter** is a filter that uses acousto-optical modulation of a reference light beam. It contains a source of coherent monochromatic light (laser), directed with an optical beam-forming system at an acousto-optical modulator, and a photoreceiver. The electrical signal is formed at the input of the photoreceiver as a result of interference of the reference light beam with the signal formed as a result of acousto-optical interaction.

The frequency characteristic of an acousto-optical filter in scale repeats the amplitude of the reference light beam in the lane of the input aperture of the diaphragm of the photoreceiver. Thus if the reference flux is a plane light wave within the bounds of the aperture, the filter realizes a rectangular frequency characteristic.

An acousto-optical filter can delay a signal, perform band filtering, calculate the convolution of a signal with some reference, perform matched filtering, and so forth.

The basic filter characteristics range from 10 MHz to 2 to 3 GHz with a band up to 30%; number of resolved frequency points depends on the time aperture of the acousto-optical modulator, and is on the order of 10^3; dynamic range exceeds 50 dB.

Acousto-optical filters are more difficult to produce than other analog filters (CCD filters, SAW filters), so their use becomes advisable for parallel processing of large masses of data in real time. *IAM*

Ref.: Kulikov (1989), p. 97; Zmuda (1994), p. 409.

An **active filter** is one that contains active elements (usually an amplifier) with *R-C* networks in its input and feedback paths. An advantage of the active filter is that use of inductors may generally be avoided. The active elements are bipolar and field-effect transistors, PIN diodes linked with the dielectric cavity of a stripline, in the centimeter band or with a filter based on waveguide-slot lines in the millimeter band. Active filers are used as bandpass or bandstop filters. The use of wideband active elements (field-effect transistors based on gallium arsenide) in a filter with varactor frequency tuning makes it possible to obtain vanishingly small losses, along with high tuning speed (less than 100 ns). *DKB, IAM*

Ref.: Fink (1982), p. 12.34; Jordan (1985), Ch. 10.

An **adaptive filter** is one that changes its parameters on the basis of current information to obtain a performance optimum in a certain sense under initial uncertainty and changing operating conditions. Input signal *S(n)* is filtered or weighted in an adaptive filter (1) to obtain output signal $y(n)$, which then is compared with useful, standard, or training signal $y(n)$ to find error signal $\varepsilon(n)$ (see Fig. F22). The error signal is used with adaptation algorithm (2) for adjustment of the weighted factors of the filter parameters (usually using an integrative method) for gradual minimization of the error (or of another cost function).

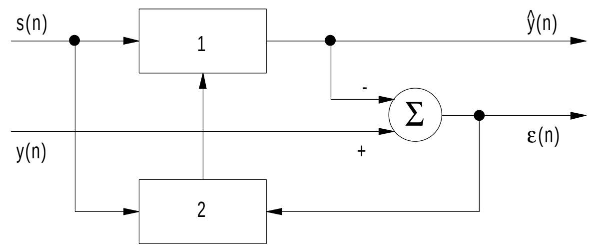


Figure F22 Adaptive filter (after Cowan, 1985, Fig. 1.1, p. 11).

Adaptive filters require a minimum volume of initial information concerning the arriving signal. Gradient methods (see **ALGORITHM, Widrow**, and **CANCELER, Howells-Applebaum**) usually are used to control filter weight parameters. IIR and FIR filters with realization in both the time and the frequency domains are used as adaptive filters.

Time to establish the adaptation mode, stability, and residual error are important adaptive filter characteristics. Adaptive filters are used widely in radar mainly as adaptive filter-compensators and filter-extrapolators. *IAM*

Ref.: Cowan (1985); Widrow (1985), p. 18.

An **analog filter** is one using analog circuit technology. Depending on the required filter response and its operating mode, an analog filter may be the simplest electronic circuit or a complex device. Analog filters are divided into the following types: filters using oscillating or aperiodic circuits, using phase and dispersive circuits; and filters using delay lines. Each type usually is used for filtering of a specific class of signals in matched processing and to obtain a different type of frequency responses (see Table F2).The types of analog filters used in modern signal processing are acousto-optical, CCD, and SAW’ filters. *IAM*

antialiasing filter (see **ALIASING**).

An **autoregressive filter** is an all-pole filter in which parameters of an observed *N*-point data sequence are modeled as those of white noise passed through a filter with $N - 1$ poles.

Ref.: Nitzberg (1992), p. 290.

Table F2
Analog Filter Types

Filter type	Type of filter implementation	Area of use	Time-bandwidth product
Based on oscillating and aperiodic circuits	LC, RC filters, quartz, piezoceramic, electromechanical filters	Filtering of coherent and noncoherent narrowband signals	$B \approx 1$
Based on phase and dispersive circuits	Cascaded S-filters, electro-acoustic dispersive circuits, connected in series or parallel	Nonlinear filtering	$B \approx 10$
Based on delay lines	Tapped delay lines	Signal filtering with large *B*	$B \approx 100$

A **bandpass filter** passes a prescribed frequency band, rejecting signal components outside this band. The frequency response of an ideal filter is unity in the passband and zero in the adjacent stopbands. In real filters, the response does not change discontinuously, but rather the response drops gradually in some transition band (Fig. F23). In the passband, a filter is characterized by the level of attenuation at the top boundary of this band. The average gain of a filter in the pass band is less than unity (as a passive element it introduces loss). Often the deviations in amplitude-frequency characteristic from ideal have the form of ripples whose amplitude must be specified. Bandpass filters are widely used in radar receivers as RF, IF, and video filters. *IAM*

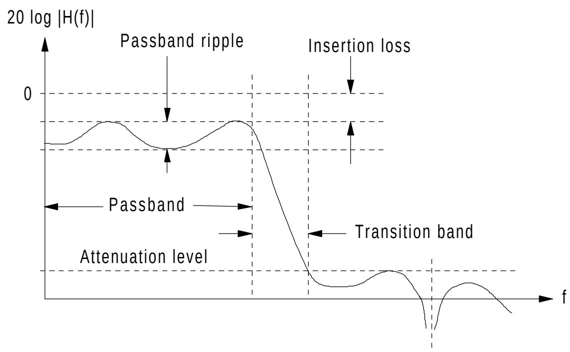


Figure F23 Deviations of characteristic of real filter from ideal characteristics (from Siebert, 1986).

Ref.: Siebert (1986)– (1988), p. 176 (in Russian); Kaganov (1981), p. 86; ITT (1975), pp. 10–11; Fink (1982), p. 12.32.

A **bandstop filter** is one with an amplitude-frequency response that has gaps in specified regions. It is used in signal filtering to suppress the most intense spectral components of interference. The frequency response of a bandstop filter is inverse to the frequency spectrum of the interference. Bandstop filters are used widely in moving target indicators (MTI), which include bandstop comb filters to form the gaps with spacings equal to the pulse repetition frequency of the pulse train. A delay-line canceler is the simplest bandstop filter. The bandstop filters can be on analog technology, but digital filters are preferred in modern radars. Adaptive digital bandstop filters fall in the filter-extrapolator class. *IAM*

Ref.: Finkel'shteyn (1983) pp. 258, 297; Sloka (1970) p. 164. Fink (1982), p. 12.33.

Filter bandwidth is a measure of the width of the frequency response, usually specified as B_3 at the half-power level. In some cases, an effective noise bandwidth B_n is specified:

$$B_n = \frac{1}{|H(f_0)|^2}\int_{-\infty}^{\infty} |H(f))| \, df$$

The power of white noise of density N_0 passing through the filter is then $N = N_0 B_n$. The ratio of noise bandwidth to half-power bandwidth for different types of filter is shown in Table F3. *DKB*

Ref.: Lawson (1950), p. 177.

Table F3
Noise and Half-Power Bandwidths

Filter type	No. of stages	B_n/B_3
Rectangular	any	1.00
Single-tuned	1	1.57
Single-tuned	2	1.22
Single-tuned	3	1.16
Single-tuned	4	1.14
Double-tuned	1	1.11
Double-tuned	2	1.04
Triple-tuned	3	1.05
Gaussian	–	1.06
Cosine^n, n = 1 to 4	–	1.05
Taylor, 30 - 50 dB sidelobes	–	1.05

Bessel filter (see **frequency-selective filter**).

Butterworth filter (see **frequency-selective filter**).

A **filter-canceler** is a filter designed for cancellation of interference on a useful signal. A filter-canceler may have constant parameters (delay-line canceler in an MTI system) or be adaptive. An adaptive filter-canceler subtracts interference from its mixture with the useful signal based on constant nulling of an error signal ε (see Fig. F24).

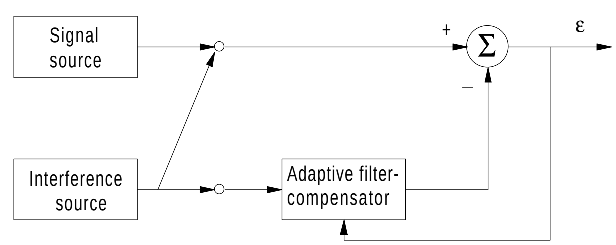


Figure F24 Filter-canceler block diagram (after Gol'denberg, 1985, Fig. 6.3, p. 164)

Adaptive filter-cancelers are used in MTI radars for passband adjustment, shaping of precise tracking zeros beyond interference frequency and phase, and in adaptive arrays for formation of nulls in the antenna response in the directions of interference action (see **ALGORITHM, Widrow** and **CANCELER, Howells-Applebaum**). *IAM*

Ref.: Gol'denberg (1985), p. 164; Nitzberg (1992), Ch. 4.

Filter characteristics. In general form an arbitrary filter can be represented as a two-port device (Fig. F25) with an input signal $x(t)$ and output signal $y(t)$, having spectra $S_x(f)$ and $S_y(f)$. The fundamental characteristics describing its operation in the time and frequency domains are the filter impulse response $h(t)$ and the filter transfer function $H(j\omega)$. The

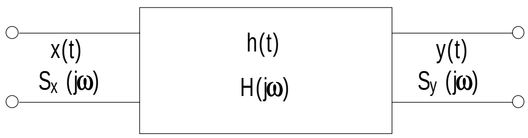


Figure F25 Generic two-port filter.

impulse response is the time response to an impulse input (δ-function). The transfer function is a complex quantity:

$$H(j\omega) = H(\omega) \cdot e^{j\phi(\omega)}$$

where $H(\omega)$ and $\phi(\omega)$ are called the amplitude-frequency response and the phase-frequency response, respectively. The complex transfer function $H(j\omega)$ is often termed the frequency response. The functions $h(t)$ and $H(j\omega)$ are a Fourier transform pair, and if they are known, the output signal and its spectrum can be found as

$$y(t) = x(t) \otimes h(t)$$

$$S_y(j\omega) = S_x(j\omega) \cdot H(j\omega)$$

where $\otimes$ is the notation for the convolution operation. For deterministic signals, $S(j\omega)$ is the signal spectrum, and for random signals, it is a signal power spectrum.

In the case of a digital filter, the transfer function $H_d(j\omega)$ becomes a periodic function, because the signal sampling at time intervals Δt:

$$H_d(j\omega) = \sum_{n=-\infty}^{\infty} H\left[j\left(\omega - n \cdot \frac{2\pi}{\Delta t}\right)\right]$$

In terms of the z-transform the transfer function of a digital filter can be represented as

$$H(z) = \frac{A(z)}{B(z)}$$

The values giving $A(z) = 0$ are called filter zeroes, and those giving $B(z) = 0$ are called filter poles. The choice of filter zeroes and poles permits design of filters with the required transfer function for different applications. *SAL*

Ref.: Fink (1982), Ch. 12; Gol'denberg (1985).

A **charge-coupled-device (CCD) filter** is an analog filter having a CCD delay line as its basic element. CCD filters are used as low-pass filters, bandpass filters, and filters for matched processing of signals with a large base. The bandwidth of a bandpass filter at the –3-dB level is 1.04 to 1.08% of the carrier frequency, and the sidelobe level is –38 to –42 dB. Selection of filter weighting coefficients determines the degree of sidelobe suppression (for Hamming weighting, the sidelobe level is –42.8 dB). The advantage of CCD bandpass filters over others is the ease in changing their resonant frequency over a broad range achieved by adjusting the clock frequency.

Programmable transversal CCD filters are widely used for matched filtering. Signal processing in CCD filters occurs in the video frequency for a bandwidth from zero to half the clock frequency, which does not exceed 20 MHz. Programmable CCD transversal filters are constructed both with control relative to the phase and amplitude of signals at delay line taps and with control only relative to phase. In the first instance, great universality is achieved; while, in the second, the base of the processed signal is maximized. For an N-stage filter, maximum base magnitude equals $N/2$ (N for a single-crystal filter reaches 512). *IAM*

Ref.: Brodersen, R. W., et al. *IEEE Trans.*, ***ED-23***, no. 2, 1976, p. 143; Hague, Y. A., and Copeland, M. A., *IEEE J. Solid-State Circuits*, ***SC-12***, 1977.

Chebyshev filter (see **frequency-selective filter**).

A **coherent integration [comb] filter** is a narrowband filter at IF or baseband that is matched to a specific target doppler frequency and whose response repeats at the pulse repetition frequency. Typically this repeated (comb) response is obtained by range-gating the input signal at the input to a narrowband filter. In search and target acquisition radar modes, a filter bank covering all doppler frequencies within the pulse repetition frequency interval is usually implemented (by fast Fourier transform or similar techniques) to detect a target at any velocity. This filter bank is an essential part of the moving target detector or pulsed doppler types of signal processing. In tracking radars and homing seekers, a single filter (or velocity gate) may be used, tracking the target doppler frequency.

Coherent integrating filters are necessarily doppler sensitive, the passband of each filter being inversely proportional to the integration time. The difficulty in direct synthesis of comb filters in the frequency domain leads to the approach in which the signal is sampled in range gates (or with range strobes, for digital filters), before passing into a bandpass filter or bank of such filters. *DKB*

Ref.: Skolnik (1980), pp. 121–125; Barton (1988), pp. 253–264; Schleher (1991), pp. 10, 73–103.

The **coherent [circulating] memory filter** is an analog processing circuit in which an IF signal is recirculated n times through a delay line, with a frequency shift that is the reciprocal of the pulse length (Fig. F26). The effect is to form, within each pulse width at the output, a series of time responses corresponding to doppler filters formed during the n-pulse integration period. This is one form of coherent integration filter.

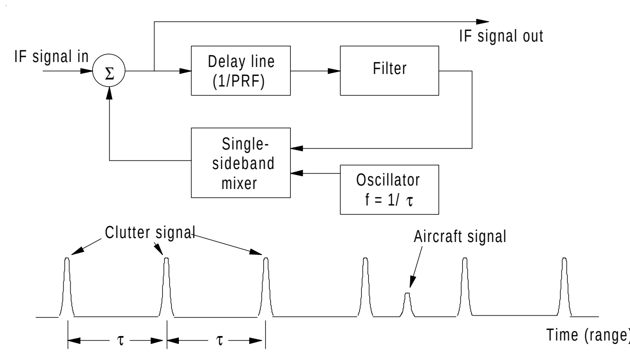


Figure F26 Coherent memory filter block diagram and output signal (after Skolnik, 1970, Figs. 62, 63, pp. 17.56–17.57).

This analog technique has been replaced in modern radar by digital filtering such as the fast Fourier transform. *DKB*

Ref.: Skolnik (1970), pp. 17.56–17.57.

A **crystal filter** is a filter whose resonant elements consist of piezoelectric crystal structures.

A **dielectric filter** is a microwave filter based on transmission lines with dielectric loading. The basis of the dielectric filter is usually dielectric cavities, which constitute dielectric bodies of a specific shape (sphere, cylinder, torus, etc.), whose relative dielectric constant is greater than that of the surroundings. The frequency characteristic of the filter is formed through reflection of waves from the boundary of the dielectric media with different constants.

They are subdivided into dielectric filters using free space as a medium, waveguide dielectric filters and strip (coaxial) dielectric filters. Filters of the first type usually consist of a number of dielectric layers with low and low dielectric constants, alternating with each other. Such filters are used for frequency and angle selection (suppression of sidelobes) of signals in the aperture of antenna systems.

Strip dielectric filters usually constitute a metal housing with microstrip line whose dielectric substrate contains dielectric cavities closely to the current conductor, usually in the form of a cylinder. Their connection to the line and with each other is through the magnetic field. The unloaded Q-factor of such filters is 1,000 or more.

Dielectric resonator-based filters are based on the phenomenon of the resonance of electromagnetic waves inside a dielectric volume. The filter is usually a cylinder with a height-to-diameter ratio of 0.3 to 0.5 on a dielectric with permeability $\varepsilon = 10$ to 40 (ceramic and so forth). The filter is distinguished for its small size (6 mm for $\varepsilon = 36$ at a frequency of 8.5 GHz), insignificant losses (inherent Q-factor 5 to 10×10^3) in the range from decimeter to centimeter and millimeter waves.

Multiresonant filters with direct links among dielectric resonators located on a flat plate at tuning distances from one another are used to obtain the necessary band and shape of the frequency response. Dielectric filters possess vitally smaller dimensions and mass than waveguide filters, but higher losses (1 to 3 dB). They are used in radar apparatus of moving objects in a frequency range of 0.1 to 100 GHz with realization of frequency characteristics both of the Chebyshev type and with a maximally flat peak. *IAM*

Ref.: Kapilevich, B. Yu., *Zarubezhnaya Radioelektronika*, no. 5, 1980, in Russian; Sazonov (1988), p. 125.

A **differentiating filter** is a high-pass filter that allows evaluation of the derivative of the input signal. The frequency response, $H(f)$ of the filter has the form $H(f) = j2\pi f$. In analog form, the filter is implemented with various transmission lines (see **differentiating circuit**), either with lumped or distributed parameters.

Digital optimum differentiating filters are characterized by the order N of the filter, the cutoff frequency F_p, which is the highest frequency at which the filter still operates, and by the maximum of the relative error, δ. Thus for $N = 16$, $\delta = 0.0136$, for $N = 32$, $\delta = 0.0062$. The cutoff frequency of the differentiating filter of an even order may reach half the sampling frequency. *IAM*

Ref.: Rabinev (1975), pp. 143, 187; Barton (1964), pp. 365–367, 422–428, (1988), p. 426.

A **diffraction filter** is an analog filter consisting of a nondispersive ultrasonic delay line of wedge shape with point transducers. In the process of transmitting the ultrasonic oscillations, diffraction occurs in the acoustic medium. One distinguishes between diffraction filters with two transducers, or with a single input-output transducer.

Pulse compression diffraction filters provide high linearity of the modulation slope during the pulse, wide frequency bands (500 MHz at a central frequency of 9,250 MHz), and high compression ratios (1,200). The high losses (60 to 75) dB are the drawback of the filters. *IAM*

Ref.: Skolnik (1970), p. 20.7; Shirman (1974), p. 137.

A **digital filter** is a nonlinear filter processing digital signals. The main source of the nonlinearity is quantization of the signal, but if the number of digits (bits) is sufficiently high the quantized signal is equivalent to a discrete signal. In this case, the basic discrete filtering equation (see **linear filter**) is applicable, resulting in the following equations that relate the output signal, $y(t)$ to the input, $x(t)$, both being digital codes at N sample points:

$$y_n = -\sum_{m=1}^{M-1} b_m \cdot y(n-j) + \sum_{k=0}^{N-1} a_k \cdot x(n-k)$$

Parameters b_m and a_k are called filter (weighting) coefficients and can be either constant or variable (dependent on n). For a nonrecursive filter, $b_m = 0$, and:

$$y_n = \sum_{k=0}^{N-1} a_k \cdot x(n-k)$$

and a_k is a set of sampled filter impulse response values, $a_k = h_k$. The parameter N is the filter length.

The first expression for y_n provides the basis for setting the digital filter configuration, as it shows that three operations are performed by the filter: (1) storing the input samples (delay), (2) multiplication (of samples by weights), and (3) summation. The complexity of filter implementation is defined by the number of memory cells (registers) required to store the input data and coefficients, and by the number of multiplication and summing operations.

Digital filtering can be realized either in the time or frequency domain. Filtering in the time domain is described by the equations above, and in radar applications a nonrecursive transversal filter is often used. In the frequency domain, the fast Fourier transform is often used in such applications as wideband pulse compression. A feature of the digital filter is that its frequency response has a periodic structure because of the sampled nature of the input signals. Digital filters are the main types used in modern signal processors. Their advantages over analog filters are in stability and repeatability of response, flexibility and reliability of operation, and convenience of implementation using integrated circuit technology. *SAL*

Ref.: Jordan (1985), pp. 28.7–28.18.

A **directional filter** is a frequency-selective directional coupler. The frequency characteristic of the directional filter in the circuit between branches corresponds to the frequency characteristic of a bandstop filter.

A waveguide directional filter with high Q-factor consists of a cylindrical three-dimensional cavity connected to two rectangular waveguides through round openings. Typical bandwidths are 0.2 to 0.4%, losses less than 0.3 dB, tuning within the limits of 10% of the band (in X-band). It is used in radars, in circuits feeding a mixer from a low-power oscillator, and also for suppression of interference at frequencies close to the carrier frequency of the radar.

Directional filters based on yttrium-iron-garnet (YIG) filters are marked by a broad adjustment and electronic tuning. Losses of such filters is 2.5 dB, minimum decoupling in the region outside the resonance band of 45 dB and maximum coupling coefficient on the order of 1.5 within the range of 8 to 12.4 GHz. *IAM*

Ref.: Skolnik (1970), p. 8.29.

A **discrete filter** is one processing discrete signals (sampled in the time domain). In radar applications a CCD filter is often used. The theory of discrete filter operation is described by the discrete filtering equation (see **linear filter**) and is the basis for the theory and implementation of digital filters. *SAL*

Ref.: Karteshev (1982), p. 11; Fink (1982),p. 12.44.

A **dispersive [dispersion] filter** is a linear device intended to obtain a delay significantly changing with frequency, or a filter with a frequency-modulated impulse response. Dispersive filters are usually used in pulse-compression radars using frequency-modulated waveforms. They are made both in analog (SAW filters, charge-coupled device filters) and in digital form. (See **digital filter.**) *IAM*

Ref.: Skolnik (1970), p. 20.6.

doppler filter (see **coherent integration filter**).

An **electromechanical filter** is one in which electrical signals are converted to acoustical waves in a solid medium, which is coupled to mechanically resonating components to provide the filtering action. The acoustical output is reconverted to electrical form for subsequent processing.

elliptic(-function) filter (see **frequency-selective filter**).

A **filter-extrapolator** is an adaptive nonrecursive filter evaluating signal y at leading moment in time n:

$$\hat{y}(n) = \sum_{k=1}^{N} h_k y(n-k)$$

$$\vec{H}^{T} = [h_1 \dots h_N],\ \vec{P}^{T} = \langle [y(n)y(n-1)\dots y(n)y(n-N)] \rangle$$

Filter coefficients h_k are selected from the condition of minimizing of the mean-square prediction error:

$$\langle e^2(n) \rangle = \langle [y(n) - \hat{y}(n)]^2 \rangle$$

and are found through solution of Wiener equation $\vec{H}_{opt} = R^{-1}\vec{P}$, where

and

$$R = \langle [y(n-1)\dots y(n-N)]^{T}[y(n-1)\dots y(n-N)] \rangle$$

is the input signal autocorrelation matrix. Figure F27 shows a diagram of the filter in z-transform terms.

Evaluation of the signal at the ($n - N$)th moment, based on preceding observations, is used along with prediction of the signal extrapolated in time. These evaluations are obtained simultaneously in a lattice filter. A filter-extrapolator is used in signal coding, noise suppression, and data smoothing. (See **Kalman filter.**) *IAM*

Ref.: Gol'denberg (1985), p. 167; Blackman (1986).

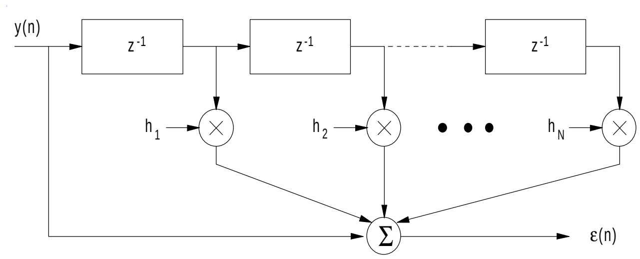


Figure F27 Filter-extrapolator diagram (after Gol'denberg, 1985, Fig. 6.5, p. 169, and Cowan, 1985, Fig. 2.2, p. 32).

feedforward filter (see **transversal filter**).

A **ferrite filter** is a small microwave filter based on ferromagnetic resonance in ferrite monocrystals. It usually is produced in the form of a polished ferroyttriferous garnet sphere located between crossed transmission lines, in an aperture in the common wall of waveguides, or at the crossing of asymmetrical striplines.

At frequencies toward resonance, a ferrite resonator acts like an isotropic magnetodielectric sample and has insignificant effect on the transmission mode, due to its small size. The transmission lines of a ferrite filter are then uncoupled. Near resonance, the ferrite link with the lines increases radically and new field components arise, which produces a strong coupling between the lines.

The inherent Q-factor of ferrite filters is 2 to 3×10^3 and the resonant frequency $f_0 \approx (3.5 \times 10^{-2}$ MHz)$\cdot H$, where H is the intensity of the magnetic biasing field (amperes/meter). The virtue of ferrite filters is their ability to change the resonant frequency over a wide band by changing the magnetic biasing field. *IAM*

Ref.: Sazonov (1988), p. 178.

finite-impulse-response (FIR) filter (see **recursive filter**).

A **frequency-selective filter** separates signals based on its frequency characteristics. The main classes are low-pass, high-pass, bandpass, and bandstop filters. Depending on the frequency of operation, these filters can be implemented either with lumped circuits (R-C filters or L-C filters) or with distributed circuits (sections of transmission lines). The latter configurations are used in microwave bands and are called microwave filters.

The main types of frequency-selective filter used in practice are the Butterworth, Chebyshev, elliptic (with the inverse hyperbolic as a special case), and the maximally linear phase-response filters. Bessel and Gaussian filters are representatives of the last type, for different parameters on n. Short descriptions and equations for frequency response of each of these types are given in Table F4, and their response curves are compared in Fig. F28. *SAL*

Ref.: Fink (1982), p. 12.32; Sazonov (1988), p. 122.

Gaussian filter (see **frequency-selective filter**)

A **harmonic filter** is used at the output of a radar transmitter to remove (frequency) harmonics that might interfere with other services or provide a source of intercepted signals for control of jammers or antiradiation missiles. A typical harmonic filter consists of a waveguide with slots that couple the several harmonics into dissipative loads located outside the walls of the transmission guide. *DKB*

Ref.: Fink (1982), p. 25-66.

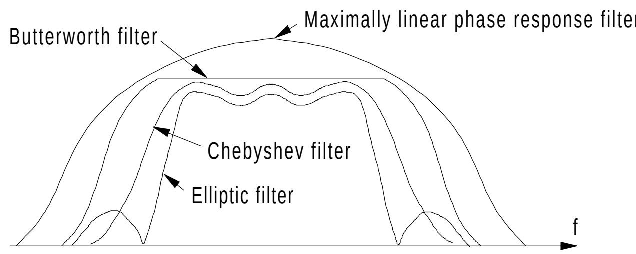


Figure F28 Comparison of response of different frequency-selective filters.

A **high-pass filter** is one passing high-frequency oscillations and blocking low-frequency oscillations. The frequency response of such a filter is given in Fig. F29.

Table F4
Frequency-Selective Filters

Type and description	Equation
Butterworth filter: "a filter whose passband frequency response has a maximally flat shape brought about by the use of Butterworth polynomials as the approximating function." The Butterworth filter is a special case of the Chebyshev filter when $V_p/V_u = 1.0$	$\left(\frac{V_p}{V}\right)^2 = 1 + \left(\frac{x}{x_3}\right)^{2n}$
Chebyshev filter: "a filter whose passband frequency response has an equal-ripple shape, brought about by the use of Chebyshev cosine polynomials as the approximating function."	$\left(\frac{V_p}{V}^2\right) = 1 + \left[\left(\frac{V_p}{V_u}\right)^2 - 1\right]\left\{\cosh\left[n\,\mathrm{acosh}\left(\frac{x}{x_v}\right)\right]\right\}^2$
Elliptic filter: a filter producing maximum rate of cutoff between bandpass and bandstop regions and giving maximum rejection in a bandstop region under given characteristics of amplitude nonuniformity in the bandpass region. Uses elliptic integrals as approximating functions.	$\left(\frac{V_p}{V}\right)^2 = 1 + \left[\left(\frac{V_p}{V_u}\right)^2 - 1\right] cd_v^2\left[n\left(\frac{K_v}{K_f}\right) cd_f^{-1}\left(\frac{x}{x_v}\right)\right]$
Inverse hyperbolic filter: an elliptic filter when $V_p/V_u = 1.0$ (or 0 dB)	$\frac{V_p}{V} = 1 + \frac{(V_p/V_h)^2 - 1}{\{\cosh[n\,\mathrm{acosh}(x_h/x)]\}^2}$
Maximally linear phase response filter: a filter retaining a linear phase characteristic in the passband and introducing minimal phase distortion. When $n = \infty$, the filter assumes Gaussian shape (Gaussian filter).	$\frac{V_p}{V} = \frac{n!}{(2n)!}\sum_{r=0}^{n}\frac{2^r}{r!}\times\frac{(2n-r)!}{(n-r)!}\left[j\left(\frac{x}{x_\beta}\right)^r\right]$
Note: V = output voltage at point x; V_p = peak output voltage in passband; V_v = valley output voltage in passband; n = number of poles (if a ladder network is used, n = number of arms); x_v and x_3 are values of x at a point on the skirt where the attenuation equals the valley attenuation or 3 dB below v_p, respectively; $cd = (cn/dn)$ = ratio of two elliptic functions, cn, dn; K_v, K_f = complete elliptic integrals of the first kind. The parameter x takes the form: (a) for a low-pass filter, $x = \omega = 2\pi f$; (b) for a high-pass filter, $x = -1/\omega$; (c) for a symmetrical bandpass filter, $x = (\omega/\omega_0) - (\omega_0/\omega)$; (d) for a symmetrical bandstop filter, $x = -1/[(\omega/\omega_0) - (\omega_0/\omega)]$.	

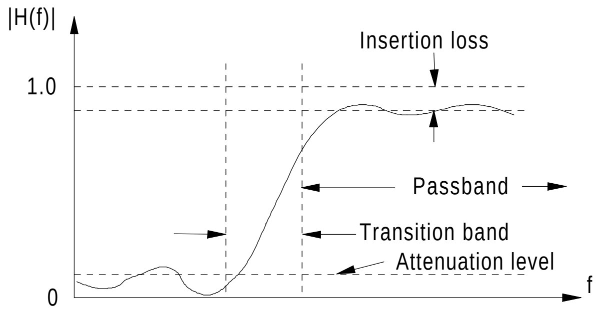


Figure F29 High-pass filter response.

The transfer function of a high-pass filter can be obtained from that of a low-pass filter by replacing s with $1/s$, so the passband and stopbands are interchanged. The main characteristics are the cutoff frequency ω_c, a threshold frequency with which attenuation of oscillations by the filter begins, the width of the transition region of the frequency response, the loss within the passband, and the magnitude of attenuation in the stopband. *IAM*

Ref.: ITT (1975), p. 10.8; Fink (1982), p. 12.32; Sazonov (1988), p. 122.

A **holographic filter** is a filter-mask printed with a holographic representation of the frequency response. It can be used in a complex matched filtering optical device because the throughput capacity and difficulty of fabrication of a holographic filter does not depend upon signal complexity. The holographic filter in a signal-processing device is illuminated such that only the wave depicting the function of the predetermined complex spectrum is propagated in the direction of the optical axis. *IAM*

Ref.: Dulevich (1978), p. 165; Zmuda (1994) p. 399.

An **inverse filter** is one that converts a complex interference input spectrum to a uniform (white) output spectrum, optimizing the output signal-to-interference ratio of subsequent signal integration circuits (Fig. F30). This filter is also known as the *Urkowitz filter* or *whitening filter. DKB*

Ref.: Barton (1988), p. 236.

inverse hyperbolic filter (see **frequency-selective filter**).

A **Kalman(-Bucy) filter** is a linear recursive filter performing optimum filtering in the sense that the *a posteriori* pdf of the smoothed estimate of the filtered parameter is maximized. This type of filter falls into the third category of radar filters (see **FILTER**) and is used primarily for data smoothing in radar trackers.

The main idea of Kalman filtering is as follows. Assume that we have a set of measurements at equally spaced times $t_1, \ldots, t_{n-1}, \ldots, t_n, t_{n+1}$ of a target state $\vec{x}$ in one coordinate. For a two-state Kalman filter, $\vec{x} = (x, \dot{x})$, where $\dot{x}$ is the first derivative of x, and for a three-state filter $\vec{x} = (x, \dot{x}, \ddot{x})$, where $\ddot{x}$ is the second derivative. According to the Kalman filter approach, a smoothed estimate of target state at t_n, (i.e., $\hat{x}_s(n)$) is computed as a weighted sum of values predicted from the previous moment of time, $\hat{x}_s(n|n-1)$, and a difference $\Delta\vec{y}(n)$ between the measured value $\vec{y}(n)$ and

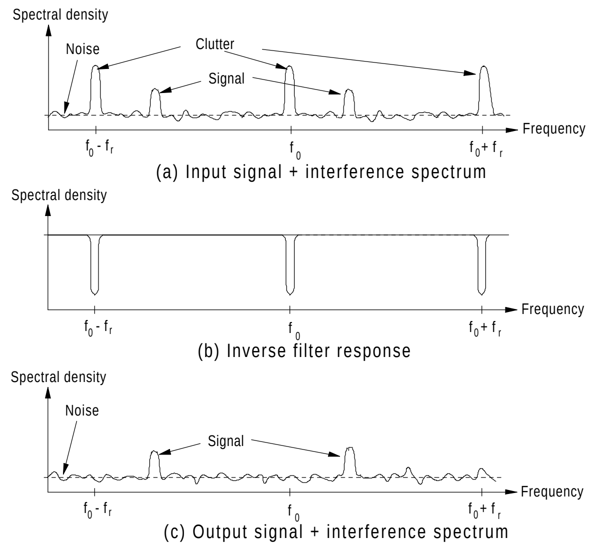


Figure F30 Inverse (Urkowitz) filter for strong clutter in thermal noise.

$\hat{x}_s(n|n-1)$. The weight is a filter gain, $G(an)$, which is a variable function of t_n. This results in a basic equation for Kalman filtering (which is a simplified form of the more general Kalman-Bucy equation):

$$\hat{x}_s(n) = \vec{x}_p(n|n-1) + G(n)\cdot\Delta\vec{y}(n) \qquad (1)$$

We see from this equation that implementation of the Kalman filter requires the following steps:

(1) The prediction model defining the relationship between $\hat{x}_s(n-1)$ and $\hat{x}_p(n|n-1)$ must be specified. This is assumed to be a discrete Markovian sequence, which for a deterministic motion model results in

$$\vec{x}_p(n|n-1) = \vec{\Phi}_p \cdot \vec{x}_s(n-1)$$

$$\vec{K}_p(n) = \vec{\Phi}_p \cdot \vec{K}_s(n-1) \cdot \vec{\Phi}_p^T$$

where $\vec{\Phi}_p$ is the prediction matrix, and $\vec{K}_p(n)$ and $\vec{K}_s(n-1)$ are prediction and smoothed estimate covariance matrices for corresponding moments of time. For a polynomial representation of the target trajectory (second-order polynomial),

$$\vec{\Phi}_p = \begin{bmatrix} 1 & T & T^2/2 \\ 0 & 1 & T \\ 0 & 0 & 1 \end{bmatrix}$$

where $T = t_n - t_{n-1}$, giving the following model of target state prediction:

$$x_p(n|n-1) = \hat{x}_s(n-1) + \hat{\dot{x}}_s(n-1)\cdot T + \hat{\ddot{x}}_s(n-1)\cdot\frac{T^2}{2}$$

$$\dot{x}_p n|n-1 = \hat{\dot{x}}_s(n-1) + \hat{\ddot{x}}_s \cdot T$$

$$\ddot{x}_p(n|n-1) = \hat{\ddot{x}}_s(n-1)$$

(2) The measurement model has to be specified. For the Kalman filter this would have the form

$$\vec{y}(n) = \vec{H} \cdot \vec{x}_s(n) + \delta\vec{y}(n)$$

where $\vec{y}(n)$ is the vector of measured coordinates at *tn*, $\vec{x}_s(n)$ is the vector of estimated trajectory parameters at *tn*, and $\delta\vec{y}(n)$ is the vector of coordinate measurement errors. The transition matrix $\vec{H}$ defines the relationship between measured and estimated vectors; for example, when $\vec{y}(n) = (R, \varepsilon, \beta)$ and $\vec{x}_s(n) = (R, \dot{R}, \varepsilon, \dot{\varepsilon}, \beta, \dot{\beta})$, where R,ε,β are range, elevation, and azimuth. The vector $\delta\vec{y}(n)$ is assumed to have a Gaussian distribution with zero mean and covariance matrix $\vec{K}m(n)$. For independent measurements with rms errors σ_r, σ_ε, and σ_β:

$$\vec{K}_m = \begin{bmatrix} \sigma_R^2 & 0 & 0 \\ 0 & \sigma_\beta^2 & 0 \\ 0 & 0 & \sigma_\varepsilon^2 \end{bmatrix}$$

In this case the second term in (1) becomes

$$\Delta\vec{y}(n) = \vec{y}(n) - \vec{H} \cdot \vec{x}_p(n|n-1)$$

and the following basic equations describing the Kalman filter can be specified:

$$\vec{x}_s(n) = \vec{x}_p(n|n-1) + \vec{G}(n) \cdot [\vec{y}(n) - \vec{H} \cdot \vec{x}_p(n|n-1)] \quad (2)$$

$$\vec{K}_s(n) = \vec{K}_p(n) - \vec{K}_p(n) \cdot \vec{H}^T \cdot \vec{S}^{-1}(n) \cdot \vec{H} \cdot \vec{K}_p(n) \quad (3)$$

$$\vec{S}(n) = \vec{H} \cdot \vec{K}_p(n) \cdot \vec{H}^T \cdot \vec{K}_m^{-1}(n) \quad (4)$$

$$\vec{G}(n) = \vec{K}_s(n) \cdot \vec{H}^T \cdot \vec{K}_m^{-1}(n) \quad (5)$$

Equation (2) provides the filtering algorithm itself (i.e., how to compute the smoothed estimate of target state at the *n*th step). Equations (3) and (4) give the errors of this estimate, defined by the covariance matrix $\vec{K}_s(n)$, which is a function of the prediction errors $\vec{K}_p(n)$ and the measurement errors $\vec{K}_m(n)$. Equation (5) gives the formula for the filter gain $\vec{G}(n)$, which is a function of the prediction and measurement errors at time t_n.

The simplified version of the Kalman filter when the filter gain is constant, $\vec{G}(n) = G$ and does not change adaptively with the errors is known as the α-β(-γ) filter. *SAL*

Ref.: Bozic (1979); Blackman (1986), p. 25; Brammer (1989).

A **Kalmus clutter filter** is a circuit intended to detect slowly moving targets whose doppler shifts are insufficient to be resolved in the presence of clutter. The circuit operates on the principle that the moving target has an average doppler shift different from zero, where vegetation and similar clutter sources move back and forth, creating a spectrum that is symmetrical about zero doppler. *DKB*

Kalmus, H. P., "Doppler Wave Recognition with High Clutter Rejection," *IEEE Trans.* ***AES-3***, no. 6 (Eastcon Suppl.), Nov. 1967, pp. 334–339; Skolnik (1980), p. 497.

A **lattice filter** is a form of realization of a nonrecursive adaptive filter in the form of a structure comprising links connected in series that are nonbifurcated extrapolation filters. (See **filter-extrapolator**.) This structure splits a signal to a set of direct $e(n)$ and inverse $e(n-N)$ samples of difference signals with delays augmented in the inverse channel (Fig. F31). Multipliers in the transverse branches of grid k are called *reflection factors* based upon a physical interpretation of lattice filters in the form of wave propagation in a stratified medium.

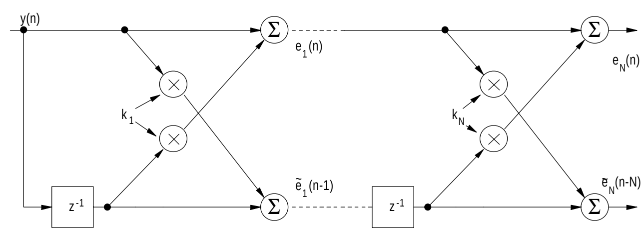


Figure F31 Lattice filter diagram (after Gol'denberg, 1985, Fig. 6.6, p. 170).

Special recursive methods are used to subtract the deflection factors (or PARCOR coefficients, from partial correlation). Advantages of the lattice structure include cascading of identical links; coefficient magnitudes that do not exceed unity, ensuring filter stability; simplicity in checking stability; and a good rounding characteristic. However, the lattice structure does not have a minimum number of multipliers and adders for the assigned transfer function. Lattice filters are used widely for adaptive processing of signals in phased arrays, for adaptive suppression of noise, and for evaluation of a spectrum with high resolution. *IAM*

Ref.: Gol'denberg (1985), p. 169; Cowan (1985), p. 123, in Russian.

A **linear filter** is one in which the output and input signals are linked by a conventional linear differential equation (linear analog filter) or by a linear difference equation (linear discrete filter). The important feature of the linear filter is that the output $y(t)$ and input $x(t)$ for an analog filter are related through the convolution integral:

$$y(t) = \int_{-\infty}^{\infty} x(t)h(t, \tau)d\tau$$

where $h(t)$ is the filter impulse response. In most cases the filter parameters are time-invariant, giving the expression that is fundamental in radar signal filter theory:

$$y(t) = \int_{-\infty}^{\infty} x(t-\tau)h(\tau)d\tau$$

For discrete filters this takes the form

$$y(n \cdot \Delta t) = \sum_{m=-\infty}^{\infty} x[(n-m) \cdot \Delta t] \cdot h(m \cdot \Delta t)$$

where Δt is the sampling interval. This expression can be represented in the form that is the basic discrete filter equation:

$$y(n \cdot \Delta t) = -\sum_{j=1}^{M-1} b_j \cdot y[(n-j) \cdot \Delta t] + \sum_{k=1}^{N-1} a_k \cdot x[(n-k) \cdot \Delta t]$$

where a_k and b_j are filter weighting coefficients. If the coefficients $a_k \geq 0$, $b_k \geq 0$, the filter is called a *recursive filter*, while if all coefficients $b_j = 0$, it is a *nonrecursive filter*. The latter type is more widely used in radar digital signal processing. In this case, assuming $h(0) = a_0$, $h(\Delta t) = a_1$, ... , $h(n\Delta t) = a_n$, this expression results in a formula basic to describing the nonrecursive digital filter:

$$y(n \cdot \Delta t) = \sum_{k=0}^{n} a_k \cdot x[(n-k) \cdot \Delta t]$$

where a_k are weighting coefficients. Strictly speaking, digital filtering is a nonlinear operation, but thanks to their simplicity, linear filters are used not only for linear, but also for quasioptimum, nonlinear filtering of radar signals. *SAL, IAM*

Ref.: Gol'denberg (1985), p. 46.

A **low-pass filter** is "a frequency-selective filter which passes low frequencies and blocks high frequencies." It has a single passband, $0 < \omega < \omega_c$, and a single stopband, $\omega > \omega_c$, where ω_c is a cutoff frequency separating the two bands (Fig. F32). The main characteristics of a low-pass filter are the cutoff frequency, the width of the transition region of response, the magnitude of the loss in the passband, and the magnitude of attenuation in the stopband. Chebyshev, Butterworth, and elliptical filters are the most commonly used low-pass filters. *IAM*

Ref.: ITT (1975), pp. 10.2–10.8; Fink (1982), p. 12.20; Sazonov (1988), p. 122.

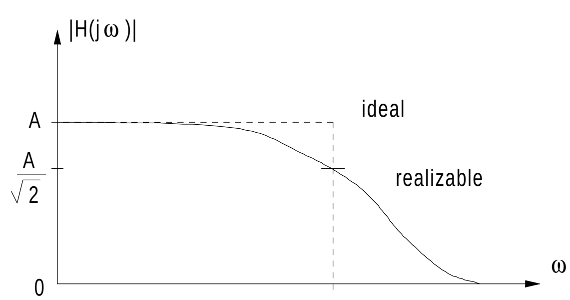


Figure F32 Low-pass filter frequency response.

A **mainlobe clutter filter** is used in an airborne pulsed doppler radar to remove the strong clutter that originates in the mainlobe of the antenna pattern of a downward-looking radar from subsequent signal-processing stages. This filter is normally applied to each range-gated channel of the radar. An analog filter may be used for each range gate in cases where the mainlobe clutter exceeds the dynamic range of the A/D converter or digital filtering circuits. *DKB*

Ref.: Schleher (1991), p. 66; Morris (1988), pp. 101–107.

A **matched filter** is the optimum filter for a given signal in a background of white noise, and as such it maximizes the output ratio of peak signal power to mean noise power. It has a frequency response $H(f)$ that is the complex conjugate of the received spectrum $A(f)$, or equivalently, has an impulse response $h(t)$ that is the time inverse of the received waveform $s(t)$:

$$H(f) = A^*(f)$$

$$h(t) = ks(t_d - t)$$

where t_d is the time delay required to make the filter realizable, and k is a dimensional normalizing factor. These relations lead to waveform-response pairs illustrated in Fig. F33.

When this filter is used, the ratio of output peak signal power to mean noise power is $2E/N_0$, where E is the received signal energy and N_0 is the noise spectral density referred to the receiver input point at which E is measured. *DKB*

Ref.: Cook (1967), pp. 5–9; DiFranco (1968), pp. 143–184.

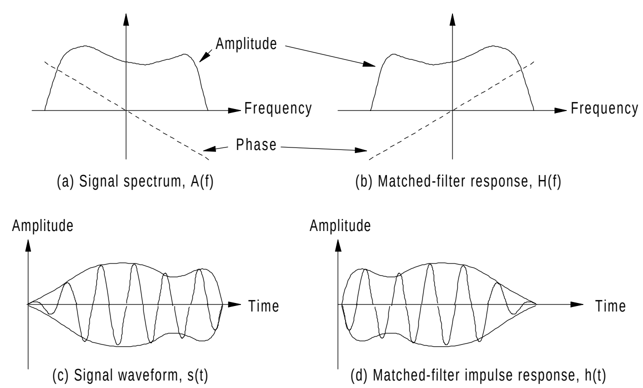


Figure F33 Signal and matched filter response functions (after Cook, 1967).

maximally linear phase response filter (see **frequency-selective filter**).

microstrip filter (see **strip filter**).

A **microwave filter** is a passive two-port device that transmits microwave oscillations to a matched load in accordance with a preset frequency response and the passband. It is usually a frequency-selective filter for microwave frequencies. To reduce losses in the pass band, the filter is generally made from reactive components. A sharp increase in attenuation outside the passband is assured through almost complete reflection of the microwave oscillations from the filter input. Wideband matching circuits (see **ADAPTER, microwave**) are similar to the filters in structure and design principle.

Microwave filters usually have the form of a cascade connection of a series of sections. The sections can be various cavities, loops, sections of linked transmission lines, and so forth. In terms of type of transmission lines, one distinguishes between waveguide, dielectric, and bandpass filters, which belong to the class of reciprocal multiports. Ferrite filters, which can be retuned within a frequency band, possess nonreciprocal properties.

Design of microwave filters is quite complex and is done in several stages, beginning with a simplified circuit, the prototype circuit, and ending with parameter adjustments from the results of experiments.

Microwave filters are used in microwave circuits and also to form inter-stage connections in receiving and transmitting devices. They are also called RF filters. *IAM*

Ref.: Saad (1971), p. 153; ITT (1975), Ch. 7–9; Sazonov (1988), p. 122; Gassanov (1988), p. 51.

A **mismatched filter** is one that fails to meet the specification of a matched filter and hence provides an output signal-to-noise ratio less than the ratio of input signal energy to noise spectral density. It thus has a mismatch loss, relative to that optimum performance. It is often used in radars to simplify the filter design, to reduce time sidelobes in a pulse-compression system, or to compensate for expected presence of non-white (colored) noise. *IAM*

Ref.: Barton (1964), p. 20, (1969), p. 84; Sloka (1970), pp. 19, 64.

A **moving-target-indicator (MTI) filter** is a filter intended to reject clutter components at and near zero frequency, or near the center frequency of the clutter spectrum, while passing as much as possible of the target signal spectrum. The frequency responses of the single- and double-delay line cancelers, often used in MTI systems, are given by

$$H_1(f) = 2\left|\sin\left(\pi\frac{f}{f_r}\right)\right|$$

$$H_2(f) = 4\left(\sin\pi\frac{f}{f_r}\right)^2$$

where $H(f)$ is the frequency response and f_r is the pulse repetition frequency. More complicated response is available when feedback is used around the cancelers, or when range-gated channels are filtered in low-pass digital or analog filters (see also **CANCELER**). *DKB*

Ref.: Barton (1988), pp. 236–238.

Multidimensional filtering is the filtering of a multidimensional signal (a signal that depends on a set of parameters), typically during integration, for the purpose of more precise estimation of signal parameters. In the simplest case, a scalar parameter linearly dependent on the multidimensional signal is evaluated; for example, when evaluating generally grouped coordinates of complex targets and when integrating uniform measurements in radio navigation. Similar problems are solved using linear filters. In the general case, nonlinear filters are used to evaluate multidimensional parameters. *IAM*

Ref.: Fal'kovich (1981), p. 237.

A **multistage filter** consists of a cascade of similar stages. The frequency response of a multistage filter is the product of frequency characteristics of the stages.

Multistage filters are used, for example, for matched filtering of pulse-compression waveforms through cascaded connection of delay lines. *IAM*

A **nonlinear filter** is one whose input and output signals are related through nonlinear differential equations. This is typical when signals or interference have non-Gaussian distribution or when interference is not additive. In practice, quasilinear filtering methods are often used. *SAL*

Ref.: Korostelev (1987), p. 260; Fal'kovich (1981), p. 203.

nonrecursive filter (see **recursive filter**).

An **optical filter** is a filter based on the phenomenon of variation of light amplitude and phase when it passes through an optically inhomogeneous medium. It has broad bandwidth and makes it easy to implement multichannel and multifunctional signal processing. Practically any frequency and phase response can be synthesized in the optical filter. The basic components of the optical filter are the optical system, units storing input signal and filter frequency response that are light modulators, and the units of photoelectron registration at the output. The advantages of optical filters are very suitable for analog filtering of pulse-compression waveforms. *IAM*

Ref.: Sloka (1970), p. 221; Zmuda (1994), p. 408.

An **optimum filter** provides the maximum attainable performance for a specified optimization criterion. Two typical examples are filters that provide maximum output signal-to-noise ratio or minimum rms error in signal fidelity. In radar applications, the term optimum filter is usually applied to a linear filter providing the maximum possible signal-to-interference ratio. In this case the filter does not retain the shape of the input signal but deliberately distorts it so that spectral components of the signal are superimposed to maximize the signal-to-interference ratio. The useful by-effect of optimum filtering in this case is a pulse compression phenomenon, widely exploited in modern radar.

In general, when the interference consists of white noise of spectral density N_0, and clutter has an energy spectrum $|S(f)|^2$, the square of the signal spectrum, the optimum filter transfer function is

$$H(f) = \frac{S^*(f)}{N_0 + k|S(f)|^2}$$

where k is a constant related to the clutter-to-noise ratio. When the interference is random white noise, the optimum filter is called a *matched filter*, and its transfer function from the equation above is simply the complex conjugate of the signal spectrum. *DKB, SAL*

Ref.: Nathanson (1969), pp. 301–306; Leonov (1988), p. 61.

A **pi-[section] cavity filter** is a microwave bandpass filter with pi-shaped bent cavity. It uses striplines and has smaller dimensions compared with other types of bandpass filters and also uses a simple production technology. It is used in the low-frequency microwave range as well as for relative bands of more than 20% through the use of its advantages in production of wideband π-cavity filters in comparison with other types of filters. *IAM*

Ref.: Veselov (1988), p. 95; Isipov, L.S., and Balandinskiy, B.B. *Antenny* ***29***, p. 160, in Russian.

A **presumming filter** is a bandpass filter used in synthetic aperture radar processing to reduce the number of samples that must be processed in the final processing.

.Ref.: Brookner (1977), p. 257.

pulse-compression filter (see **PULSE COMPRESSION**).

The **filter Q-factor** is equal to the ratio of the average electromagnetic energy stored in it to the energy lost in the period of oscillations at the resonance frequency. One distinguishes between the external Q-factor, Q_e and the unloaded (internal) Q-factor Q_i. In determining the former, one allows for losses only in external active loads, and in determining the unloaded Q-factor, only for ohmic losses due to the nonideal nature of the conductors and dielectrics inside the resonance system. The full (loaded) Q-factor of a filter Q is found by the formula

$$\frac{1}{Q} = \frac{1}{Q_e} + \frac{1}{Q_i}$$

IAM

Ref.:Gardiol (1984); p. 138; Sazonov (1981), p. 170.

A **quasioptimum filter** is a mismatched filter that introduces minimum losses relative to the optimum filter. Most radar signal-processing filters fall into this class (e.g., doppler filters operating in a background of nonwhite noise). *IAM*

Ref.: Sloka (1970), p. 19; Barton (1988), p. 268.

A **range-gated filter** is an MTI or doppler filter operating on range-sampled signals. The range gating has the effect of producing a comb filter with repeated responses at intervals of the pulse repetition frequency, thereby matching the spectrum of the coherent pulse train. All signals are reduced in bandwidth to lie within the pulse repetition frequency, but the total system bandwidth remains matched to the pulse bandwidth if multiple range-gated channels cover the entire pulse repetition interval.

Range-gated MTI filters can readily be implemented with either broad or narrow rejection notches of arbitrary depth and with steep transitions from stopband to passband. *DKB*

Ref.: Skolnik (1980), pp. 117–119.

The **rectangular filter** has an idealized bandpass frequency response:

$$H(f) = 1, |f - f_0| \le B/2,$$
$$H(f) = 0, |f - f_0| > B/2$$

where f_0 is the center frequency and B is the bandwidth. This response is approximated by a multiple-stage, stagger-tuned receiver. *DKB*

A **recursive filter** is a digital filter including multipliers that provide feedforward (weighting coefficients a) and feedback (weighting coefficients b) (see **filter characteristics**). It is the most general form of digital filter. Its transfer function has both zeroes and poles (see filter characteristics), as opposed to a nonrecursive filter, which has only zeroes (because it uses only feedforward weightings). When implemented with a tapped delay line, the recursive filter theoretically has infinite memory and is termed an *infinite impulse response (IIR) filter*, as opposed to the nonrecursive type, which is a *finite impulse response (FIR)*. A comparison of the main features of the two types is given in Table F5.

The nonrecursive filter is more complicated and less flexible, and so to obtain a specified frequency response it requires more components than the recursive filter, but it is always stable and has no transient period. This is especially important in radar applications, where strong interference (clutter or jamming) can cause prolonged ringing in the recursive filter, masking targets until the ringing damps out. In surveillance radars, the number of pulses is restricted, so operation of processors such as the moving target detector will take place during the transient period if a recursive filter is used. On the other hand, the recursive filter is much more flexible because poles and zeroes are available for shaping the frequency response, offering practically any desired frequency response even without use of staggered PRF. *SAL*

Ref.: Skolnik (1980), pp. 110–114; Cowan (1988), p. 17.

Table F5
Comparison of Recursive and Nonrecursive Filters

Feature	Nonrecursive	Recursive
Absolutely linear phase response	Yes	No
Internal noise power	Much less	Much greater
Complexity of calculating coefficients	Simpler	More complicated
Transient period	No	Yes
Number of operations	Much greater	Much less
Design flexibility	Less	Greater

A **ripple filter** is a device accomplishing data smoothing by linear combination of data based on a given criterion. The least squares method usually is used as the criterion. Ripple filters are differentiated by the shape of the curve used for smoothing (smoothing with a straight line, with parabolas of the second, fourth degree, etc.), as well as by the number of data added with certain weights defining filter length. Filter coefficients are numerically equal to the weights of the added data and are defined by filter type.

The transfer function of a ripple filter has maximum value at zero frequency and drops with increasing frequency. The frequency response drops more quickly as filter length is increased, while the magnitude of the subsequent oscillations drops slightly.

Along with smoothing of measurement data, ripple filters are used for compensation of truncated series pulsations. Filters with impulse response providing windowing in the time domain are used for this purpose. *IAM*

Ref.: Hamming (1977), p. 36, in Russian.

A **robust filter** is an analog or digital filter that performs robust signal processing. There are robust filters for detection of a pulse train observed against the background of additive interference, independent of the signal, with a known probability distribution function. Robust processing algorithms depend only on the differences of random phases ψ_N of the observed pulse train, referred to the interval $[0, 2\pi]$, and known phases of the pulses of the transmitted pulses.

Algorithms for processing the input samples of $\vec{x}_N, x_i = \psi_i/(2\pi)$ consist of calculating the specified criterion (measure) and comparing the measure with the threshold (table). The pdf of the output signal of robust filters, when there is only interference at the input, does not depend on the statistical characteristics of the interference. Depending on the analog or digital form of the implementation, one distinguishes between analog and digital robust filters. In terms of type of measure for determining the threshold, one distinguishes between filters of the Neyman-Pearson criterion, David criterion, and the criteria based on construction of the confidence for an empirical pdf $F(x, \vec{x}_N)$, constructed from the observed sample (types 3 through 6 in Table F6).

A robust compression filter is analogous to the known compression algorithm of pseudo-noise phase-coded waveforms, but applies to the case of random phase code.

It is impossible to choose the best filter from the robust class, since this would require information about the statistical characteristics both of the interference and of the signal, but robust filters are intentionally designed to avoid this requirement. *IAM*

Ref.: Ovodenko (1981), pp. 41–86.

A **single-band filter** is a bandpass filter with one bandpass in the entire possible frequency band. An example of use is the single-band filter of a pulsed-doppler radar, which does not pass side frequencies differing from the central frequency of the radar by a multiple of the pulse repetition frequencies. *IAM*

Ref.: Skolnik (1970), p. 19.4.

A **single-pole filter** uses a single *L-C* or *R-C* section to form an appropriate bandpass response.

A **smoothing filter** is one that smooths (averages) a set of radar measurements, typically to reduce the fast fluctuating random component of measurement error (noise error). The most widely used smoothing filters are the α-β(-γ) filter and the Kalman filter. *SAL*

A **solid-state filter** is a filter implemented using solid-state technology (semiconductor components and integrated circuits).

Table F6
Classification of Robust Filters

Type of Filter	Measure
1. Neyman-Pearson	$\frac{m}{N}\sum_{i=1}^{m}\nu_i^2 - N$ where ν_i is the number of elements x_N in the interval $S_i = [(i-1)/m, i/m]$
2. David (digital)	where S_0 is the number of intervals $S_1, \ldots, S_m$ that do not contain a single element x_n
3. Kolmogorov-Smirnov (analog)	$\sup_{(-\infty < x < \infty)} \left\|F(x, \vec{x}_N) - F_0(x)\right\|$ $F_0(x) = \begin{cases} 1, x > 1 \\ x, 0 \le x \le 1 \\ 0, x < 0 \end{cases}$
4. Shirman (analog)	$\frac{1}{2}\sum_{i=1}^{N+1}\left\|x'_i - x'_{i-1} - \frac{1}{N+1}\right\|$ where x'_i are normalized elements
5. Morgan-Kimboll (analog)	$\sum_{i=1}^{N+1}(x'_i - x'_{i-1})^2$
6. Kramer-Mieses (analog)	$\frac{1}{N}\sum_{i=1}^{N}\left(x_i - \frac{2i-1}{2N}\right)^2$
7. Compression (analog)	$\sqrt{\sum_{i=1}^{N}\sum_{j=1}^{N}\cos[2\pi(x_i - x_j)]}$

Space-time filtering is filtering of spatial-temporal signals. Signal dependence on spatial coordinates may be stipulated both by the difference in coordinates of receiving elements and by signal dependence on the spatial parameters of the target. In both instances, by virtue of the complete similarity of the input signal in the spatial and in temporal characteristics, spatial-temporal filtering is accomplished using conventional linear and nonlinear filtering methods considering the specifics of concrete problems. Spatial-temporal filtering is used in phased arrays, navigation, terrain mapping, and guidance radars. *IAM*

Ref.: Korostelev (1987), p. 80; Baklitskiy (1986), p. 26; Skolnik (1990), p. 16.23.

Spatial filtering is the filtering of a signal, the evaluated parameters of which depend upon the spatial coordinates of the target. By virtue of the complete similarity of the spatial and temporary characteristics, generally spatial filter methods are the same as both temporal and space-time filtering. Spatial filtering is realized in the spatial section of phased-array radars, forming in the absence of interference a matched spatial filter. The sole difference between spatial and temporary filtering is presence of an additional operation for integration of spatially filtered signals relative to all coordinates of the antenna aperture, which actually is an inverse Fourier transform. The term *spatial filtering* also is used in terrain-mapping airborne radars, understood to mean simultaneous (parallel) processing of separate image elements. *IAM*

Ref.: Baklitskiy (1986), p. 11.

Spatial-adaptive filtering is spatial filtering with a change over time of spatial filter parameters in accordance with the variation in radar signal inputs. Spatial-adaptive filtering is used widely in phased-array radars for spatial selection of a useful signal observed against the background of an interfering signal. Filtering involves optimum evaluation of the weighted vector of phased-array elements performed by a filter-compensator in accordance with the algorithm selected. (See **ALGORITHM, Widrow**; **CANCELER, Howells-Applebaum,** etc.) The weighted vector is controlled adaptively by feedback from output to input. The latter is realized during correlation of total output voltage and the voltages of individual channels. This leads to automatic formation of notches in the directivity performance toward sources of interference. Spatial-adaptive filtering is realized both in digital and analog forms. *IAM*

Ref.: Shirman (1981), p. 352; Nitzberg (1992).

Spatial-extent filtering is a technique in which clutter is rejected by filtering all returns which extend over more than n range cells, when targets are known to extend over fewer than n cells.

Ref.: Currie (1987), p. 283.

Filter stability. A filter is considered stable if under any conditions of operation an arbitrary limited input signal $|x(t)| \le A$ results in a limited output signal $|y(t)| \le B$. Practically, it means that

$$\left|Z_k^{(P)}\right| < 1,\ k = 1, 2, \ldots, M-1$$

where $\left|Z_k^{(P)}\right|$ is a filter pole, or

$$\sum_{n=0}^{\infty} |h(n \cdot \Delta t)| \le C$$

where $h(n\Delta t)$ is the impulse response of a signal sampled at rate Δt, and C is a constant.

These criteria show that a nonrecursive filter is always stable, and to be always stable a filter must have a finite impulse response. The recursive filter can be either stable or unstable depending on its transfer function. The unstable filter cannot operate when an input signal has infinite duration but can be used in some cases when the signal has limited time duration. *SAL*

Ref.: Gol'denberg (1985), pp. 54, 59.

A **stagger-tuned filter** is one in which successive stages in a receiver are tuned to slightly different frequencies, broadening the bandwidth and achieving an approximation to the rectangular filter response.

A **strip(line) filter** is based on sections of strip transmission lines. Stripline filters are used as high-pass and low-pass filters, providing different frequency responses (e.g., Chebyshev filter, elliptical filter). In terms of type of line sections, one distinguishes between filters with distributed and quasi-lumped parameters. The first use sections of a stripline that, depending on the wave resistance, perform capacitive or inductive functions. Filters with quasi-lumped parameters use line sections in the form of a coil (inductance coil), lines with opposing-pin structure (capacitor), or inductive-capacitive loops.

The manufacturing technology of striplines limits the passband of the filters to 20 to 60% depending on the structure. To increase the bandwidth (with the corresponding increase in losses), strip dielectric and ferrite filters are used. *IAM*

Ref.: Veselov (1988), p. 87.

A **surface-acoustic-wave (SAW) filter** is an analog filter, the basic element of which is a surface-acoustic-wave delay line. Prior to the development of digital technology, it was widely used as a bandpass filter for matched filtering of pulse-compression waveforms. SAW filters are used in the 10-MHz to 2-GHz band with a 100-kHz bandwidth all the way to 50% of the central frequency. Minimum bandwidth is constrained by physical dimensions. The level of sidelobes reaches –60 dB, bandwidth ratio (at the –40 dB level compared with the –3 dB level) reaches 1:1, and response realization accuracy in the bandwidth is about 0.7 dB.

A SAW filter for matched filtering of pulse-compression waveforms comprises a tapped delay line, phase shifters, and an adder. Programmable transversal SAW filters with signal phase and amplitude control at output of each tap have found wide use. The time-bandwidth product of the signal the filter processes usually is limited to a value of 1,000 due to an increase in insertion losses. The level of the sidelobes may be brought to –45 dB. *IAM*

Ref.: Matthews (1977); Jordan (1985), p. 28.26.

A **synchronously tuned filter** is one in which successive stages are tuned to the same frequency, rather than being stagger-tuned. The bandwidth for m synchronously tuned stages, each of bandwidth B_1, is

$$B = B_1\sqrt{\frac{\ln 2}{m}}$$

tapped delay-line filter (see **transversal filter**).

Temporal [time] filtering is filtering of signals, the evaluated parameters of which are functions of time. Temporal filtering is the main type of filtering in surveillance radars and is used in all types of non-phased-array, tracking, and scanning radars. In scanning radars with temporal filtering, the assigned coverage volume is monitored sequentially by beam scanning. As a result, the observed image is a function of time. *IAM*

Ref.: Baklitskiy (1986), p. 12.

A **tracking (loop) filter** is a closed nonlinear filter designed for evaluation of a parameter nonlinearly coded in a signal. A tracking loop filter nonlinear with respect to input signal $S(\Lambda(t))$ is similar to a linear filter with respect to its parameter $\Lambda(t)$. Basic tracking loop filter assemblies are a discriminator generating an error signal relative to the measured parameter in the linear sector of its response, estimator, synthesizer, and seek-and-capture circuit (Fig. F34). The synthesizer provides the requisite discriminator response by shaping the correlator reference signal (or filter impulse response).

A tracking loop filter operates in two modes: acquisition and track (main mode). In the acquisition mode, the tracking loop is open and the acquisition circuit changes reference value Λ_0 of the detector and discriminator. When Λ_0 becomes close to actual, the trap relay trips and the seek circuit ceases operation. At that moment, the input signal reaches the discriminator operating sector and the tracking loop switches to its main mode. Variable-purpose tracking loop filters are used in radars: range, speed, and azimuth tracking loop estimators, as well as those measuring several signal parameters. *IAM*

Ref.: Korostelev (1987), p. 265.

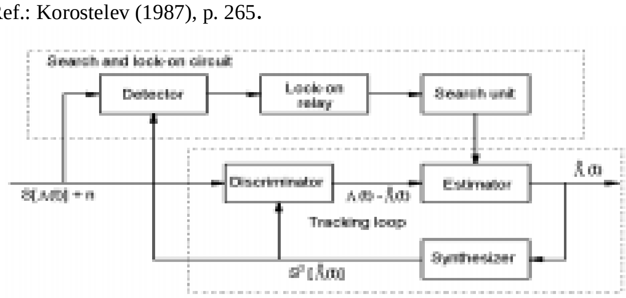


Figure F34 Tracking loop filter diagram (after Korostelev, 1987, Fig. 8.4, p. 267).

track-while-scan filter (see **α–β filter**).

filter transfer functions (see **filter characteristics**).

A **transversal filter** is a nonrecursive filter using a tapped delay line to implement the basic filter equation. (See **linear filter.**) It can be realized either as an analog filter (using SAW lines, CCDs, etc.) or as a digital filter using digital delays. A generic form of transversal filter is shown in Fig. F35, in which the inputs, $x(t)$, consisting of $m + 1$ pulses, are weighted to form the output, $y(t)$.

Digital transversal filters are widely used in radar digital signal processing for clutter suppression in moving target indicators (see **CANCELER**), pulse compression, and in other applications. This type of filter is also known as the tapped delay-line filter, feedforward filter, finite memory filter, finite impulse response filter, or simply the nonrecursive filter, although other forms of nonrecursive filter are possible). *DKB, SAL*

Ref.: Skolnik (1980), p. 110.

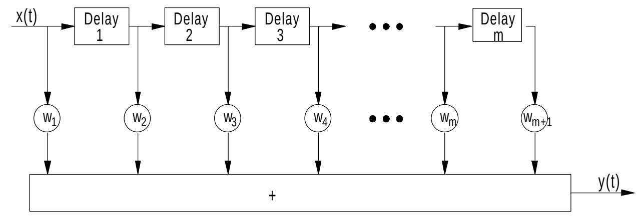


Figure F35 General form of a transversal filter for signal processing (after Schleher, 1991).

Urkowitz filter (see **inverse filter**).

A **waveguide filter** is a microwave filter based on waveguide transmission lines. Waveguide filters are used as bandpass, low-pass, or high-pass filters. For narrowband systems (10%), the waveguide filter realizes the equivalent circuit in the form of a ladder circuit: a circuit of parallel resonant loops (or individual reactive components) connected by serial loops (or reactive elements). At low power, ladder circuits are realized by mounting resonant irises at quarter-wave distance in the waveguide. In powerful devices, a waveguide cavity in the form of a volume bounded by pairs of inductance pins is used in place of irises.

Stepped filters, which constitute a circuit of like waveguide lines of identical electrical length but different wave resistance, possess wideband properties. The amplitude-frequency characteristic of a stepped filter constitutes a number of alternating pass and elimination bands. Stepped and smooth microwave adapters are examples of waveguide filters. *IAM*

Ref.: Fel'dsteyn, (1963), pp. 259, 289; Saad (1971), p. 153.

A **waveguide-dielectric filter** is a microwave dielectric filter that constitutes serial or parallel connections of waveguide-dielectric cavities. The connection between the cavities is made through both propagating fields and attenuating fields (limit mode). In the latter case, filters of significantly smaller size are realized. A round waveguide with periodically arranged dielectric disks is a typical design for a filter of the first type.

Waveguide-dielectric filters are used as low-pass, bandpass, and stopband filters. To produce bandpass filters, the connection between the waveguide-dielectric cavities is reduced by increasing the distance between them.

Depending on the types of waves used, and the shape of the dielectric elements, the unloaded Q-factor is of the order of 6,000, and losses in the pass band are 0.1 to 1 dB. *IAM*

Ref.: Chung-Li Ren, and Han-Chiu Wand, *IEEE Trans* ***MTT-2***, no. 12, Dec. 1974, pp. 1,202–1,209.

filter weighting (see WEIGHTING).

Wiener filter is a linear optimum filter (with respect to the criterion of minimum rms error) used to extrapolate the signal received in a noise background. It can be implemented either in analog and digital variants and is characterized by the impulse response $h(\tau)$, the correlation function B_x, of the signal $\vec{x}$, and cross-correlation function B_{sx} of the signal $\vec{x}$ and

$\vec{S} = \vec{x} + \vec{n}$, where n is the noise vector. The impulse response of the Wiener filter is derived from the solution of the Wiener-Hopf integral equation:

$$\int_0^t h(\tau' + t_e)B_x(\tau - \tau')d\tau' = B_{sx}(\tau + t_e)$$

where t_e is extrapolation time.

The impulse response of the Wiener filter is

$$h_w = \begin{cases} h(\tau), \tau \geq 0 \\ 0, \tau < 0 \end{cases}$$

The frequency response is

$$H(f) = \frac{\tilde{B}_{sx}(f)}{\tilde{B}_x(f)} \exp(j2\pi f t_e)$$

where $B_{sf}(f)$ and $B_x(f)$ are Fourier transforms of $B_{sx}(\tau)$ and $B_x(\tau)$. The Wiener filter is a filter-extrapolator. *IAM*

Ref.: Farina (1985), vol. 1, p. 83; Korostelev (1987), p. 219.

The **yttrium-iron-garnet (YIG) filter** uses a monocrystal of iron-yttrium garnet, which possesses ferromagnetic properties and low losses. They are used as dispersion and directional filters (see also **ferrite filter**).

Dispersion properties arise in the presence of an external nonhomogeneous magnetic field in electromagnetic, spin, and acoustic waves having different types of propagation. The required linearity of dispersion is obtained through rational distribution of intensity of the constant magnetic field along the length of the crystal, through control of propagation of the acoustic wave using toroidal plates of aluminum-yttrium garnet, by changing the tilt of the edges of the crystal, and so forth.

YIG filters are used for analog matched filtering of linear frequency-modulated waveforms with bandwidths of hundreds of megahertz, providing a compression factor in the hundreds. *IAM*

Ref.: Shirman (1974), p. 149.

FLIP-FLOP (see **TRIGGER**).

FLUCTUATION. The term fluctuation is applied to a unsystematic (typically random) deviation of any physical quantity from its nominal value. In radar it is applied to radar echoes to describe the changes in RCS of complex targets (RCS fluctuations) and effects of irregularities in atmospheric refractive index (angle-of-arrival fluctuations). Other types of fluctuations can include instability of carrier frequency, antenna scanning, rotation of the polarization plane, and so forth. The main issue is the RCS fluctuation, usually described in terms of the Swerling target models (see **RCS fluctuation**). These lead to degradation of probability of detection and to errors in radar measurement. *SAL*

Ref.: Skolnik (1962), p. 50; Popov (1980), p. 457; Vasin (1977), pp. 84–88.

Angle-of-arrival fluctuations of radar signals are the result of irregularities in the refractive index of the atmosphere, which cause the path of the echo wave to vary from the geometric line of sight. In the troposphere, these fluctuations are primarily caused by irregularities in the water-vapor content, especially in the vicinity of clouds but also in the clear air. The rms value of angle-of-arrival fluctuations in clear air is a few tens of microradians, increasing to over a hundred microradians on paths through clouds. These fluctuations are independent of radar wavelength. Ionospheric fluctuations also occur, and these vary as the square of wavelength. *DKB*

Ref.: Barton (1969), App. D, (1988), pp. 309–313.

antenna-scanning fluctuation (see **MTI, limitations to performance**).

Path-length fluctuations occur along atmospheric paths as a result of random variations in refractive index of the troposphere and ionosphere. Tropospheric fluctuations are independent of radar wavelength and typically amount to a few millimeters when observed over periods of minutes, with greater drifts over longer periods. Ionospheric fluctuations are in the order of centimeters for 0.3m wavelength, varying with the square of wavelength. *DKB*

Ref.: Barton (1969), App. D, (1988), p. 313.

FOLLOWER

A **cathode follower** is an electron tube amplifier with a load in a cathode circuit, which repeats the shape and phase of the input voltage. The cathode follower has a high input resistance, low input capacitance, low output resistance, and high stability. The gain coefficient of the cathode follower is less than one.

The cathode follower is used in pulsed amplifiers (video amplifiers) operating with a low-ohm active-capacitance load; and also for matching of a high-ohm source with a low-ohm load; for example, when a transmission line matched at the terminal is connected. *IAM*

Ref.: Terman (1955), p. 357; Popov (1980), p. 171; Fradkin (1969), p. 70.

An **emitter follower circuit** is a transmitter amplifier having the common-collector configuration. The performance and usage is analogous to the cathode follower. *IAM*

Ref.: Zherebtsov (1989); Fradkin (1969), p. 184.

FREQUENCY is the rate at which a periodic phenomenon is repeated. In radar, one considers first the radio frequency of the transmitted carrier, which is the rate at which the sine wave representing the electromagnetic field is repeated. Other frequencies used to describe radar operation are the intermediate frequency, referring to the downconverted signal in a superheterodyne receiver; the video frequency components, which represent the baseband pulse; and the pulse repetition frequency, which is the rate at which transmitter pulses are repeated. *DKB*

Frequency agility refers to the capability of a radar to change its transmitted carrier frequency on a pulse-to-pulse or burst-to-burst basis, as opposed to frequency diversity, which refers to a radar's use of several complementary transmissions at different frequencies. Frequency agility and diversity are basically ECCM techniques that force a potential jammer to spread the available energy across a wider bandwidth, thereby decreasing the jamming energy density (W/Hz). *PCH*

Ref.: Skolnik (1990), p. 9.17.

Angular frequency is frequency expressed in radians per second, equal to the frequency in hertz multiplied by 2π. *PCH*

Ref.: Van Nostrand (1983), p. 1,276.

Automatic frequency control (AFC) is the automatic maintenance of signal frequencies within the various stages of a receiver. Absolute AFC maintains a constant absolute frequency, while difference AFC maintains a constant difference between the signal frequency and a local oscillator. An AFC system comprises a discriminator which determines the magnitude and sign of required tuning adjustment, a filter that limits the bandwidth of the adjustment, and a controller to vary the frequency of the local oscillator.

AFC systems may be based on frequency or phase loops. A frequency feedback system uses a frequency detector and has a steady-state residual mistuning. A phase-locked loop (PLL) uses a phase detector and provides exact tuning.

AFC systems in pulsed radars may also be categorized as either fast or slow, with the former providing tuning adjustments in a time shorter than the pulse duration, while the latter adjusts the tuning over several pulse repetition intervals.

Another differentiating feature is the type of tuning control. An electronic control system uses a voltage-controlled oscillator, klystron, traveling-wave tube, IMPATT diode oscillator, or Gunn-diode oscillator, and provides fast response but a narrow tuning band. The tuning loop or cavity may also be adjusted mechanically, in which case the AFC system provides a wide tuning band at the expense of slow response. The oscillator may also be adjusted thermally, or with a combination of electronic, mechanical, and thermal means.

An AFC system may operate in two modes: search mode, in which the local oscillator frequency is controlled by the search program, and track (locked) mode. Not all AFC systems provide the search mode.

The performance of an AFC system is characterized by the width of its lock-on band (the band within which the system responds to slow changes in the input frequency difference), the capture band (the band within which the system can acquire the signal frequency, with rapidly-changing mistuning), and the duration of the tuning process.

Automatic frequency control systems are implemented in analog or digital circuitry, or a combination, depending upon the requirements for accuracy, rapid response and reliability.

Automatic frequency control with frequency feedback uses a frequency detector to determine the magnitude and sign of the frequency difference between an input signal and a reference. When operating in steady state, there is a residual mistuning, which results in AFC being used mainly in noncoherent pulsed radars. The performance of an AFC system is characterized by the autotuning coefficient, which is the ratio of the initial frequency difference to the steady state value, as well as other parameters commonly used to describe tuning systems.

Automatic frequency control using frequency and phase uses two feedback loops, providing both frequency and phase tuning. An automatic frequency-phase control system contains both a frequency detector and a phase detector. In comparison with a phase-locked loop, a frequency-phase loop is able to provide a wide capture band, the width of which is determined by the system's frequency channel. The width of the lock-on band is determined by the phase channel.

Automatic frequency control with a PLL is a form of automatic frequency control in which a phase detector is used to determine the magnitude and sign of the frequency difference between an input signal and a reference. The reference signal generated by a local reference oscillator is presented to the controlling input of the phase detector. There is no residual mistuning in a phase-locked loop, as there is in a frequency-feedback AFC system. This makes phase-locked loops suitable for use in coherent radar receivers.

The use of digital circuitry increases the accuracy, reliability, and stability and permits automatic transitions between the search and lock modes. *IAM*

Ref.: Sokolov (1984), pp. 170–174; Skolnik (1980), p. 157; Neri (1991), p. 142; Morris (1988), p. 257.

The **beat frequency** is formed when two ac signals of different frequencies are combined to form a third signal whose frequency is equal to the difference between the frequencies of the two original signals, the "beat frequency." Beat frequencies can be produced by passing the sum of the two signals through a nonlinear circuit, such as a rectifier, or by mixing the two signals, a process called *heterodyning*.

The doppler beat frequency is the frequency generated by mixing a doppler-shifted echo with the transmitted carrier frequency or other reference signal. In a coherent MTI or doppler system, it is equal to the doppler frequency shift. In a noncoherent MTI, the reference signal may originate in moving clutter, and the beat frequency will then be the difference in doppler shifts of the target and the clutter, removing the first-order effect of radar platform velocity. *PCH*

Ref.: Stimson (1983), p. 595.

Combination frequencies are the frequencies of mixer products defined by the expression

$$f_k = nf_1 \pm mf_2$$

where n, m are integers; f_1 and f_2 are the frequencies of interacting oscillations. For example, in a superheterodyne receiver, the difference combination frequency $f_s - f_{lo}$, where f_s is the signal frequency and f_{lo} is the local oscillator frequency, is typically used as an intermediate frequency.

Ref.: Skolnik (1970), p. 5.7; Popov (1980), p. 182.

The **critical (ionospheric) frequency** is the highest frequency that will be reflected, at vertical incidence, by the ionosphere. For electron density N_e per m^3

$$f_c = 9\sqrt{N_e}$$

where f_c is in hertz. For typical daytime conditions, $N_e \approx 10^{12}$, $f_c \approx 9$ MHz. *AIL*

Ref.: Barton (1988), pp. 309–313.

The **critical frequency shift** is the frequency separation necessary to obtain independent samples of target or clutter signals in a frequency agility or diversity radar system.

The **cutoff frequency** is (1) the frequency defining the boundaries of frequency-selective filter passband (if the passband is restricted from both ends, the higher boundary frequency in termed the upper cutoff frequency and the lower is lower cutoff frequency); or (2) the frequency in waveguide transmission lines below which no significant propagation occurs. Energy propagates, without significant loss, in various "modes" at frequencies above cutoff. These modes are defined in terms of the type of the propagating electromagnetic wave, transverse electric, which has no electric field component in the *z*-direction, and $TE_{m,n}$ or transverse magnetic, $TM_{m,n}$, having no magnetic field component in the *z*-direction (Fig. F36). The subscripts *m* and *n* indicate, for a rectangular guide, the number of half-wave variations of the transverse fields in the *x*- or *y*-directions, respectively.

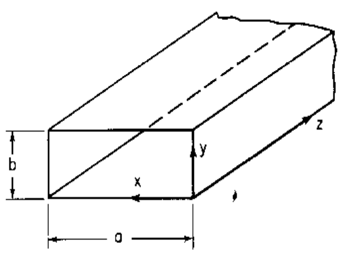


Figure F36 Rectangular waveguide dimensions and coordinate system (after Skolnik,1970, p. 8.7).

The cutoff frequency of a rectangular waveguide depends on the mode of operation and is equal to $c/2a$ for the dominant mode, TE_{10}. Tables of cutoff frequencies for standard waveguides are available in the radar literature. *PCH, AIL*

Ref.: Skolnik (1970), pp. 8.7–8.11; Popov (1980), p. 97.

Frequency deviation is a maximum deviation of the carrier frequency within the modulation interval t_r for frequency modulation (Fig. F37). The typical notation is Δf. *AIL*

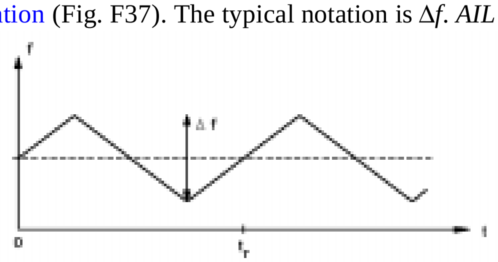


Figure F37 Frequency deviation of sawtooth FM waveform.

Ref.: Terman (1955), p. 586; Popov (1980), p. 103.

Frequency diversity refers to the technique of transmitting and receiving on two or more carrier frequencies, to overcome target or propagation path fading, narrowband jamming, or clutter having pulse-to-pulse correlation or non-Rayleigh distribution. Diversity is distinguished from frequency agility in that the latter implies pulse-to-pulse changes in frequency, while diversity implies change on a burst-to-burst or scan-to-scan basis or use of parallel channels at different frequencies. To be effective, the spacing between channels must exceed the jamming bandwidth or a critical frequency shift (or correlation frequency), generally given by

$$f_c = \frac{c}{2L_r}$$

where L_r is the radial length of the target or clutter source, or the difference in pathlengths between direct and reflected components.

One frequency diversity option uses *k* sequential bursts, each consisting of *m* constant-frequency pulses. Coherent processing over the *m* pulses is possible, followed by noncoherent integration of the *k* outputs during the dwell of $n = km$ pulses. Another option uses scan-to-scan frequency change, with all *n* pulses from a given scan at the same frequency. Other options divide each transmitted pulse of width τ into *k* subpulses transmitted at different frequencies, giving *kn* subpulses per dwell with coherent processing available over *n* subpulses per channel. In this last case, a broadband receiver front-end passes all *k* frequency channels, which are then processed in separate narrowband amplifiers and aligned in time by incremental delays τ/k for postdetection integration. *DKB*

Ref.: Barton (1988), pp. 85–87.

doppler frequency shift (see **DOPPLER EFFECT**).

frequency drift (see **frequency stability**).

Frequency hopping refers to frequency agility or sequential frequency diversity. The term is normally applied to communications systems rather than to radar.

image frequency (see **RECEIVER**).

frequency instability (see **frequency stability**).

Instantaneous frequency is $1/2\pi$ times the rate of change of phase. In the expression for a sine wave, $e(t) = \cos\theta(t) = \cos(\omega_c t + \theta_0)$, where ω_c, the derivative of $\theta(t)$, is the radian frequency and θ_0 is some initial phase angle. When $\theta(t)$ varies with time, as in the case of an FM carrier, the value ω_c in the equation above no longer represents the frequency of the waveform, and it is necessary to define an instantaneous frequency ω_i such that $\omega_i = d\theta/dt$. *PCH*

Ref.: Schwartz (1959), p.113.

Intermediate frequency (IF), in a superheterodyne receiver, is the frequency that results when the received RF signal at f_{rf} is mixed with a signal of frequency f_{lo} from the local oscillator, and filtered to eliminate frequency components at or above f_{rf}. Thus, $f_{if} = f_{rf} - f_{lo}$. It is formed by frequency conversion in a mixer. *AIL*

Ref.: Terman (1955), p. 568; Popov (1980), p. 324.

A **frequency multiplier** is a device for increasing the frequency of a basic oscillator by some integer factor. The operating principle of multipliers lies in distortion of the shape of the sinusoidal input signal by a nonlinear circuit with subsequent selection of the necessary harmonic by resonant circuits.

Multipliers are implemented by diode and transistor circuits. Their operation is based on the use of nonlinear active resistance (point contact, PN, and tunnel diodes and transistors) or nonlinear capacitance (varactors, storage, diodes, bipolar transistors). In the microwave band, varactor and transistor multipliers are the most used.

Transistor multipliers, compared with diode ones, are distinguished by their better separation of input and output and their capability of producing amplification with a low multiplication factor. To obtain high values of multiplication factor, transistor-varactor multiplication circuits are used with transistor stages for intermediate multiplication-amplification, and varactor stages at the output. This is due to the low drop in power with an increase in frequency in varactor multipliers in comparison with transistors.

The basic parameters of frequency multipliers are output power of the required harmonic, efficiency, operating frequency band, level of suppression of secondary oscillations, and instability of phase of output oscillations.

Frequency multipliers are used in radar exciters with master crystal oscillators. *IAM*

Ref.: Terman (1955), p. 473; Gassanov (1988), p. 199; Faber (1995).

Radar frequency bands. Figure F38 shows that most radars occupy the microwave region of the electromagnetic spectrum, but the range of frequencies over which different radars have been designed to operate covers a spectrum nearly 17 octaves in width: from 2 MHz in the case of HF groundwave radars to 300 GHz or higher for laser radars. In theory, there is no limit to the operational frequency of a radar, but there are a number of practical considerations that strongly influence the choice of radar frequency, including atmospheric properties, physical constraints, and availability of radar devices. For example, the inverse relationship between the physical size of RF components, such as waveguide and antenna) and the frequency of operation, plays a significant part in frequency selection for airborne systems, where small size, efficient packaging, and limited prime power are the driving considerations. For most surface-based radar systems, however, the effects of the earth's atmosphere and its weather usually dominate frequency selection.

During World War II, for reasons of security, radar frequencies were divided into letter bands, and for convenience and by tradition, the practice has continued. In 1984, the Institute of Electrical and Electronics Engineers (IEEE) officially adopted the nomenclature of Table F7, and these designations are universally used by radar engineers world-wide. The International Telecommunications Union (ITU) is responsible for assigning specific frequency bands, generally as subsets of the bands shown in Fig. F38, for radar use on a geographical basis. Slight differences in the specific frequencies assigned exist in each of the three ITU regions that comprise the worldwide radar frequency allocation.

The IEEE has adopted standard letter designation for the frequency bands used by radars (Table F7). These letter designations permit the operating frequency of a radar to be identified closely enough to indicate its sensitivity to various fundamental and environmental factors, without providing exact tuning limits that might be militarily sensitive. The frequencies corresponding to each band, and the corresponding assignments by the ITU, applicable to Region 2 (North and South America), are shown in the table.

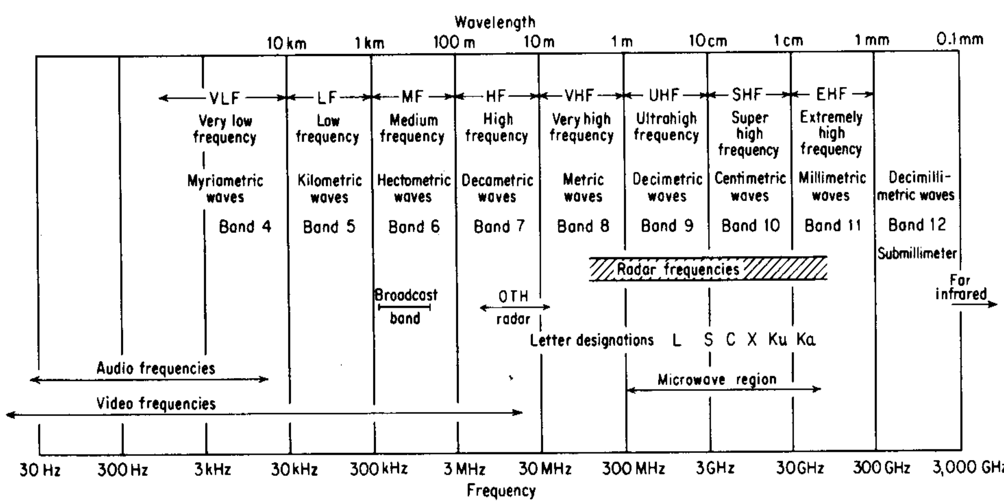


Figure F38 Radar portion of the electromagnetic spectrum (from Skolnik, 1980, Fig. 1.4, p. 7, reprinted by permission of McGraw-Hill).

Table F7

Standard Radar-Frequency Letter Bands

Band letter	Nominal frequency range	Specific frequency ranges for radar based on ITU assignments for Region 2
HF	3–30 MHz	No special radar bands assigned
VHF	30–300 MHz	138–144 and 216–225 MHz
UHF	300–1000 MHz	420–450 and 890–942 MHz
L	1–2 GHz	1.215–1.4 GHz
S	2–4 GHz	2.3–2.5 and 2.7–3.7 GHz
C	4–8 GHz	5.25–5.925 GHz
X	8–12 GHz	8.5–10.68 GHz
K_u	12–18 GHz	13.4–14.0 and 15.7–17.7 GHz
K	18–27 GHz	24.05–24.25 GHz
K_a	27–40 GHz	33.4–36.0 GHz
V	40–75 GHz	59–64 GHz
W	75–110 GHz	76–81 and 92–100 GHz
mm	110–300 GHz	126–142, 144–149, 231–235, 238–248 GHz

The designation mm is derived from millimeter-wave radar and is also used to refer to radar frequencies above 40 GHz and occasionally also to K_a-band. *PCH, DKB*

Ref.: IEEE Standard 521-1984; Skolnik (1990), p. 1.14

The **high-frequency (HF) region** of the radar spectrum extends from 3 to 30 MHz, and as a consequence of the relatively long wavelengths associated with this band (10 to 100m), the RF hardware is physically large, and very large antenna structures are required to obtain narrow beams for high angular resolution. Early use of this frequency band, in the Chain Home network of air-surveillance radars, was made by the British before World War II because in 1938 HF was the highest frequency for which high-power RF components were available. An advantage of HF lies in the fact that at these frequencies, ionospheric refraction occurs, extending the potential detection range far beyond the radar line-of-sight. This over-the-horizon (OTH) feature makes HF attractive for surveillance of large areas of the earth's surface, usually, however, at the price of good target resolution. *PCH*

The **very-high-frequency (VHF) region** shares many of the characteristics of HF, including high occupancy (by nonradar communications), potentially high external noise interference, and relative immunity to weather effects. Since the wavelengths at VHF are ten times shorter than HF, high-power transmitters, along with antennas yielding acceptable angular resolution, are more readily achievable in practical sizes for field deployment. Under smooth-surface conditions, VHF radars can take advantage of ground or sea-bounce multipath constructive interference effects to extend the detection range to nearly twice the free-space range, but then must contend with the resulting nulls in coverage due to the accompanying destructive interference effects. VHF is also a region of the spectrum suited to the detection of low-RCS targets, in that in this region, so-called *stealth* RCS reduction techniques are least effective. VHF radars can be relatively inexpensive solutions to wide-area air surveillance problems, and have seen service in this role. In general, however, radar use of this RF region carries with it major compromises in terms of achieving uninterrupted spatial coverage and good target resolution. *PCH*

The **ultra-high-frequency (UHF) region** covers a range from 300 to 1000 MHz, and the wavelengths (1 to 0.1m) make it suitable for use in physically constrained platforms, such as in airborne early warning (AEW) aircraft. Beamwidths can be made reasonably compatible with aircraft vectoring requirements, and radars at these frequencies are little affected by atmospheric attenuation and rain. *PCH*

L-band covers from 1.0 to 2.0 GHz, and its wavelengths make it suitable for higher resolution radars associated with airspace surveillance and en route air traffic control. Weather effects, although more severe than those suffered by radars at lower frequencies, are still of minor consequence for ranges out to 400 km. L-band radars have application in surface-based, seaborne, airborne (AEW), and space radar roles and may lend themselves to uses where 3D, multifunctions (search and track) are required. *PCH*

S-band, 2.0 to 4.0 GHz, holds many advantages for medium-range radar applications. Weather radars at S-band provide accurate data on rainfall rate, and the superior beamwidths achievable with moderate-sized antennas make this frequency band suitable for multifunction radars and specialized tracking/instrumentation radars as well. Most of the airport surveillance radars (ASR) operate at S-band, as do many military search radar whose requirements include accurate target designation to subsidiary radar-directed fire control systems. *PCH*

At **C-band** (4.0 to 8.0 GHz), the effects of atmospheric attenuation and weather become serious impediments to its use for long-range search. C-band represents a middle ground between S-band and X-band and shares the advantages, as well as the disadvantages, of each. Given sufficient power-aperture product and clear-air conditions, C-band can function well in the range instrumentation radar role, and, in addition, is probably the lowest frequency at which a precision multifunction (target acquisition, track, and missile guidance support) can be seriously entertained. *PCH*

The **X-band** extends from 8.0 to 12.5 GHz, making this frequency regime suitable for high-resolution radar applications such as weapon-system fire control (either as separate search and track radars, or in 3D multifunction radars), precision tracking, shipboard navigation, weather avoidance by aircraft, and other miscellaneous short-range roles, such as vehicle speed measurement. The combination of wide available bandwidth and narrow beamwidths in small dimension antennas yield high resolution capabilities in range, doppler, and angle, with manageable hardware configurations. Atmospheric attenuation at X-band can be overcome, if necessary, by a high (but expensive) power-aperture product, but the rain attenuation and backscatter effects are severe, even at moderate rain rates. For this reason, X-band is generally used only over short-to-medium ranges, typically 150 km or less. *PCH*

K_u, K, and K_a Bands. The so-called K-band, defined by a 24-GHz radar developed during World War II, was subdivided into two bands on either side of this frequency after it was recognized that the original frequency lay near a water vapor absorption band. The lower region, K_u-band, is centered approximately at 17 GHz, and the upper region, K_a-band, lies nominally at 35 GHz. As one would expect, radars in these frequencies, especially those at K_a-band, are suitable only for short range radar, and where high angle resolution is required. To date, it has been difficult to generate high power at K_a-band and applications at this frequency are limited, in general, to radars requiring a few hundred watts of average power in a small physical package (e.g., active radar missile seekers and other airborne applications). The use of K_u-band has been limited to special-purpose radars, such as slaved range-only radars and has seen service, in a similar role, in a few airborne intercept (AI) and ground-based antiaircraft fire control radars (AAA). *PCH*

The **millimeter band**, roughly from 40 GHz to 300 GHz, encompasses the V-, W-, and mm-band designations shown in Table F7. Because of the variability of atmospheric absorption characteristics in this region, only a relatively few frequencies are suitable as serious candidates for radar application. One of these lies at 94 GHz, although the attenuation in the vicinity of this frequency is greater than the water-absorption line at the original K-band (24 GHz), which precluded further consideration of the latter's use. Some development has been accomplished in the design of an active radar seeker at this frequency, but there is no trend in this direction. *PCH*

Ref.: Button (1981); Currie (1987).

Summary of frequency by application. Table F8 summarizes the use of the radar frequency spectrum by application. Note that although this chart is, in general, representative of actual radar developments and deployments, notable exceptions do exist. *PCH*

Ref.: Skolnik (1980), p. 7.12; Skolnik (1990), pp.1.13–1.18; Barton (1991), App. K.

Rated frequency is the frequency of oscillator under the normal conditions of its operation in the absence of frequency drift. *AIL*

Ref.: Popov (1980), p. 251.

Table F8
Use of Radar Frequencies by Application

Band	Application
HF	Over-the-horizon radar, combining very long range with low resolution and accuracy
VHF and UHF	Long-range, line-of-sight surveillance with low to medium resolution and accuracy, free of weather effects
L-band	Long-range surveillance with medium resolution, slight weather effects
S-band	Short-range surveillance, long-range tracking with medium accuracy, subject to moderate weather effects
C-band	Short-range surveillance, long-range tracking with high accuracy, subject to increasing weather effects in light to medium rain
X-band	Short-range surveillance in clear weather or light rain; long-range tracking with high accuracy in clear weather, reduced to short range in rain
K_u- and K_a-band	Short-range tracking, used especially when antenna size is very limited and when all-weather operation is not required; wider airborne use at altitudes above most weather
mm-wave bands	Limited to short ranges in atmosphere, very short ranges in rain; used generally for tracking with very small antennas; possible airborne and space use.

The **frequency shift between direct and multipath** components of a target signal is the result of different target velocity components projected on the lines of sight to the radar and to the surface reflection region. Referring to the geometry of Fig. F39, the frequency shift for specular reflection is

$$\Delta f = \frac{v_t}{\lambda}(\cos\theta_t - \cos\theta_r) \approx \frac{v_t}{2\lambda}(\theta_t^2 - \theta_r^2) = \frac{v_t h_t h_r}{\lambda R^2}$$

For a typical subsonic, low-altitude aircraft (v_t = 250 m/s, h_t = 100 m) observed by a surface-based, X-band radar (λ = 0.03m, h_r = 10 m) at R = 10 km, the resulting frequency shift is very small: Δf = 0.083 Hz. *DKB*

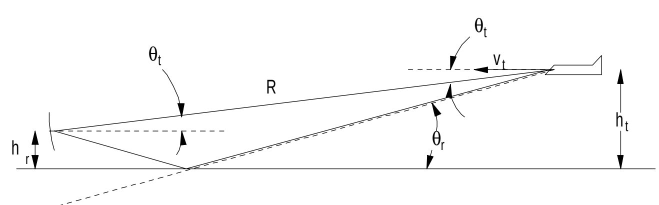


Figure F39 Low-altitude target geometry.

Ref.: Barton (1988), p. 522.

Frequency stability refers to the stability of reference frequency sources within a radar system. For example, in modern radars using the master oscillator power amplifier (MOPA) technique, the master oscillator is the source of RF signals for both the transmitter and receiver. Short-term frequency stability (i.e., from pulse-to-pulse for a pulse radar, or within the signal integration time for a continuous wave (CW) radar), is a principal issue in modern radar design. In CW, MTI, and pulsed doppler radar, short-term frequency variations generate spurious modulation sidebands on main-beam clutter, causing serious degradation of radar performance against targets in clutter. Long-term instability, or frequency drift, can affect the accuracy of target parameter measurements, and, under some conditions, create spurious targets from PRF harmonics, but long-term frequency drift specifications that prevent these problems are easily met for monostatic radars.

Frequency stability requirements for coherent radar are largely driven by the magnitude of the received clutter-power-to-noise spectral density ratio (C/N_0) in which the radar is expected to operate. For a properly designed coherent radar, the clutter level at the output of the doppler filters covering the frequency band of expected targets would be below thermal noise, if all of the radar frequency sources were ideal. Any actual source, such as a master oscillator, will experience short-term frequency fluctuations that give rise to AM and FM noise spectra which are impressed on the received clutter signal. Some components of the resultant clutter-reflected noise spectrum spread into the radar's target doppler filters, reducing the signal-to-interference (S/I) ratio and, under some circumstances, presenting false targets.

For components such as master oscillators, which supply signals to both transmitter and receiver, a mixing process permits cancellation of a large part of the undesired short-term frequency fluctuation effects, especially at short ranges where the clutter return is strongest. This "common mode" cancellation is not available to several other radar components such as the transmitter power amplifier, waveform generator, or any other oscillators in the system not associated with the master oscillator. All such sources must be included in an overall noise error budget for the system. Noise specifications for frequency sources and other noise sources within the radar are generally given in terms of allowable AM and FM noise power density per Hertz relative to carrier power (dBc), as a function of frequency offset from the carrier. *PCH*

Ref.: Skolnik (1970), pp. 19.18–19.22; N. Slawsby, "Frequency Control Requirements of Radar," *Proc. of the 1994 IEEE International Frequency Control Symposium*, 1–3 June 1994.

Video frequency is a term understood to mean the frequencies of the components of the video (envelope-detected or baseband) signal spectrum. The video signal (video pulse) in a radar is shaped at receiver detector output and occupies a band from tenths to several megahertz. *AIL*

FRESNEL-FRAUNHOFER TRANSITION POINT (see **FIELD, antenna**).

FRIIS TRANSMISSION FORMULA. This formula, originally published by Friis in 1946, gives the relationship between transmitted power P_t and receiver power P_r in a one-way, free-space radio link:

$$P_r = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 R^2}$$

where G_t and G_r are the transmitting and receiving antenna gains, λ is the wavelength, and R is the distance between antennas. This formula serves as a basis for deriving the radar range equation. *SAL*

Ref.: Johnson (1984), p. 1.12.

"FRUIT" is the type of interference caused by beacon replies to interrogation asynchronous with the observer's interrogator. The acronym stands for false replies unsynchronized with interrogator transmission. *SAL*

Ref.: Johnston (1979), p. 60; Stevens (1988), p. 288.

FUNCTION, random. A function $f(\vec{q})$ is random when x exhibits a random dependence on its arguments, $\vec{q}$. The vector $\vec{q}$ can represent a set of parameters (e.g., coordinates of the target, x, y, z) or a single parameters (e.g., time t). In the latter case, when $\vec{q} = t$ the random function is called a *random process*. It is a set of random variables with time t running through all real numbers, and in this case the random function $\xi(t)$ is represented by the set of sample functions $x(t)$ with associated probabilities.

The complete probabilistic description of an arbitrary random function requires knowledge of the nth-order probability distribution function, which, for a random process, is defined as the probability that at times $t_1, t_2, \ldots, t_n$ the function $\xi(t)$ will be less than the corresponding levels defined by a sample function, $x(t)$:

$$P\{\xi(t_1) \le x(t_1) = x_1, \ldots, \xi(t_n) \le x(t_n) = x_n\}$$
$$= F_n x_1, \ldots, x_n, t_1, \ldots, t_n$$

If the probability distribution function has a derivative

$$\frac{\partial^n}{\partial(x_1, \ldots, \partial x_n)}(x_1, \ldots, x_n, t_1, \ldots, t_n) = f_n(x_1, \ldots, x_n, t_1, \ldots, t_n)$$

this derivative is called the *probability density function (pdf)*.

In practical cases it is difficult to obtain the nth-order statistics of a random function, and its consideration is limited to first- and second-order statistics, the mean value or mathematical expectation:

$$m_x(t) = \langle x(t) \rangle = \int_{-\infty}^{\infty} x f_1(x, t) dx$$

the variance:

$$\sigma_x^2(t) = \langle [x(t) - m_x(t)]^2 \rangle = \int_{-\infty}^{\infty} [x(t) - m_x(t)]^2 f_1(x, t) dx$$

and the correlation function:

$$K_x(t_1, t_2) = \langle x(t_1) \cdot x(t_2) \rangle = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} x_1 x_2 f_2(x_1, x_2, t_1, t_2) dx_1 dx_2$$

Here, <y> denotes statistical averaging of y.

These definitions show that determining the mean value and variance of a random function requires specification only of the first-order pdf, $f_1(x_1,x_2,t_1,t_2)$. A further simplification can be achieved if one assumes that the random function $\xi(t)$ is a stationary random process. In this case the mean value m_x, and variance, $\sigma_x{}^2$, become consistent values, invariant to the time shift $\tau = |t_1 - t_2|$ and the correlation function depends only on the parameter τ:

$$m_x(t) = m_x; \; \sigma_x^2(t) = \sigma_x^2; \; K_x(t_1, t_2) = K_x(\tau)$$

Since the basic task of any radar is target detection and measurement in a background of random interference (noise, clutter, jamming), the concept of random functions is fundamental to the theory of radar operation. It is widely used in the theory of radar detection, measurement, recognition, signal processing, and so forth. *SAL*

Ref.: Barkat (1991), Ch. 2.

FUSION (see **DATA fusion**).

FUZE, radar. A radar fuze is a noncontact fuse mounted on a missile (bomb, shell) based on compact, simple radar and ensuring the explosion of missile warhead at the required distance from the target, using the information in radar returns. The basic proximity fuze is a CW homodyne device in which commonly a single element is used as both oscillator and mixer-detector. Typically, radar fuzes employ solid-state technology and they have to meet basic requirements of small size, low cost, and reliability under high acceleration. *SAL*

Ref.: Skolnik (1990), p. 14.20.

G

GAIN is "the increase in signal power in transmission from one point to another under stated conditions." Gain is often expressed in decibels,

$$G_{\mathrm{dB}} = 10\log\frac{P_2}{P_1} = 20\log\frac{V_2}{V_1}$$

where (P_2/P_1) is power gain a (V_2/V_1) is voltage gain in dimensionless units. *SAL*

Ref.: Johnston (1979), p. 61.

Automatic gain control (AGC) is a circuit that maintains output signals of a receiver at a given level, without distortion of their shape, regardless of significant changes in amplitude of the input signals. These changes can occur as a result of change in the distance between the radar and the target, or from fading of the signal. The operation of AGC is based on a change in the gain coefficient of the receiver depending on the value of the output signals under the action of a regulating voltage which is generated by the AGC system as a result of amplitude detection of output signals. The voltage of strong signals of powerful interference (instantaneous gain adjustment) may be used as the regulating voltage. In this case, a bias voltage, which moves the operating point to a region where there is no overload of the receiver, is applied practically instantaneously to the controlled stages of the receiver. Such control is practically noninertial.

In terms of speed we distinguish between fast (or low-inertial) and inertial systems of automatic gain control. In fast systems, the time constant is comparable with the duration of the signal, while in inertial ones it significantly exceeds that. Inertial systems of automatic gain control the gain depending on the value of the signals from one selected target through strobing of signals in the receiver. Such systems can be used in fire-control and missile-guidance radars.

Depending on the type of controlled stages, automatic gain control systems are divided into "forward" and "back" systems. In the former, the gain coefficient of the low-frequency amplification stages following the detector in the absence of feedback is controlled. A "back" system provides great depth of adjustment as a result of control of the gain coefficient of preceding stages of intermediate frequency. Drawbacks of such adjustment include some inertia, so some time is required to restore the sensitivity of the receiver after exposure to strong interference. In addition, such control does not protect the receiving circuit from saturation by pulsed interference. For purposes of stabilization of false alarms with an increase in the intensity of noise at the input of the receiver, noise AGC is used. The regulating voltage is formed on the basis of smoothing of the noise voltage in the range at which the appearance of useful signals is eliminated. Such an automatic gain control with feedback is the most widely used in practically all types of radars.

In search radars, gain control may be used independently of the value or presence of received signals. The gain is varied as a function of time (generated in accordance with a specific time program) and assures uniform amplification of signals reflected from targets which are at different distances from the radar. This is called *sensitivity time control*. (See **SENSITIVITY**.) *IAM*

Ref.: Skolnik (1970), p. 5.19; Popov (1980), pp. 14, 70, 222.

diversity gain (see **DIVERSITY**).

Integration gain is that obtained from pulse integration, defined as a ratio of detectability factor of a single pulse to that for each of n pulses. (See **INTEGRATION gain**). *SAL*

Ref.: Barton (1989), p. 71.

Inverse gain refers to a jamming technique providing deception or noise jamming with amplitude modulation in phase opposition to that generated by the scanning radar beam. In the simplest case the modulation is of the on-off type with the off period coinciding with the maximum radar echo signal (Fig. G1). *SAL*

Ref.: Neri (1991), p. 380.

Manual gain control (MGC) is the mode in which the gain of a receiver is controlled by the radar operator to make the dynamic range of the received signals match the dynamic range of the display. *SAL*

Ref.: Currie (1987), p. 489.

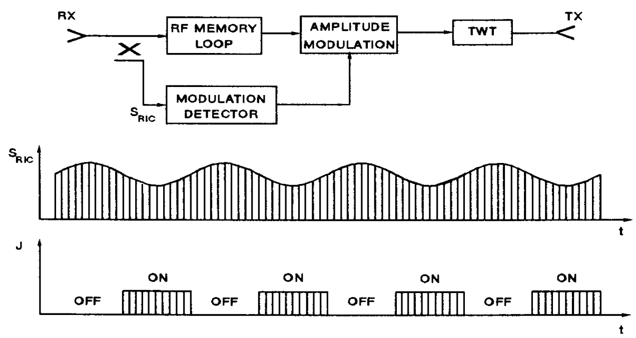


Figure G1 Inverse-gain jamming waveform (from Neri, 1991, Fig. 5.33, p. 381).

Obstacle gain is the increase in received field over an obstructed path resulting from knife-edge diffraction. It is defined as the ratio of the power density beyond the obstacle to the density that would occur in the absence of the obstacle. *DKB*

Ref,: Meeks (1982), p. 34.

swept gain (see **SENSITIVITY time control**).

GARBLE in a secondary radar system is "a term applied to chance overlapping of two replies so that the pulse positions of one reply fall close to the pulse positions of the other reply, thereby making the decoding of reply data prone to error." The situation arising and persisting when two aircraft are within the same 3-km slant range interval from an interrogator and at similar bearings, causing replies to overlap in the receiver is called *synchronous garble*. *SAL*

Ref.: Stevens (1988), pp. 288, 293.

GATE, GATING. A gate is "(1) an interval of time during which some portion of circuit or display is allowed to be operative, or (2) the circuit that provides gating," such as an electronically controlled switch having the capability of passing or inhibiting a signal. The gate is normally controlled as a function of time (see **range gate**) to select signals appearing within a given interval of time delay, but gating is also used to select signals during a given interval in a scan, or in a given angle or doppler frequency interval. (See **angle gate**, **velocity gate**.) *DKB*

Ref.: IEEE (1993), p. 549.

An **angle gate** is a circuit capable of selecting and passing signals arriving from a given angle sector. In a sector-scanning radar, the selection may be based on the timing of the signal relative to the start of the scan. In a tracking radar, the amplitude of the error signal provides the data used in controlling the gate. *DKB*

Ref.: Barton (1988), p. 391.

A **correlation gate** is used in track-while-scan radars to select a specific target signal in range and azimuth after initial detection, providing validation of the target and inputs to the tracking filter from successive scans. The size of this gate must be such as to accommodate prediction errors caused by uncertainties in target velocity, while minimizing the opportunities for passing clutter or echoes from other targets into the tracking channel. *DKB*

Ref.: Blackman (1986), pp. 4–10.

early gate (see **split gate**).

A **guard gate** is a range or velocity gate placed adjacent to the signal gate to detect the approach of an interfering signal. When the interference enters the guard gate, it initiates a change in tracking mode (e.g., entering the coast mode) designed to prevent loss or contamination of the track. This alternative mode is used for a predetermined time or until the interference is detected in a guard gate on the other side of the signal gate. Multiple guard gates may be used to develop information optimizing the alternative tracking and coast modes. *DKB*

late gate (see **split gate**).

A **range gate** is (1) a circuit passing signals during a specific interval of range delay after a transmission, or (2) the signal channel corresponding to that range delay interval. The latter usage is synonymous with the term *range cell*. In digital processing systems, where a short strobe samples the signal in an analog-to-digital converter, the equivalent range gate is the convolution of the strobe width with the impulse response of the prior receiver circuits, which should be approximately matched to the signal waveform. In FMCW doppler systems and pulse-compression systems using the stretch processing technique, the equivalent range gate is a filter channel passing echo signals within a specific delay interval.

In search radar, fixed range gates may cover the entire pulse repetition interval, each passing signals from a given range delay to an integrator or pulsed doppler filter bank. In a tracking radar, a single gate (often divided into a pair of gates for split-gate tracking) often suffice, the single or sum gate passing signals for gain control and angle tracking while the split gate provides a range error channel. The range gate is usually matched to the processed pulse width, and when a simple pulse (without pulse compression) is used the matched filter may be implemented with a wideband IF amplifier followed by such a matched gate and a bandpass (or low-pass) filter.

To reduce cost and complexity, range gates wider than the processed pulse may be used, at the expense of mismatch or collapsing loss. The matching loss is proportional to the square of gate width when a gate wider than the pulse width is used in IF stages preceding a narrowband filter and envelope detector. When such a gate is used following an IF matched filter or at video, a collapsing loss results. *DKB*

Ref.: IEEE (1993), p. 549; Barton (1964), p. 9.

range gate pull-off (see **ECM, range-measurement**).

A **rectangular gate** is used in range or angle tracking when it is unnecessary to provide controlled weighting of signals received from different portions of the gating interval. For example, a rectangular range gate following a wideband IF amplifier provides matching to a rectangular pulse of equal width. When an angle gate is used to select and integrate input pulses for tracking in a sector-scanning radar, its optimum shape is that of the received pulse envelope (the two-way beam shape), but the loss from use of a rectangular gate may not be large enough to warrant the more complicated matched gate shape. *DKB*

A **sampling gate** is a time gate use to select a given group of pulses or signals for subsequent processing. It may be implemented in range-delay time, scan-cycle time, or any other domain in which signals are available for sampling.

A **split gate** is a tracking gate designed to sense the departure of the centroid of a signal pulse from the center of the gate. The two halves of the gate are generated with opposite polarity and followed by an integrating circuit (narrowband or low-pass filter). The portion of the input pulse received during the leading half (early gate) is passed with a positive polarity (or 0 phase), and that received during the lagging half (late gate) is passed with negative polarity (or 180° phase). The integrating circuit, at the end of the split-gate, then has a residual voltage proportional to the position of the signal centroid relative to the center of the gate. This voltage is used as the input to a tracking loop, which holds the gate centered on the target. (See **DISCRIMINATOR, time**).

A split angle gate is used in locating the centroid of the envelope of echo pulses received in a sector-scanning radar, which represents the time at which the beam is pointing at the target. The circuit operates in the same way as the split gate for ranging, except that its input is the envelope of pulses received during the scan, and the duration of the gate is matched to the time required to scan through one beamwidth. Ideally, to minimize effects of thermal noise, the shape of the gate will be matched to the derivative of the two-way beam pattern. To minimize the errors caused by target fluctuation, the gate may be extended toward the skirts of the beam pattern. *DKB*

Barton (1988), pp. 390–397, (1988), pp. 435–437; Skolnik (1990), pp. 18.27–18.30.

A **tracking gate** is a range, angle, or velocity gate used in tracking a particular target, either in a radar dedicated to that target or in a track-while-scan radar.

A **velocity gate** is actually a filter used to track targets in the doppler frequency domain. This filter may be a bandpass filter to select target as inputs to angle tracking or AGC circuits, or a discriminator (analogous to the split gate in range tracking) to develop a doppler loop error signal.

velocity gate pull-off (see **ECM, velocity measurement**).

GENERATOR

boxcar generator (see **CIRCUIT, sample-and-hold**).

A **diffraction radiation generator** is a device producing radiation in the millimeter or submillimeter bands with very low levels of secondary oscillations. The generator contains a diffraction grid over which the electron flux is distributed, and an open cavity. The first spatial harmonic is synchronized with an electron beam. The produced resonance structure is capable of holding only the operating oscillation and maximally suppressing (to a level of –100 to –120 dB) the secondary types. *IAM*

Ref.: Nefedov (1986), p. 157.

A **mark generator** is a device for generating and shaping scale marks on radar displays. The generator is started at the moment of radiation of the transmitted pulse. The mark pulses exist only during the direct track of the beam of the cathode ray tube. The range-mark generator consists usually of a shock-excited oscillator and a diode limiter, which converts the sinusoidal signals into pulses, a differentiating stage that synchronizes the blocking oscillators generating the marks of the different scales (for example 5 km and 10 km range marks), and a mark mixer.

The generation of azimuth marks of a radar with rotating antenna is controlled by a photoelectric modulator located in the pedestal of the antenna system. Such a generator constitutes a circuit for processing the pulses of the photomodulator (change of leading edge, reduction of duration for their conversion into marks of the required parameters). *IAM*

Ref.: Penrose (1959), p. 225; Perevezentsev (1981), pp. 357, 359.

A **pulse generator** is an electronic circuit creating video pulses of various amplitude, length, and relative pulse duration. In terms of operating principle they are divided basically into blocking oscillators (single-stage oscillators with transformer feedback), multivibrators, and flip-flops (two-stage oscillators with capacitive and potentiometric connections respectively between stages).

It may also be an instrument belonging to a group of pulse test oscillators. It is widely used in tuning and testing of radar equipment, for time measurement, for modeling of periodic processes, and so forth. *IAM*

Ref.: Popov (1980), p. 82.

A **signal generator** is a radio-frequency generator in which the frequency, amplitude, and depth of modulation can be changed over broad limits. Signal generators include sweep generators, test generators of pulses of various shapes, random-signal generators, and so forth. Signal generators are often used as simulators of radar signals and for testing and tuning of radar circuits. *IAM*

Ref.: Van Voorhis (1948), p. 278; Gold (1969, 1973), p. 169, in Russian.

A **sweep generator** is a voltage or current generator of a special form intended to produce a time sweep in radar displays. The voltage generators are used in cathode-ray tubes with electrostatic control, and current generators with electromagnetic control. Depending on the type of sweep, sawtooth voltage (current) generators are used to form linear coordinate sweeps, and generators of signals proportional to the sine and cosine of the turning angle are used for a circular sweep (sweep in polar coordinates).

Digital generators are used along with analog sweep generators (see **pulse generator**). The digital circular sweep generator contains a generator of pulses of frequency f, which is started by the synchronizing pulse of the radar station, a circuit for shaping pulses of frequency f cos ϕ, f sin ϕ, a pulse counter, and digital-to-analog converters. The latter shape the signal modulated by sinϕ and cosϕ arriving at the perpendicular coils of the cathode-ray tube. *IAM*

Ref.: Puckle (1944); Druzhinin (1967), p. 249; Finkel'shteyn (1983), p. 524.

A **swept-frequency generator** is a signal generator in which the frequency of oscillations changes periodically within certain limits, which speeds up the reading of frequency characteristics of various targets. A typical frequency range of sweep-frequency generators of devices for measuring the standing-wave ratio and complex parameters of microwave components is 1.5 to 17.0 GHz. The necessary signals are formed through amplitude and frequency modulation of oscillations of the generator, which in microwave measurement devices are backward-wave tubes. *IAM*

Ref.: Van Voorhis (1948), p. 286; Bondarenko (1969), p. 251.

A **synch(ronization)-pulse generator** is a device that generates pulses used for synchronizing operation of individual radar devices. The generator pulses have set high-stable parameters: duration, amplitude, and relative time position. *IAM*

Ref.: Popov (1980), p. 83.

A **voltage generator** is a source of current in the idle mode. In this mode, voltage drop at the internal resistor of the generator is insignificant, and the voltage at the load is constant (depends little on the current and resistance of the load).

Generators of voltages of various shapes are used: sawtooth (linear in individual sections), triangular, and other shapes. Sawtooth voltage generators, for example, are used to produce a time sweep (see **sweep generators**) in cathode-ray tubes with electrostatic deflection and for other purposes. *IAM*

Ref.: Chance (1949); Popov (1980), p. 82; Druzhinin (1967), p. 249.

A **waveform generator** is a device used to generate radar waveforms. In older types of radars it was a low-power analog oscillator generating the waveforms either at the carrier frequency or at intermediate frequency (with subsequent upconversion), and the desired output waveform was shaped and amplified by transmitter circuits. In modern radars digital waveform generators are used. The simplified block diagram of such a unit is shown in Fig. G2. It consists of a digital counter defining moments of time when the waveform is generated, a waveform coefficient storage unit (typically a programmable read-only memory, PROM), and a digital-to-analog converter (DAC). The signal is generated at baseband or at the lowest intermediate frequency, and from the output of the DAC it passes to a sample-and-hold circuit (to remove transients due to the nonzero transition tie of the DAC), to a low-pass filter that smooths (interpolates) the analog signal components between waveform samples, and then through an upconverter to the transmitter driver. The advantage of a digital waveform generator is its high stability with well-defined distortion characteristics and complete repeatability of waveforms, very important in pulse-compression radars. *SAL*

Ref.: Skolnik (1990), p. 10.7.

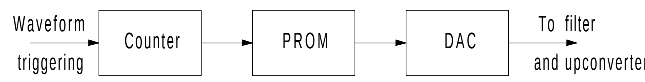


Figure G2 Digital waveform generator.

GHOST. A ghost is "(1) An unwanted signal appearing on the screen of a radar indicator, caused by echoes which experience multiple reflections before reaching the receiver. (2) In passive detection, the intersection points of lines of position which do not represent actual targets but are only crossover points of multiple plotted lines of position from two or more detection stations." *SAL*

Ref.: Johnston (1979), p. 61.

GLINT is "the inherent component of error in measurement of position and/or doppler frequency of a complex target due to interference of the reflections from different elements of the target." Sometimes the peak values of glint can lie beyond the target dimensions in measured coordinates and be the cause of considerable tracking errors in all radar coordinates. (See also **ERROR, radar.**) *SAL*

Ref.: IEEE (1993), p. 559.

A **GONIOMETER** is an electrical device that consists of a transformer with two fixed stator windings and a moving rotor (search) coil. Essentially obsolete, a goniometer is one of the technical implementations of a phase detector.

The goniometer was used in tracking radio range-finders of old models of close-in navigation systems. Data about direction, arriving at the input of the goniometer in the form of amplitude-modulated signals, are converted into an angle of turn of a rotor, which tracks the direction equal to zero on its winding. Signals go to the goniometer input from two mutually perpendicular frame antennas. The rotor, which is in the field of the stator coils, turns to an angle that is equal to the angle in the horizontal plane between the plane of one of the frame antennas and the direction of the source of radiation. *IAM*

Ref.: Gething (1978), p. 5; Sosnovskiy (1990), p. 134.

"GRASS" is the colloquial term for noise as it is seen on some displays (for example, A-display). *SAL*

Ref.: IEEE (1993), p. 563.

GUIDANCE, radar. Radar has been used, both directly and indirectly, for guiding aircraft, antiaircraft artillery (AAA), and missiles of various sorts since the last years of WWII. The term "radar guidance" can have a broad range of application. To the extent that an air traffic controller directs the pilot of an airliner in response to radar tracking data, that aircraft may, in some sense, be construed to be using radar guidance. The modern meaning, however, refers more specifically to the guidance of unmanned vehicles or missiles.

Radar guidance of missiles can be further differentiated on the basis of whether or not the guided missile is equipped with an on-board target sensor or "seeker," in which case it is said to use radar homing guidance. Nonhoming radar guidance techniques (i.e., no onboard seeker), include command guidance and radar beam-riding guidance. In a command guidance system, one or more radars track both the target and the missile under control, and the tracking data are used to compute missile autopilot commands that are transmitted (uplinked) to the missile either continuously or at a high data rate. In a beam-rider guidance scheme, the target is tracked by an external radar and the missile, equipped with its own antenna and receiver system, senses its position relative to the center of the beam. Missile course corrections are made in proportion to the amount of lateral deviation of the missile from beam center, thus the name *beam rider.*

Radar guidance using missileborne seekers can be categorized in terms of the source of the sensed radar energy, and all such seekers are said to use homing guidance. Passive radar seekers, also referred to as *antiradiation homing seekers* sense the energy emitted from a target radar. An active seeker is a self-contained radar that transmits its own radar signal, detects and tracks the target, and homes in on the reflected energy from the target. In a semiactive guidance system, an external radar, the illuminator radar, tracks and illuminates the target. Part of the target-reflected energy is received by a passive antenna and receiver system on board the missile, which uses the signal for homing guidance to intercept.

Modern medium-to-long-range defensive missiles (e.g., 50 to 100 km), may use a combination of guidance modes. For example, a typical multimode guided missile might employ inertial or command-inertial guidance for the initial and midcourse phases of flight, with a transition to semiactive or active radar homing for the terminal phase. Such a combination takes advantage of the best features of each mode. Command or command-inertial guidance accuracy deteriorates with range (or flight time), while radar homing accuracy improves as the missile-to-target range decreases. (See **RADAR, missile guidance.**) *PCH*

Ref.: Skolnik (1990), Ch. 19; James (1986).

A **GULL** is "a floating radar reflector used to simulate surface targets for deceptive purposes." *SAL*

Ref.: Johnston (1979), p. 61.

GUNN EFFECT. The Gunn effect, experimentally discovered in 1963 by J. Gunn, entails the onset of microwave oscillations in a gallium arsenide (GaAs) conductor in the presence of a high field intensity (on the order of 3 kW/cm). It is explained by the effect of the field on the mobility of the carriers, conditioned by the presence of two energetic zones of electronic conductivity. Under the effect of the critical values of the field intensity the electrons begin to move to the upper zone, where their mobility is decreased as a result of the strong interaction with the field of the GaAs crystal lattice. When the field intensity is less than critical but close to it, as a result of the heterogeneity of the material of the semiconductor, a separate area of electrons with less mobility occurs. This is a so-called domain, a thin layer of negative space charge, slowly moving toward the positive electrode. In the remaining part of the semiconductor, the intensity remains less than critical and other domains are not formed. The domains occur in series in the vicinity of the cathode and determine the current pulses in the external circuit with a period approximately equal to the transit time of the electrons in the sample. The Gunn effect serves as the basis of the Gunn diode. (See **DIODE.**) *IAM*

Ref.: Fink (1982), p. 9.72; Gassanov (1988), p. 186.

GYROTRON. A gyrotron is a microwave device using a relativistic electron beam and converting constant electron energy to microwave energies in an intense electromagnetic field. It has good prospects for production; at millimeter-wave frequencies, where peak powers much higher than those can be obtained with conventional millimeter-wave tubes using previous techniques. Achievable and predicted gyrotron power levels over a range of frequencies are given in Table G1 and Fig. G3. There are different types of gyrotron amplifiers configurations, the main being the gyromonotron, gyroklystron, gyro-TWT, gyro-BWO, and gyrotwystron based on the corresponding parent configurations, which are shown in Fig. G4. Other terms used for gyrotron are cyclotron resonance maser or electron-cyclotron maser. *SAL*

Ref.: Ewell (1981), pp. 76–79; Currie (1987), pp. 466–470; Brookner (1988), p. 334; Flyagin, V. A., *IEEE Trans.* ***MTT-25***, No 6, 1977.

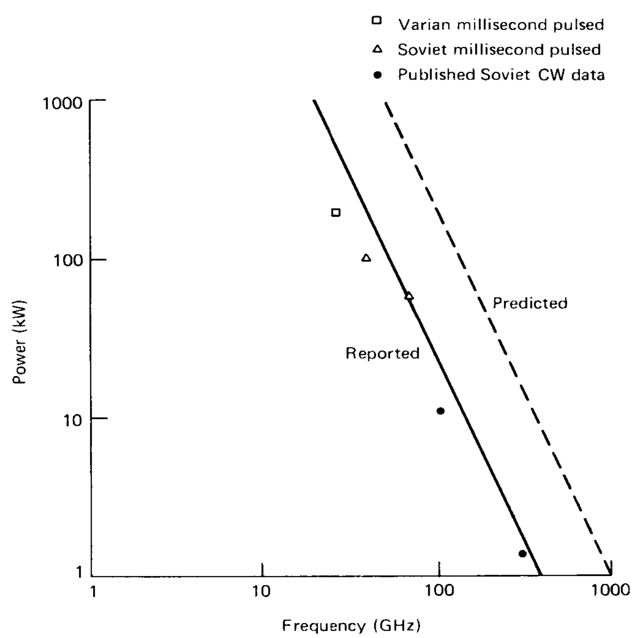


Figure G3 Gyrotron state-of-the-art capabilities (from Ewell, 1981, Fig. 2-43, p. 77).

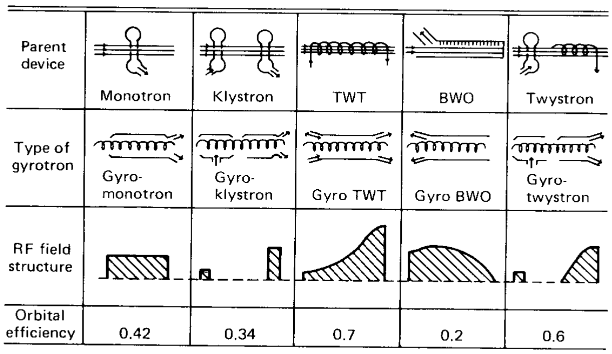


Figure G4 Types of gyrotron amplifier configurations (from Ewell, 1981, Fig. 2-44, p. 79, reprinted by permission of McGraw-Hill).

Table G1
Peak Power Levels from Cyclotron Masers Driven by Intense Relativistic Electron Beams

Wavelength (cm)	Peak microwave power (MW)	Accelerating voltage (MV)	Diode current (kA)
4.0	900	3.3	80
2.0	350	2.6	40
0.8	8	0.6	15
0.4	2	0.6	15

(from Ewell, 1981, p. 77, reprinted by permission of McGraw-Hill).

H

HAMMING WINDOW (see **WEIGHTING**).

HEIGHT FINDER. Height finders are radars designed to measure the elevation angle of targets in a surveillance system, permitting target altitude to be calculated from measured range. The methods by which elevation angle and hence altitude is determined include:

(1) Assignment of a specialized radar that performs sector scan in elevation for measurement in that coordinate, on targets designated by a 2D search radar.

(2) Search with a scanning-beam 3D radar, in which a narrow beam is scanned over a raster covering both azimuth and elevation and providing measurement of both angles, along with range, on detected targets.

(3) Search with a stacked-beam 3D radar, in which multiple beams cover the elevation sector as the antenna scans in azimuth, providing monopulse measurement in elevation.

(4) Measurement of multipath time delay on targets detected in a 2D search radar, such that target altitude may be calculated from known target range, radar antenna altitude, and multipath delay.

(5) Measurement of the ranges at which a target passes through multipath lobes of a 2D search radar antenna pattern, from which a constant target altitude may be calculated.

(6) Measurement of relative amplitude or phase of target echoes in two antennas displaced in altitude, leading to a monopulse estimate of elevation angle.

Some types of height finders are described below. (See also **RADAR, '3D.**) *DKB*

The **nodding-beam height finder** is the specialized radar (1) in the basic height-finding methods. A 2D search radar maintains surveillance over the volume of interest, detecting targets and measuring their ranges and azimuth angles. Upon detection and establishment of a track on a new target, a request is sent to the height finder for elevation measurement. The height finder slews its antenna to the azimuth designated by the search radar and performs a scan over an elevation sector appropriate to the range designated by the 2D radar. Target echoes are displayed on a range-height indicator (RHI), and elevation angle θ_t is measured by an operator or an angle-gate circuit. Range R is also measured to an accuracy better than that provided by the search radar. The target height above the horizontal plane at the radar site is then calculated as

$$h_t = R \sin \theta_t$$

This height is corrected as necessary for site altitude, atmospheric refraction, and curvature of Earth to give target altitude above sea level, as needed for ground-controlled intercept of the target by a fighter aircraft. A typical nodding-beam height finder is shown in Fig. H1. The antenna is elongated in the vertical direction to provide a narrow elevation beam for accurate measurement, while the azimuth beamwidth is wide enough to accommodate errors in designation from the search radar. *DKB*

Ref.: Skolnik (1990), p. 20.3.

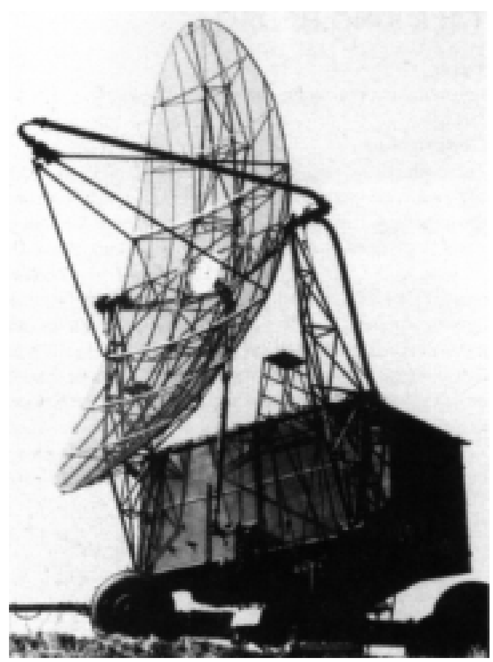


Figure H1 A Russian S-band nodding height-finder radar (from *Jane's Radar and Electronic Warfare Systems*, 1993-94).

A **squinted-beam height finder** uses a modified monopulse beam pair, pointed far enough above the horizon to minimize multipath reflection errors. The technique is similar to using off-axis monopulse tracking and offers reduction but not elimination of the troublesome multipath errors on low-elevation targets. Figure H2 compares conventional monopulse and squinted-beam antenna patterns. *DKB*

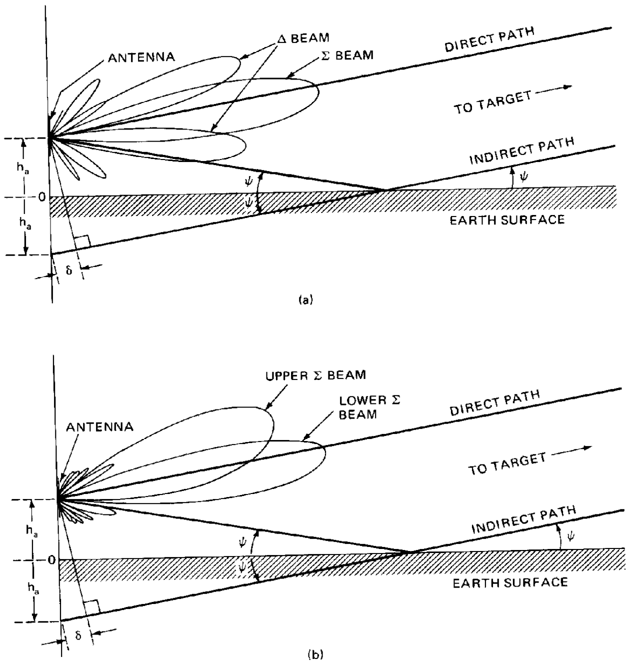


Figure H2 Multipath effects on monopulse and squinted-beam height finders: (a) conventional Σ,Δ monopulse, (b) squinted-beam low-angle technique (from Skolnik, 1990, Fig. 20.12, p. 20.37, reprinted by permission of McGraw-Hill).

Time-difference height finding uses the multipath propagation phenomenon to obtain additional target data needed for estimation of altitude. The time delay between the direct path and the specularly reflected multipath, over a flat earth, is given by

$$\delta_0 = R\left[\frac{\cos\theta_t}{\cos\psi} - 1\right] \approx \frac{2h_r h_t}{R}$$

where R is target range, θ_t is elevation angle, ψ is grazing angle at the surface, h_r is radar antenna altitude, and h_t is target altitude. If the range, radar altitude, and one-way (receiving) multipath delay δ_0/c are known, target altitude can be calculated. The calculation is more complicated over a spherical earth, but it can be performed easily with modern computers.

The multipath delay is very small for surface-based radars, so the technique is restricted in practice to airborne radars and those on high mountain sites, where the delay commonly exceeds the processed pulse width. When the surface is too rough to support specular reflection, a spread multipath return is received, but the delay to its leading edge is still useful for target height measurement. *DKB*

Ref.: Long (1992), p. 349.

A **V-beam height finder**, depicted in Fig. H3, uses two broad radar beams from a single radar scan in azimuth. One of the beams is tilted in the elevation plane so as to produce an output whose delay, in the azimuth scan cycle, is proportional to elevation angle.

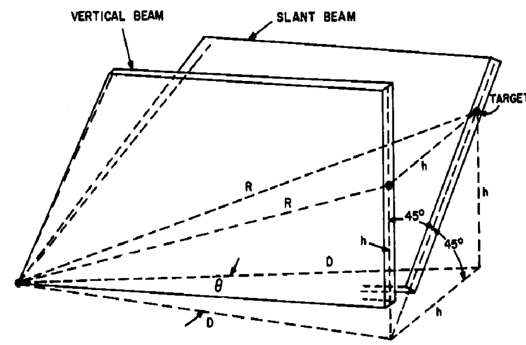


Figure H3 V-beam geometry (from Fink, 1989, p. 25.49, reprinted by permission of McGraw-Hill).

The elevation error is proportional to the difference between the two azimuth readings multiplied by the cotangent of the tilt angle. (See also **PATTERN, V-beam**).*PCH*

Ref.: Fink (1989), p. 25.49.

HETERODYNING is the process of the conversion of modulated radio-frequency oscillations into modulated oscillations of a lower intermediate frequency to make subsequent amplification more efficient. *SAL*

Ref.: Terman (1955), p. 568; Popov (1980), p. 84.

A **HIT** is “a target echo from one single pulse.”

Ref.: IEEE (1990), p. 18.

HOLE, radar. A radar hole is region with reduced coverage, caused by the extension of the radar range in other directions due to the effect of ducting. For example, air targets above a surface duct, within which the radar range against surface targets is enhanced, might be missed, though they would be normally detected when there is no ducting. *SAL*

Ref.: Skolnik (1980), p. 451.

HOLOGRAM, HOLOGRAPHY, microwave. Radio-frequency (RF) holography is a method of restoration of the wave front of a radio wave in which the indicator, the radio hologram, registers information both about the amplitude and about the phase of the field dispersed by the object. As a result of subsequent illumination of the radio hologram, the restoring wave forms an image of the object. The processes of formation and restoration of images constitute virtual forward and reverse integral Fourier and Fresnel transforms depending on the curvature of the front of the object and reference waves.

An RF hologram is a fixed picture of interference between the field scattered by the object (object wave) and a coherent reference wave. From the picture received, the image of the object is restored by the methods of RF holography. RF holograms are formed by two basic methods. The first is associated with the creation of a real aperture (quasioptical RF hologram), while the second presupposes the use of a synthesized aperture (synthesized RF hologram). In both cases, the reference wave is created artificially. It is introduced to the receive circuit from the generator of the transmitted signal. Combined methods are also used. Thus in holographic earth-scanning radars they use a combination of synthesis along the track line and a real aperture (in the form of a one-dimensional antenna array) in the transverse direction. The hologram is represented in the form of the sum of the data components and the background. The data components are the product of the integral Fourier transform or Fresnel transform of the scattering function of the object and the function describing the field created by the reference source.

The structure of the interference picture depends on the curvature of the leading edge of the object and reference waves and is described in the form of a Fourier, Fresnel, or Fraunhofer radio hologram. Based on the type of registration, one distinguishes between quadratic and multiplicative radio Algerians obtained using a quadratic detector or a decoder-multiplier, respectively.

The basic characteristic of the RF hologram is its resolution, which is defined, for wavelength λ, distance r to the object, and size d of the hologram, in a plane parallel to the hologram as

$$\delta l_p \geq \frac{\lambda r}{d}$$

and in the transverse plane as

$$\delta l_t \geq \frac{2\lambda r^2}{d^2}$$

A single image is restored from a multiplicative RF hologram, while when a quadratic RF hologram is used, an interfering background also arises, in addition to the image. Restoration of images is possible optically or digitally. In the first case, the same circuits are used for imaging in visible light as are used in optical holography. Digital processing makes it possible to raise the speed of measurement and the dynamic range of the measurement equipment, and to optimize the parameters of image reconstruction. In the optical method of restoration, a visible image of the radar object is obtained. Due to the difference in the frequency of light and the radio frequency used in forming the radio hologram, there is a change in the dimensions of the image compared with the size of the object. The scale of the image depends on the type of hologram (Fourier, Fresnel, or Fraunhofer radio hologram) and the curvature of the front of the restoring wave and may be increased or reduced. When restoring the images obtained, for example, using antenna arrays, a large number of images arise. Exceptions to the overlay of images are achieved through selection of a specific period of quantization.

Radio holographic methods are widely used in radar imaging, terrain mapping, discrimination and identification, and radio astronomy for precise determination of the coordinates of celestial objects. *IAM*

Ref.: Safronov (1973), pp. 45, 166; Bakhrakh (1980); Tuchkov (1985), p. 130.

A **complex hologram** is a form of recording of a radio hologram when it is represented by a complex function of field coordinates. The complex form is convenient in an analytical description of the process of image reconstruction. Single- and two-dimensional practical registration circuits correspond to a complex radio hologram. In the first case, the amplitude and phase of the result field are registered; in the second, the quadratic components of the complex radio hologram. *IAM*

Ref.: Tuchkov (1985), p. 132.

A **Fourier(-transform) hologram** is one that registers the interference picture formed during interaction of an object wave with a spherical front in the plan of the hologram with a spherical reference wave having a radius of curvature equal to the mean radius of curvature of the object wave.

Fourier radio holograms are the sum of the products of complexly linked Fourier transforms of the scattering function of an object and reference waves. A reverse Fourier transform of the function describing the radio hologram is used for reconstruction of the image.

The Fourier radio hologram in particular is implemented in synthesized radio holograms by turning of the object about the center of mass with a fixed transmitting and receiving system. *IAM*

Ref.: Safronov (1973), p. 42; Tuchkov (1985), p. 133.

A **Fraunhofer (diffraction) hologram** registers the interference picture from the interaction of planar or spherical reference waves with planar waves from the object. With a planar reference wave, a Fraunhofer radio hologram has the same mathematical structure as the Fourier hologram. In the case of a spherical reference wave, the Fraunhofer hologram is the sum of products of the complexly-linked Fourier transforms of the scattering function of the object and the Fresnel transform of the reference wave. *IAM*

Ref.: Safronov (1973), p. 41.

A **Fresnel(-diffraction) hologram** registers the interference picture formed during interaction of planar or spherical reference waves with an object wave having a spherical phase front. The Fresnel radio hologram is the sum of complexly linked products of the Fresnel transform of a scattering function of an object and a Fourier transform (Fresnel) of a planar (spherical) reference wave.

In practice, Fresnel radio holograms formed using reference oscillations with planar phase fronts are most widely used. Restoration of images by such radio hologram is done by a wave with a planar phase front, parallel to the plane of the radio hologram. *IAM*

Ref.: Safronov (1973), p. 56.

A **mosaic hologram** is obtained through synthesis of individual radio holograms formed under various conditions at various moments of time in the same or different points in space.

This type of hologram is used for increasing the contrast and improving the quality of restored images and also to increase the angle of view when inspecting optical holograms transformed from radio holograms. *IAM*

Ref.: Safronov (1973), p. 204.

A **multiplicative hologram** is registered using detection of the product of the object signal U_{ob} and reference signal U_{ref} waves at a high or intermediate frequency:

$$h\,(x,y) = \mathrm{Re}[\; U_{ob}(x,y)\; U_{ref}\,(x,y)\;]$$

where x, y are the spatial coordinates, and Re(x) is the real part of the quantity x.

The multiplicative radio hologram, like the quadratic hologram, can be registered in complex form. The multiplicative radio hologram does not have an optical analog. *IAM*

Ref.: Tuchkov (1985), p. 131.

A **one-dimensional (1D) hologram** is one in which the values of the scattering function of an object are re-created from one coordinate, and from the other its integral transform. An image received by a radar with a linear antenna and transmitted pulses that are short compared with the length of the target can serve as an example of a one-dimensional radio hologram. The values of the scattering function are restored along the line of sight in the form of amplitudes of the interference relief of the radio hologram.

During restoration of images from one-dimensional radio holograms, special circuits are used in which an integral transform is taken in accordance with one coordinate, and the scattering function of the object is reproduced in accordance with the other. *IAM*

Ref.: Safronov (1973), pp. 174, 189.

A **planar hologram** is one in which the registered interference relief is an integral transform (Fourier or Fresnel) in two coordinates.

Planar radio holograms are constructed on the basis of a real aperture in two coordinates (quasioptical hologram), on the basis of a synthesized aperture in two coordinates (using small antennas and a “point” aperture, or antennas with a linear aperture and range synthesis through linear frequency modulation of the signal), and also through a combination of a real and synthesized aperture in different coordinates. *IAM*

Ref.: Safronov (1973), p. 173.

A **quadratic hologram** is registered using quadratic detection of the subject of the object U_{ob} and reference waves U_{op} at a RF or intermediate frequency:

$$h\,(x,y) = |\; U_{ob}\,(x,y) + U_{op}\,(x,y)\; |^2$$

where x, y are the coordinates of the antenna aperture.

Recording of the quadratic hologram is done in quadrature form or in the form of amplitude and phase components. Quadratic radio holograms are the analog to optical holograms. *IAM*

Ref.: Tuchkov (1985), p. 131.

A **quasioptical radio hologram** is formed using antenna arrays or a system of separated antennas. In contrast to synthesized holograms, a quasioptical radio hologram uses a real aperture, the field is measured at discrete point, and the radio hologram is recorded through commutation or parallel processing of signals at the output of the receiver elements. The properties of quasioptical radio holograms are similar to the properties of optical holograms, which possess similar ratios of apertures to lengths of the illuminating waves.

High-quality images of objects with a number of resolved elements on the order of a hundred using an antenna array may be obtained at ranges which in the best case amount to tens of linear dimensions of the object. For this reason such radio holograms are used effectively in close-in radars.

Coherent radio holograms using a system of antennas separated by a distance on the order of tens or hundreds of kilometers makes it possible to produce quality images of aircraft at ranges of 100 to 1,000 km, respectively. However the realization of such systems is associated with great technical difficulties in supporting the coherent functioning of many receiver elements in large areas. *IAM*

Ref.: Safronov (1973), p. 218.

A **synthesized [synthetic] hologram** is one formed by registration of a field reflected from an object at various points of space successively in time using one receiving antenna. The synthesis can take place in one or several coordinates, in two in the picture plane (perpendicular line of sight) and a third along the line of sight. In synthesis in the picture plane, any relative motion of the object, the source of radiation, and the receiver in accordance with a known law may be used. The resolution of radio holograms in this case is determined by the size of the synthesized aperture (length of synthesis) (see **hologram**). In synthesis along the line of sight, wideband frequency-modulated waveforms are used, and the resolution δl, is determined by the width of the spectrum ΔF according to $\delta l = \pi c/\Delta F$, where c is the speed of light.

Synthesized radio holograms are widely used in radars mounted on moving objects and having antennas of limited size, on ground radars for raising the accuracy of coordinate determination and receipt of target images, and also for measurement of local characteristics of scattering of radar targets in anechoic chambers. *IAM*

Ref.: Safronov (1973), p. 172; Tuchkov (1985), p. 130.

A **three-dimensional (3D) hologram** is one that makes it possible to form a 3D image of an object. The three-dimensional radio hologram can be realized by three methods: (1) through successive arrangement at specific distances from one another of planar hologram; (2) through formation of a radio hologram in 3D structures which permit fixation of the changes in parameters of the electromagnetic field in the three dimensions; and (3) by synthesis from linear and point elements or their combinations.

As a rule, radio holograms are formed in digital form using algorithms of fast Fourier transforms. *IAM*

Ref.: Safronov (1973), p. 204.

two-dimensional (2D) hologram (see **planar hologram**).

HOMING, radar. Use by a seeker-equipped missile of the radar energy received from a target for missile guidance is called *radar homing*. The energy source may originate at the target itself (e.g., from an airborne radar), in which case the guidance is called *passive radar homing*. If the missile seeker is the energy source, i.e., an on-board radar, the guidance is referred to as active radar homing, and if the energy source is an external radar, the guidance mode is called *semiactive radar homing*. (See **GUIDANCE, radar**; **RADAR, missile guidance**; **SEEKER, radar**.) *PCH*

Home-on-jam (HOJ) refers to a missile radar homing seeker guidance mode in which the seeker tracks, and the missile guides on, an active jammer signal rather than the target "skin" return. Radar-guided air-defense missiles, both surface-to-air (SAM) and air-to-air (AAM), typically incorporate a HOJ mode as a guidance option, which is primarily intended to defeat self-screening noise jammers (SSJs). HOJ may sometimes be effective against poorly implemented deceptive SSJs as well. The decision to transition from skin track to HOJ may be made by an external tracking radar, or it may be decided autonomously on board the missile. If the decision is made without target range information, however, the missile is subject to capture by an out-of-range jammer. (see **RADAR, missile guidance**). *PCH*

HOMODYNE (see **RECEPTION, homodyne**).

HORIZON, radar. The radar horizon is the imaginary line beyond which targets located below some height h_t cannot be detected because the radio waves in free space, like optical rays, are straight lines (Fig. H4). From the figure, it is easy to compute the distance R_{ht} from the target to the radar horizon

$$R_h = \sqrt{(ka + h_t)^2 - (ka)^2} \approx \sqrt{2kah_t}$$

where ka is the effective earth radius. From a radar site at altitude h_r above the earth, there will be an additional range R_{hr} between the radar and the horizon, and the total horizon range will be.

$$R_h = R_{ht} + R_{hr} \approx \sqrt{2ka(h_t + h_r)}$$

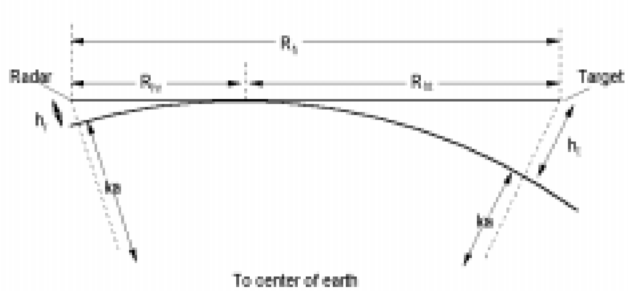


Figure H4 Radar horizon.

Passing through the real atmosphere, in which varying index of refraction typically decreases with height, radar rays curve downward and the radar horizon is slightly beyond the optical horizon. This effect is approximated by using, in the equation, an effective earth's radius ka equal to 4/3 times the actual radius, or 8.5×10^6m instead of 6.5×10^6m. The resulting calculations are adequate for radar and targets at low and medium altitudes. For high-altitude radar or targets, more accurate results can be obtained using range-height-angle charts plotted according to methods originated by Blake. (See **CHART, range-height-angle.**)

To observe the targets lying far behind the radar horizon over-the horizon radars can be used. *SAL*

Ref.: Skolnik (1970), pp. 2.47, 2.48.

clutter horizon (see **CLUTTER**).

HUYGENS SOURCE. The Huygens source is an elementary source of electric and magnetic current used to construct field distribution in aperture antennas. *SAL*

Ref.: Fink (1975), p. 18.5.

The **Huygens-Fresnel principle** is the set of physical assumptions for the approximate decision for a set of diffraction problems. According to this principle, the wave radiated by some source can be represented as an arbitrary point as a superposition of coherent secondary waves of imaginary sources. These sources are considered to be continuously distributed along the auxiliary arbitrary surface surrounding the source. This principle, initially devised in optics theory, is widely used in the theory of aperture antennas. *IAM*

Ref.: Silver (1949), p. 108; Kobak (1975), p. 90.

HYBRID (junction), microwave. A microwave hybrid junction is "a waveguide or transmission line arrangement with four ports, which when the ports have reflectionless terminations has the property that energy entering at one port is transferred (usually equally) to two of the remaining three ports." Actually it is a simple power divider (or adder) of one channel into two others. Hybrids are used in various transmission lines: coaxial, waveguide, etc. A coaxial hybrid is usually a U-shaped symmetrical connection of three coaxial lines. If two arms of this connection are loaded to resistances equal to the wave resistance, then the third arm, the supply arm, is mismatched, with a standing-wave ratio of 0.5.

Usually the hybrids provide equal power division. Like any nonuniformity, the branching introduces additional reactivity of a volume or inductance nature depending on the type of hybrid and its parameters. Compensation of reflections from hybrids, arising due to the connection of the arms and due to the nonuniformity, which is the branching, is provided by four-wave transformers and short-circuited loops for the coaxial hybrid, and inductive irises and stepped transformers for the waveguide hybrid.

At microwave frequencies, waveguide hybrids are most common. They are divided on the basis of shape into T-shaped (magic tee) or Y-shaped hybrids. The latter is distinguished by the somewhat greater passband and the capability of changing the division factor with a replaceable metal insert inside the waveguide.

Hybrid tees are used with a branch in the H-plane (from the thin wall) and in the E-plane (from the wide wall). When there are identical distances from the axis of the splitter, the power arriving from the input arm is divided equally between these loads. The supply arm is matched then. An E-plane hybrid is distinguished by the fact that the intensity vectors of the electrical field in the output sections, which are at identical distances from the splitting axis, are opposite.

The H-branch can pass more power, while the E-branch changes the reactance less in the frequency band, i.e. it has a somewhat larger passband. Hybrid tees are narrow-band elements. *IAM*

Ref.: IEEE (1993), p. 610; Lavrov (1974), p. 348; Gardiol (1984), p. 282.

I

IDENTIFICATION, radar (see **TARGET RECOGNITION AND IDENTIFICATION**).

ILLUMINATION (see **APERTURE illumination**).

ILLUMINATOR. An illuminator is "a system designed to impose electromagnetic radiation on a designated target so that the reflections can be used by another sensor, typically for purposes of homing." It is often the tracking radar in an air defense system that "illuminates" a target, thus allowing a radar-guided semiactive missile to use the target-reflected energy as the source for terminal homing guidance. Illuminator radars are either narrow-beam, single-target tracking radars, or are systems that transmit the energy and are slaved to another radar that tracks the target. (See also **RADAR, continuous-wave**; **SEEKER**). *PCH*

Ref.: IEEE (1993), p. 616.

IMAGE, IMAGING, radar. A radar image is the spatial distribution of the scattering sources of an object, obtained as a result of analysis of the electromagnetic field scattered by it. Such characteristics may include for example the reflective capabilities in a given frequency range, the geometric shape of the object, the dielectric constant of the surface material, and so forth.

The radar image may be constructed by various methods. When the dimensions of the object significantly exceed the size of the resolving element of the radar, the image is formed through serial or parallel viewing of resolution cells of the radar (usually parallel scanning for range and serial scanning for angles). The use of aperture synthesis methods (or inverse synthesis) makes it possible to significantly increase the quality of the image by processing the coherently summed signals accumulated during observation of the object moving relative to the radar. This is usually implemented on the basis of doppler analysis of signals, the method of radar holograms, which ensure that the spatial amplitude-phase distribution of the reflectivity of the target is obtained. When the signal bandwidth is broadened to improve range resolution, the elements of the radar echo from the target may be compared with the elements of the geometrical structure of the target. This provides a range image using a wideband waveform and produces a geometric target image, its profile function (cross section vs. range) depending on the use of ultrawideband signals.

The radar image may be observed visually on the display screen (radar imaging) or formed and processed without visualization. The production of high-information images demands the use of a high-speed components and computers.

When the aspect of the object is fixed, the received signal is virtually a projection of the complete image of the object, formed by the methods of microwave tomography.

The formation of radar images is widely used in radar target recognition and mapping of terrain and extraterrestrial objects. (See also **HOLOGRAM**; **TOMOGRAPHY, microwave.**) *IAM*

Ref.: Boerner (1985), part 1; Andreyev, G. A., *Zarubezhnaya Radioelektronika*, no. 6, 1989, in Russian; Rihaczek (1996); Steinberg (1991), Mensa (1991); Trevett (1986).

Angle-angle imaging is imaging based on the properties of radar angle resolution. To form such an image a radar has to have very good angle resolving capability, so practically only millimeter-wave radars and especially laser radars are useful to create angle-angle images. In laser radar. the receiver aperture resolving capability is used to image the target backscattered illumination onto the receiver detector, where the optical photon energy is then converted to electrons and processed to provide an image of the target field. To optimize the image, the illumination format (pencil, floodlight, etc.) is provided by the transmitter aperture. Such images can be used in discrimination of large targets at short ranges (planes, ships, etc.). One-dimensional angle images are also used, and combined images in angle-velocity and angle-range coordinates. *SAL, IAM*

Ref.: Bernard, D. S., *Proc. IEEE* ***77***, no. 5, 1989; Jelalian (1992), p. 48.

Conversion of a radar image to a television image is done to raise the brightness of the image observed on the television screen compared with the image on the cathode-ray tube of the radar display. This is done through repeated registration of the signal recorded in one scanning period of the radar, and display of the regenerated images with a high repetition frequency on television indicators. The increase in the apparent average brightness is estimated by the ratio of the period of the television frame sweep to the period of scan of the radar, which is on the order of 500. The conversion of radar signals to television signals is done using storage cathode-ray-tubes or graphecons. *IAM*

Ref.: Poole (1966), p. 223; Finkel'shteyn (1983), p. 514.

Image correction is applied to a radar image of an surface-mapping radar when there are distortions in the configuration of area and elongated objects and the distances between them. The distortions are usually due to disturbance of scales of the image along and across the line of movement occurring because of instability of flight trajectory and failure to correct for the difference between the slant range and the horizontal range due to the sphericity of the wave front.

The basic methods of image correction are allowance for data regarding trajectory deviations when recording the image (control of recording speed and registration delay), stabilization of the antenna in space, and electrical control of the antenna beam.

Image correction is necessary both in conventional radars with a conventional antenna, and in radars with a synthesized aperture, since in the latter case for the above reasons the dynamic range of output signals is reduced, along with resolution; and with angular oscillations in the radar platform, also the image contrast. *IAM*

Ref.: Kondratenkov (1983), p. 113; Curlander (1991), Ch. 8.

Image decoding is the process of detection, discrimination, and determination of location of various objects, and also determination of the nature of the terrain and its elements from their radar image. Detection and discrimination of objects is based on analysis of the tone, shape, and size of the radar image of the object, the shape of its shadow, and other features. The coordinates of objects are determined by various methods from the registered coordinates of the radar platform, and the known size of the image with its scale marks, and also by the methods of topographic survey. The latter method is based on measurement of object coordinates of relatively known terrain elements in the image, whose coordinates are determined from a topographic map. The method of reference to a topographic map has great precision, since it is not associated with the errors of the navigational system of the radar platform.

The basic problem of automatic decoding is the processing of object discrimination. For this reason, it is usually limited to automation of certain operations (processing of a large number of images, search for frames with given objects, large changes in density, or returns from moving objects, etc.). *IAM*

Ref.: Kondratenkov (1983), p. 133; Curlander (1991), p. 412.

Focused-beam imaging is three-dimensional imaging performed using a radar with a focused antenna. Such imaging can be used, for example, in high-resolution RCS measurement. In Fig. I1, a 3D image received with a focused Cassegrainian antenna is shown. In this case, the antenna beam was raster-scanned across the target at different elevation angles to develop the image, the range dimension resolution was about 30 cm, and a spot less than 30 cm in diameter was developed at a range about 75m. *SAL*

Ref.: Currie (1989), p. 396.

holographic imaging (see **HOLOGRAM**).

Imaging, with monopulse radar, is based on wideband monopulse radar processing that makes it possible to measure the position of an isolated point target in two orthogonal dimensions of cross-range. The general process of the imaging with monopulse radar is illustrated in Fig. I2. Differential error signals are produced in the azimuth and elevation channels of a monopulse radar, and orthogonal cross-range dimensions are obtained from these error signals. To resolve targets in slant range, pulse-compression or stepped-frequency waveforms can be used. Three channels are required or 3D monopulse imaging: a sum channel and two difference channels. *SAL*

Ref.: Wehner (1987), pp. 341–369; Currie (1989), p. 397.

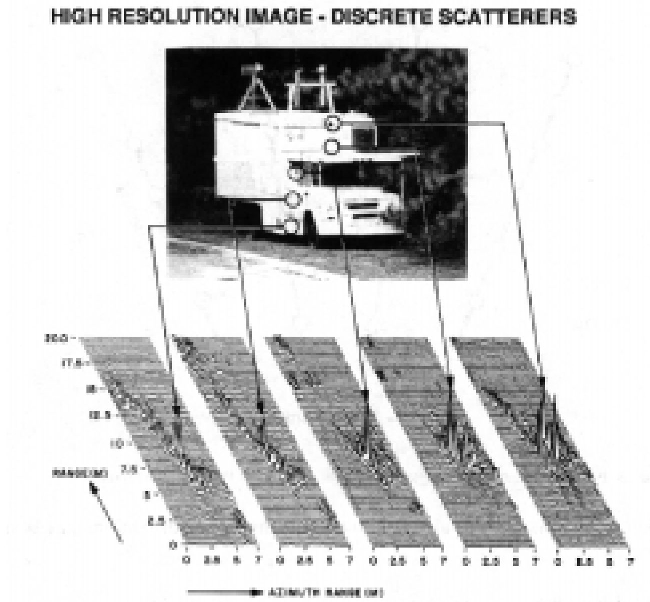


Figure I1 Image created using focused beam antenna (from Currie, 1989, Fig. 10.16, p. 398).

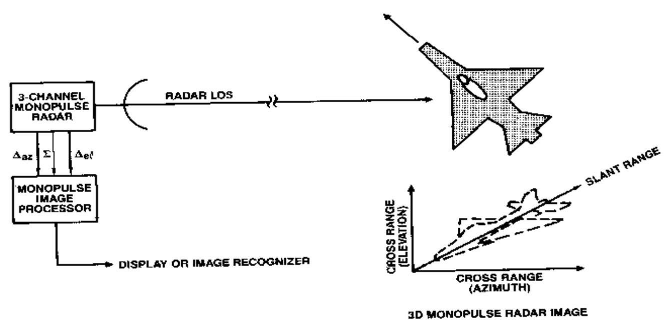


Figure I2 Three-dimensional imaging with monopulse radar (from Wehner, 1987, Fig. 8.2, p. 343).

Range imaging uses the distribution of target scattering sources along one coordinate: range. The basic type of such an image is distribution of amplitudes or RCS of the target in range that is sometimes termed a *target range profile*. To obtain it, pulse-compression waveforms are used to increase the range resolution. For target dimensions that significantly exceed the wavelength, the range profile is represented in the form of signal amplitudes reflected by individual illuminated points of the target. When ultrawideband signals are used, it is possible to obtain an image in the form of a profiled target function that characterizes the distribution of area of the target section along the radar beam.

The target range profile is widely used for recognition of aerospace targets (aircraft, missiles, and spacecraft). *IAM*

Ref.: Nebabin (1984), p. 103, (1995), p. 110; Wehner (1987), p. 148; Astanin (1989), p. 173.

Range-doppler imaging produces a two-dimensional radar image of a target that characterizes the distribution of amplitude or RCS in range and doppler coordinates. Waveforms with thumbtack ambiguity functions (phase-coded waveforms or pulse trains) are typically used to produce such images. The latter have an advantage associated with the absence of side lobes with respect to velocity near the central peak.

To increase the detail of the image, doppler range-finding processing is used. This provides high range resolution through pulse compression and coherent summing of signals accumulated in the process of observing a moving target (the principle of reverse synthesis of the antenna aperture). The algorithm for obtaining the image boils down here to a two-dimensional Fourier transform of a fixed echo frequency. *IAM*

Ref.: Mensa (1981); Munson, D.C., *Proc. IEEE* ***71***, no. 8, 1983.

A **synthesized image** can be either

(1) The simultaneous display on the plan position indicator (cathode-ray tube) of signals received in the active channel and target returns discriminated during several preceding rotations of the antenna. The latter returns are generated by a digital computer which stores the data at a rate no less than 20 Hz to exclude the flicker effect.

(2) An image of an object obtained by synthetic aperture method when there is relative movement of the object and the radar. The image is formed as a result of a Fourier transform of stored radar data using a computer. *IAM*

Ref.: Finkel'shteyn (1983), p. 520.

Three-dimensional imaging presents the distribution of scattering sources of a target depending on the values in three coordinates. Images in coordinates of range and two angles, in the coordinates of range, doppler velocity, and angle are three-dimensional images.

The three-dimensional image may be formed from the total of two-dimensional images by the methods of tomography. The production of three-dimensional geometric images of small targets (with dimensions close to the wavelength) using ultrawideband waveforms is promising. Such methods are based on Lewis-Boyars equations, which make it possible to relate a three-dimensional geometric image and its scattering characteristics. *IAM*

Ref.: Wehner (1987), Ch. 8; Astanin (1989), p. 177; Reedy, E. K. *Proc. IEEE Nat. Radar Conf.*, Los Angeles, CA, Mar. 12–13, 1986, pp. 7–11.

Image transmission is the transmission of radar information in the form of signals forming radar images. Surveillance radar image transmissions consist of range data (video or synthetic video) and antenna angular data, along with synchronization signals. Antenna angular position is usually represented by a binary code, in the interval between the video from maximum range and the next transmission.

In the simplest method of direct transmission of data, various carriers are used with frequency division multiplexing, either by RF cable or radio channel. A drawback of the method of direct transmission is the complexity and wide bandwidth of the multichannel communications line. Transmission of the radar image converted into a television image with a graphecon also requires a wide band. For this reason, transmission methods often use digital encoding with compression, taking advantage of the inherent redundancy to fit the information within a narrowband channel.

Transmission of the radar image is used when automated collection and imaging of data received by separate radars are required, at a central control point, for example at the air-traffic control center of an airport. *IAM*

Ref.: Ridenour (1947), Ch. 17; Finkel'shteyn (1983), p. 525.

Two-dimensional imaging shows the distribution of scattering sources of a target on a two-coordinate plane. Two-dimensional images include images in two angular coordinates and in range and doppler velocity coordinates. A two-dimensional image is virtually a projection of the basic three-dimensional image of a solid target and therefore has a significantly smaller volume of information about the target in comparison with a three-dimensional image.

A two-dimensional image depends on the target aspect, which usually requires obtaining many images in the interest of target discrimination, measurement of their scattering characteristics, and restoration of the three-dimensional image.

In observation of surface-distributed targets, for example in radar scanning of the earth, two-dimensional images provide significant information about the distribution of characteristics over the surface. *IAM*

Ref.: Kuchkov (1985), p. 126; Wehner (1987).

IMPEDANCE. Impedance is “the ratio of the phasor equivalent to a steady-state sine-wave voltage or voltage-like quantity (driving force) to the phasor equivalent of a steady-state sine-wave current or current-like quantity (response).” In a two-conductor transmission line it is: the ratio of the complex voltage between the conductors to the complex current on one conductor in the same transverse plane; while in a waveguide it is “a nonuniquely defined complex ratio of voltage and current at a given transverse plane in the waveguide, which depends on the choice of representation of the characteristic impedance.”

IEEE (1993), p. 620.

Impedance matching is “the control of impedance for the purpose of obtaining maximum power transfer or minimum reflection.” The maximum possible power from the source is received when a load impedance connected to a source is adjusted to be equal to the complex conjugate of the impedance of the source. Impedance matching for minimal reflection is typically used in transmission lines with the aid of various impedance-matching circuits. At microwaves the main impedance-matching circuits are matching stubs and transformers. *SAL*

Ref.: Johnson (1984), pp. 43.1–43.27.

The **intrinsic impedance of the medium** is the ratio of the effective or root-mean-square values of the electric *E*- and magnetic *H*-field intensities. The usual notation is $Z_0 = E/H$. *SAL*

Ref.: Fink (1975), p. 1.43.

INDUCTANCE is "the property of an electric circuit by virtue of which a varying current induces an electromotive force in that circuit or in a neighboring circuit." Essentially it is the rate of increase in magnetic "linkage" with an increase in current in an electric circuit such as a coil, where linkage is defined as the product of the magnetic flux through the circuit (coil) and the number of turns. The inductance of a coil is given by

$$L = \frac{X}{2\pi f}$$

where X is the reactance and f the frequency of the alternating current. *PCH*

Ref.: Van Nostrand (1983), p. 1,592.

INFORMATION MEASURE [METRIC]. An information measure describes the amount of information, determining the unit and method of measurement of information in the process of its receipt. This consists of "transformation" of some *a priori* distribution, $P_n = \{p_k, k=1, n\}$, describing observed objects and their assumed properties in the empirical $Q_n = \{q_k, k = 1, n\}$ after processing of the received data.

As the information measure we use the difference

$$I_k = H(p_k) - H(q_k), k = \overline{1, N}$$

where $H(x)$ is entropy. Depending on the type of function describing entropy, one can use Shannon, Kulbak, Kotel'nikov, Bayes, or Fisher measures.

The Bayes measure corresponds to entropy of the type

$$H(x) = 1 - x, x \in (0,1)$$

The amount of Bayes information is determined by the expression

$$I_k = q_k - p_k.$$

The Fisher measure corresponds to entropy of the type $H(x) = \ln^2 x, x \in (0,1)$. The quantity of Fisher information is determined by the expression

$$k = \ln^2 p_k - \ln^2 q_k.$$

The Kotel'nikov measure corresponds to entropy of the type $H(P_n) = 1 - \max\{p_1, \ldots, p_n\}$, and determined by the expression

$$I = \max\{q_1, \ldots, q_n\} - \max\{p_1, \ldots, P_n\}.$$

The Kulbak measure corresponds to entropy of the type

$$H(x) = \log\frac{1-x}{x}, x \in (0, 1)$$

The amount of Kulbak information is determined by the formula

$$I_k = \log\frac{q_k}{p_k} - \log\frac{1-p_k}{1-q_k}$$

The Shannon measure corresponds to entropy of the type

$$H(x) = \log(1/x), x \in (0,1)$$

The amount of Shannon information is determined by the expression

$$I_k = \log(q_k/p_k).$$

If the logarithm base 2 is used, then the amount of Shannon information is measured in bits.

In radar applications, information measures are used for information description of a radar channel. *IAM*

Ref.: DiFranco (1968), Ch. 7; Kosenko (1982), pp. 30–33.

INTEGRATED CIRCUIT. An integrated circuit is a solid-state circuit that is "the combination of interconnecting circuit elements inseparably associated or within a continuous substrate." Integrated circuits can be of hybrid and monolithic types (Table I1).

Table I1
Comparison of Microwave Integrated Circuits

Characteristics	Hybrid integrated circuits	Monolithic integrated circuits
Possibility of replacement and tuning of components	Yes	No
Possibility of producing significant output power	Yes	No
Possibility of creating built-in nonreciprocal devices and microwave filters	Yes	Limited
Reliability	Lower	Higher
Yield	High	Lower
Cost	Lower	Higher

Integrated circuits make possible the microminiaturization of electronic equipment (increasing the density of components per unit volume by a factor of 10 to 100), improvement in speed by a significant factor, increase in reliability, and reduction in power consumption. Such circuits are widely used in radar signal processors, receivers, and lower power portions of radar transmitters. *IAM*

Ref.: IEEE (1993), p. 662; Frey (1985); Jordan (1985), Ch. 20; Hoffman (1987); Gassanov (1988), p. 24; Nikolaev (1992), pp. 9, 273.

A **bipolar integrated circuit** is based on bipolar transistors of the NPN or PNP type. Such circuits have a speed that is an order of magnitude higher than integrated circuits on an MOS structure. By a criterion such as power-to-speed ratio, a bipolar integrated circuit also surpasses integrated MOS circuits. Joint use of bipolar and MOS instruments in integrated circuits makes it possible to improve the characteristics of the devices made on their basis. *IAM*

Ref.: Popov (1980), p 51; Fink (1982), p. 820; Nikolaev (1992), p. 10.

A **digital integrated circuit** is used in digital signal and data processing. The quality of digital integrated circuits is evaluated by speed and the specific delay time of signal propagation, which is no more than 2.5 ns/logic component. In terms of technology of production, digital integrated circuits are usually classed as *semiconductor integrated circuits*.

Digital integrated circuits are widely used in computer systems with microprogram control and high productivity (tens and hundreds of millions of operations per second) and in digital signal processors. *IAM*

Ref.: Fink (1982), p. 8.80; Efimov (1983), p. 22; Nikolaev (1992), p. 11.

A **distributed IC** is a microwave microelectronic circuit with distributed parameters. The most commonly used form of distributed circuit is the microstrip transmission line. *SAL*

Ref.: Fink (1975), p. 8.76.

A **germanium integrated circuit** is a type of semiconductor integrated circuit. In some cases, owing to a number of advantages, the use of germanium makes it possible to create faster logic circuits. In addition, germanium transistors are marked by a lower dependence of the gain coefficient on the temperature, which makes it possible to use such logic circuits at reduced temperatures. *IAM*

Ref.: Popov (1980), p. 84.

In a **hybrid integrated circuit (HIC)** some elements have an independent structural formulation. The technology of producing hybrid microcircuits is based on successive application of thin-film passive components (microstrip transmission lines, oscillators, microwave components with distributed and lumped parameters) onto an insulating substrate with subsequent connection of active and some passive components and assemblies (bridges, directional couplers, filters, ferrite devices, etc.) in the form of individually mounted parts. Semiconductor instruments in HIC are primarily ungrounded, with a low lead inductance. Integrated circuits are connected to external devices by coaxial-strip adapters and electrical connectors.

HICs are produced by industry in the form of parametric and transistorized amplifiers, switches, mixers, oscillators, phase inverters, transistorized multiplication circuits, elements of phased-array antennas, and so forth. *IAM*

Ref.: Popov (1980), p. 85; Nikolaev (1992), p. 12; Gassanov (1988), p. 24.

An **IGFET-structure integrated circuit** is designed on the basis of IGFET transistors. Integrated IGFET circuits can be used to make microwave complex integrated circuits that have high or low speed, including integrated matrices (systems in the form of a large number of standard structural cells). Integrated IGFET circuits are marked by their variety. This may be explained by the universality and multiplicity of the combinations that may be produced within a single silicon crystal. Joint use of bipolar transistors and IGFET transistors makes it possible to improve the characteristics of the devices produced on their basis. *IAM*

Ref.: Popov (1980), p. 157; Fink (1982), p. 8.32.

A **large-scale integrated (LSI) circuit** is an integrated circuit with a high degree of integration of components, performing the functions of an assembly of electronic apparatus (e.g., an arithmetic-logic device or operational storage device). A large-scale integrated circuit in an IGFET structure can contain 103 to 105 elements, and a bipolar one 500 to 2,000. Circuits with greater integration are called very-large-scale integrated circuits. *IAM*

Ref.: Popov (1980), p. 54; Jordan (1985), p. 20.7; Nikolaev (1992), p. 10.

A **linear integrated circuit** implements a low- or high-frequency amplifier. Base standard crystals containing a specific number of elements are used for these integrated circuits: for example, one of them contains 12 elements, another 24 elements (e.g., 7 transistors, 10 resistors, 7 “tunnels” for making connections). *IAM*

Ref.: Popov (1980), p. 207; Fink (1982), p. 8.45.

A **lumped integrated circuit** is a microwave microelectronic circuit using lumped circuit elements such as resistors, capacitors, or inductors. For electrical components to behave as lumped elements, their physical dimensions have to be much smaller than the wavelength of the signal. Integrated circuit components can maintain their lumped characteristic up to much higher frequencies than their discrete counterparts. *SAL*

Ref.: Fink (1975), p. 8.96.

A **metal-insulator-semiconductor- (MIS-) structure integrated circuit** is based on an MIS-transistor structure. This type of IC is flexible because of the variety of combinations that can be realized within a single silicon crystal and are typically used for producing complex microwave circuits of low- and medium-speed capabilities. Sometimes mutual use of bipolar transistors and MIS-transistors can enhance IC performance. *AIL*

Ref.: Popov (1980), p. 157; Fink (1982), p. 8.49.

A **monolithic integrated circuit** is produced as a single whole from a whole semiconductor crystal (see semiconductor integrated circuit) or other solid-state material that controls the characteristics of the circuit. The elements of a monolithic integrated circuit are arranged within it, with some on the surface. The term *monolithic microwave integrated circuit (MMIC)* applies not only to a material of homogeneous structure, but also to a nonhomogeneous material, for example silicon on sapphire. The semiconductor material (monostatic film) may be in small local sectors of the substrate of the integrated circuit. Such circuits have the highest degree of integration. *IAM*

Ref.: Popov (1980), p. 238; Fink (1982), p. 8.54.

A **MOS-structure integrated circuit** is made on the basis of metal-oxide semiconductor transistors, which are a variety of IGFET transistor. (See **IGFET-structure integrated circuit**.) *IAM*

A **passive integrated circuit** does not contain active amplifying or rectifying elements. It is usually made by successive application of passive elements (distributed, lumped inductance coils, capacitors, intracircuit connectors) onto a single insulating substrate. *IAM*

Ref.: Popov (1980), p. 278; Nikolaev (1992), p. 12.

A **semiconductor integrated circuit** consists of active and passive components produced in the same semiconductor monocrystal. Some of the connections in such a circuit may be in three dimensions, and some on the protective layer of the crystal. At frequencies up to 10 GHz, structures of the silicon-on-sapphire type (silicon integrated circuits) are used for semiconductor integrated circuits. In the higher frequency range, gallium-arsenide (GaAs) is used. Passive components are created by the methods of diffusion onto a substrate or deposition of thin or thick films on it. Active components (most often field-effect transistors, see **IGFET-structure integrated circuit.**), are grown on a high-ohm GaAs substrate by the methods of ion epitaxy or ion doping.

Large capital investments, highly pure materials, and automation of design and manufacture are required for organization of production of semiconductor integrated circuits. *IAM*

Ref.: Gassanov (1988), p. 26.

A **silicon integrated circuit** is the most common variety of semiconductor integrated circuit. The use of silicon offers a number of technological advantages and makes it possible to create integrated circuits with fewer parasitic connections between components operating at higher temperatures. Silicon circuits are also used in micropower (1 to 300W) integrated logic circuits, making it possible to reduce the power consumed by onboard computer devices, to reduce their weight and dimensions, and extend operating lives. *IAM*

Ref.: Popov (1980), p. 199.

A **solid-state integrated circuit** is a microwave module that is produced in a combination of different waveguide lines spatially arranged in layers of the dielectric and connected to one another by capacitive, inductive, galvanic, or electromagnetic connections. Individual functional components (base elements) of a solid circuit are based on optimal types of transmission lines for each of them, with optimal junctions between them. Solid integrated circuits are a promising trend in the development of microwave integrated circuits. At present multistage circuits are used in which the base components are arranged in different stages on individual substrates that are little associated electrically. Suspended and coaxial-waveguide-strip connectors are widely used as the junction components. Examples of use of such circuits include solid FAR transceiver modules, IFF beacons, frequency mixers, and modulators. *IAM*

Ref.: Nagihara E., *IEEE Tran.s **MTT-30***, no. 3, 1982, pp. 235–242.

A **very-large-scale integrated circuit (VLSI)** is one with a very large degree of integration, which constitutes a finished item capable of performing the functions of an apparatus (for example a microprocessor). In degree of integration, semiconductor microcircuits with bipolar transistors are inferior to on IGFET-structure integrated circuits. (See Table I2.) *IAM*

Ref.: Jordan (1985), p. 20.12; Nikolaev (1992), p. 10.

INTEGRATION, INTEGRATOR. Integration is the process of combining *n* samples of a signal, each accompanied by an independent sample of noise or interference, to improve the signal-to-noise ratio. The samples may be successive pulses in a train or pulses received in parallel channels with independent noise. In the case of successive signal samples having phase coherence, the integration may be performed by adding voltages at intermediate frequency or at baseband on I- and Q-samples, before envelope detection. This is called *coherent* or *predetection integration*. If phase coherence cannot be obtained, or if the complexity and doppler sensitivity of the coherent integrator is not desired, the signals may be passed through an envelope detector prior to addition of voltages. This is called *video*, *noncoherent*, or *postdetection integration*. Envelope-detected signals may be converted to digital form and integrated in accumulating registers. If the conversion is to one-bit digital form, the process is called *binary integration*. Signal information may also be combined using a simple process of accumulating probabilities of detection over *n* trials. This may be termed *cumulative integration*.

The signal-to-noise ratio is improved by integration, greatest improvement being obtained in coherent integration, where the integrated $(S/N)_i = n(S/N)$, and S/N is the signal-to-noise ratio per sample at the input. For video integration, the improvement is less: $(S/N)_i = n(S/N)/L_i(n)$, where $L_i(n)$ is the integration loss. The binary has an additional loss of about 1.6 dB relative to video integration. (See also **DETECTION**). *DKB*

Skolnik (1980), pp. 29–33; Barton (1988), pp. 69–74.

An **accumulator (integrator)** is a digital summing device, such as that used in an *n*-pulse video integrator, either forming the sum of *n* pulse amplitudes or counting up to *m* threshold crossings before generating an output alarm. *SAL*

Ref.: Barton (1991), p. 4.13.

An **analog integrator** is the analog unit integrating the pulse train to improve the signal-to-noise-ratio. The basic components of analog integrator are delay line with the delay equal or multiple to the pulse repetition interval t_r and adder (Σ) (Fig. I3). In the configuration of the one-cycle integrator shown in the figure, here is positive feedback with the feedback coefficient β less then unity, but close to it (β ≈ 1). When the signal and noise appear at the input of such an integrator, the multiple circulation (recirculation) arises and a set of signals appears at the output that are delayed by the time $k{\cdot}t_r$ (k is an integer) and multiplied by the factor β^k. The sum of these signals is taken from the output 1. Such type of an integrator is often called a *recirculator.*

Table I2
Degree of Integration of Very-Large-Scale Integrated Circuits

Functional purpose	Type of integrated circuit	Number of elements or components on crystal
Digital	On IGFET structure	More than 10,000
Digital	Bipolar	More than 2,000
Analog	Combined (on bipolar and IGFET transistors	More than 300

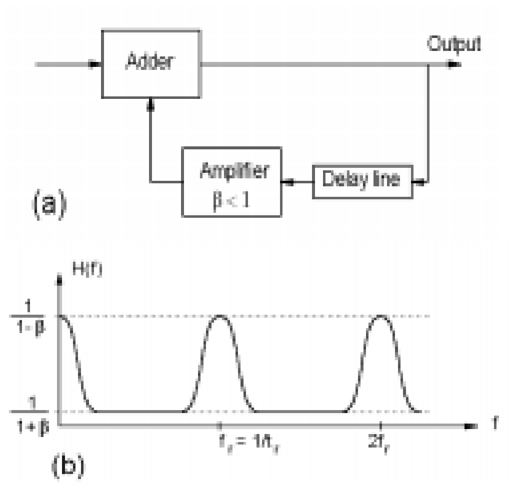


Figure I3 Delay-line based integrator: (a) block diagram, (b) frequency response.

Depending on the type of delay line, analog integrators are divided into *dynamic* and *static* integrators. The latter are often termed *synchronous integrators*. In dynamic integrators, typically ultrasonic delay lines or surface-acoustic-wave delay lines are employed. In static integrators, the delay of the pulses is implemented through its recording in magnetic tape, disk, or cathode-ray tube. The reading is done in the desired moment of time. These integrators have poor performance. The general disadvantage of one-cycle integrators is a comparatively small improvement of the signal-to-noise ratio. It can be increased by using a two-cycle integrator (Fig. I4) that can have a signal-to-noise ratio improvement about twice that of the one-cycle integrator. *AIL*

Ref.: Skolnik (1970), p. 17.27; Finkel'shteyn (1983), pp. 265–280; Lezin (1969), pp. 256–276.

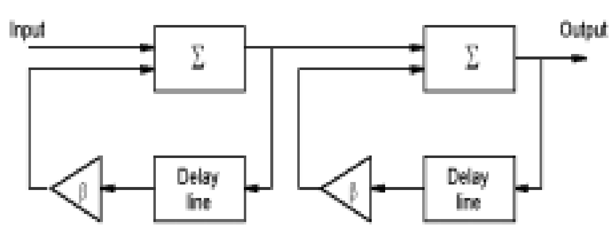


Figure I4 Two-cycle integrator.

Batch integration is the process of collecting n successive returns, performing the integration, and then discarding these returns before collecting the next batch. This is in contrast to the moving-window integrator. *DKB*

Binary integration is a noncoherent integration process in which envelope-detected signals are quantized by an initial threshold into one-bit binary signals, before being passed to an accumulator. When the accumulator count reaches a second threshold level m, detection is declared. When n pulses are integrated in this way, the optimal second threshold is in the order of $m_{opt} = 1.5\sqrt{n}$, and the integration gain is approximately 1.6 dB lower than that of the n-pulse video integrator. (see **integration gain**). The probability P of detection or false alarm at the output of a binary integrator can be found as a function of the corresponding probability p at its input as

$$P(m/n) = \sum_{j=m}^{n} \frac{n!}{j!(n-j)!} p^j (1-p)^{n-j}$$

Several common cases are:

$P(1/1) = p$
$P(1/2) = p^2 + 2p(1-p)$
$P(2/3) = p^3 + 3p^2(1-p)$
$P(2/4) = p^4 + 4p^3(1-p) + 6p^2(1-p)$

The general block diagram of the binary integrator is shown in Fig. I5. If the binary counter is periodically decremented to maintain a low false-alarm probability on a continuous stream of input signals, the integrator is known as a *moving-window* (or *continuous*) *binary integrator*. Otherwise, it is a *batch integrator*, in which the counter is set to zero after each group of n pulses.

Although the binary integrator introduces an additional loss relative to the ideal noncoherent integrator, it is much less sensitive to the effects of large, random interference pulses because the energy in a single pulse contributes no more than a single "one" in the binary counter, rather than its large voltage in the linear integrator.

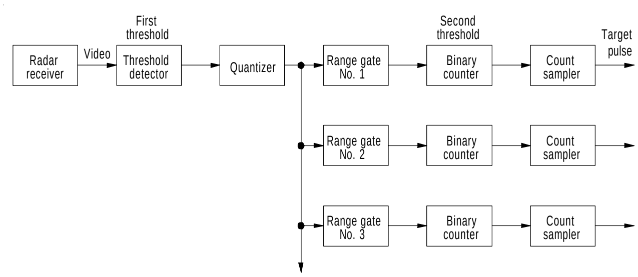


Figure I5 Block diagram of binary integrator (after Skolnik (1980), Fig. 10.7, p. 388).

This type of integrator is also known as the double-threshold detector, m-out-of-n detector, or coincidence detector, and is widely used in radar signal processors. *DKB*

Coherent [predetection] integration occurs when all of the radar pulses n received from a target during the observation time are added in phase before envelope detection. The signal-to-noise ratio (SNR) is enhanced by the factor $n = f_r t_o$ over that of a single pulse, where f_r is the pulse repetition frequency and t_o is the integration time. Similarly, for a continuous wave (CW) radar, during the observation time t_o, there will be $n = B_n t_o$ samples of signal and independent noise added in a coherent integration process, where B_n is the noise bandwidth of the filter. In either case, coherent integration requires that the signal have a predictable phase relationship (i.e., coherence) and that the phase response of the filter be such as to bring all of the signal components into the same phase during the integration process. In an ideal coherent integration scheme, the coherent integration gain is exactly n. Coherent integration is sometimes referred to as *predetection* *integration*, in that it occurs in the radar receiver before the second (envelope) detector. *PCH*

Ref.: Barton (1991), pp. 4–11.

Continuous integration is an n-pulse integration process in which a new integrated value is formed after reception of each pulse. The oldest pulse is discarded in moving-window integration, while in recursive integrators the weighting of the older pulses is reduced by the weight assigned to the new pulse. The process is the opposite of batch integration (integrate-and-dump), in which a given group of n pulses is integrated once and then discarded. Continuous integration avoids the angle straddling loss characterizing batch integration with a continuously scanning beam. *DKB*

Ref.: Skolnik (1980), p. 390.

Cumulative integration refers to the process in which a detection decision with probability P_1 is made on each pulse, resulting in a cumulative detection probability P_c after n pulses:

$$P_c = 1 - (1 - P_1)^n$$

This process gives far less gain than other video integration techniques (see integration gain), but may represent the only option if the pulses are separated in time by an interval that permits targets to move from one resolution cell to the next, as in scan-to-scan integration. *DKB*

Ref.: Barton (1988), p. 74.

A delay-line integrator is a continuous integrator in which pulses are recirculated through a delay line, the recirculation loop having a gain < 1 to preserve stability. The resulting weighting function is approximately exponential (see **analog integrator**). Common configurations include the recirculating integrator and the tapped-delay-line integrator. *DKB*

Ref.: Skolnik (1980), p. 390.

A **digital integrator** is one in which the signal is converted to digital form before being processed in digital memory, weighting, and summing circuits. Video integrators of the delay-line (recursive) type pass the digitized signal through a single shift register, recirculating the output with a loop gain < 1 as with analog delay-line integrators. The digital implementation has the advantage of inherent gain stability, permitting gains near unity (many integrated pulses) to be realized. Batch-process integrators may also use a single shift register as a memory, accumulating n pulses with unity weights and clearing the register after each batch. Digital integrators of the moving-window type use n shift registers to store outputs from n pulse repetition intervals, forming sums in each range cell during each interval. *DKB*

exponential integrator (see **recirculator integrator**).

feedback integrator (see **delay-line integrator**).

Integration gain is the net improvement in received signal-to-noise ratio (SNR) due to the addition of independent samples of signal and noise available during the radar observation time (time-on-target). There are four distinct ways, listed in order of declining integration efficiency and declining complexity of implementation, in which the information from n pulses or samples may be processed to improve the radar detection performance: (1) coherent integration, (2) noncoherent (or video) integration, (3) binary integration, and (4) cumulative integration. Figure I6 compares the integration

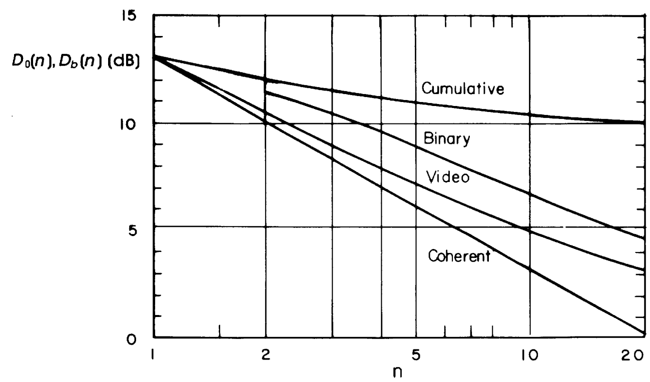


Figure I6 Comparison of detectability factors for four methods of integration (from Barton, 1988, Fig. 2.3.3, p. 75).

efficiency of each of the four methods in terms of the resultant detectability factor (or SNR) required as a function of n for the conditions noted. *PCH*

Ref.: Barton (1991), pp. 4–14.

In-phase/quadrature channel integration is the integration of coherent pulse train after synchronous detection in two channels by phase detectors with the reference voltages proportional to $\cos 2\pi f_0 t$ and $\sin 2\pi f_0 t$; that is, shifted in phase by $\pi/2$, where f_0 is carrier frequency. Each channel has an integrator and a squaring device. After the summing these two channel and extracting the square root from the sum, the signal passes to thresholding unit. The advantage of such an integration is that the data about the phase are not lost, but video (baseband) pulses are integrated, instead of RF frequency pulses, simplifying the design of the optimum receiver. *AIL*

Ref.: Finkel'shteyn (1983), pp. 237–239; Schleher (1991), p. 608.

A **moving-window integrator** processes the incoming signals continuously, dropping the oldest return when adding the most recent. This is contrasted to the batch integrator. *DKB*

A **multichannel integrator** is one used in multichannel radar. Because multichannel coherent integrator design is complicated, noncoherent integration is employed, making it possible to reduce the number of channels in the integrator and to reduce the required phase stability. When the number of integrated pulses is not too large (less than 20), amplitude recirculators with delay lines in the positive feedback circuit with feedback coefficient β can be efficient (Fig. I7). This integrator has a bank of doppler filters ($F_1 .... F_n$), n detectors (D) and switches, and a common noncoherent integration unit (adder Σ, modulator M, delay line LN, an amplifier with positive feedback coefficient β, and detector D) for all channels. *AIL*

Ref.: Lukoshkin (1983), pp. 284–287; Nitzberg (1992), p. 240.

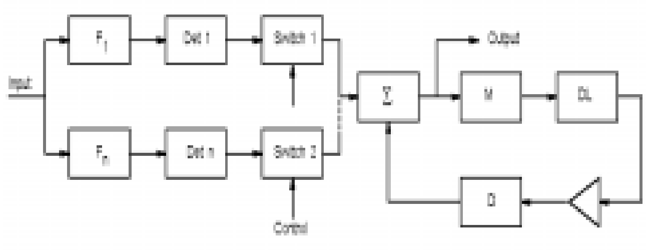


Figure I7 Multichannel recirculating integrator with time multiplexing of channels.

Noncoherent [video, or postdetection] integration occurs when n signal samples or pulses are added after passing through an envelope detector which removes any phase relationships. The integrated SNR is increased by a factor approaching n, but decreased by a detector loss due to a loss in information for target detection purposes. As the signal input to the envelope detector decreases, the small signal suppression loss increases as shown in Fig. I8, and further video integration cannot restore the SNR represented by the total signal energy ratio. As a result, there is a requirement for more energy per pulse, which can be considered as integration loss, which is the ratio of total signal energy required of the n-pulse train to that which would have been required if a single pulse had been transmitted and processed, or if a matched filter for the n-pulse train had been used.

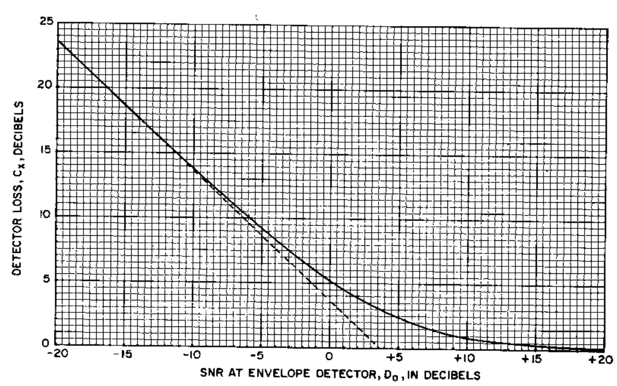


Figure I8 Envelope detector loss vs. SNR (from Barton, 1988, Fig. 2.2.4, p. 64).

Figure I9 gives the magnitude of this integration loss L_i versus the number of pulse n, noncoherently integrated, as a function of the detectability factor D_0 (1) for a steady target. *DKB*

Ref.: Barton (1988), pp. 69–76.

predetection integration (see **coherent integration**).

A **recirculator integrator** is an integrator employing delay units in a positive feedback circuit. It can be done of analog or digital configuration. An analog recirculator consists of an adder and a delay line in a feedback circuit (Fig. I10). To prevent a self-oscillation mode, the feedback coefficient β is set less then unity, or a feedback circuit regulated by a special

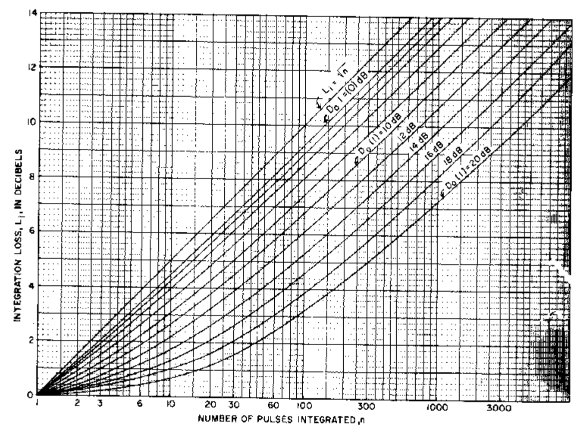


Figure I9 Integration loss vs. number of pulses integrated after envelope detection (from Barton 1988, Fig. 2.3.2, p. 72).

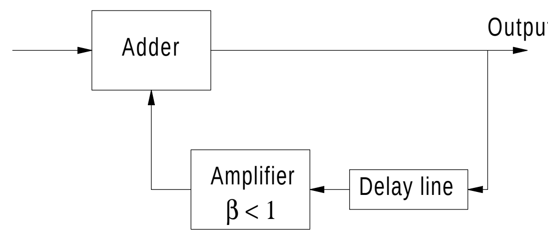


Figure I10 Recirculator integrator.

control unit is employed. A digital recirculator in every cycle evaluates the sum:**,**

$$\sum_{j=0}^{\infty} \beta^j y_j$$

where y_j is quantized input voltage at the moments $t - jT$, where T is the delay interval. The response of a circulator to the unity pulse is a grid function with an exponential decreasing envelope. Consequently, a recirculator is also termed an *exponential* or feedback integrator. *IAM*

Ref.: Dulevich (1978), pp. 157, 171; Skolnik (1990), pp. 8.5–8.8.

Integration time, also referred to as *observation time* or *time-on-target*, is the time during which the target is illuminated by the radar on each scan, that is

$$t_o = \frac{\theta_{3az}}{\omega_s} = \frac{t_s \Psi_b}{\Psi_s}$$

where θ_{3az} is the 3-dB azimuth beamwidth of the radar, ω_s is the radar scan rate, t_s is the radar frame time, ψ_s is the radar's solid angle of search (in steradians), and ψ_b is the solid angle of the radar beam. For a scanning pulse radar, there will be $n = t_o f_r$ pulses or samples available for integration during the time-on-target, where f_r is the pulse repetition frequency. For a continuous-wave (CW) radar, or a coherent pulse radar, the coherent integration time is not the same as the time-on-target but is the integration time of the predetection filter, approximately equal to the reciprocal of the filter bandwidth. Subsequent noncoherent integration is possible over the time-on-target. *PCH*

Ref.: Barton (1988), pp. 17, 24.

video integrator (see **noncoherent integration**).

INTELLIGENCE, radar. There are two different interpretations of the term "radar intelligence": (1) the process through which information about a particular "target" radar is obtained and (2) use of radar in the collection of intelligence information concerning the operations, technical capabilities, and intent of the target subject.

The first definition includes radiation intelligence (RADINT), and other means, such as visual observation and photography, that may reveal significant information about the subject radar or radars. Passive reception of an operating radar's RF emission can yield data such as radar frequency, effective radiated power (ERP), azimuth and elevation beamwidths, antenna scan rate and time-on-target, volumetric coverage, antenna sidelobe pattern, and transmitted waveform characteristics. From these, other radar characteristics independently obtained through visual means, and some assumptions concerning technology-driven nonobservables such as internal radar losses and receiver noise figure, it is possible to establish the role and mission of the radar, to estimate its nominal detection range, to determine whether the radar was designed to detect targets in land and weather clutter, and to provide some insight into the potential vulnerability of the radar to ECM.

Radar intelligence in the alternative sense includes the use of space or airborne radar to observe areas of the earth's surface for evidence of operational military activity, ground mapping to establish industrial capability, the amount and location of natural resources, the environmental health of certain regions, and the monitoring of fishing activity, as well as support to the interdiction of illegal drug activity, and so forth. *PCH*

INTERFERENCE. Electromagnetic interference is the reception of waves other than produced by the transmitter or target of interest. The main sources of interference are clutter, noise, jamming, and other factors discussed under **ELECTROMAGNETIC COMPATIBILITY**. Interference typically has a random nature and it is the main factor limiting the detection performance of a radar. *SAL*

Wave interference occurs when two or more waves from different sources are superimposed, resulting in intensification or weakening of the resultant electromagnetic field intensity, depending on the relative phases of the waves. In radar applications, wave interference manifests itself in distortion of the coverage pattern due to multipath propagation. (See **PROPAGATION**.) An application of the interference phenomenon is the interferometer used in precision phase-based angular measurement. (See **INTERFEROMETER'**). *SAL*

Ref.: Mayzel's (1972), p. 9; Meeks (1982), p. 43.

INTERFEROMETER, radar. A radar interferometer is "a receiving system that determines the angle of arrival of a wave by a phase comparison of the signals received at separate antennas or separate points on the same antenna." The resolution is inversely proportional to the baseline d of the interferometer (the distance between antennas) and can be estimated by the width of the main interference lobe λ/d. In terms of the method of practical realization, interferometers are subdivided into *additive*, which perform coherent addition of signals and detection of resultant signal, and *multiplicative*, which multiply the signals and perform integration. The drawback of additive interferometers is the presence of a constant component of the output signal, which is determined by the power of the received signals. This degrades the detection characteristics against a background of interference.

The basic difficulty in realization of radar interferometers is elimination in the ambiguity of angular coordinates, caused by the multilobe structure of the dependence of the output signal on the angle (interference pattern). To eliminate this ambiguity, wideband waveforms are often used, reducing the lobes other than the main lobe in the interference pattern. Additional antennas, spaced within the bounds of the base are also used.

Interferometers are used in radar surveillance of the earth, in radiometry of the earth's surface, and in radio astronomy.

A synthetic aperture interferometer uses the principle of a synthetic aperture to receive the signals. In contrast to a conventional interferometer, it permits measuring the range of the targets, along with the angles, if the range is approximately of the same order as the interferometer base. When the base is oriented along the target velocity vector and the beamwidth of the interferometer beams θ is comparatively small, the resolution along the path of target direction δL is

$$\delta L = \frac{4\lambda}{\theta^2} \cdot \frac{R_0}{b}$$

and range resolution

$$\delta R = \frac{\lambda}{2\theta} \cdot \frac{R_0}{b}$$

where R_0 is the target range.

These values differ from corresponding resolution parameters for active synthetic aperture radars by factor of R_0/b specific for interferometers only and indicate that in contrast to SAR, the resolution in this case is higher for range than along the target motion direction. Synthetic aperture interferometers are used in terrain observation radars and radar astronomy where because of large bases and motion due to the earth's rotation, it is possible to determine the angular location of a sky object with high accuracy. *IAM*

Ref.: Barton (1964), p. 513; Reutov (1970), p. 83; Dulevich (1978), p. 551; Mel'nikov (1980), p. 116.

INTERPOLATION is the process by which target position is measured within a fraction of the radar resolution cell width. In angle, the process is also known as *beam-splitting*. Interpolation of position within the resolution cell is fundamental to the process of measurement of target coordinates.

The interpolation ratio is the ratio of the width of the radar resolution cell to the accuracy of the target position measurement in that coordinate. In angle this is also known as the *beam-splitting ratio*. *DKB*

Ref.: Barton (1988), p. 377.

INTERROGATION, INTERROGATOR. Interrogation "in a transponder system is the signal or combination of signals intended tom trigger a response." The interrogator is the transmitting part of a secondary surveillance radar (SSR). The transmitter-receiver of such a system, properly called an *interrogator-responser*, is sometimes referred to simply as an *interrogator*. The typical format of interrogation is shown in Fig. I11. Pulse P1 and P3 are transmitted through the interrogate beam, and P2 through the control beam. The spacing

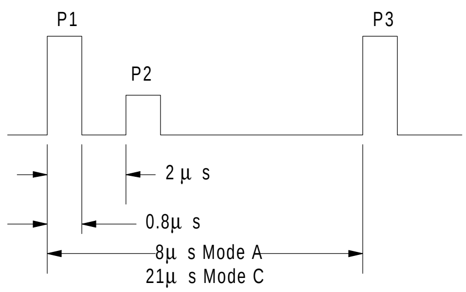


Figure I11 Interrogation signal format (after Stevens, 1988), Fig. 3.3, p. 23.

between P1 and P3 determines the data content of the transponder reply, and comparison of amplitudes of P1 and P2 permits the transponder to reject sidelobe interrogations. The following basic interrogation modes are used in aircraft SSR systems:

Modes 1 to 3: Military IFF identity mod.;

Mode 4:Military IFF identity mode (using different interrogation and reply formats than other modes, with encryption techniques).

Mode A:Common military and civil identification.

Mode B:Once used as a second SSR identity mode, currently not in use.

Mode C:Flight level reporting mode.

Mode D:SSR mode reserved for possible future use.

Mode S:Enhanced SSR mode, in which a unique 24-bit address is included to identify the specific aircraft being interrogated. In this case the problem of garble is avoided and additional data can be obtained in the reply.

The main characteristics of interrogation modes are given in Table I3. *SAL*

Ref.: IEEE (1993), p. 677; Stevens (1988), pp. 67–106; Skolnik (1970), p. 38.1; Vasin (1977), p. 66.

INTRUSION (in electronic warfare usage) is "the intentional insertion of electromagnetic energy into transmission paths in any manner with the objective of deceiving operators or of causing confusion." *SAL*

Ref.: Johnston (1979), p. 62.

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Table I3
Interrogator Mode Pulse Specifications

P1, P2, P3 pulses:	Pulse duration: Rise time: Decay time:	0.8 ± 0.1 μs ≥ 0.05 μs, ≤ 0.01μs ≥ 0.05 μs, ≤ 0.02μs
Mode spacing, Mode:	P1–P2 Separation	P1–P3 separation
1	2 ± 0.01 μs	3 + 0.01 μs – 0.05 μs
2	2 ± 0.01 μs	5 ± 0.01 μs
3/A	2 ± 0.01 μs	8 ± 0.01 μs
B	2 ± 0.01 μs	17 ± 0.01 μs
C	2 ± 0.01 μs	21 ± 0.01 μs
D	2 ± 0.01 μs	25 ± 0.01 μs

IONOSPHERE. The ionosphere is the ionized region of the atmosphere at altitudes exceeding 60 km. Ionization of the atmosphere occurs mainly due to the effect of solar radiation (mainly ultraviolet). The vertical section of the ionosphere comprises four ionized layers characterized by a specific number of electrons in a unit of volume. These layers conditionally are designated D, E, F_1, and F_2. The D layer manifests itself only during daylight hours at an altitude of 60 to 90 km, stratum E at 100 to 120 km, while the F_1 and F_2 layers occupy regions at an altitude ranging approximately from 200 to 450 km from the earth's surface. The presence of ionized layers significantly alters the picture of the electromagnetic field during propagation of radio waves from a transmitting to a receiving antenna. The atmosphere exerts the greatest impact on propagation of HF and VHF waves. They are reflected off the ionized layers (see **WAVE, sky**), serving as the basis for OTH radar operation. The ionosphere does not exert a noticeable impact on propagation of UHF and shorter wave bands.

Irregular solar radiation exerts a strong impact on the ionosphere. There is a radical change in the parameters of the ionosphere, referred to as an *ionospheric storm*. During such a storm, F-layer ionization decreases such that the layer loses its ability to reflect a wave exceeding 10m in length. Here, passage of HF waves ceases completely at high latitudes. This is explained by onset of aurora inhomogeneities of electronic concentration. Radar research into aurora inhomogeneities is aimed mainly at refinement of structures and properties and establishment of the reasons for the onset and dynamic of the development of polar ionospheric storms, as well as at checking the validity of different theoretical hypotheses.

Both active and passive radar methods are used in polar ionospheric research. Cosmic radio-frequency radiation or the signals of multifrequency radio transmitters installed aboard special satellites serve as passive radar signal sources. When active radar methods are used to determine the characteristics of the ionosphere, artificial ionized formations of a type of barium clouds are used widely, along with direct probing of the researched regions. An ionospheric storm mainly impacts operation of decimetric waveband OTH radars and a portion of the operation of metric band radars, as well as communications equipment operating at wavelengths greater than 10m. *AIL*

Ref.: Blake (1982); Kolosov (1984), p. 112; Dolukhanov (1972), pp. 184–235.

IRIS, matching. A matching iris is a waveguide capacitive or inductive iris used for matching circuits by the compensation method. The iris performs the function of the source of reflection with the coefficient of reflection at the point of connection, so that the total reflection from the load and the iris is minimal. The matching irises are used basically for narrowband matching.

They are used, for example, in waveguides to feed the radiating element for matching of phased arrays. Irises may also be used for matching waveguides of differing cross-sections and coaxial circuits. *IAM*

Ref.: Montgomery (1947), Ch. 6; Rakov (1970) vol. 2, p. 248; Voskresenskiy (1981) p. 219.

A **capacitive iris** is a matching iris that reduces the spacing between the wide walls of a waveguide. The field is concentrated between the edges of the iris, and a reserve of electrical energy forms. For this reason, in the equivalent circuit such an iris is represented by a capacitor connected in parallel to the transmission line. The capacitive iris greatly reduces the breakdown voltage of the waveguide. *IAM*

Ref.: Montgomery (1947), p. 166; Sazonov (1988), p. 65; Rakov (1970), Vol 2, p. 249.

An **inductive iris** is a matching iris that reduces the spacing between the narrow walls of a waveguide. Transverse currents on the wide walls of the waveguide are partially closed through the plates that connect these walls. In the magnetic field of the currents flowing over the plates of the iris, magnetic energy is stored.

The equivalent circuit of the iris is an inductance coil connected in parallel to the transmission line. *IAM*

Ref.: Montgomery (1947), p. 164; Rakov (1970), vol. 2, p. 248; Sazonov (1988), p. 65.

A **resonant iris** is a metal plate with rectangular or oval opening covering the cross-section of the waveguide and containing elements of inductive or capacitive irises. The dimensions of the opening of the resonant iris may be selected so that at a given resonance frequency, the iris does not affect the dissemination of the wave H_{10} in the waveguide (i.e., it has zero conductivity). In selection of the size and shape of the iris, the external quality level is also allowed for. An equivalent circuit of a resonant iris has the shape of a parallel resonance circuit that shunts the transmission line. *IAM*

Ref.: Montgomery (1947), p. 169; Sazonov (1988), p. 66.

A **waveguide iris** is a matching iris with a thin metal partition placed in the waveguide circuit to cover part of its cross-section. In a rectangular waveguide, the most common are the symmetrical inductive, symmetrical capacitive, and resonant irises. The first two are used as matching devices (see **IRIS, matching**). For precise calculations of parameters of irises, special graphs or computer programs are used. *IAM*

Ref.: Montgomery (1947), Ch. 6; Sazonov (1988), p. 65.

J

JAFF (jammer + chaff) is the technique of using jammer-illuminated chaff, where previously ejected chaff is illuminated with a coherent noise or deception signal to impose a suitable doppler frequency on the radar return. In the normal situation, chaff is strongly attenuated by MTI, but when illuminated by the jamming signal the resulting echo falls outside the doppler rejection notch. Such a technique can be used for self-screening and its aim is to produce relatively cheap off-board decoys endowing chaff cloud reflection with a proper doppler shift (Fig. J1). Sometimes this technique is termed CHILL (chaff, illuminated).

SAL

Ref.: Neri (1991), p. 396.

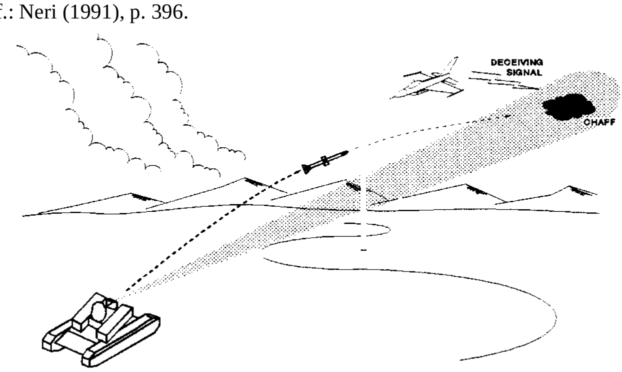


Figure J1 Jammer-illuminated chaff (from Neri, 1991, p. 397).

JAMMER, JAMMING. Jamming is (1) "a form of electronic countermeasures (ECM) in which interfering signals, typically noise-like, are transmitted at frequencies in the receiving band of a radar to obscure or distort the radar signal." (2) The deliberate radiation, reradiation, or reflection of electromagnetic energy with the object of impairing the use of electronic devices, equipment, or systems by an enemy. A jammer is the transmitter used to jam radio or radar channels.

Ref.: IEEE (1993), p. 691; Johnston (1979), p. 62.

Accidental jamming is the term sometimes used to denote interference due to transmission by friendly equipment.

Ref.: Johnston (1979), p. 56.

Active jamming is jamming using intentional radiation of electromagnetic energy to jam the victim radar (as opposed to passive jamming when the energy is not radiated but only reradiated). The main types of active jamming are noise jamming, deception jamming, and active decoys.

Ref.: Schleher (1986), p. 9.

Angle jamming is an "ECM technique, when azimuth and elevation information from a scanning fire control radar present in the modulation components on the returning echo pulse is jammed by transmitting a jamming pulse similar to the radar pulse but with modulation information out of phase with the returning target angle modulation information." *SAL*

Ref.: Johnston (1979), p. 52.

Barrage jamming is jamming with a bandwidth much wider than radar signal bandwidth. The most common form is barrage noise jamming; that is, noise jamming in which the jamming bandwidth is designed to cover the entire tuning range of a victim radar, or to encompass several radars operating at different frequency. This wideband feature distinguishes the barrage noise jammer from the spot noise jammer, whose bandwidth is designed to cover the radar's instantaneous signal bandwidth.

As does noise jamming in general, barrage noise jamming realizes its effect by raising the receiver noise level in the victim radar. While the radar's system noise temperature T_s was, in the absence of jamming, determined only by the receiver and the natural environment, the jammer raises this to a new value $T_i = T_s + T_j$, with the jammer temperature given by

$$T_j = \frac{P_j G_j A_r F_j^2}{4\pi k B_j R_j^2 L_{aj}}$$

where $P_j G_j$ is the effective radiated power of the jammer, F_j is the pattern propagation factor of the radar receiving antenna in the jammer direction, B_j is the bandwidth of the jamming signal, R_j is the jammer range from the radar, and L_{aj} is the one-way propagation loss for the jammer.

Barrage jamming can be very effective in screening quiet (nonjamming) penetrator aircraft from detection. A typical tactic is for one or more high-power barrage jammers to stand off at a range beyond the engagement range of the defensive weapon system, and to screen a high-speed, relatively low-RCS penetrator into its weapon release range. The range at which the victim radar can detect the screened target, called the *burnthrough range* is given by the equation:

$$R_{bt} = \left(\frac{P_{av} t_o G_t G_{rj} \sigma F^2 R_j^2 L_{\alpha j}}{4\pi \left(\frac{P_j G_j}{B_j} \right) D_0(1) L_s} \right)^{\frac{1}{4}}$$

where P_{av} is the radar average power, t_o is the radar observation time, G_t is the radar transmit antenna gain, G_{rj} is the radar receive antenna gain in the direction of the jammer, σ is the penetrating target's RCS, F is the radar propagation factor, $L_{\alpha j}$ is the one-way atmospheric attenuation suffered by the jammer, $P_j G_j / B_j$ is the jammer power spectral density (W/Hz), $D_0(1)$ is the SNR required for the radar to detect the Swerling Case 1 target with a given probability of detection and probability of false alarm, and L_s is the total of the radar system losses.

In angle jamming, the barrage scan-frequency jamming technique is used, in which noise or repeater jamming is amplitude-modulated with a band of audio frequencies covering the possible conical-scan rate of the victim radar. A portion of the resulting AM spectrum, at the radar's actual scan frequency, then enters the tracking servo to cause tracking noise. The technique is less effective than inverse-gain jamming or AM at the actual scan frequency because the noise within the servo bandwidth is only a fraction of the total. However, when the radar uses conical scan on the receiving beam only (COSRO), the actual scan frequency may be unknown to the jammer, and barrage scan-frequency jamming may be the only practical countermeasure. *DKB, PCH*

Ref.: Barton (1988), pp.139–140, 494; Goj (1993), p. 45.

Bistatic jamming is created by reflection from sources displaced in space from the jamming transmitter. It can be created from a single jammer platform by illuminating an external object with a directional jamming beam. Then the radar sees the jamming source at or near the illuminated objects. If these objects lie within the same resolution cell as a target, accurate tracking of the target becomes impossible. Possible external objects are chaff, towed or expendable decoys, or the surface of the earth. The special case of bistatic jamming is use of the surface of the ground that results in ground-bounce jamming. The radar ECCM against bistatic jamming is basically better resolution. *SAL*

Ref.: Barton (1989), p. 502.

Blinking jamming is noncoherent multiple-source jamming done with several (basically two) jamming transmitters by turning them on one at a time. This kind of jammer is considered an effective electronic countermeasure against guided missiles using radar seekers. When the jamming transmitters are located on different platforms the main aim of this jamming is to make a missile wander from one target to another, as a result going between the platforms without hitting them. To ensure the effectiveness of this jamming, the correct choice of the commutation rate must be made. Too high a rate causes tracking radar to average the data (and, therefore, angular error is minimized), and too low a rate allows the radar to determine the real position of each target. The best solution is when the rate is of the order of the tracking servo's bandwidth (about 0.1 to 10 Hz). *SAL*

Ref.: Schleher (1986), p. 155; Maksimov (1979), p. 55.

Chaff jamming is passive jamming using chaff as a deceptive measure. (See **CHAFF**.) *SAL*

Cooperative jamming is jamming that requires the cooperation of two platforms, each possessing either a deception or noise jammer. The typical example of cooperative jamming is blinking jamming. Cooperative jamming requires a radio link between two platforms to assure that the emissions are synchronized. *SAL*

Ref.: Neri (1991), p. 385.

Count-down jamming is jamming using an on–off type jamming signal with frequency and duty cycles to prevent the AGC of radar receiver from being at the right level. To increase the effectiveness, the jamming duty cycle is changed periodically (this fact is reflected in the name of the jamming, as originally a counter was used to perform a count-down to determine the period of variation of the duty cycle). *SAL*

Ref.: Neri (1991), p. 381.

Cover-pulse jamming is a self-defense noise-jamming technique in which the noise envelope is increased slowly from zero at a time estimated to precede the arrival of the radar pulse at the target, forcing the radar CFAR threshold (or AGC level) to increase enough to suppress the target echo. The level is decreased again to zero after arrival of the pulse. When properly implemented, the radar operator may not recognize that jamming has occurred, and no jam strobe is generated. The duration of the cover pulse must be sufficient to counter the effects of variable pulse repetition interval, since initiation of the jamming is based on the time of arrival of the previous pulse. *DKB*

Ref.: Schleher (1986), p. 145.

Cross-eye jamming is coherent multisource jamming "whereby two beams of energy are transmitted in the direction of the target; in a manner so that the beams are crossed between the target and the transmitters. With such a technique it is difficult for the target to determine the points from which the transmissions are originating" (Fig. J2).

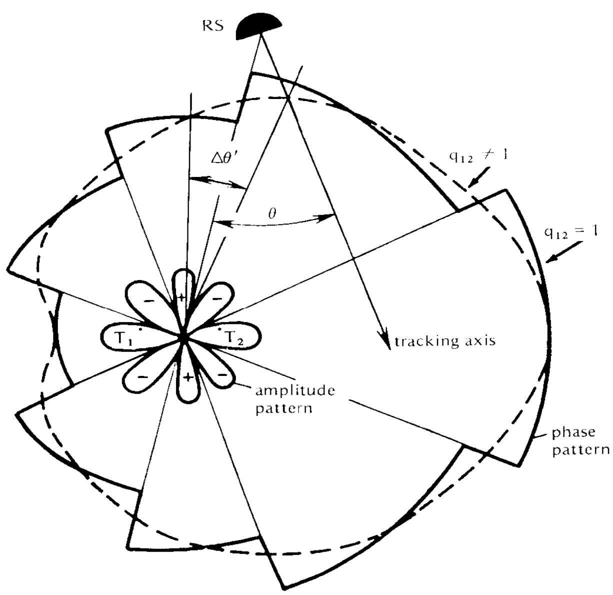


Figure J2 Wavefronts produced by a cross-eye jammer (from Maksimov, 1979, Fig. 2.16, p. 60).

In principle the cross-eye jammer is a source of enhanced glint error, and it attempts to create the situation when signals from two equal target sources arrive at the radar antenna in phase opposition. It can be done by radiating coherent signals from the single platform but from two separated antennas (e.g., on the wingtips of the aircraft). Properly designed, this kind of jamming can be very effective against all types of tracking radars. Sometimes this technique is called *two-point coherent jamming*. (See also **ECM, cross-eye**). *SAL*

Ref.: Johnston (1979), p. 57; Barton (1989), p. 501; Schleher (1986), p. 158.

Cross-polarization jamming is jamming based on the effect of cross-polarization and retransmitting a signal which is polarized orthogonally to the polarization of the radar transmission. The cross-polarized signal, when received by a parabolic or shaped reflector antenna, introduces large angle errors in tracking radars. (See also **ECM, cross-polarization**). *SAL*

Ref.: Barton (1991), p. 12.7; Leonov (1986), pp. 238–253.

Continuous-wave jamming is jamming using a CW carrier waveform. It can be either unmodulated or modulated. Unmodulated CW jamming is typically used against bandpass radars with limited tuning capability and employs the transmission of a high-power carrier frequency with the aim of overloading the radar receiver. Modulated CW jamming is a carrier waveform modulated with some other signals such as *noise* or signals with high, medium, or low frequencies, typically to be used as deceptive active jamming. Amplitude, frequency, phase, or pulse modulation can be used. *SAL*

Ref.: Johnston (1979), pp. 60, 63, 68.

Deception [deceptive] jamming is intended to insert false information into a victim radar. It is typically classified either as active deception jamming, wherein the erroneous information is provided by the generation of false signals that are similar to the signals that radar expects, or passive deception jamming, wherein erroneous information is provided by reflections from false targets. Deception jammers can be CW deception jammers (coherent or noncoherent types, Fig. J3) or pulse deception jammers (Fig. J4) depending on the type of the waveforms produced. If the jammer receives, modulates and retransmits each radar pulse, it is called a *repeater jammer*.

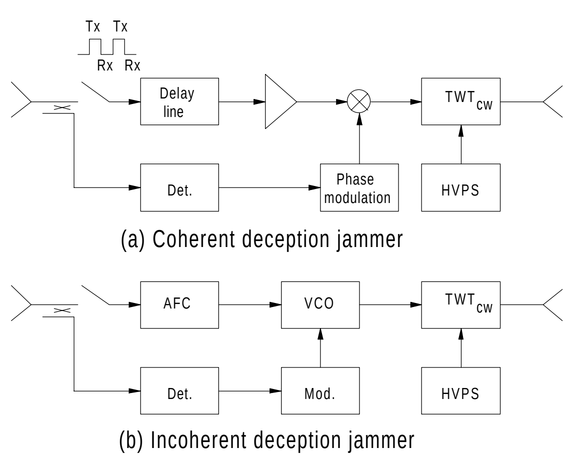


Figure J3 CW deception jammers (after Neri, 1991, Fig. 5.16, p. 359).

Active deception jamming against radars usually is divided into deception jamming against search, angle track-

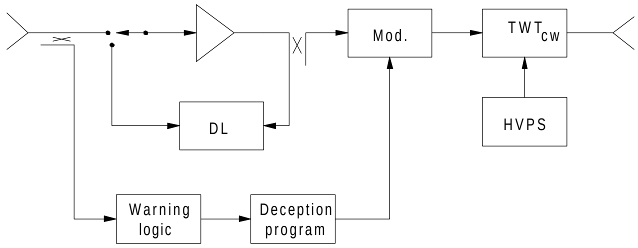


Figure J4 Pulse deception jammer (after Neri, 1991, Fig. 5.12a, p. 353).

ing, range tracking, or velocity tracking systems. Against search systems deception jamming has to imitate the multiple false targets to oversaturate the radar output. Against automatic tracking systems the intended effects are to cause incorrect resolution of signals, introduce the unacceptably large error in radar measurements, and finally to disrupt the operation in the automatic tracking mode (in angles, range or velocity coordinates). The main active deception jamming techniques against angle tracking systems are blinking jamming, cross-polarization jamming, cross-eye jamming, intermittent jamming, scan-frequency jamming, and surface-bounce jamming, and, against range tracking systems, range-gate pull-off, and against velocity tracking systems is velocity-gate pull-off.

Active deception jammers are usually more sophisticated than noise jammers as the jammer's performance characteristics must be more closely matched to the type of the system to be jammed. Passive deception jamming is performed with decoys and chaff. (See also **ECM, deception**). *SAL*

Ref.: Barton (1988), p. 140; Maksimov (1979), pp. 46–76; Neri (1991), pp. 353–365.

Doppler radar jamming is jamming in doppler (relative velocity), which is basically achieved by modulating a signal in phase (or frequency) or in amplitude to generate multiple sidebands that represent false dopplers when processed in the radar receiver. *SAL*

Ref.: Chrzanowski (1990), p. 84.

Down-link jamming is jamming that is intended to screen the missile beacon signal from the view of the radar tracking the missile. (See also **ECM, down-link**). *SAL*

Ref.: Johnston (1979), p. 58; Chrzanowski (1990), p. 162.

jamming equation (see **RANGE EQUATION, with jamming**).

Escort(-screening) jamming (ESJ) is jamming in which the vehicle with a jammer accompanies the strike vehicles (typically both vehicles are aircraft) and jams radars to protect the strike vehicles. The electronic device providing escort jamming is called an *escort jammer*. Typically, it is a high-power emitter on board a tactical aircraft which accompanies and screens a group of penetrating attack aircraft. This kind of jammer usually employs deception jamming or barrage noise jamming techniques. *SAL*

Ref.: Skolnik (1990), p. 9.6; Lothes (1990), p. 3.

False target jamming is jamming that transmits replicas of the radar pulses so the received pulses appear to be targets. If the replicas are received through the antenna sidelobes, the angular location of the false target appears to be very different from that of the jamming source and many different targets can be created with different angular locations. The main methods to reduce the effectiveness of false target jamming are to use waveforms that are rather difficult to repeat or to use an auxiliary antenna and compare the energies received to determine whether the source of the jamming within or outside the antenna mainlobe. *SAL*

Ref.: Nitzberg (1992), p. 22.

jammer finder (see **STROBE**).

Frequency-swept jamming is used when the victim radar's instantaneous bandwidth is such that the noise jammer bandwidth cannot continuously cover it, or when more than one frequency is used by one or more victim radars, either simultaneously or at a rate that the jammer cannot follow. The jammer may then be forced to "sweep" the jamming energy, in a "spot" bandwidth B_j, over the potential range of radar operational frequencies. The resultant frequency-swept jamming will have a duty cycle of less than 100-percent, and the jammer effectiveness will be reduced in proportion to the enforceable jamming duty cycle.

The term "frequency-swept jamming" may also apply to a deceptive ECM technique used against coherent doppler radars, missile seekers, and fuzing systems. Referred to as velocity gate pull-off (VGPO), or a velocity gate stealer, the self-screening jammer (SSJ) repeats the victim's amplified signal, developing a high jammer-to-signal (J/S) ratio, "captures" the victim's velocity gate, and pulls it away from the true target doppler. The jammer thus causes a false doppler measurement or forces the victim radar or seeker to break-lock and initiate a reacquisition process. (See **ECM, velocity measurement**.) *PCH*

Ground-bounce jamming is jamming based on multipath propagation to prevent the radar from obtaining adequate data about the jamming platform. In this case the jammer located on the protected vehicle (typically aircraft) transmits the energy toward the surface of the earth in a manner that reflects the energy from the surface toward vehicle with the victim radar (typically an interceptor) (Fig. J5). This technique is primarily useful against low-flying vehicles. Some-

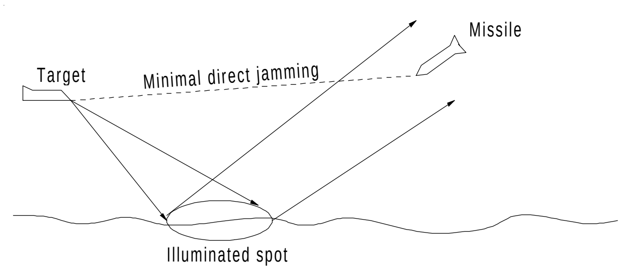


Figure J5 Ground-bounce jamming.

times this kind of countermeasure is called *surface-bounce jamming* or *terrain-bounce jamming*. *SAL*

Ref.: Barton (1989), p. 530; Schleher (1986), p. 160; Chrzanowski (1990), p. 155.

Image-frequency jamming is jamming penetrating a receiver through the radar image channel, separated by twice the intermediate frequency from the intended response. As an ECCM, use of RF preselector is successful. *SAL*

Ref.: Barton (1988), p. 499.

Imitative jamming is "the jamming technique of transmitting a signal identical to the original guidance signal." This is accomplished through the use of a repeater jammer. *SAL*

Ref.: Johnston (1979), p. 61.

Intermittent jamming is jamming that is not continuous. Intermittent jamming may occur unintentionally as a consequence of electromagnetic interference (EMI) from one or several sources of RF energy at the victim radar frequency or one of its harmonics. Intentional jamming sources that may be intermittent in nature include frequency-swept jamming and other jamming techniques designed to attack one or more functions of a fire control radar that may be sequential or periodic in nature, such as those functions related to weapon system guidance and control.

One type of intermittent jamming is single-source blinking jamming, consisting of periodic bursts of rectangular pulses radiating by the same jamming transmitter. This type of jamming affects the receiver with AGC using the transient processes that take place in an AGC system when its input is acted upon by strong pulse signal. It overloads the receiver and interrupts the transmission of information to the tracking channel reducing the average gain of an angle tracking system. The main ECCM solution to this type of jamming is to extend the dynamic range of the receiver to the value over which useful output can be obtained, and to use fast AGC or logarithmic amplifiers to minimize the time spent in saturation mode. *PCH, SAL*

Ref.: Barton (1989), p. 496; Maksimov (1979), p. 57.

Intermediate frequency jamming is "the form of continuous-wave jamming that is accomplished by transmitting two CW signals separated by a frequency equal to the center frequency of the radar receiver IF amplifier." *SAL*

Ref.: Johnston (1979), p. 62.

Inverse-gain jamming is jamming that repeats the replica of the received signal with an induced amplitude modulation that is the inverse of the victim radar's combined transmitting and receiving antenna scan patterns. It was designed to be used against conical-scan radars where it inserts an effect similar to positive feedback, and this effect pushes antenna away from the target rather than to the target. The jammer employing such a technique is called an *inverse-gain repeater jammer*. (See also **GAIN, inverse**). *SAL*

Ref.: Skolnik (1990), p. 9.6; Johnston (1979), p. 62; Schleher (1986), p. 11.

Jammer look-through is the mode of jammer operation whereby periodic monitoring of the threat environment is provided with an objective of determining its own effectiveness, for example, correct tuning and sufficient power level. This is usually accomplished by using an intercept receiver that can be an integral part of or complement the jammer itself. *SAL*

Ref.: Schleher (1986), p. 133.

Mechanical jamming is the term sometimes used for chaff. *SAL*

Ref.: Johnston (1979), p. 63.

Modulated jamming is a jamming signal obtained by modulating a CW signal. The main types of modulation used in jamming are pulse modulation, sinusoidal modulation, and sawtooth modulation. *SAL*

Ref.: Johnston (1979), pp. 63, 65, 66.

Multiple-source jamming is jamming employing spatially dispersed jamming sources. The phase characteristics of these jammers may be either coherently or noncoherently related. In the first case a deterministic or synchronized relationship between the phase of multiple jammers is implied, and the term *multiple-source coherent jamming* is used. The representative of such a technique is cross-eye jamming. When the relationship among jamming sources is random, the technique is termed multiple-source noncoherent jamming. The major representatives of this technique are blinking jamming and formation jamming. In Russian radar literature, this type of jamming is called *multipoint jamming*. *SAL*

Ref.: Schleher (1986), pp. 154–159; Maksimov (1979), p. 54.

Multitarget generator jamming refers to use of a special generator to produce numerous targets designed to saturate the tracking capability of a radar. It is controlled by computing when to transmit false targets at the radar's frequency on the basis of information about a radar's pulse repetition frequency, scan rate, and antenna pattern. *SAL*

Ref.: Johnston (1979), p. 63.

Noise jamming has the objective of masking the actual signal. In principle, noise jamming increases the radar receiver noise level and the optimum jamming signal has the characteristics of receiver noise (that in practice may be difficult to achieve). From the point of view of the occupied bandwidth, noise jamming is classified as *barrage noise jamming* when the bandwidth extends over the entire tuning band of the victim radar, and *spot noise jamming* when the bandwidth is less than that of the radar and power is concentrated at the radar signal frequency. From the viewpoint of waveform, noise jamming is divided into continuous noise jamming and pulse noise jamming. Continuous noise jamming in its turn can be categorized as direct noise jamming (when random noise is amplified and radiated directly), and modulated noise jamming (amplitude- or frequency-modulated). Pulse jamming may be represented as a burst of RF pulses with a given duty factor and random change of pulse amplitude, duration and the spaces between the adjacent pulses (random pulse noise jamming which is difficult to produce in practice), and jamming pulses of constant amplitude and repetition frequency (regular pulse noise jamming).

Noise jamming is the most common type of ECM technique, and it is effective against most types of radars. One of the advantages of this type of jamming is that, in comparison with deception jamming, very little needs to be known about the characteristics of the victim radar to be effective. The quality of jamming noise and its effectiveness against a radar depends on the methods used for generation, modulation and amplification of transmitted power. The electronic devices emitting noise jamming are called noise jammers. The most common types of such jammers are frequency-swept noise jammers, narrowband spot jammers, barrage noise jammers, and blinking noise jammers. This type of jamming is more effective against search radars than against tracking radars. A major operational form of noise jamming is stand-off jamming. Sometimes this kind of jamming is also called *denial jamming* or *obscuration jamming*. *SAL*

Ref.: Barton (1988), p. 139; Barton (1991), pp. 5.22, 12.5; Maksimov (1979), pp. 30–46; Johnston (1979), p. 63; Schleher (1986), pp. 117–137.

The **jamming-to-noise ratio (JNR),** or J/N, is the ratio of jamming power received by the radar to background noise level in the absence of the jammer:

$$\frac{J}{N} = \frac{P_j G_j A_r F_j^2 F_p^2}{4\pi R_j^2 k T_s L}$$

where P_j is the jammer power, G_j is the jammer antenna gain, A_r is the radar receiving aperture area, F_j is the jammer pattern-propagation factor (including allowance for radar sidelobe level), F_p is the polarization factor, R_j is the jammer range, k is Boltzmann's constant, T_s is the radar system receiving temperature, and L is the total loss between the point where P_j is measured and the point where T_s is measured. *DKB*

obscuration jamming (see **noise jamming**).

Off-line jamming is "the jamming from a station away from the vicinity of the target." Sometimes this jamming is called *off-target* or *off-board jamming*. *SAL*

Ref.: Johnston (1979), p. 64.

On-line jamming is "the use of the jamming device in line with the target and the radar set." This type of jamming may be either active or passive. *SAL*

Ref.: Johnston (1979), p. 64.

On-off jamming is pulse jamming that attacks mainly the AGC function of radar receiver and forces the AGC to adjust gain on the basis of jamming bursts rather than on the target return signal. The jammer burst typically consists of a group of transponder or repeater pulses coinciding with target return pulses. Jamming bursts of duration T_{on} followed by quiet periods of duration T_{off}, and this causes the gain-controlled amplifier to cycle between the state when the amplifier output is at limit level (then all signal modulation is lost as a result of output limiting) and the state when amplifier gain is so low that the output signal is too small to be of any use (Fig. J6).

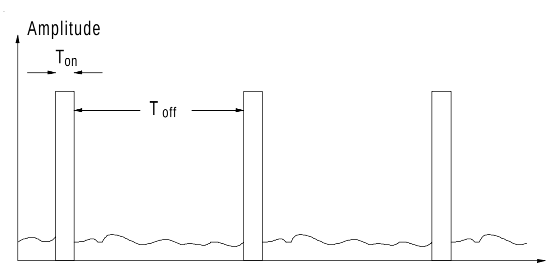


Figure J6 On-off jamming plus signal at receiver input (after Lothes, 1990, Fig. 6.6, p. 182).

Sometimes this kind of jamming is called *AGC deception*. *SAL*

Ref.: Lothes (1990), pp. 167–204.

Passive jamming is jamming employing confusion reflectors to deceive a radar by returning spurious and confusing signals. Practically passive jamming is performed with chaff and passive decoys. *SAL*

Ref.: Johnston (1979), p. 64; Maksimov (1979), pp. 63–76.

Pulse jamming is jamming using the modulation of a carrier with pulses of various widths and repetition rates. Depending on whether the law of pulse parameters variations is random or regular random pulse jamming and regular pulse jamming is distinguished. *SAL*

Ref.: Johnston (1979), p. 65; Barton (1988), p. 490.

Repeater jamming is a deception jamming technique based on confusion or deception of a victim radar, causing its equipment to present false information. Typically, it sends back an amplified version of the signal received from the radar that is stronger than the radar signal and so captures the radar's range gate. Then the deception signal is progressively delayed and the range gate walks off the actual target. The electronic device employing such a technique is called a *repeater jammer*. If the jammer uses phase information of the received signal in creating false targets, it is called a *coherent repeater jammer*. *SAL*

Ref.: Barton (1988), p. 490; Johnston (1979), pp. 56,62.; Schleher (1986), p. 11.

Scan-frequency jamming is jamming against conical-scan and other sequential scanning radars in which the errors are introduced by amplitude modulation of a noise or repeater jammer. When the scanning frequency is exactly known, the jamming modulation is approximately the same as the antenna-scanning frequency, and this kind of jamming is termed *scan-frequency selective jamming*. When the scanning frequency is not known, *scan-frequency barrage jamming* can be employed, which can be either noise or swept scan-frequency jamming. Against the monopulse radar, this type of jamming is ineffective because it uses an amplitude modulation that does not much affect this type of radar. *SAL*

Ref.: Barton (1989), pp. 493–496; Maksimov (1979), pp. 49–53.

Self-screening [-protection] jamming is designed to protect the vehicle or platform that carries the jammer from engagement by air-defense weapons systems. Self-protection jamming may use noise jamming to deny the victim radar or missile seeker range information, or it may employ a repertoire of deceptive jamming techniques to frustrate the victim's angle tracking and homing guidance system. *PCH*

Sidelobe jamming is "jamming through a sidelobe of the receiving antenna in an attempt to obliterate the desired signal received through the mainlobe of the receiving antenna at fixed points." *SAL*

Ref.: Johnston (1979), p. 66.

The **jamming-to-signal ratio (JSR** or ***J/S*)**, is ratio of the power received from a jammer (or jammers) to the target signal power received, as measured at a common point in the victim radar (e.g., at the radar antenna terminals). For the general noise-jamming case, this ratio can be expressed as

$$\frac{J}{S} = \frac{\left(\dfrac{P_j G_j A_r B_r F_j^2}{4\pi R_j^2 B_j L_{\alpha j}}\right)}{\left(\dfrac{P_t G_t A_r \sigma F^4}{(4\pi)^2 R^4 L_\alpha}\right)} = \left(\frac{P_j G_j}{P_t G_t}\right)\left(\frac{F_j^2}{F^4}\right)\left(\frac{B_r}{B_j}\right)\left(\frac{R^4}{R_j^2}\right)\left(\frac{L_\alpha}{L_{\alpha j}}\right)\left(\frac{1}{\sigma}\right)$$

where the radar parameters are identified in accordance with the radar range equation. Here the subscript j applies to the jammer parameters, r applies to the victim radar parameters, F_j includes the relative radar antenna gain in the direction of the jammer, and L_α represents the signal loss due to atmospheric attenuation (one-way for the jammer, two-way for the radar signal). (See also **barrage noise jamming**, **RANGE EQUATION, with jamming**.) *PCH*

Skirt-frequency jamming is jamming that can be used against monopulse radar based on the fact that sum and difference channels must be well matched in phase so that the error detector can preserve the polarity of the target error voltage. An error in receiver phase matching causes a shift in the tracking point, and if error exceeds 90° the tracking loop may settle on the previously unstable null, more than one beamwidth away from the desired tracking axis. A jamming technique that exploits this problem is called *skirt-frequency jamming*, because it produces a signal lying on the skirts (edges) of the IF passband. *DKB, SAL*

Ref.: Barton (1989), p. 497.

A **smart noise jammer** is a hybrid type of jammer that incorporates the features of both noise and deception jamming. In this case, repeater-type jammer generates responsive noise in a transponder mode. A noise burst occurs before and after the actual target return and covers the true echo. The span of ranges where the noise is radiated is synchronized to the victim radar and the jammer may adapt its directive beam to the angles of arrivals of radar illumination. *SAL*

Ref.: Barton (1988), p. 367; Schleher (1986), p. 11.

Spot jamming is "the jamming of a specific channel or frequency." *SAL*

Ref.: Johnston (1979), p. 67.

Stand-forward [stand-in] jamming is jamming emitted from a platform located between the weapon system and the strike vehicle. Since the stand-forward jammer usually is located within the lethal range of defensive weapon systems for a considerable time, the use of remotely piloted vehicles as a platform for jammer is practical. *SAL*

Ref.: Skolnik (1990), p. 9.6; Chrzanowski (1990), p. 82.

Standoff jamming is jamming emitted from a support platform (typically an aircraft) orbiting within line-of-sight of a victim radar but beyond the range of the defensive system. When the protected vehicles start their penetration of the combat area, the support vehicle can direct jamming against all significant radars in this area. In principle it is a generic term applied to standoff ECM, escort ECM, and cooperative ECM. The term *support jamming* can be applied to this or any other offboard jamming technique. *SAL*

Ref.: Johnston (1979), p. 67.

straddle jamming (see **two-signal jamming**).

Support jamming is any type of jamming originating on a platform other than the radar target.

surface-bounce jamming (see **ground-bounce jamming**).

A **swept jammer** is a jammer that sweeps a narrow band of electronic energy over a broad bandwidth. A transmitter in which a narrowband signal can be tuned over a broad frequency ban and then locked on a particular frequency is called a *sweep-lock-on jammer*. *SAL*

Ref.: Johnston (1979), p. 67.

Two-frequency jamming is jamming emitted as two jamming carriers separated by the IF of the victim radar. In this case radar receivers that lack RF preselection are subject to generation of spurious outputs. The effectiveness of this method depends on the jamming level and receiver performance, in particular on the nonlinearity of input mixers. Basically the technique is applied against radar seekers that must operate at very short ranges. The main ECCM is to use RF preselection or amplitude comparison after envelope detection. *DKB, SAL*

Ref.: Barton (1989), p. 501.

two-point coherent jamming (see **cross-eye jamming**).

Two-signal jamming "is a method of jamming whereby two signals are transmitted on two frequencies only slightly separated. Effective against certain types of radars where receiver bandwidth is narrow enough to defeat noise jamming." *SAL*

Ref.: Johnston (1979), p. 67.

JITTER. (1) The noise-like fluctuations due to target echo characteristics, propagation, or receiver thermal noise that degrade the tracking accuracy and result in zero-mean random errors in successive target position measurements. Sometimes is termed also tracking noise. (2) "Intentional variation of a radar parameter, for example, pulse interval." *SAL*

Ref.: IEEE (1990), p. 19; Skolnik (1980), p. 167.

JOINT, microwave. Microwave joints are connections of two transmission lines of the same type, providing reliable electrical contact. The joints must ensure retention of matching and power-handling capability of the circuit with minimum induced damping of power and without parasitic radiation.

In high-frequency joints (connectors) of coaxial cables, the contacts are provided using spring clamps and plugs held in the connection by external threaded or other connections.

The sections of the waveguides are connected using flanges of two types: contact and choke. The contact flanges require careful working and strict parallelness of the touching surfaces. In the choke flange the contact is made through a serial short-circuited loop (see **CHOKE, microwave**).

For connecting transmission lines that rotate relative to one another, rotary and rolling joints are used. *IAM*

Ref.: Sazonov (1988), p. 53.

A **rolling joint** is an element of a transmission line intended for turning one part of a circuit relative to another without disturbance of the electrical contact and the quality of matching. If the turning angle is great, then the design of the rolling joint is analogous to that of the rotary joint. When the turning angle is small, a set of several sections of rectangular waveguides is used. They automatically assume intermediate angular positions relative to the input and output sections of the joint. The smooth transition formed here does not cause a change in the standing-wave ratio at the point of twist of the waveguides. *IAM*

Ref.: Perevezentsev (1981), p. 214.

A **rotary joint** is an element of a transmission line intended for rotating of one part of a circuit relative to another without disturbance of the electrical contact and the quality of matching. A wave with axial symmetry is excited in rotating adapters, so either round waveguides or coaxial lines are used in such joints. Rotating and fixed parts of a line are connected using choke-flange (for waveguide lines) and coaxial choke connections (see **CHOKE, microwave**), which provide the electrical contact between moving surfaces. *IAM*

Ref.: Sazonov (1988), p. 55; Pereverzentsev (1981), p. 213.

JOSEPHSON EFFECT. The Josephson effect is the tunneling of linked pairs of electrons through an insulating barrier between two superconductors. The accumulation of current in the presence of critical bias voltage has an extremely pronounced nature, which determines the pronounced inflection of the volt-ampere characteristic. The Josephson effect is used in millimeter signal mixers. The Josephson mixer is distinguished by a low level of shot noise (see **NOISE, shot**) and low power of the control oscillator (less than 1 μW). *IAM*

Ref.: Tucker, J. R., *Appl. Phys. Lett.* ***36***, no. 6, 1980.

K

The **KABANOV EFFECT** is the phenomenon wherein HF radio waves, after reflection from the ionosphere, are dispersed after they have reached Earth's surface. Some part of the dispersed radiation is returned by the reverse path to the source of the radiation where it can be detected. The reverse-dispersed signals can be picked up at ranges from several hundreds to thousands of kilometers. Over-the-horizon (OTH) radar operation is based on the Kabanov effect. *IAM*

Ref.: Druzhinin (1967), p. 114.

KLYSTRON. The klystron is a velocity-modulated microwave linear-beam tube using one or a series of cavities and a slow-wave structure to achieve power generation and amplification. It was invented in 1939 and was initially relegated to the role of a local oscillator in superheterodyne receivers.

The klystron has proven to be quite important for radar applications. The main advantages of this device are high average and peak power (tens of megawatts of peak power are available); high gain (26 to 60 dB is conventional, and even 90 dB was reported); high efficiency (35 to 75%); low interpulse noise; and good compatibility with sophisticated pulse-compression waveforms. The main disadvantages are relatively low bandwidth, compared with traveling-wave tubes (10 to 12% for multicavity klystrons); relatively high operating voltages; and large physical size.

Klystrons are manufactured for wavelengths from 70 to 80 cm to several millimeters and are used primarily as high-power oscillators and amplifiers in radar transmitters, and for low-power local oscillators in receivers (typically in the centimeter wave bands).

Some types of high-power klystrons are listed in Table K1. *IAM*

Ref.: Lavrov (1974), p. 348; Andrushko (1981), pp. 18, 23; Gilmour (1986), hap. 9; Ewell (1981), pp. 54–64.

The **antiklystron** is an electronic vacuum microwave device that uses centrifugal electrostatic focusing of the electron beam. One feature of the antiklystron is the use of a system with multiple interaction of the rotating stream of electrons in the HF field (the beam passes several times through one and the same region of interaction with the field). Thanks to the multiple handling of the electronic beam, “accumulation of modulation” occurs in the system. In the antiklystron, the preceding turns are directly connected to subsequent ones through the beam itself, and there is a feedback circuit of the subsequent turns with the preceding ones through the HF field. The feedback circuit is possible when the grouped electrons enter the braking phase of the HF field. Here there will be an increase in the energy of the HF signal due to the energy of the electron beam.

The antiklystron can operate as an oscillator with the ability to tune over a broad band. *IAM*

Ref. Popov (1980), p. 39.

Table K1
Some Commercially Available High-Power Klystrons

Tube type	Center frequency (GHz)	Peak power (MW)	Frequency range (GHz)	Max duty cycle	Peak		Gain (dB)
					Voltage (kV)	Current (A)	
VA-1513	0.4–0.45	20	0.015	≈0.015	230	280	40
TV2023	1.3	40	1.2–1.4	0.001	300	230	53
F2049	2.856	30	Fixed	0.008	290	295	50
VKS-8262	2.9	5.5	0.005	0.001	125	88	50
SAC42	5.65	3.3	5.4–5.9	0.002	135	112	23.5
SAX191	9.015	1.25	8.83–9.2	0.0048	85	50	50

(from Ewell, 1981, Table 2-6, p. 61, reprinted by permission of McGraw-Hill).

An **extended interaction klystron** uses an output circuit with extended interaction, which has several interaction spaces for extraction of beam energy. Such an output circuit may be viewed as a series of successive and interrelated cavities, each of which is associated with the beam. As a result, the efficiency and bandwidth are increased. However, since there is a connection between several zones in a chain of extended interaction cavities, and each of them is connected to the beam, it is hard to obtain oscillations in the traveling-wave mode. *IAM*

Ref.: Gilmour (1986), pp. 203, 235, 315-316; Skolnik (1970), p. 7-34.

A **floating-drift klystron** uses two or more cavities (Fig. K1). Operation of a double-cavity klystron consists of forming a beam of electrons with a electron gun and passing the beam through the gap between the walls of the input cavity, the drift space, and the output cavity, after which the electrons go to the collector. In the first cavity, RF oscillations are excited by an external signal. In the drift space, the electrons are grouped into clusters through velocity modulation. During passage of the second cavity, the clusters of electrons impart RF oscillations of the same velocity, and these are transmitted over the line to the load. When there is positive feedback between the input and output cavities, the system operates as an oscillator with self-excitation.

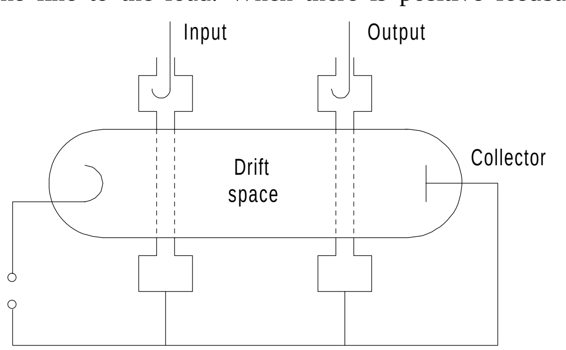


Figure K1 Double-cavity, floating-drift klystron (after Rakov, 1970, Fig. 1.19, p. 49).

To increase the output power, the floating-drift klystron is made as a multicavity klystron, with several parallel electron beams, each of which is formed by its electron gun.

Floating-drift klystrons can operate either in the continuous or in the pulse mode. Powerful pulsed floating-drift klystrons are used in radar amplifiers and oscillators, charged-particle accelerators, and so forth. Continuous-wave floating-drift klystrons are used on microwave communications systems. *IAM*

Ref.: Popov (1980), p. 324; Andrushko (1981), p. 36; Rakov (1970), vol. 2, p. 51.

A **gyroklystron** is a gyrotron amplifier consisting of several resonant cavities in which cyclotron resonance interaction occurs. At the input cavity the electrons are modulated by the input RF signal resulting in bunching, and further interaction and amplification occurs in the following cavities. Typically, gyroklystrons are designed to operate in the millimeter-wave band and have higher gain, higher efficiency, and higher power than the gyro-TWT but less bandwidth. At 28 GHz, peak power of 65 kW, gain of 30 dB, 10% efficiency, and 0.2% bandwidth were achieved.

Drawbacks of gyroklystrons include the self-excitation of parasitic modes and the relatively small amplification band (no more than 1%). The output power in the stable amplification ode does not exceed 65 kW of continuous power with a gain coefficient of 40 dB and 9% efficiency in a range of 28 GHz. *SAL*, *IAM*

Ref.: Currie (1987), p. 469; Gilmour (1986), p. 430.

A **multiple-beam klystron** uses several beams and cavities in the form of a periodic structure and was developed for producing high power at a given operating voltage.

To broaden the bandwidth of a multibeam klystron, the resonance nonperiodic structure is replaced with a traveling-wave structure. In contrast to the multibeam traveling-wave tube, the structure of the traveling wave in such a klystron is at right angles to the beams rather than along them.

The low supply power of the solenoid, which creates the focusing field, is another advantage of the multibeam klystron. Its drawbacks include its complexity and low reliability.

Multibeam klystrons are usually used for supplying a group of phased-array antenna elements. *IAM*

Ref.: Skolnik (1970), p. 7-36; Andrushko (1981), p. 36.

A **multicavity [multiresonator] klystron** is a floating-drift klystron with several (usually four or six) cavities. It is marked by a high gain coefficient and high efficiency.

In the three-cavity klystron, in comparison with the two-cavity one, the efficiency is improved from 15 to 35%. A further increase in the number of cavities does not lead to a significant increase in the electronic efficiency, but increases the gain coefficient (to 70 dB in the four-cavity klystron) and changes the amplitude-frequency characteristic. Passbands up to 10% are achieved by detuning the cavities.

Multibeam klystrons are also often referred to as *multicavity klystrons*. *IAM*

Ref.: Andrushko (1981), p. 31; Rakov (1970), vol. 2, p. 54; Gilmour (1986), pp. 209, 235.

A **multiplier klystron** is a floating-drift klystron whose output cavity is tuned to a higher frequency than the input. The current of the electron beam in the klystron contains a high number of harmonics, thanks to which effective multiplication by a factor of 10 to 20 is possible. The efficiency of the multiplying klystrons usually amounts to units of percentage, and the output power is on the order of 1W. *IAM*

Ref.: Rakov (1970), vol. 2, p. 56.

A **reflex klystron** uses one cavity and has a special electrode-reflector that has a negative potential (Fig. K2). The cavity performs two functions simultaneously, modulation of the electron beam in speed for formation of clusters of electrons, and conversion of its kinetic energy into RF electromagnetic energy. The electrons emitted by the cathode pass through cavity grids and are modulated in speed, then move first in the direction of the reflector, and then under the influence of its braking field, in the opposite direction, and intersect the gab between the resonator grids a second time. By this time the electrons are grouped in clusters and in the braking phase of the variable voltage between the grids of the resonator, they give up their kinetic energy to the HF field of the cavity, and through the output to the load.

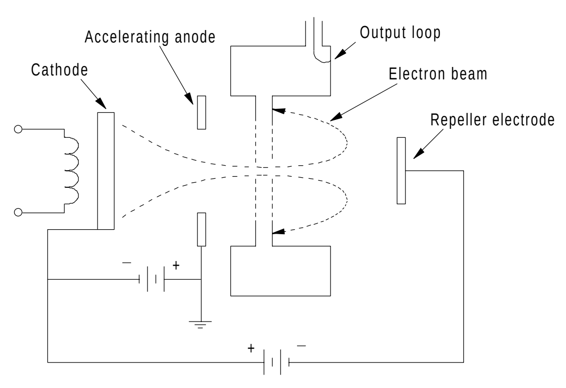


Figure K2 Reflex klystron principles (after Gilmour, 1986, Fig. 9.45, p. 236).

Reflex klystrons are used in the centimeter and millimeter bands, usually as local oscillators of microwave receivers, as measurement oscillators (see **OSCILLATOR, klystron**). An advantage of the reflex klystron is its electronic tunability (i.e., by changing the voltage at the reflector). *IAM*

Ref.: Popov (1980), p. 272; Andrushko (1981), p. 37; Gilmour (1986), pp. 235-238.

L

A **LENS** is a body of a specific geometric form made of a dielectric via which an electromagnetic wave is propagated with phase velocity v_ϕ differing from that of the identical wave in free space c (speed of light). It is known from optics that, if a beam strikes the surface of the division of two media with permittivities ε_1 and ε_2, respectively, then angle of refraction β can be found from this ratio (Fig. L1):

$$\sin\beta = \frac{n_1}{n_2}\sin\alpha$$

where n_1 and n_2 are the indices of refraction, $n_1 = \sqrt{\varepsilon_1}$, $n_2 = \sqrt{\varepsilon_2}$, and α is the angle of incidence.

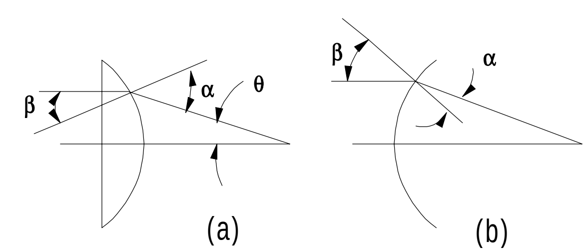


Figure L1 Ray directions in a lens: (a) convex lens, (b) concave lens.

Lenses, along with reflectors, are used as collimating elements in microwave antennas. A comparison of the basic features of lenses and reflectors is given in Table L1. In general, lenses are used in applications when reflectors cannot provide the required performance, and reflectors are used when possible. In phased arrays, a lens is more versatile than a reflector and can be competitive especially when wide-angle scanning is required.

In antenna technology, lenses with this index of refraction are used:

Table L1
Comparison of Reflector and Lens Collimators

Feature	Reflector	Lens
Degrees of freedom	1	Up to four*
Aperture blockage	Present	Absent
Losses	Lower	Higher
Chromatic aberration	No	Yes
Supporting structures	Simpler	More complex
Weight and size	Less	More

* (1) inner surface, (2) outer surface, (3) index of refraction, (4) inner-vs.-outer radiator position (in constrained lens).

$$n_2 = \sqrt{\varepsilon_2} = c/v_\phi > 1, \text{ and } \quad n_2 = c/v_\phi < 1$$

The surface of a lens facing toward a radiation source is referred to as being illuminated, while one facing away is referred to as being in shadow. It is evident from the layout of Fig. L1 that the illuminated side must be convex when $n_2 > 1$ and concave when $n_2 < 1$, to obtain a plane wave in the lens. Lenses where $n_2 > 1$ are called delay lenses (Fig. L1a); those where $n < 1$ acceleration lenses (Fig. L1b). Basic lens configurations are shown in Fig. L2.

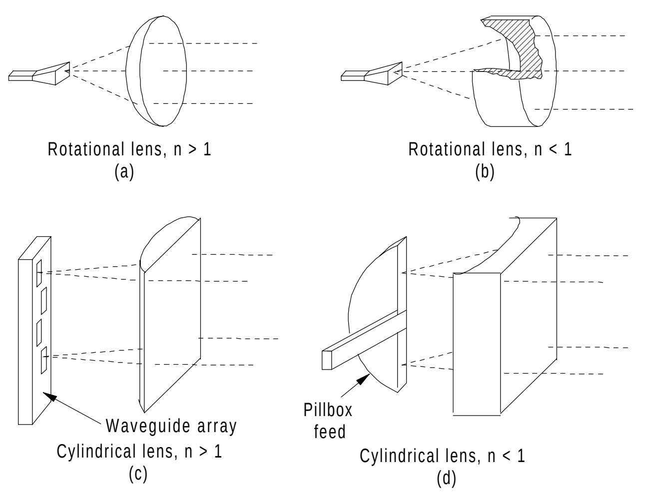


Figure L2 Basic lens configurations (after Johnson, 1984, Fig. 16.1, p. 16.3).

Lenses are classified based on a series of features. They are categorized as homogeneous and heterogeneous based on their dielectric property. If the index of refraction in the entire volume of the lens is identical, such a lens is called homogeneous. In the opposite case, it is referred to as heterogeneous. If the lens index of refraction changes under the effect of any control signals, the lens is referred to as a controllable lens. Lenses are referred to as dielectric, metal-plate, and metal-air, depending upon the lens material used. Lenses are categorized as focusing, reflection, and divergent based on the task accomplished. Focusing lenses are those converting a wave with a nonplanar front into one with a plane phase front. Those that rereflect electromagnetic waves striking them in a given direction are referred to as reflection lenses. Lenses with a heterogeneous dielectric with special profile converting a spherical wave into a wave with any form of phase front are called *divergent lenses*. Lenses are used widely in variable-purpose radar antennas. Their use in millimeter-wave-band antennas is very promising. (See **ANTENNA, lens**).

Phased array antennas in the form of transmission lenses made up of phase-shifting elements are also used in multitarget and multifunction radars. (See **FEED, antenna.**) *AIL*

Ref.: Zelkin (1974), pp. 27–191; Johnson (1984), Ch. 16.

The **Archer lens** is a variant of the Rotman lens in which the parallel-plate region between the feeds is filled with a dielectric material to reduce the linear dimensions. Stripline or microstrip lens and feed-port connecting lines are printed on an extension of the dielectric material. The beam angle can differ from the feed angle, providing large beam scan angles with practical feed positions or, for limited scan angles, reducing the lens size. An example of a 20-radiator, 16-beam microstrip Archer lens is shown in Fig. L3. *SAL*

Ref.: Johnson (1984), pp. 20–21.

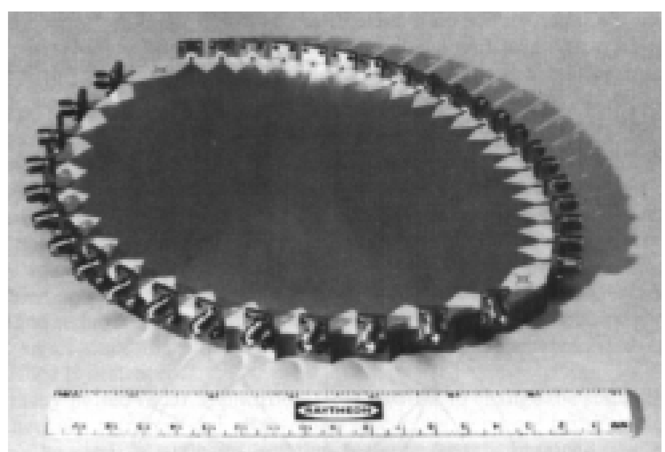


Figure L3 Archer lens (from Johnson, 1984, Fig. 16.17, p. 16.21, reprinted by permission of McGraw-Hill).

A **bootlace-type lens** is one having four degrees of freedom, providing four focal points to obtain wide-angle performance with multiple beams or a scanning beam. Most designs fix the outer surface and use three focal points. Bootlace lenses can be two- or three-dimensional, and they can produce fan beams or multiple pencil beams. The most common types are the Rotman and Archer lenses. *SAL*

Ref.: Johnson (1984), pp. 16.19–16.22.

Computing lens is a term sometimes used for a lens that forms beams at low power that will be transmitted at high power from amplifiers placed at the output (element) ports of the lens. It can take the form of a Luneburg lens or a bootlace lens. *SAL*

Ref.: Skolnik (1980), p. 316.

A **dielectric lens** is made of homogeneous dielectric with a curvilinear refracting surface. Figure L4 depicts a dielectric lens illuminated by a spherical wave from source F (feed).

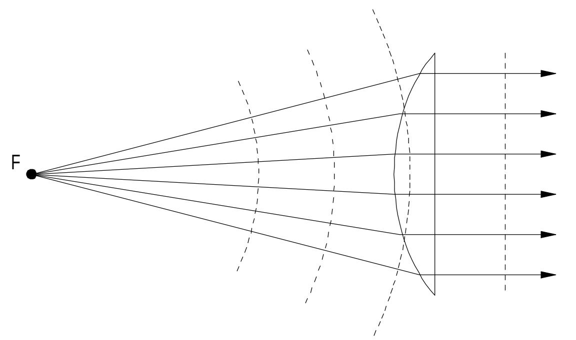


Figure L4 Dielectric lens (after Leonov, 1986, Fig. 2.6, p. 19).

The index of refraction of such a lens is $n_2 = c/v_\phi > 1$, where v_ϕ is the phase velocity of the wave in the lens and c is the speed of light. The path the wave travels in the lens is longer in the center of the lens and shorter at its edges. Consequently, a plane front is obtained by virtue of the movement of the wavefront in the direction from the feed to the dark surface. Dielectric lens shortcomings, which include their great weight and expense, limit their use in radars. *AIL*

Ref.:Skolnik (1970), pp. 10.19–10.22; Leonov (1984), p. 19.

A **dome-lens** is a metal-plate waveguide lens with round waveguide sections filled with dielectric, each of which is calculated to create a specific magnitude of phase shift. The lens is a hemispherical dome over a flat phased-array antenna with circular aperture. Waveguide sections are inserted into the holes on the dome surface. The magnitude of the phase shift in the dome-lens changes from section to section and depends on the combination of the magnitudes of the permittivity of the filler and the depth of packing of the round waveguide section. Given 20 discrete phase increments (20 types of phase shifters) in the lens under examination, 80 such combinations were used to cover the phase angle interval from 0 to 360°. A dome-lens makes it possible to scan an entire hemisphere using only one phased-array antenna (i.e., the scan sector is expanded significantly). Moreover, use of such a dome makes it possible to greatly reduce antenna system cost. Without such a dome, at least three phased array antennas would be required for scanning in the upper hemisphere. A shortcoming of a dome-lens antenna is rotation of the polarization plane of the transmitted signal with change in azimuth scanning angle. This is observed when signal polarization is linear. Use of circular polarization can eliminate this effect. *AIL*

Ref.: Johnson (1984), pp. 16.23–16.25.

A **Luneburg lens** is made in the form of a hemisphere (Fig. L5) at the base of which is installed a flat reflecting surface ensuring a mirror reflection of the radiation source from point O_u to point O_u'. Thanks to spherical symmetry, its focusing ability does not depend on the direction of wave arrival. The main properties of a Luneburg lens include the fact that a plane wave falling on it is focused at a point lying on the opposite side of the surface and, consequently, a wave from point source O_u' located on the surface of the lens is converted into a plane wave during passage through it.

A Luneburg lens has a variable index of refraction. The index of refraction has the greatest value in the center of the lens and equals $\sqrt{2}$; it equals 1 on the surface of the lens. Source displacement along the surface of the lens causes corresponding displacement of the beam in the opposite direction. There are two ways to rock the beam: either by displacement of a single feed along the surface of the lens or using a large number of feeds and switching a transmitter or receiver from one feed to another. A Luneburg lens can be used in wide-angle scanning antennas. *AIL*

Ref.: Leonov (1984), p. 20; Johnson (1984), pp. 16.26–16.27.

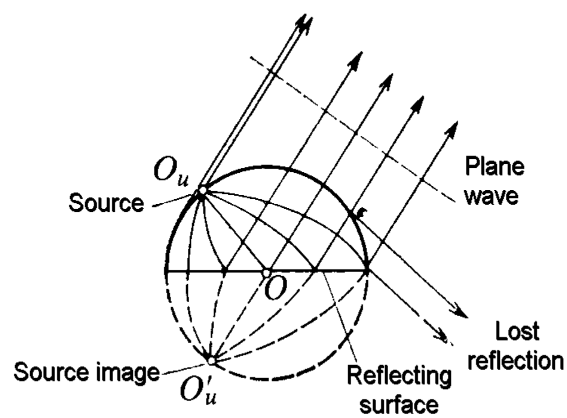


Figure L5 Hemispheric Luneburg lens (from Leonov, 1986, Fig. 2.8, p. 20).

A **metal-plate lens** is a lens made of metal plates parallel to the vector of the incident electrical field and located at distance $L_1 > \lambda/2$ from each other. In such a lens (Fig. L6a), a wave is propagated medium with phase velocity:

$$v_\phi = \frac{c}{\sqrt{1-(\lambda/2d)^2}} \qquad (1)$$

where c is the speed of light. Consequently, always $v_\phi > c$, and the lens index of refraction is $n_2 = c/v_\phi < 1$ (i. e., it is an acceleration lens).

Given appropriate selection of the concave shape of the lens, all beams exiting the radiator placed at focus F_1 will reach the aperture at the same time, and the field in the aperture will be cophasal. A shortcoming of this lens is its significant thickness, weight, and losses. Stepped metal-plate lenses (Fig. L6b) are used to decrease thickness. Thanks to the stepped shape of the profile, the length of the path of the

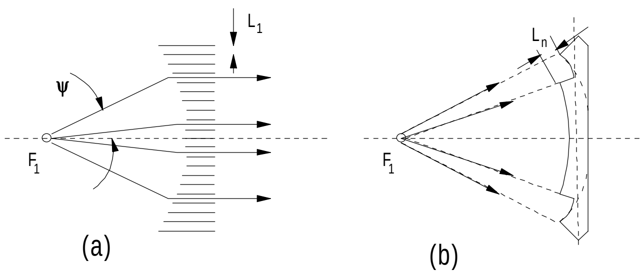


Figure L6 Metal-plate lenses: (a) waveguide, (b) stepped Fresnel (from Leonov, 1986, Fig. 2.7, p. 19).

beams in the lens is reduced. Without disruption of the phase coincidence of the wavefront at lens output, plate thickness may be decreased to the magnitude in the center of the lens and, consequently, to decrease weight and losses in the lens. Fresnel lens shortcomings include energy losses, increased level of sidelobes caused by shading due to the steps, and increased lens sensitivity to a change in frequency. *AIL*

Ref.: Leonov (1984), p. 19; Johnson (1984), pp. 16.12–16.17.

A **Rotman lens** is a metal-plate waveguide lens with several feeds. Flat waveguides are used in which the wave is propagated with a phase velocity, v_ϕ, given in. (1) above, where L_1 is the distance between waveguides ($\lambda/2 < L_1 < \lambda$). Consequently, always $v_\phi > c$ and the lens index of refraction is $n = c/v_\phi < 1$, (i. e., the lens is an acceleration lens). A Rotman lens (Fig. L7) is a device comprising an input radiator array and a waveguide lens (output array). Movement of the radiators relative to lens axis makes it possible to shape several beams at different angles to the lens axis.

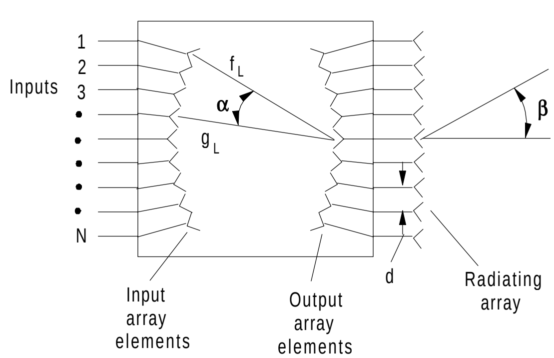


Figure L7 Rotman lens.

A Rotman lens can be used as the array feed when forming several beams to ensure a broad scanning sector. Figure L7 schematically depicts an antenna system using a Rotman lens. In this figure, f_L is the focal length off the axis, g_L is the central focal length, N is the number of radiators, d is the radiating array spacing, α is the angular position of the displaced focal points, and β is the angular direction of the phase front corresponding to the displaced focal points. The first Rotman lens models were created in microstrip form on substrates made of different materials. It always is very important to ensure the minimum degree of coupling between input and output arrays. Element coupling depends on the curve and shape of the array surface. The greater the curve, the greater the link between elements. The ratio g_L/f_L is the measure of the input grid curve, while the ratio of the depth of the output array to its length is the measure of the output array curve. (See also **ARRAY, Rotman.**) *AIL*

Ref.: Johnson (1984), p. 16.19; Barton (1988), pp. 178–179; Mailloux (1994), pp. 505–511.

The ***R-KR* lens** is a two-dimensional bootlace lens with the feedpoints and the inner surface on a circle of radius *R* and the outer surface on a radius *KR*, with equal line length between the inner and outer surface radiators. It provides nearly linear wavefront over an arc of approximately 120°. Beams covering 360° can be obtained simultaneously by adding a circulator in each line. A special case is the *R*-2*R lens. SAL*

Ref.: Johnson (1984), p. 16.22.

The **Schmidt lens** is a scanning antenna configuration in which a hemispherical reflector is combined with a correcting lens that passes through the center of the sphere from which the hemisphere was derived. *SAL*

Ref.: Johnson (1984), p. 18.13.

A **waveguide lens** is one using waveguide as the propagation medium to constrain rays to paths parallel to the waveguide axis. *SAL*

Ref.: Johnson (1984), p. 16.12.

LIKELIHOOD

The likelihood function is the conditional functional of the probability density $W(y(t)|H_i)$, determined on the condition that the hypothesis H_i is true, and viewed as a function of the number of the hypothesis i with fixed samples of $y(t)$. Presence of a class of target from which the signal $y(t)$ is received, or the value of the measured target parameter, may serve as the hypothesis.

The ratio of the values of the likelihood functions for two analyzed hypotheses is called the *likelihood ratio*. The decision is made on the basis of a comparison of the likelihood ratio with the threshold. Typically, this procedure is the basis for design of optimal (or quasioptimal) devices for measurement, detection, and discrimination of radar targets. *IAM*

Ref.: Urkowitz (1983), p. 536; DiFranco (198), p. 232; Kazarinov (1990), p. 30.

The **likelihood ratio** (Λ) is the ratio of the probability density function $W_1(y|\vartheta)$ of the received data $y(t)$ when the signal is present to $W_0(y/\vartheta)$, when the signal is absent (ϑ is an estimated parameter):

$$\Lambda(\vartheta) = \frac{W_1(y|\vartheta)}{W_0(y|\vartheta)}$$

It is used in the problems of radar detection and measurement. *(S*ee **DETECTION, likelihood ratio criterion.**) *AIL*

Ref.: DiFranco (1968), p. 232; Kulikov (1978), pp. 5–10.

LIMITER. A limiter is a "transducer whose output is constant for all inputs above a critical value." It is used to limit both RF and video pulse amplitudes. Based on type of active element, they are subdivided into diode and transistor limiters. Transistor limiters are usually used as controllable microwave amplifier-limiters. *IAM*

Ref.: IEEE (1993), p. 716; Gassanov (1988), pp. 135, 142, 143; Skolnik (1990), p. 3.30.

An **avalanche-multiplying limiter** is an uncontrolled power limiter based on avalanche microwave breakdown in the semiconductor. It is made in the shape of an element with surface-oriented electrodes in the form of counter-teeth on a semiconductor wafer. Such a planar-integrated element has a low threshold microwave power and low recovery time due to the short life span of the charge carriers in the surface region. The typical value of the output power in the limitation mode is in an integral of 0.5 to 1W with an input power that reaches 10 to 15W. The recovery time does not exceed 0.1 μs.

The avalanche-multiplying limiter is a solid-state analog of the protective resonance crowbar. Such limiters are useful for work in protective devices in the centimeter and millimeter bands. *IAM*

Ref.: Lebedev, I. V., *Radioelektronika* ***33***, no. 10, 1990, pp. 58–61.

A **bandpass limiter** is a limiting amplifier having a narrowband filter at its output, producing a sinusoidal output of controlled amplitude independent of the input amplitude. It normally consists of a symmetrical clipper followed by a linear filter that passes only the fundamental of the applied waveform. A block diagram of a receiver in which a bandpass limiter is used is shown in Fig. L8. The purpose of the limiter is to reduce the dynamic range of received echoes. *SAL*

Ref.: Nathanson (1969), pp. 119–130.

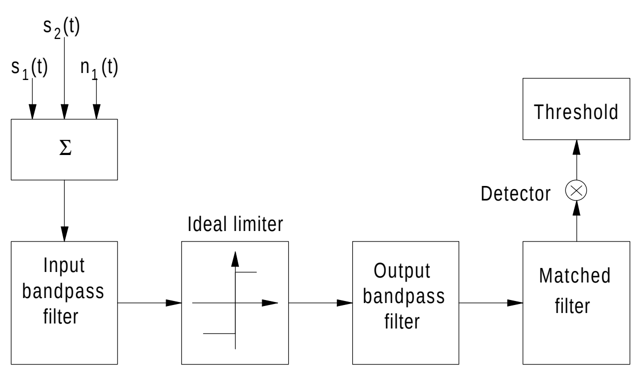


Figure L8 Block diagram of bandpass limited receiver (after Nathanson, 1969, Fig. 4.5, p. 120).

A **diode limiter** uses a diode to pass signals of low power while attenuating those above some threshold. Two different implementations are used, depending on whether the diode operates unbiased or with forward-bias current. The unbiased operation is known as *passive* and can be used for low-power applications. Biased operation is *active* and is capable of handling higher power. Diode limiters are used in radar receivers to compress the dynamic range of signals, and in receiver protectors as solid-state replacements for or supplements to gas-tube or circulator-type duplexers.

The operating principle of a diode limiter is based on the change in resistance of the diode. Depending on whether the diode is connected in serial or parallel relative to the load resistance, the diode limiter is called a *serial* or *parallel limiter.* The choice is determined by the load resistance (serial limiters should be connected to circuits with a low output resistance), and by the type of microwave transmission line.

The series-diode limiter (Fig. L9) has a more uniform amplitude characteristic than a limiter with a shunting diode. It is used in a band that does not exceed several tens of megahertz.

A tunnel diode limiter consists of a transistorized amplification stage containing a circuit in the form of one or two counter-serial diodes with resistors connected in parallel to them.

The resistors are selected such that in the range where the tunnel diodes have a negative resistance, the conductivity of the parallel connection of the diode and resistor is close to zero. In this region, the limiting circuit is practically opened and the instantaneous amplification of the stage is sharply limited. For signal levels above or below this region of negative resistance of the diode, the amplifier has a large instantaneous amplification.

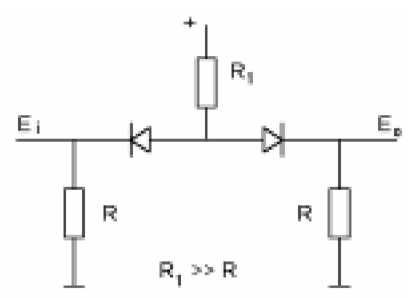


Figure L9 Limiter based on serial diodes (after Skolnik, 1970, Fig. 33, p. 5.38).

The range of levels of the input signal within which the amplification is limited is equal to 6 dB and may be increased using several limiter stages.

Limiters based on PN diodes usually have a higher speed (units of nanoseconds), while those based on PIN diodes have a high level of limitation, low speed (units of microseconds), and higher power-handling capability. Limiters based on Schottky-barrier diodes are marked by their almost complete lack of accumulation of charges of minority carriers and the associated reactivity. Thanks to this, Schottky diode limiters are used in circuits with phase-coded and frequency-modulated waveforms, where the conversion of amplitude modulation to phase modulation is undesirable; at the input of receivers, before the demodulator; or in a transmitter at the output of the frequency (phase) modulator.

Diode limiters are both for low- and high-power levels (e.g., as protectors at the input of the receiver of a monostatic radar). In the latter case, multidiode limiters or avalanche-multiplication limiters are used.

The general shortcoming of diode limiters is their considerable thermal noise level, which must be taken into consideration when calculating the noise temperature of the low-noise input circuit of the receiver. *IAM*

Ref.: Gassanov (1988), p. 142; Skolnik (1970), pp. 5.37–5.38.

The **ferrite limiter** is based on the properties of ferrite materials to suppress RF power passing in different directions and can be used in receiver protectors. In this application the ferrite limiter is followed by a diode limiter, needed to reduce the spike leakage to a safe level, since the ferrite by itself does not give sufficient spike suppression at high peak power. *SAL*

Ref.: Skolnik (1970), p. 8.35, (1980), p. 364.

A **hard limiter** is one in which the gain is sufficient to place the noise level in the limiting region of response, such that the output for any input condition is a rectangular wave. When a signal with high signal-to-noise ratio is passed through a hard limiter, the loss in detection performance is approximately 1 dB. However, when the noise (or other interference) exceeds the signal a small-signal suppression effect occurs, and the loss reaches or exceeds 6 dB. *DKB*

Ref.: Nathanson (1969), p. 113.

An **instantaneous value limiter** is a limiter of pulse amplitudes. Instantaneous value limiters are divided into maximum limiters (or top limiters) and minimum limiters (or bottom limiters).

Pulsed radar devices used them for formation of a flat peak of pulses, for reduction in the duration of leading edges, and for stabilization of pulse amplitudes; (e.g., in circuits used for maintaining a constant false alarm rate). Limiters are also used for selection of pulses based on amplitude or polarity. *IAM*

Ref.: Grigor'yants (1981), p. 46.

A **microstrip limiter** is used up to the shortwave part of the centimeter band, where resonant circuits designed for a relatively narrow band of operating frequencies are used to improve the electrical parameters of the diode limiters. Resonant circuits are based on nonhomogeneous microstrip lines and the reactive parameters of diodes.

Microstrip limiters are used as protective devices at the input of microwave receivers, in antenna switches, and in radio measuring apparatus. *IAM*

Ref.: Lazunin, Yu. A., *Radiotekhnika*, no. 10, 1987, p. 61, in Russian.

A **microwave power limiter** is a circuit whose output power is constant over a broad range of power levels of the input signal. Controlled power limiters include attenuators whose level of limiting power is regulated by the control voltage (see also **transistor limiter**). An uncontrolled limiter is a section of a transmission line in which a limiting diode with a p-n or p-i-n structure in the short-circuit mode with constant current is connected in parallel to the line (see **diode limiter**). At low power levels, the diode is closed. At high power levels, during the positive half-period of the microwave field, the diode is shunted by the low forward resistance. Through the charge accumulation effect of the minority carriers, the resistance of the diode remains low also during the negative half-period. As a result of reflection and partial absorption of energy, power is limited in the diode at the level of 3 to 10 mW. Counterparallel connection of the diodes is more effective.

Microwave power limiters are used for protection of low-noise amplifiers of microwave receivers from loads during seepage of the power of its own transmitter to the input, or of another interfering signal, for equalization of the power of the local oscillators, exciters, low-power tunable oscillators in the operating frequency band, and so forth. (See also **varactor limiter**.) *IAM*

Ref.: Gassanov (1988), p. 142.

A **multidiode limiter** is a limiter of high microwave power, using uniform distribution of microwave power between all diode structures with identical removal of heat from them (see **diode limiter**). In the decimeter waveband, the limiter consists of a large number of diodes connected in parallel to a coaxial line with increased size of cross-section. In the centimeter band, a serial-parallel arrangement of diodes covering the waveguide section, placement of diodes in the slot of a metal iris, and connection of diodes in a waveguide resonance grid of metal bars placed within the section of the waveguide are used.

The increase in input power is limited by the number of diodes in the section of the transmission line, which is decreased proportionally to the increase in frequency.

Multidiode limiters are used to protect the input circuits of the receiver at high power levels of the input signal, as occurs for example during operation of the radar receiver and transmitter with a common antenna. *IAM*

Ref.: Lebedev, I. V., et al., *Radiotekhnika*, no. 8, 1982, pp. 3–5.

A **transistor limiter** is a limiter using a transistor as the active element. Usually controlled transistorized limiters are used. In the simplest limiter, the transistor operates in the strong-signal mode and the operating point moves beyond the limits of the linear region of the characteristic: on one side to the saturation region, and on the other to the cutoff region. In microwave circuits, single- and double-gated field-effect transistors with a Schottky barrier are used. Microwave power limiters using field-effect transistors have been developed basically in the form of wideband amplifier-limiters with automatic gain adjustment. Limiters based on field-effect tetrodes have automatic gain regulation at the second gate. Automatic gain regulation of multistage amplifier-limiters covers several stages, which prevents their saturation and distribution of signal shape, in contrast to diode limiters.

The amplitude characteristic of the amplifier-limiter based on field-effect Schottky transistors differs from the characteristic of diode limiters by the steeper break with the transition to saturation and the higher level of output power.

Power limiters are designed for a power on the order of a Watt, and their output stages can be built in a balance circuit.

These limiters are used to equalize power in the frequency band in exciter-local oscillators and microwave circuits of special radio receivers. *IAM*

Ref.: Skolnik (1970), p. 5.38; Gassanov (1988), p. 145; Grigor'yants (1981), p. 65.

A **varactor limiter** uses the variation of capacitance with voltage to realize power limiting at microwave frequencies. The input signal experiences low loss until it reaches some threshold value, after which the output will remain constant until a second threshold is reached. Beyond the second threshold the output power increases with input, but the loss is large. *DKB*

Ref.: Gardiol (1984), p. 310.

LINE

delay line (see **DELAY LINE**).

The **long-line effect** is the effect of distortions that can be produced by multiple reflections along transmission-line paths. In wideband radars, a single mismatch in nondispersive transmission line reduces the amount of transmitted signal without introducing distortion, and two or more mismatches produce a nonlinear phase variation with frequency because of the interference caused by superposition of resulting waves traveling in the same direction. *SAL*

Ref.: Skolnik (1970), p. 7.13; Wehner (1987), pp. 60–64.

A **pulse-forming line** is a network used in radar modulators to store energy and release it to the transmitting tube in a pulse of predetermined width and constant amplitude.

Ref.: Glasoe (1948), Ch. 7.

transmission line (see **TRANSMISSION LINE**).

LOAD. A load is "a power consuming device connected to a circuit." If any element of the circuit is removed (temporarily or permanently) the artificial termination that replaces it is termed a *dummy load*, which is defined as "a dissipative but essentially nonradiating substitute device having impedance characteristics simulating those of the substituted device."

Ref.: IEEE (1993), pp. 393, 730.

LOBE, LOBING. A lobe is a response peak, generally having symmetric curvature, in any radar coordinate. The term is most often applied to antenna patterns, but can also be used to describe the output waveform of a receiver, especially one in which pulse compression is used.

antenna (pattern) lobe (see **PATTERN, antenna**).

grating lobes (see **PATTERN, array**).

The **mainlobe** is an antenna lobe in the direction of maximum gain. *SAL*

Ref.: Johnson (1984), p. 1.13.

multipath (propagation) lobing (see **PROPAGATION**).

Lobe-on-receive-only is an angle-tracking technique in which the receiving beam pattern is switched between two positions offset from the axis of the transmitted beam (Fig. 10a). The target position relative to the axis is sensed by

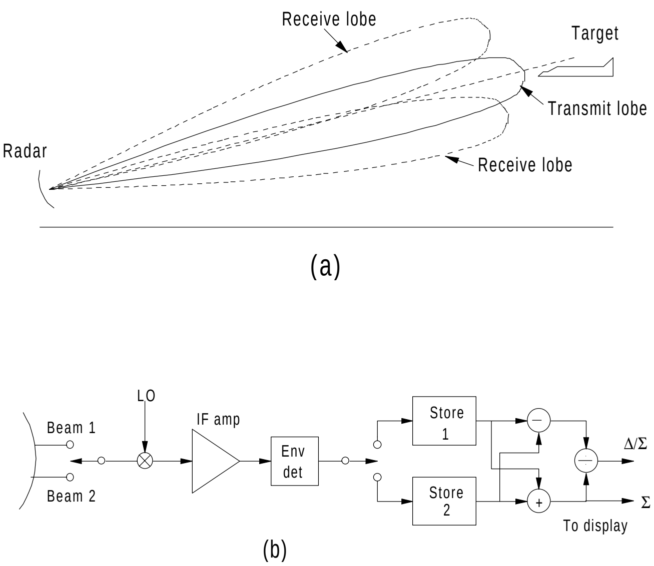


Figure L10 Lobe-on-receive-only tracking system: (a) transmit and receive beam patterns, (b) processing system for received signals.

comparing the amplitudes of signals received successively on the two lobes (Fig. L10b). The technique can be applied to a conically scanning receive beam, in which case it is sometimes called *conopulse*. The advantage of lobing-on-receive-only is that the lobing rate cannot be sensed by a jamming system, and hence the jammer must use a technique such as barrage scan-frequency jamming, dispersing its modulation power over a band of audio frequencies to cover the actual scan frequency with at least some AM components. *DKB*

Ref. Barton (1988), p. 420.

Lobe [sidelobe] reduction is the technique by which undesirable lobes in response of an antenna or pulse-compression system are reduced. Tapering of the aperture illumination and weighting of the signal spectrum are the principal techniques used for sidelobe reduction. (See **APERTURE illumination; WEIGHTING**.)

Sequential lobing is the angle-tracking technique in which the antenna lobe is moved between two or more positions in rapid sequence to sense target position. Lobe switching and conical scan are the principal types of sequential lobing. *DKB*

Ref.: Barton (1988), p. 383.

sidelobe (see **SIDELOBE**).

silent lobing (see **lobe-on-receive-only**; **TRACKING**).

simultaneous lobing (see **MONOPULSE**; **TRACKING**).

Lobe switching is an angle-tracking technique in which antenna lobes are switched sequentially between two positions offset from the tracking axis. The switched lobes may be used for both transmitting and receiving, or for receiving only (see **lobe-on-receive-only**). *DKB*

Ref.: Barton (1988), p. 420.

time sidelobes (see **SIDELOBE, range**).

LOCAL OSCILLATOR (LO). A local oscillator is one used in a superheterodyne receiver to convert the frequency of a received signal to a given intermediate frequency, and sometimes to generate oscillations necessary for matched filtering of the received signals. The latter requires high phase stability in the transmitting and receiving circuits of the radar. In a superheterodyne radar receiver, one or more local oscillators is used. The first local oscillator makes it possible to tune the receiver without touching the intermediate frequency unit. In a coherent system, the parameters of the first local oscillator (stable local oscillator, or STALO) affect the result of signal processing more than the parameters of the transmitter. The last local oscillator (coherent local oscillator, or COHO) is often used to introduce phase adjustments to compensate for the change of phase of the transmitter, the motion of the radar platform, and so forth. A radar with an amplifier-type transmitter uses a single local oscillator, which also generates the carrier frequency of the radar with the required shift relative to the frequency of the first local oscillator. In pulse transmitters with self-excitation, automatic frequency control (AFC) is used to maintain the required frequency shift of the carrier frequency and the frequency of the first local oscillator.

The requirements on stability of the first local oscillator are usually defined in the form of acceptable spectrum of phase modulation. Klystron resonator-stabilized local oscillators with electronic stabilization are widely used as local oscillators. The use of a frequency-multiplier circuit, especially one with low levels of frequency multiplication, provides the lowest level of near-in sidebands and is recommended for stabilization of klystron oscillators of noncoherent and pseudocoherent radars. A local oscillator is a receiving device, but in truly coherent systems it is also part of the transmitter. *IAM*

Ref.: Skolnik (1970), pp. 5.12–5.19.

A **coherent local oscillator (COHO)** is one whose frequency and phase of oscillations are strictly associated with the RF oscillations of the transmitter. Phasing of a coherent local oscillator is done at an intermediate frequency at the start of each repetition period of the transmitted pulses during the time of their emission. After the conclusion of the synchronization pulses, the local oscillator must operate in the mode of free continuous oscillations at least for the time corresponding to the maximum range of the radar. A COHO includes a master oscillator, which has short-term stability, and a phase-tuning circuit. Examples of practical COHO include the phase-locked local oscillator and the phase-shift local oscillator. The former is marked by a pulse mode of operation of the master oscillator, while the latter is a high-stability master oscillator in the continuous mode.

A COHO is used usually as a second (and sometimes subsequent) radar local oscillator. *IAM*

Ref.: Ridenour (1947), p. 632; Skolnik (1962), p. 117; Pereverzentsev (1981), p. 271.

A **crystal local oscillator** uses crystal stabilization of frequency, implemented through the use of a crystal resonator. The frequency of the generated oscillations coincides with the frequency of the first (up to 30 MHz) or higher (up to 200 MHz) mechanical harmonics of the crystal. To produce oscillations of the necessary range, several frequency multiplication stages are used with a common coefficient of multiplication reaching 100 and more. (See **frequency-multiplier STALO.**) Frequency retuning of a crystal local oscillation is usually done by changing the crystals. The frequency stability of a crystal local oscillator reaches 10^{-7} to 10^{-8}. *IAM*

Ref.: Pereverzentsev (1981), p. 242; Fink (1982), p. 7.32; Rakov (1970), vol. 2, p. 30.

A **frequency-multiplier STALO** is one in which all frequencies, including the carrier, are obtained from a high-stable frequency generator (see **OSCILLATOR, crystal**) through multiplication. A local oscillator of this type is used to form frequencies of the first and second local oscillators of the radar (Fig. L11) and is called a *frequency synthesizer*. The frequency of the first local oscillator is obtained from that of the second by a multiplication circuit that uses, for example, a charge-accumulation diode, which provides a multiplication factor of 100 or more. The first multiplier creates a series of frequencies that are separated from one another by the frequency of the reference oscillator and are symmetrical with respect to the frequency of the second local oscillator.

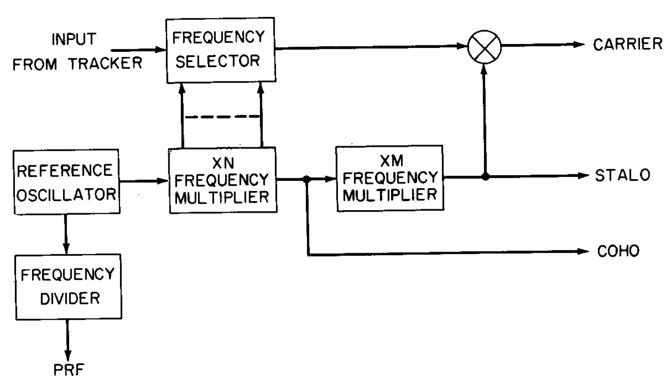


Figure L11 Local oscillator in frequency multiplication circuit (from Skolnik, 1970, Fig. 11, p. 5-15, reprinted by permission of McGraw-Hill).

Control signals arriving from the tracking system connect the signal of the frequency that most closely adjusts the carrier, with allowance for the doppler shift. The pulse repetition frequency may also be obtained from a general reference oscillator using a frequency divider. This makes it possible to get rid of parasitic combination components of the signal at the input of the mixer. *IAM*

Ref.: Skolnik (1970), pp. 5.15–5.16.

A **phase-locked COHO** is a second local oscillator with phase locking implemented through synchronization with a downconverted sample of the transmitter pulse in the resonant circuit of the oscillator. It is used as the second (coherent) local oscillator to compensate for phase fluctuation of a pulsed-oscillator type of radar transmitter (e.g., magnetron).

Just before each transmission, the COHO oscillations are quenched, destroying the phase information from the previous transmission. The reference phase for the new transmission is obtained at the intermediate frequency by mixing part of the transmitter RF with the first local oscillator in a mixer controlled by a lock pulse. The pulse voltage of the difference (intermediate) frequency is amplified, and then is sent to the resonant circuit of the second local oscillator (Fig. L11). The phase of oscillations of the second local oscillator is then matched to the phase difference between the first local oscillator and radar transmitter, and the continuing COHO output preserves this random phase for the remainder of the pulse repetition interval. *IAM*

Ref.: Skolnik (1970), p. 5.16

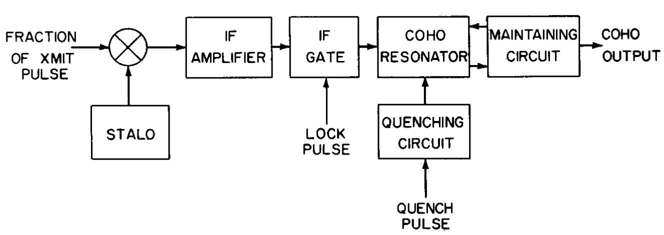


Figure L12 Keyed COHO (from Skolnik, 1970, Fig. 12, p. 5.16, reprinted by permission of McGraw-Hill).

A **phase shift COHO** is a second local oscillator with automatic phase tuning of oscillations of crystal local oscillator with phase of the synchronization pulse. The tuning is done using a phase-shifting device with electronic control (Fig. L13). It is used as a second loop of a radar.

The circuit for automatic phase tuning is turned on with an electronic switch only for the time of operation of the synchronization pulse. In the interval between the successive pulses, the voltage of the phase correction is retained in the integrating capacitor *C*. The quadratic phase detector and the high-speed phase-shifting device with an interval of 180° are required to maintain the phase delay between 0 and 180°.

The process of phase strobing and storage does not influence the frequency of the second local oscillator and the requirements on stability are minimal. The use of a crystal resonator assures higher stability of a local oscillator with phase shift than of a phase-locked COHO. *IAM*

Ref.: Skolnik (1970), pp. 5.16–5.17.

A **phase-locked STALO** is a local oscillator with phase autotuning of frequency, which has the capability of retuning. It is used as a first local oscillator of a radar (Fig. L14). The local

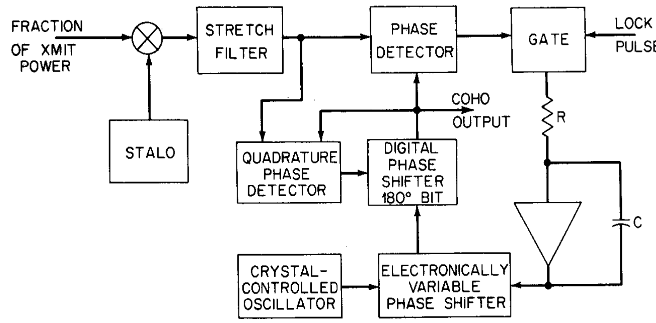


Figure L13 Phase-shift COHO (from Skolnik, 1970, Fig. 14, p. 5-17, reprinted by permission of McGraw-Hill).

oscillator in the figure creates a linear spectrum in the microwave band with a frequency interval of Δf using a snap diode. (See **DIODE, charge storage.**)

Some of the output power of the klystron is mixed with the linear spectrum and an oscillation with a beat frequency of $f/2$ is applied, after amplification, to the $f/2$ frequency phase detector. An operational amplifier provides the required transient characteristics of the circuit for phase autotuning of the frequency. The klystron is synchronized in phase at the discrete frequencies $f(N \pm 1/2)$. The stability of the klystron is determined by that of the reference frequencies. *IAM*

Ref.: Skolnik (1970), p. 5.14.

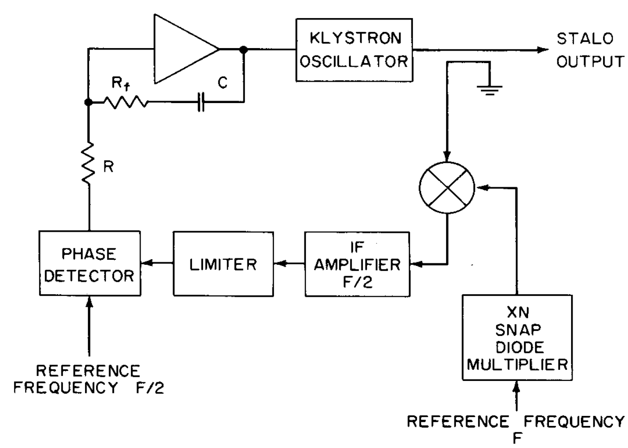


Figure L14 Phase-locked STALO (from Skolnik, 1970, Fig. 10, p. 5-14, reprinted by permission of McGraw-Hill).

A **quartz local oscillator** uses a quartz resonator for frequency stabilization. The oscillator frequency coincides with the frequency of the first (up to 30 MHz) or higher (up to 200 MHz) mechanical harmonics of the quartz. Multiple frequency multiplications can be employed to get oscillations with the required frequency. Frequency agility typically is obtained by changing the quartz slab. The frequency stability is about 10^{-7} to 10^{-8}. *IAM*

Ref.: Pereverzentsev (1981), p. 242; Fink (1982), p. 7.32.

A **reflex klystron local oscillator** is a local oscillator in the centimeter or millimeter waveband that uses a reflex klystron with an external or internal three-dimensional cavity as its active component. Klystrons with external cavities are used in the lower microwave and UHF bands (4 to 60 cm wavelength) and allow frequency retuning of ±15%. Local oscillators based on klystrons with an external cavity are distinguished by high resistance to mechanical influences, but they have a smaller retuning range, not exceeding ±10%.

Stabilized power sources in the circuits of the reflector, cavity, and filament are used for stabilization of the frequency of klystron local oscillators. A high-quality three-dimensional cavity is connected to the klystron. The local oscillator requires reliable shock isolation.

Klystron local oscillators are used in landing and navigational noncoherent radars, test equipment, and other apparatus. *IAM*

Ref.: Ridenour (1947), pp. 414–416; Pereverzentsev (1981), p. 242; Zherebtsov (1989), p. 330.

A **resonator-stabilized LO** uses high-quality microwave resonators. Depending on the frequency band and the requirements for frequency stability, the following may be used: surface-acoustic-wave resonators (up to 1 to 2 GHz); three-dimensional resonators of material with a low temperature coefficient of linear expansion, for example invar (in the decimeter and centimeter bands); or dielectric resonators (in a band of 1 to 20 GHz), which have become common thanks to their high stability of characteristics and compatibility of design with hybrid-integrated technology.

The resonators are connected to the frequency-generating circuit of the local oscillator; for example, in the gate circuit in an oscillator based on a field-effect transistor in a circuit with a common source, or in a frequency autotuning circuit of a klystron oscillator. In the last case, the autotuning circuit contains a microwave discriminator based on a hollow resonator, an operational amplifier-integrator that provides the transient characteristics necessary for keeping the frequency of the klystron very close to the average frequency of the discriminator. *IAM*

Ref.: Gassanov (1988), p. 176; Skolnik (1970), pp. 5.13–5.15.

A **semiconductor LO** uses a Gunn diode, field-effect transistor, or tunnel diode for generation of low power. Stabilization of oscillations is done with an external high-quality resonator with a Q-factor up to 20,000. Synchronization of semiconductor local oscillators can be done in various modes: at input operating frequency, or at a harmonic or subharmonic of input frequency. To increase stability of oscillations of diode local oscillators, their self-synchronization is used. Toward this end, some of the output power is sent to the oscillator through an additional feedback circuit. Semiconductor local oscillators are used in transmitting and receiving modules of antenna arrays and in modern solid-state receivers. *IAM*

Ref.:Fink (9182), p. 13.112; Gassanov (1988), pp. 177, 180.

A **stable LO (STALO)** is the first local oscillator in a coherent radar system, providing downconversion to the first IF with minimum accompanying phase noise. *DKB*

Ref.: IEEE (1993), p. 1,273; Skolnik (980), p. 105, (1990), p. 3.12.

A **tracking LO** is one whose frequency is automatically tuned to the frequency of the received (analyzed) signal. It is used, for example, in doppler trackers. In this case, the frequency of the low-frequency modulation of a signal of the local oscillator is tuned to a value close to the doppler shift. The tuning is done by the control circuit, which measures the frequency difference of the local oscillator and the doppler signal and contains the low-pass filter, an amplitude and phase detector, and an integrator. *IAM*

Ref.: Supryaga (1974), p. 109; Skolnik (1990), p. 19.13.

A **vacuum-tube LO** uses a vacuum tube with grid control, or a microwave tube. Usage of the former (special types of microwave triodes and tetrodes) is limited to frequencies of 2 to 3 GHz. Because of the high phase instability and frequency restrictions, their use is limited in the local oscillator of modern radars. Klystron sources with cavity stabilization are more suitable for coherent radar systems. *IAM*

Ref.: Ridenour (1947), p. 414; Druzhinin (1967), p. 373.

LOCK-ON, radar (see **TARGET lock-on**).

LOSS, in radar. A loss is a value opposite to a gain in a radar channel. It shows the degree by which the power P_2 of a signal in the presence of a specified degrading factor is less than the power P_1 without that factor, or the amount by which the input signal must be increased to overcome the factor. The usual notation is L, and when expressed in decibels it is

$$L_{\mathrm{dB}} = 10\log\frac{P_1}{P_2}$$

In this encyclopedia, values and equations for loss will be in terms of the power ratio, rather than the decibel value, unless the latter is specifically noted. The loss may be defined in terms of signal-to-noise ratio, rather than signal power only (e.g., see **matching loss**)

The total radar loss consists of many components originating throughout the radar channel, that consists of the target, the propagation medium, and the radar itself. Proper calculation of the loss budget is important for accurate prediction of radar performance. Although each component of radar loss can be small, the total can degrade performance considerably. For that reason, the origin of radar losses must be understood and their value minimized by proper radar design.

Radar losses are of four distinct types:

(Type 1) Antenna and propagation loss factors, resulting from effects other than attenuation, and from target location in the nonscanning coordinate.

(Type 2) Transmit RF loss factors, which attenuate the signal passing from the transmitter to the target, relative to ideal equipment operating over a free-space path.

(Type 3) Receive RF loss factors, which attenuate the signal passing from the target to the receiver, relative to ideal equipment operating over a free-space path.

(Type 4) Receiver-processor loss factors, which reduce the ability of the receiver-processor to detect the target echo, relative to ideal processing of echoes received from a steady target, centered in the scanning beam and in a range-doppler cell.

Several components within each type are listed in Tables L2 through L5. Some losses of types 1, 2, and 3 may apply differently to signals and clutter when the latter appears in different range ambiguities or different portions of the beam, and these must be evaluated carefully in equations for signal-to-clutter ratio. Note that type 2 losses apply to the radar range equation and the beacon interrogation equation, but not to the beacon response or to jamming equations. Type 3 losses contribute both to signal attenuation and to system noise temperature, and apply along with Types 1 and 4 to all radar equations for signal-to-noise ratio and jamming-to-noise ratio. Most type 4 losses apply to equations for signal-to-clutter ratio as well as to those for signal-to-noise and signal-to-jamming ratios. Type 4 losses are, in many cases, functions of detection probability and hence increase the effective detectability factor, relative to the theoretical or basic detectability factor.

Table L2
Antenna and Propagation Loss Factors (Type 1)

Component	Symbol	Notes
Antenna blockage loss [factor]	$1/\eta_b$	1
Antenna dissipative loss (includes feed, phase shifters)	L_a	1, 2
Antenna pattern factor	f_t, f_r	3, 7
Antenna spillover loss	$1/\eta_s$	1
Antenna surface tolerance loss	$1/\eta_{st}$	1
Array bandwidth loss	L_z	1
Array phase error loss	L_ϕ	1
Array phase quantization loss	L_q	1
Atmospheric lens loss	F_a	3
Beamwidth factor	L_n	10
Crossover loss	L_k	7
Faraday rotation loss	F_f	3
Polarization loss	F_p	3, 7
Pattern-propagation factor	F_t, F_r	7
Scanning loss	L_{sc}	3
Squint loss	L_{sq}	1
Taper (illumination) loss	$1/\eta_i$	1

Table L3
Transmit RF Loss Factors (Type 2)

Component	Symbol	Notes
Atmospheric loss [attenuation] (1-way)	$L_{\alpha 1}$	7, 10
Transmit line loss (includes line, RF filter, duplexer, radome)	L_t	4

Table L4
Receive RF Loss Factors (Type 3)

Component	Symbol	Notes
Atmospheric loss[attenuation] (1-way)	$L_{\alpha 1}$	7, 10
Receive line loss (includes line, RF filter, duplexer, radome)	L_r	2

Table L5
Receiver-Processor Loss Factors (Type 4)

Component	Symbol	Note
Beamshape loss	L_p	5, 8
Binary integration loss	L_b	6, 8
Clutter-related losses:		
Clutter distribution loss	L_{cd}	6, 8
Clutter-reference loss	$1/F_c$	6, 8
Transient gating loss	L_{eg}	6, 8
Collapsing loss	L_c	6, 8
CFAR loss	L_g	6, 8
Detector loss	C_x, C_a	4, 5, 8
Eclipsing loss	L_{ec}	6
Fluctuation loss	L_f	5, 8
Integration loss	L_i	5, 8
Integrator weighting loss	L_w	6, 8
Limiting loss	L_l	6, 8
Matching losses [factors] (see discussion of M factor):		
Doppler filter matching loss	L_{mf}	6, 8
Filter (IF) matching loss	L_m	5, 7
Range gate mismatch loss	L_m	5, 8
Miscellaneous signal-processing losses	L_x	5, 7
MTI, doppler-processing losses:		

Table L5
Receiver-Processor Loss Factors (Type 4)

Component	Symbol	Note
Blind phase loss	$L_{mti(b)}$	6, 8
Noise correlation loss	$L_{mti(a)}$	6, 8
Velocity response loss	$L_{mti(c)}$	6, 8
Operator [display] loss	L_o	6, 8
Pulse width loss	L_{pw}	6, 8
Quantizing loss	L_q	6, 8
Range cusping loss	L_{er}	6, 8
Scan distribution loss	L_d	10
Straddling losses:		
Angle straddling loss	L_{ea}	6, 8
Doppler filter straddling loss	L_{ef}	6, 8
Range gate (or strobe) straddling loss	L_{er}	6, 8

Notes for Tables 2 through 5:	
1	Included in calculation of antenna gain G
2	Included in calculation of system noise temperature T_s
3	Included in calculation of pattern-propagation factor F
4	Used in calculation of tracking error
5	Included in calculation of detectability factor D
6	Included in miscellaneous signal processing loss L_x
7	Consider possible differences between losses for signal and clutter; effective S/C may or may not be affected
8	Losses do not apply to clutter power; effective S/C is reduced by these losses
9	Included as separate term in conventional range equation
10	Loss included only in search loss factor of search radar equation

The notes indicate how the losses are accounted for in the range equations. In the Blake chart, the detectability factor D includes integration and fluctuation losses, L_i and L_f, while matching, beamshape, and miscellaneous signal-processing loss are entered as separate factors M, L_p, and L_x. These are included in the effective detectability factor in the standard range equation:

$$D_x = DML_pL_x$$

where D is the basic detectability factor for integration of n pulses using the fluctuation model for the target. The transmit line loss L_t, pattern-propagation factor F, and atmospheric attenuation L_α, are included separately in the standard range equation and in the Blake chart. (See **RANGE EQUATION**.) *DKB, SAL*

Ref.: Blake (1980), Ch. 2; Skolnik (1980), p. 56; Barton (1988), pp. 574–575, (1993), pp. 130–137.

Angle straddling loss is the result of batch processing in the integration process of pulse or samples received by a scanning radar, such that the available target energy is split between two contiguous batches. The loss is defined as the increase in input signal energy required to achieve a given detection performance on targets appearing randomly in angle, compared with that required for a target echo centered within a single batch. The loss depends on the ratio of the batch duration (integration time t_{sb}) to the observation time (time-on-target) t_o, the target fluctuation model, and on the required detection probability. Figure L15 shows the loss for

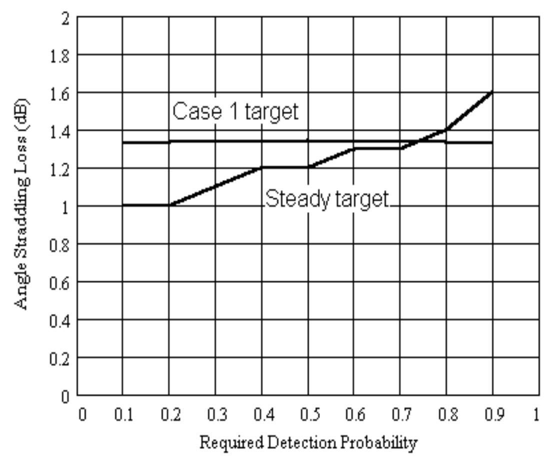


Figure L15 Angle straddling loss vs. required detection probability, for batch video integration.

a uniformly weighted video integrator using contiguous windows matched to the observation time. An approximate equation for this loss is

$$L_{ea} = 1.25P_d^{\frac{1}{3}}\left(\frac{t_{sb}}{t_o}\right)^2 \text{ (dB)}$$

The loss may be avoided if a moving-window integrator, in which overlapping batches are started with each received pulse, is used. For coherent integration, this loss is accounted for in the beamshape loss, and L_{ea} = 0 dB. *DKB*

Ref.: Barton (1993), pp. 104, 133.

Antenna losses listed in Table L2 are of three basic types:

(1) Ohmic losses, describing the dissipation of RF energy into heat within the antenna components (note 2). These losses act directly to reduce effective radiated power (accounting for the difference between directive gain and power gain) and received signal power, and also act as a source of additional noise in reception.

(2) Nonohmic losses in mainlobe (on-axis) gain, describing factors that spread the radiation pattern (notes 1, 10). These affect both directive and power gain.

(3) Factors that prevent the full on-axis gain of the antenna from being brought to bear on the target (notes 3, 4). These factors are normally included in the pattern-propagation factor and are dependent on the target position relative to the beam (and on target and antenna polarizations).

The treatment of these losses in the radar equation is indicated by the notes to the tables. In the case of the beamwidth factor, it applies only to the search radar equation, and results from the assumption there that the transmit antenna gain can be replaced with the term $4\pi/\theta_a\theta_e$.

The total antenna loss describes the reduction in antenna power gain G relative to G_0 for the ideal (uniformly illuminated (lossless) antenna of the same aperture area A. This loss is normally expressed by its reciprocal, the aperture efficiency η_a:

$$G = G_0\eta_a = \frac{4\pi A}{\lambda^2}\eta_a$$

where the aperture efficiency is the product of several components:

$$\eta_a = \eta_d\eta_i\eta_f\eta_b\eta_s\eta_{st}$$

The components are defined in Table L6. In some cases, the transmit and receive antennas have different characteristics, and the components for the two must be stated separately (often using additional subscripts t and r for each component)

The dissipative (ohmic) portion of receive antenna loss must be included in calculation of system noise temperature, in addition to including its effect on antenna gain. In the search radar equation, the antenna dissipative loss must be included in the system loss factor (doubled for two-way loss if the physical receive aperture is used instead of effective aperture). *DKB*

Ref.: Johnson (1984), pp. 1.6–1.9.

The **antenna pattern factor [loss]**, denoted by $f(\theta,\phi)$, represents the one-way voltage gain pattern of the antenna, as a function of angles θ and ϕ (normalized to its maximum, on-axis, gain). For a radar that scans in one coordinate (e.g., ϕ), $f(\theta - \theta_b)$ for the nonscanning coordinate is included as a component of the pattern-propagation factor, accounting for the position of the target relative to the beam axis angle θ_b in that coordinate (usually elevation). Thus, in the typical radar coverage plot (Fig. L16), the dashed curve for free-space coverage represents the antenna elevation pattern factor, relative to unity value at the peak of the beam ($\theta_b = 1°$in this case). For a discussion of the lobing pattern, see **PROPAGATION, pattern-propagation factor**.

To account for random target positions in the scanning coordinate, the beamshape loss L_p for targets at random locations relative to the peak of the beam, is used in the denominator of the range equation, replacing the antenna pattern factor $f^4(\theta)$ as part of the pattern-propagation factor F^4 in the numerator, since L_p represents the average value of $[1/f(\theta)]^4$

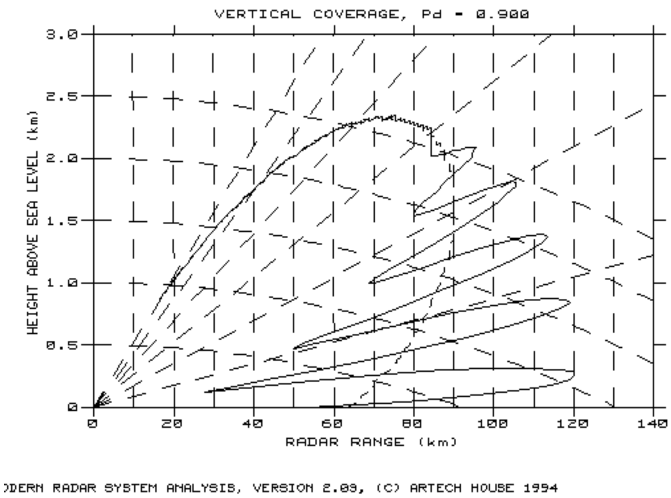


Figure L16 Typical search radar vertical coverage plot, showing the antenna patter factor as the dashed curve for free-space coverage (from Barton, 1993, Fig. 2.36, p. 69).

(see **beamshape loss**). For a raster scan, L_p^2 is used to account for random target positions in both coordinates, relative to the beam axis positions, and the pattern factor is used only in calculation of reflection lobing. In a nonscanning radar $f(\theta,\phi)$ would be included in the pattern-propagation factor to account for the reduced gain on a target at those known coordinates.

Table L6
Antenna Efficiency Components

Component	Symbol	Brief description
Dissipative [ohmic]	$\eta_d = \eta_f\eta_f = 1/L_a$	Ohmic loss in RF components
Feed	η_f	Ohmic loss in feed
Phase shifter	η_ϕ	Ohmic loss in phase shifters
Taper [illumination]	η_i	Design taper for reduced sidelobes
Blockage	η_b	Reflector blockage by feed and supports
Spillover	η_s	Fraction of power not intercepted by reflector
Surface tolerance	η_{st}	Effect of phase error relative to precise surface

In summary, off-axis target position is accounted for statistically, in the scanned coordinate(s), through the beamshape loss. In the nonscanned coordinate, when a coverage pattern is prepared, it is accounted for through the systematic change in pattern-propagation factor F (and hence in coverage range), as a function of that coordinate. In that case, the range calculation is normally carried out for the angle of the beam axis, and scaled directly by F for other angles. If a single detection range number is to be derived, applicable to all targets within a given sector, the statistical beamshape loss should be used in both coordinates, modified if necessary for the effects of reflection lobing (e.g., and average value of F can be used in the radar equation).

As with other antenna gain and loss factors, radars having different transmit and receive patterns will use subscripts (f_t and f_r) to represent the two patterns. *DKB*

Barton (1988), pp. 290–296.

Array bandwidth loss occurs when the bandwidth B_a of an array antenna is insufficient to pass all signal components in their proper phase relationship. For a signal of half-power bandwidth B, the loss in decibels is

$$L_z = 10\log\left(1 + \frac{2B^2}{B_a^2}\right)$$

DKB

Barton (1988), pp. 185–187.

Array phase error loss is the reduction in antenna gain caused by errors in phase shifters, feed networks, or amplifier modules in a phased-array antenna. The loss as a power ratio is given by

$$L_\phi = \exp(\sigma_\phi^2) \approx 1 + \sigma_\phi^2$$

and in decibels is $L_{\phi\mathrm{dB}} \approx 4.3\sigma_\phi{}^2$, where σ_ϕ is the rms phase error in radians, assumed << 1. For the case of phase error caused by quantization, see the next entry. If there are amplitude errors as well as phase errors, a second variance $\sigma_a{}^2$ representing the amplitude error should be added to σ_f^2 in calculating loss caused by the feed network, phase shifters, and amplifier modules. *DKB*

Ref.: Mailloux (1994), p. 399.

Array phase quantization loss is the reduction in antenna gain caused by phase quantization error in phase shifters of a phased-array antenna. The loss as a power ratio is given by

$$L_q = \exp(\sigma_q^2) \approx 1 + \sigma_q^2 = 1 + \frac{\pi^2}{3 \cdot 2^{2p}}$$

and in decibels is $L_{q\mathrm{dB}} \approx 4.3\sigma_q{}^2 .= 14.15/2^{2p}$, where σ_q is the rms phase quantization error and p is the number of bits in the quantized command. For the common case $p = 4$, the loss is 0.05 dB. Since quantization error is uncorrelated with other phase errors (the phase variances add directly), this loss may be calculated separately from other components of phase error loss, and the two loss values multiplied (added in decibels) to arrive at the total phase error loss. *DKB*

Ref.: Barton (1969), p. 193.

Atmospheric (attenuation) loss is experienced by the radar signal in passing from the transmit antenna to the target and back to the receive antenna, as a result of interaction with atmospheric molecules and precipitation particles. In the radar equation, the two-way attenuation is applicable; in the beacon or jamming equations, one-way loss is applicable. (See also **ATTENUATION.**)

The one-way loss $L_{\alpha 1}$ is a function of range R and elevation angle θ for the path and can be found from
where h is the height determined by R and θ, $n(h)$ is the

$$L_{\alpha 1}(R, \theta) = \int_0^{h(R)} \frac{[V_0(h) + V_w(h)]}{\sqrt{1 - \left[\dfrac{n_0 \cos\theta}{n(h)(1 + h/r_0)}\right]^2}} dh \quad \text{dB} \quad (1)$$

refractive index at height h, n_0 is the surface refractive index, r_0 is the earth's radius (6.5×10^6m), and $V_o(h)$ and $V_w(h)$ are frequency-dependent models of the absorption coefficients, in decibels per unit length, for oxygen and water vapor. Based on this equation, the total attenuation for earth-based radar, on a path extending through the entire atmosphere (beyond 30 km in altitude) are as shown in Fig. L17. Plots for attenuation indifferent frequency bands, as functions of range and elevation are available in Blake (1980).

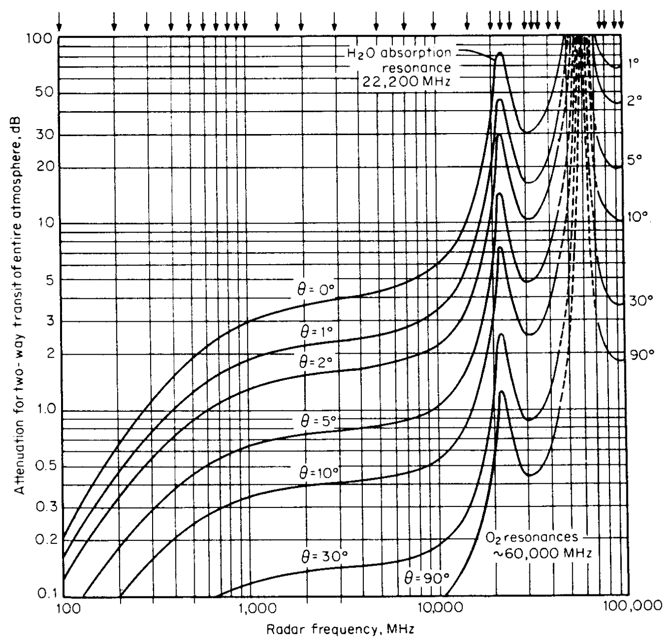


Figure L17 Total earth-based radar attenuation (two-way) vs. frequency for target outside the troposphere for different elevation angles in standard atmosphere (from Blake, 1980), Fig. 5.25, p. 219).

The use of (1) requires complicated computations of absorption coefficients, $V_o(h)$ and $V_w(h)$, which are functions of atmospheric pressure and temperature, height, and radar frequency. Simplified approximations to calculate two-way atmospheric loss are

$$L_\alpha(R, \theta) = k_\alpha R_a[1 - \exp(-R/R_a)] \text{ (dB)}$$

where k_α is the sea-level attenuation coefficient and

$$R_a = \frac{3000}{\sin\left(\theta + \frac{2.5 \times 10^{-4}}{\theta + 0.028}\right)} \text{ m}$$

is the effective sea-level pathlength. Frequency dependence is accounted for by the coefficient k_α, shown in Table L7.

If part of the path R_{pr} is occupied by precipitation, there will be an additional loss:

$$L_\alpha(R_{pr}, \theta) = k_{\alpha pr} R_a \left[1 - \exp\left(-\frac{R_{pr}}{R_a}\right)\right] \quad \text{(dB)}$$

where $k_{\alpha pr}$ is the precipitation attenuation coefficient shown in the table. The total attenuation will include that of the air and the embedded precipitation:

$$L_a(R, \theta) = L_\alpha(R, \theta) + L_{\alpha pr}(R, \theta) \quad \text{(dB)}$$

DKB, SAL

Ref.: Blake (1980), pp. 197–221; Barton (1993), p. 113.

Table L7
Two-Way Atmospheric Attenuation Coefficients (dB/km)

Band	Freq. (GHz)	Atmospheric conditions		
		Clear, k_α	Rain, k_α/r	Snow, k_α/r
UHF	0.4	0.01	0	0
L	1.3	0.012	0.0003	0.0003
S	3.0	0.015	0.0013	0.0013
C	5.5	0.017	0.008	0.008
X	10	0.024	0.037	0.002
Ku	15	0.055	0.083	0.004
K	22	0.30	0.23	0.008
Ka	35	0.14	0.57	0.015
V	60	35	1.3	0.03
W	95	0.80	2	0.06
	140	1.0	2.3	0.06
	240	15	2.2	0.08

r is the precipitation rate in mm/h

Atmospheric lens loss is a propagation loss at low elevation angles, in which rays are refracted downward (according to the 4/3 effective earth's radius model, under normal conditions), diluting the power density at the target relative to that calculated from antenna gain and free-space propagation theory. The loss is nondissipative, and hence should be included as a reduction in the propagation factor, rather than as a component of atmospheric attenuation. The loss is reciprocal, and the two-way values are shown in Fig. L18, As a function of range, for different elevation angles. The term *lens-effect loss* is sometimes used interchangeably. *DKB*

Ref.: Weil, T. A., "Atmospheric Lens Effect: Another Loss for the Radar Equation," *IEEE Trans.* ***AES-9***, no. 1, Jan. 1973, pp. 51–54; Blake (1980), p. 188.

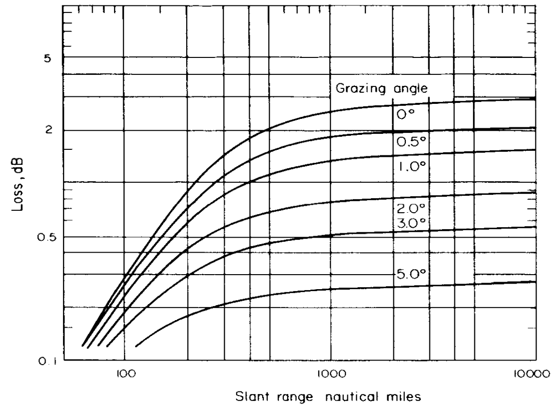


Figure L18 Atmospheric lens loss (two-way) vs. range, for different elevation angles (after Weil).

Beamshape loss is the result of covering the search sector with beams having the typical (approximately Gaussian) mainlobe shape, rather than with (idealized) contiguous rectangular beams. This loss is defined as the ratio of echo power required at the peak of a scanning pattern, to achieve a given detection performance when the signal integration is matched to that pattern, relative to the power that would have been required for a uniform signal envelope existing over the time required to scan one beamwidth (Fig. L19). For a two-coordinate scan, it is the ratio of transmitted energy required for given detection performance for targets uniformly distributed over the search solid angle to the energy that would have been required using contiguous, rectangular beams.

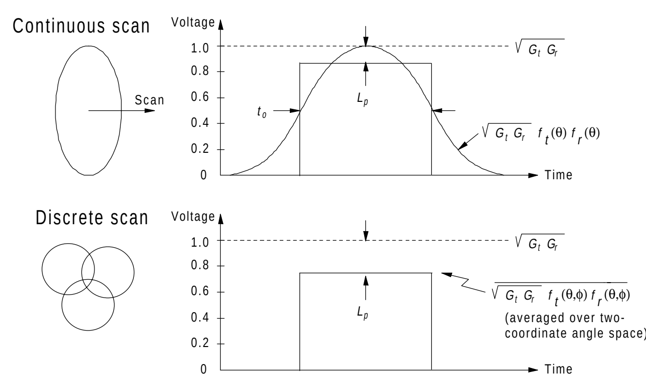


Figure L19 Definition of beamshape loss for one-coordinate continuous scan and discrete-position raster scan.

For both continuously scanning beams and step-scanned beams with optimum spacing, the beamshape loss for $P_d \approx 0.5$ is 1.23 dB for one-coordinate scans, and 2.5 dB for two-coordinate (raster or spiral) scans. (In Blake's early work, the value of 1.6 dB included nonoptimum integrator weighting as well as beamshape loss.) The loss increases for higher values of P_d, because the increased P_d (e.g., from 0.9 toward 1.0) near centers of the beams cannot compensate for the decrease (e.g, from 0.9 toward 0) near the beam-overlap regions. When scan occurs in one coordinate, the random off-axis position of the target in the other coordinate appears as a reduction in the antenna pattern factor.

The beamshape loss was originally calculated as the reciprocal of the integrated power received by the antenna for scatterers distributed uniformly in angle, relative to that of an idealized rectangular beam matched to the one-way, half-power beamwidths of the actual antenna:

$$\frac{1}{L_p^2} = \frac{1}{\theta_a\theta_e}\int\int_{4\pi}[f(\theta,\phi)]^4 d\theta d\phi$$

In that form, it accounts for the clutter power received from a uniform cloud filling the region in both coordinates around the beam, as in radar meteorology, and that is the appropriate definition for volume clutter power calculations. When applied to the range equation for target detection by scanning radar, however, it is not sufficient to increase the on-axis power by this constant loss unless detection probabilities near 50% are being considered.

The beamshape loss is approximately constant for small beam motion between samples (pulses, in the case of a continuous rapid scan in one coordinate, scan lines or beam spacing for the other coordinate and for discrete, or step, scanning). Approximations for L_p as a function of P_d and number of samples per beamwidth n are as follows, for a steady target:

$$L_p(n) = 1.2 + 4.5(P_d - 0.4)\cdot n^{-3} \text{ (dB)}, \quad P_d \geq 0.4$$
$$= 1.2 + 1.35(P_d - 0.4)\cdot n^{-3} \text{ (dB)}, \quad P_d < 0.4$$

and for the fluctuating (case 1) target:

$$L_p'(n) = 10^{-0.03L_f'}\cdot L_p(n) + 0.4L_f' \text{ (dB)}$$

where the fluctuation loss approximation is

$$L_f' = -8.4\log(1-P_d) - 3.0(1-P_d)^2 \text{ (dB)}$$

Using these approximations, the beamshape loss is given in Figs. L20 and L21 as a function of the beam motion between pulses (or scan lines), for different probabilities of detection.

It can be seen that the loss increases rapidly for a steady target as P_d and beam motion increase, but that for the fluctuating target the loss actually decreases with increasing P_d. This is because of the contributions made by adjacent beam positions to detection of the fluctuating target. *DKB*

Ref.: Blake, L. V., "The Effective Number of Pulses per Beamwidth for a Scanning Radar," *Proc. IRE* ***41***, no. 6, June 1953, pp. 770–774; Hall, W. M., and Barton, D. K., "Antenna Pattern Loss Factor for Scanning Radars," *Proc. IEEE* ***53***, no. 9, Sept. 1965, pp. 1257–1258; Hall, W. M., "Antenna Beam-Shape Factor in Scanning Radars," *IEEE Trans.* ***AES-4***, no. 3, May 1968, pp. 402–409.

The **beamwidth factor [loss]** is a loss used in the search radar equation to account for the fact that the beam solid angle $\psi_b = \theta_e\theta_a$ does not include all the radiated power of the pattern. This leads to a gain relationship

$$G_t = \frac{4\pi}{\theta_e\theta_a L_n}$$

where $L_n \approx 1.2 = 0.8$ dB for many array antennas and may approach $1.6 = 2$ dB for reflector systems having significant spillover and blockage. *DKB*

Ref.: Barton (1969), p. 334.

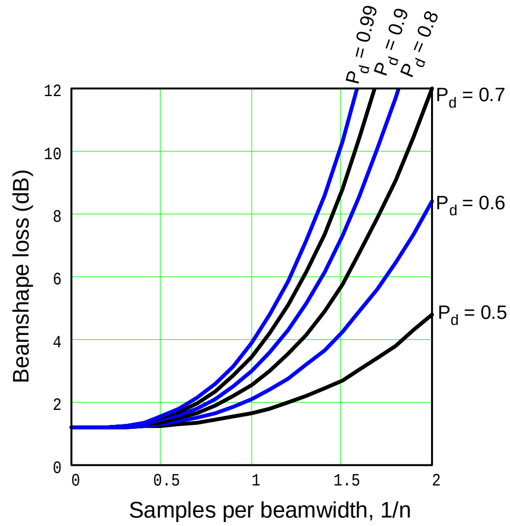


Figure L20 Beamshape loss for a steady target as a function of the beam motion between pulses or scan lines.

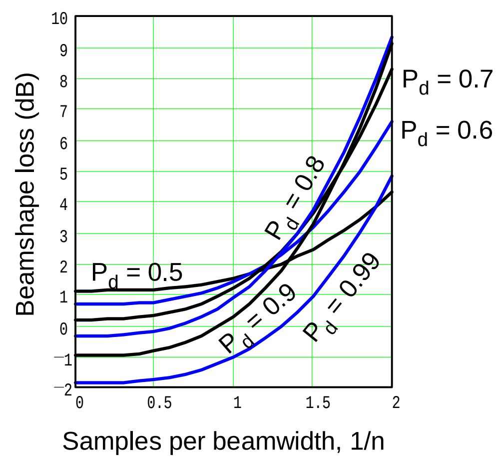


Figure L21 Beamshape loss for a case 1 target as a function of the beam motion between pulses or scan lines.

binary integration loss (see **integration loss**).

Blind-phase loss is the result of using a single phase detector and canceler channel in an MTI system, rejecting the quadrature components of signal and noise. It is defined as the increase in signal power required to achieve a given detection performance when the single channel is used instead of an I-Q channel pair (vector MTI processing). Although the single channel has, on average, the same SNR as the vector canceler, the number of degrees of freedom for both signal and noise is reduced from two to one per pulse (the distributions are changed from Rayleigh to Gaussian). Unless the processor integrates over a period long enough to receive both components of a fluctuating signal, the fluctuation loss (in decibels) is doubled, a serious factor for high values of P_d (see Fig. L26). The increase in fluctuation loss is the blind phase loss. *DKB*

Ref.: Barton (1988), p. 251; Nathanson (1991), p. 393.

A **loss budget** is a listing of all loss factors applicable to a given radar system operating in a given mode and environment. The loss should be divided into the four classes shown in Tables L2 to L5 to permit proper values to be used in different forms of the radar equation. A typical loss budget, applicable to a short-range search radar using MTI, is shown in Table L8. *DKB*

Ref.: Barton (1988), p. 28..

Table L8
Typical Search Radar Loss Budget

Component	Symbol	Loss in dB
Atmospheric loss	L_α	1.2
Beamshape loss	L_p	1.3
Beamwidth factor*	Ln	1.2
Filter matching loss	L_m	0.8
Fluctuation loss (for $P_d = 0.9$)	L_f	8.4
Integration loss	L_i	3.2
Miscellaneous signal-processing loss	L_x	3.0
Receive line loss	L_r	1.0
Transmit line loss	L_t	1.0
Total system loss	L_s	21.1
* This factor is included only in the search radar equation		

Clutter distribution loss results from clutter having a probability density function (pdf) broader than the Rayleigh distribution. The broader pdf requires a higher threshold, relative to the average clutter level, to achieve a given probability of false alarm, whether this threshold is set manually or by an adaptive CFAR process. Figure L22 shows this loss for Weibull and log-normal pdfs. The loss can be minimized by using a limiter or clutter map gain control to remove clutter peaks before the CFAR process. *DKB*

Ref.: Barton (1988), pp. 92–94

The **clutter-reference loss function** expresses the reduction in output signal-to-noise ratio $(S/N)_o$ in a noncoherent MTI

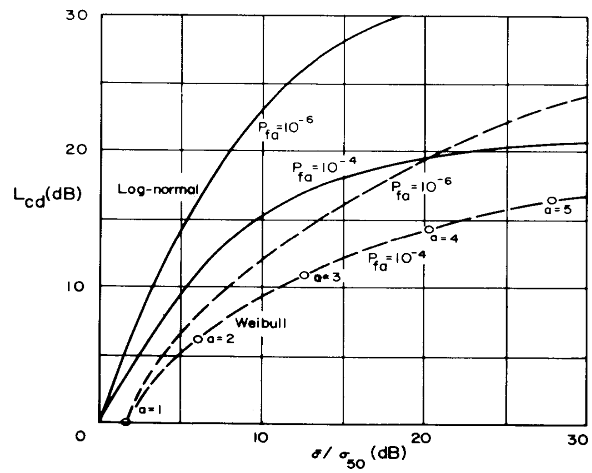


Figure L22 Clutter distribution loss for Weibull and log-normal clutter (from Barton, 1988, Fig. 2.5.4, p. 93).

system. The output ratio is related to input S/N and clutter-to-noise ratio C/N by

$$\left(\frac{S}{N}\right)_o = \frac{2(C/N)}{1+2(S/N)+2(C/N)}\left(\frac{S}{N}\right)$$

where the factor multiplying S/N is the clutter-reference loss function, expressed as an efficiency factor (value < 1, multiplying S/N). *DKB*

Ref.: Currie (1987), p. 266.

Collapsing loss is the result of video integration that includes extra noise samples along with the intended signal-plus-noise samples. It can occur when additional noise samples are combined with signal-plus-noise samples in an integration cell with the actual target because receiver bandwidth is wider than optimum, when the cell extends beyond the pulse width, or for other reasons. For a square-law detector followed by an integrator processing n samples of signal plus noise and m additional noise samples, a collapsing ratio is defined as $\rho = 1 + m/n$. The collapsing loss is then calculated as the increase in integration loss:

$$L_c(\rho n) = \frac{L_i(\rho n)}{L_i(n)}$$

where $L_i(\cdot)$ is the integration loss (Fig. L27). For $n > 10$, $L_c \approx \sqrt{\rho}$.

As in the basic video integration process, the loss is reduced slightly relative to that shown in Fig. L28 if the integration of more samples reduces the number of opportunities for generation of a false alarm, permitting P_{fa} to be increased for a given false-alarm time. Table L9 shows several sources of collapsing loss, indicating those for which P_{fa} must be held constant (larger L_c as given directly by Fig. L28), and those for which a new $P_{fa}' = \rho P_{fa}$ is acceptable. In the latter case, L_c is reduced by the ratio of detectability factors, $D(P_d,P_{fa})/D(P_d,P_{fa}')$, which may be determined from curves given under **DETECTION curves**.

In Table L9, use of an uncoded, rectangular pulse is assumed. With pulse compression the value τ may be replaced with $\tau_n = 1/B$, and collapsing loss will not depend on the pulse compression ratio but only on the ability of the subsequent processing to retain the resulting resolution. In pulsed doppler systems, the effect of mismatched range gate or receiver bandwidth appears directly as a matching loss, rather than as the lower collapsing loss applicable to systems using noncoherent integration. *DKB*

Ref.: Blake (1980), p. 49; Barton (1988), pp. 77–79.

Table L9
Equations for Collapsing Ratio

Cases for which P_{fa}/ρ remains constant:	
(a) Restricted CRT sweep speed, s, where d is spot diameter and τ is pulse width	$\rho = \dfrac{d + s\tau}{s\tau}$
(b) Restricted video bandwidth, B_v, where B is IF signal bandwidth	$\rho = 1 + \dfrac{1}{2B_v\tau} = 1 + \dfrac{B}{2B_v}$
(c) Collapsing of coordinates onto the display:	
$2\Delta_r/c$ is time-delay interval per display cell	$\rho = \dfrac{2\Delta_r}{c\tau}$
$\omega_e t_v$ is elevation scan sector, θ_e is elevation beamwidth	$\rho = \dfrac{\omega_e t_v}{\theta_e}$
$\omega_a t_v$ is elevation scan sector, θ_a is elevation beamwidth	$\rho = \dfrac{\omega_a t_v}{\theta_a}$
Cases for which P_{fa} remains constant:	
(d) Excessive IF bandwidth, $B >$ $1/\tau$, followed by matched video	$\rho = 1 + \dfrac{1}{B\tau}$
(e) Receiver outputs mixed at video, m is number of receivers	$\rho = m$
(f) IF filter followed by gate of width τ_g and by video integration	$\rho = \dfrac{1}{B\tau} + \dfrac{\tau_g}{\tau}$

Constant-false-alarm-rate (CFAR) loss is the result of using a CFAR circuit to establish the detection threshold, rather than using a fixed threshold based on exact knowledge of the noise (or interference) level and statistics. It is defined as the increase in signal power required to achieve a given detection performance using the CFAR process on noise (or interference) of unknown level and with Rayleigh distributed, relative to that required for a fixed threshold on known level. For a class of conventional, cell-averaging CFAR, the loss is given by Fig. L23, where the parameter x is the negative exponent of P_{fa} (e.g., $x = 6$ for $P_{fa} = 10^{-6}$), and the CFAR ratio is x/m_e, where m_e is the effective number of reference samples.

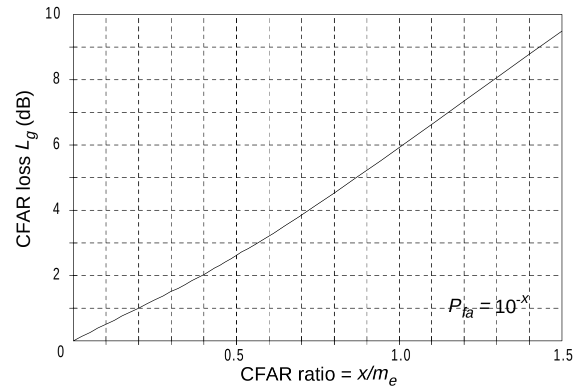


Figure L23 Universal curve for CFAR loss in single-hit detection, for steady or Rayleigh target (after Gregers-Hansen).

The general equation for m_e is

$$m_e = \frac{m + k}{1 + k}$$

where values of k are shown in Table L10, and m is the number of reference cells used to form the threshold. *DKB*

Table L10
Constants Determining Number of Effective CFAR Reference Samples

Square-law detector	$k = 1$
Linear envelope detector	$k = 0.09$
Log detector	$k = 0.65$
Greatest-of CFAR, square-law detector	$k = 0.37$
Greatest-of CFAR, linear envelope detector	$k = 0.5$
Greatest-of CFAR, log detector	$k = 1.26$
Hard-limiting CFAR with Dicke fix (add 1-dB limiter loss)	$m_e = \dfrac{B_w}{B_n} - 1$
Hard-limiting dispersive or pulse-compression CFAR (add 1-dB limiter loss)	$m_e = B\tau - 1$

Ref.: Gregers-Hansen, W., “Constant False-Alarm Rate Processing in Search Radars,” *IEE Intl. Radar Conf. Radar-73*, Oct. 1973, pp. 325–332; Barton (1988), pp. 88–92.

Conversion loss refers to the loss in signal-to-noise ratio S/N in passing through the mixer in a superheterodyne receiver. The loss does not appear directly in the radar equation but is included in the receiver noise factor. *DKB*

Ref.: Van Voorhis (1948), p. 18.

Crossover loss results from the squinting of the beam axis from the tracking axis in conical scan or other types of lobe-switching trackers. It appears as a component of pattern-propagation factor in the radar equation, and directly in the equation for tracking error of a conical scan radar (see **accuracy of sequential lobing**). *DKB*

Ref.: Barton (1988), p. 387.

Detector loss results from passing a signal at finite S/N through an envelope detector, or in a monopulse tracker from using the sum channel signal at finite S/N as the reference in a phase-sensitive error detector. For detection calculations, the detector loss may be modeled as $C_x = 1 + 2.3N/S$. The video integration loss is a direct result of detector loss, and may be calculated as $L_i(n) = C_x(n)/C_x(1)$, where $C_x(n)$ applies to the reduced S/N value made possible by integration, and $C_x(1)$ is the value that would apply for the S/N of a single pulse giving the same detection performance. The loss $C_x(1)$ reflects the difference between detection performance with the envelope detector and that with a coherent detector (known signal phase). *DKB*

Ref.: Barton (1988), pp. 64, 467.

Display loss describes the loss in detection performance, relative to the theoretical performance of an optimum detector, when target detection is performed by an operator viewing a CRT display. Factors causing this loss are (1) insufficient dynamic range in the receiver or video circuits, (2) inadequate brightness of the display, (3) defocusing of the CRT (a form of collapsing loss), (4) nonoptimum displayed pulse widths, (5) insufficient noise deflection or intensity on the display, (6) excessive numbers of resolution elements on the display, (7) crowding of data in the region near the center of a PPI display, and (8) operator fatigue (usually considered separately as operator loss. *DKB*

Ref.: Nathanson (1969), pp. 100–106.

Doppler filter matching loss results from failure to match the narrowband filters of a doppler radar to the entire waveform envelope (over the observation time t_o). This mismatch may be intentional, as when inputs to an FFT processor are subject to weighting to reduce filter sidelobes), or the result of equipment economies. The efficiency (reciprocal of loss) for a weighted filter is shown in Fig. L24, as a function of sidelobe level, for different families of weighting. In this case the loss is also termed weighting loss. *DKB*

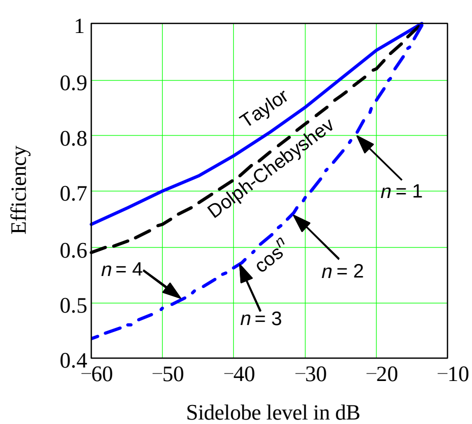


Figure L24 Doppler filter weighting efficiency vs. sidelobe level for different families of weighting function.

Ref.: Barton (1969), App. B.

Duplexer loss is a component of the transmit and receive line losses, resulting from attenuation in passage of the signal through the duplexer.

Eclipsing loss is the result of signal arrival during the transmission of a pulse, either from targets beyond the first range interval $R_u = c/2f_r$ in a medium- or high-PRF radar, or within that interval at ranges within one pulse length $c\tau/2$ of zero range or R_u. The loss is defined as the increase in signal energy required to obtain a given detection performance, relative to that required for target signals that are received during times the transmitter is off. In medium- and high-PRF radars using PRF diversity to avoid eclipsing and blind regions, the average loss in received energy, indicative of the eclipsing loss for $P_d \approx 0.5$, may be estimated as the reciprocal of the fraction of the pulse repetition interval which is not occupied by the transmission:

$$L_{ec} = \frac{1}{1 - \tau f_r}$$

where τ is the transmitted pulse width and f_r is the PRF. For any given PRF, pulse width, and signal delay, the loss may be calculated as the ratio of processor signal output power to that obtained on a pulse in the clear region. *DKB*

Barton (1988), p. 270.

Faraday rotation loss occurs when the radar wave travels through the ionosphere, interacting with the earth's magnetic field (see **ANGLE, Faraday rotation**). Because the rotation is nonreciprocal, the total change in polarization angle of the wave is twice that experienced in the one-way path. The resulting polarization factor, representing the received voltage relative to what would have been received under free-space conditions, will be, for transmit and receive antennas using the same linear polarization, $F_p = \cos\phi_f$, where ϕ_f is the Faraday rotation angle. The corresponding loss is defined as the increase in signal energy required to achieve a given detection probability, compared with that required for matched polarization. For a specific rotation angle ϕ, this loss is

$$L_{fp}(\phi_f) = \frac{1}{F_p^2} = \frac{1}{(\cos\phi_f)^2}$$

For radars in the VHF region, the rotation may be large enough to give a uniform distribution over 2π radians, and the Faraday rotation loss will be a function of detection probability. The loss may be avoided by using circular polarization for both transmit and receive antennas. If polarizations of opposite sense are used, the full target RCS will be observed, while use of same-sense polarizations will cause a polarization loss. *DKB*

The **field degradation loss** is a factor sometimes included in the radar equation to account for deterioration of radar performance and imperfect maintenance subsequent to manufacture and installation of the radar. Its value may be determined by subjective estimation of the quality of maintenance personnel, or by flight testing. *DKB*

Filter matching loss results from departure of the actual receiver filter from the matched filter for the transmitted waveform, on a single-pulse basis. It is defined as the ratio of receiver output signal-to-noise ratio to that available from a filter matched to the individual pulse. For simple rectangular pulses (having no phase modulation), the loss for different types of filter is shown in Fig. L25. For linear-FM pulse compression signals, the loss as a function of sidelobe level, for different families of weighting function, is identical to that for doppler filters, shown in Fig. L10. *DKB*

Ref.: Barton (1969), pp. 56, 84.

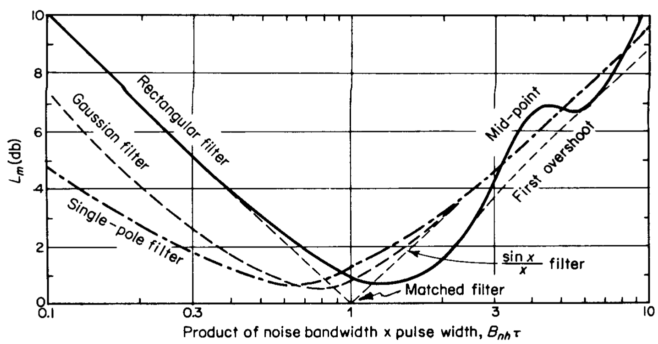


Figure L25 Filter matching loss vs. filter noise bandwidth B_{nh} for rectangular pulse of with τ with different filters (from Barton, 1969, Fig. 3.17, p. 84).

The **filter matching [loss] factor** M differs slightly from the filter matching loss in two cases:

(1) A narrowband mismatch reduces the number of opportunities for false alarm by a factor x, equal to the ratio of signal bandwidth to filter bandwidth. This permits the threshold to be lowered (false-alarm probability increased from P_f to xP_f) while still maintaining the required false-alarm time. The filter matching factor is then equal to the filter matching loss multiplied by the ratio (< 1) of detectability factors for the two false-alarm probabilities:

$$M = L_m \frac{D(xP_f)}{D(P_f)}$$

(2) A wideband mismatch at IF if followed by a video bandwidth less than half the IF bandwidth (or an equivalent broad range gate) is accounted for by calculating a collapsing loss (Table L9) rather than taking the full loss in IF SNR. The value of collapsing loss L_c may be substituted for M, in which case L_c is excluded from the miscellaneous signal-processing loss.

The difference between L_m and M is seldom more than a fraction of one decibel, but it is M that is used to calculate the effective detectability factor used in the radar equation. In the Blake chart, the value M is included to describe the net effect of a filter mismatch. *DKB*

Ref.: Barton (1988), p. 78.

Filter straddling loss results from the use of a fixed doppler filter bank to process signals from targets having arbitrary doppler shifts, some of which fall in regions having response lower than the maximum. It is defined as the increase in signal energy required to achieve given probability of detection for target doppler shifts uniformly distributed over the filter bank response, compared with a target centered in a doppler filter. The loss is minimized by use of closely spaced filters, such as those formed in an FFT filter bank when low-sidelobe weighting is used, or use of zero-padding of the FFT input (a process that adds strings of zeroes at each end of the n-pulse data sample, decreasing the filter spacing while leaving the filter bandwidth constant).

An approximation for L_{ef} is

$$L_{ef} = 1.25P_d^{\frac{1}{3}}\left(\frac{\Delta f}{B_f}\right)^2 \quad \text{(dB)}$$

where Δ_f is the filter spacing, B_f is filter bandwidth, and P_d is detection probability. For the FFT processor, $\Delta f/Bf = 1/t_f B_f$, and the straddling loss will be a function of the weighting function (Table L11) *DKB*

Barton (1993), p. 133.

Table L11
Bandwidth Constants for Weighted FFT

Weighting function	Bandwidth constant t_fB_f
Uniform (rectangular)	0.886
Cosine	1.19
Cosine-squared	1.44
Taylor	$0.9 - 0.0135(G_s + 15)$
Dolph-Chebyshev	$0.8 - 0.0135(G_s + 10)$
G_s is sidelobe level in dB relative to mainlobe.	

Fluctuation loss results when targets have other than steady RCS and depends on the fluctuation model. It is defined as the increase in average signal energy required to achieve a given detection probability, compared to that required for a steady target. The loss is greater for high detection probabilities, where target fading can reduce detection probability more than the increase for upwards fluctuations. The fluctuation loss $L_f(1)$ for cases in which a single sample from a Rayleigh target is available is shown in Fig. L26, as a function of detection probability P_d for different false-alarm probabilities P_{fa}. It is defined as

$$L_f(1) = \frac{D_1(1)}{D_0(1)}$$

where $D_0(1)$ is the detectability factor for a steady target and $D_1(1) = (\ln P_{fa}/\ln P_d) - 1$ is the detectability factor for the Swerling case 1 fluctuating target.

This case applies to signals resulting from multiple scattering centers on the target whose relative phase does not change significantly during the observation time t_o (giving correlation time $t_c >> t_o$). Fluctuation loss also depends slightly on the number of pulses integrated. For n-pulse integration, the loss is approximated as

$$10\log L_f(n, 1) = (10 + 0.03n)\log L_f(1)$$

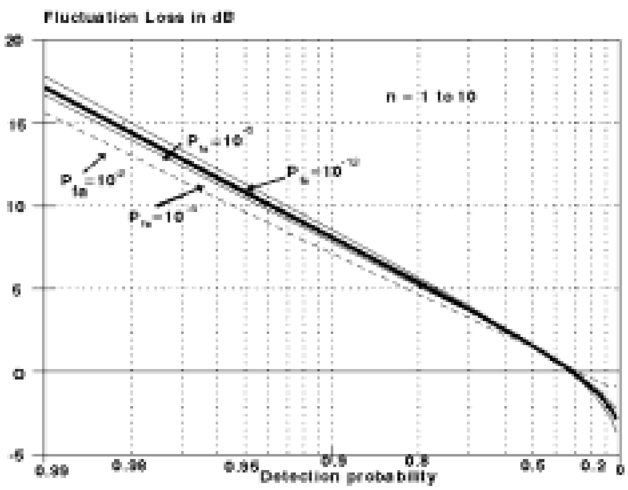


Figure L26 Fluctuation loss for a slowly fluctuating Rayleigh (Swerling case 1) target, as a function of detection probability for different false-alarm probabilities.

The fluctuation loss for other target fluctuation cases modeled by the chi-square probability density function can be derived from $L_f(n,1)$ using

$$10\log L_f(n, Kn_e) = \frac{10}{Kn_e}\log L_f(n, 1) \quad \text{(dB)}$$

where n_e is the number of independent target samples available for integration and K is one-half the number of degrees of freedom of a chi-square distribution describing the target pdf. The four Swerling cases correspond to the values of K shown in Table L12:.

Table L12
Number of Target Samples for Swerling Target Models

Case	K	n_e	Kn_e
1	1	1	1
2	1	n	n
3	2	1	2
4	2	n	$2n$

The use of diversity (in time, frequency, space, or polarization) is effective in reducing fluctuation loss, since n_e is the number of independent signal samples. The number of independent target samples available during observation time t_o is

$$n_e = 1 + \frac{t_o}{t_c} \leq n$$

while that provided by frequency diversity is

$$n_e = 1 + \frac{\Delta f}{f_c} \leq n$$

where Δf is the diversity bandwidth, $f_c = c/2L_r$ is the target correlation frequency, and L_r is the radial length of the target. Polarization diversity can provide a factor of two increase in n_e. *DKB*

Ref.: Barton (1988), p. 84.

Insertion loss refers to the attenuation inserted by a passive component into a signal path (see **ATTENUATION**).

Integration loss refers to the loss, relative to ideal (coherent) integration of signal samples, resulting from integration after envelope detection. The loss results from the increase in detector loss as input SNR is reduced and may be calculated as a function of n, the number of pulse integrated and the basic, single-pulse detectability factor $D_0(1)$:

$$L_i(n) = \frac{1 + \sqrt{1 + \dfrac{9.2nD_0(1) + 2.3}{[D_0(1)]^2}}}{1 + \sqrt{1 + \dfrac{9.2D_0(1) + 2.3}{[D_0(1)]^2}}}$$

This loss is plotted in Fig. L27. Note that the loss is moderate for small n, increasing for large n as $\sqrt{n}$. When the integration is performed digitally with binary representation of the signal amplitude, there will be an additional *binary integrator loss* of about 1.6 dB. *DKB*

Ref.: Barton (1988), pp. 71–73.

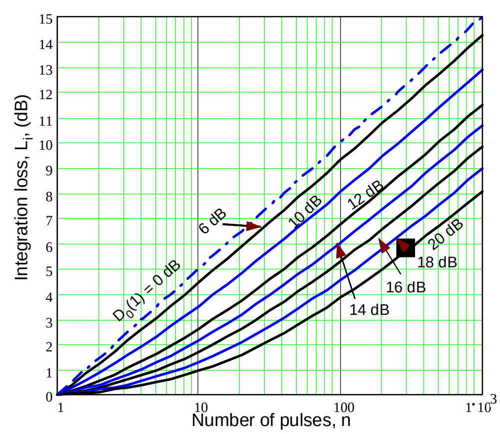


Figure L27 Integration loss vs. number of pulses for different values of basic detectability factor.

Integrator weighting loss is the result of failure to match the integrator weighting function to the signal envelope. For example, if an approximately Gaussian signal envelope results from a scanning beam, use of a rectangular weighting function extending over $t_i = 0.88t_o$ (the optimum value for rectangular weighting) causes a loss $L_{iw} \approx 0.35$ dB. This is the difference between Blake's beamshape loss of 1.6 dB and the actual beamshape loss $L_p = 1.25$ dB for a matched integrator. (See **beamshape loss**.) *DKB*

Ref.: Barton (1988), p. 76.

lens(-effect) loss (see **atmospheric lens loss**).

Limiter [limiting] loss results from passing the signal through a hard limiter at intermediate frequency before further processing. This loss is applicable to certain types of CFAR processor, such as the Dicke fix. The loss is equal to 4/π or 1.05 dB. *DKB*

Ref.: Barton (1988), p. 89.

The **loss tangent** is the ratio of the imaginary part of the complex dielectric constant to the real part. The loss in passing through a material is proportional to this loss tangent and depends in a complex way on the thickness of the material and the wavelength. *DKB*

Ref.: IEEE (1993), p. 745.

Miscellaneous signal-processing loss is the product of receiver-processor losses (type 4 in Table L5), excluding beamshape loss, filter matching loss, fluctuation loss, and integration loss. Since it depends on detection probability, it should be included in the equation for effective detectability factor, rather than as a constant parameter in the radar equation. *DKB*

Ref.: Blake (1980), p. 378; Barton (1988), pp. 31, 250, 268.

MTI processing loss is the loss resulting from passing the signal through an MTI processor, compared with the normal video channel. It is defined as the increase in required input signal energy for the actual process relative to that required for processing without the MTI. It can be divided into three components:

$$L_{mti} = L_{mti(a)} L_{mti(b)} L_{mti(c)}$$

where $L_{mti(a)}$ results from correlation of noise at the MTI filter output, $L_{mti(b)}$ is the blind-phase loss, and $L_{mti(c)}$ is the velocity response loss. It is dependent on detection probability. The effect of noise correlation can be expressed in terms of a reduction in number of pulses integrated from n to an, where

a = 2/3 for a single canceler using I and Q processing.
a = 18/35 for a dual canceler using I and Q processing.
a = 20/47 for a triple canceler using I and Q processing.
a = 1/m for a batch-process MTI using I and Q processing on m pulses per batch.
a = 1/2 times the above values for systems using I-only cancelers.

The resulting loss can be expressed as the ratio of basic detectability factors:

$$L_{mti(a)} = \frac{D_0(an)}{D_0(n)}$$

DKB

Ref.: Barton (1988), p. 250.

Operator loss refers to the inability of the operator to detect targets according to statistical detection theory. While an alert operator, observing an optimized CRT display, can closely approach the theoretical performance described by the detectability factor, this performance is degraded by fatigue, distraction by external factors, nonoptimum displays and lighting, and similar effects. No precise data on operator loss exists, but estimates from 3 to 10 dB are often applied. *DKB*

pattern-propagation [loss] factor (see **PROPAGATION, pattern-propagation factor**).

Polarization loss refers to the reduction in signal power, relative to the use of the same linear polarization for transmit and receive antennas, resulting from use of other polarizations. For spheres and typical aircraft targets observed with different polarizations, Table L13 shows the loss. This loss is sometimes included in the range equation as a polarization factor F_p^4 in the numerator. *DKB*

Ref.: Johnson (1984), p. 23.8.

Table L13
Polarization Loss Estimates

Polarizations (transmit/receive)	Loss in dB	
	Typical aircraft	Sphere
H/H or V/V	0	0
H/V or V/H	6	∞
H/R, H/L, V/R, or V/L	3	3
R/H, R/V, L/H, or L/V	3	3
R/R or L/L	3	∞
R/L or L/R	0	0
H = horizontal, V = vertical, R = right-hand circular, L = left-hand circular		

processing loss (see **miscellaneous signal-processing loss**).

propagation loss (see **atmospheric loss**).

pulse width loss (see **array bandwidth loss**).

Quantization [quantizing] loss refers to the loss in signal detection performance resulting from quantizing the receiver output in an A/D converter before digital processing. An extreme case is the binary integration loss, $L_b \approx 1.6$ dB, when one-bit quantization is used. When quantizing is carried out to m bits, the loss is approximately

$$10 \log L_q = \frac{1.6}{m^2} \text{ (dB)}$$

DKB

Ref.: Nathanson (1991), p. 662.

Radome loss refers to the reduction in power received from a target as a result of attenuation through the radome material, scattering from radome structure, effects of precipitation on the radome surface, and possible distortion of the beam due to refraction in the radome. Typical two-way losses for dry radomes are less than 1 dB (the average figure for L-band is 0.2 dB and for S-band is 0.5 dB). *DKB*

Ref.: Skolnik (1970), Ch.14.

Range cusping loss (see **range straddling loss**).

Range-gate matching loss is the equivalent of filter matching loss when a wideband IF receiver is used with range gates followed by narrowband filtering to form the correlator equivalent of an approximately matched filter. For a rectangular pulse of width τ passed through a gate of width $\tau\gamma$, the loss is

$$L_m = (\tau_g/\tau)^2, \tau_g \geq \tau$$

$$L_m = (\tau/\tau_g)^2, \tau_g \leq \tau$$

In general, the frequency response of the cascaded receiver filter, $H_1(f)$, and range gate is

$$H(f) = H_1(f)\frac{\sin \pi f \tau_g}{\pi f \tau_g}$$

Thus, for $H_1(f)$ approximately matched to the pulse spectrum, a gate of any nonzero width causes a mismatch (too narrow a system bandwidth). The optimum range gate following a matched IF filter is a sampling impulse. *DKB*

Ref.: Barton (1969), p. 85.

Range straddling loss results from reception of signals not centered in a range gate or on a sampling strobe. The loss is defined as the increase in signal energy required to achieve a given detection probability, for signals centered at random points over the gate, relative to that for a centered signal. The loss may be minimized by using overlapping gates or strobes spaced at intervals less than the processed pulse width.

The loss can be approximated as

$$L_{er} = 1.25 P_d^{\frac{1}{3}}\left(\frac{\Delta t}{\tau}\right)^2 \qquad \text{(dB)}$$

where Δt is the gate or strobe spacing (between leading edges), τ is the gate or pulse width. *DKB*

Ref.: Barton (1993), p. 133.

receiver matching [mismatch] loss (see **filter matching loss**).

sampling loss (see **straddling loss**).

Scan distribution loss is a loss used in the search radar equation to describe the effects of distributing signal energy, available during the maximum allowable frame, over more than a single scan. When compared with integration of all n pulses available from a single scan, the use of k scans gives integration gain on $n' = n/k$ pulses, with the results of k detection attempts combined using the cumulative probability of detection:

$$P_c = 1 - (1 - P_d)^k$$

The cumulative detection process is far less efficient than video integration, as shown in Fig. L28.

If video integration gives a detectability factor $D_0(n)$ for n hits in a single scan, and the cumulative process gives $D_{cum}(n,k)$, when the n hits are distributed over k scans, the scan distribution loss can be calculated as

$$L_d(n,k) = k\frac{D_{cum}(k)}{D_0(1)}\frac{L(n/k)}{L_i(n)}$$

where $L_i(.)$ is the integration loss. For example, with $n = 20$, $k = 4$, from Fig. L27 with $P_d = 0.9$, we find $D_0(1) = 13$ dB, $D_{cum}(4) = 11.2$ dB. The integration loss is the difference between the curve for coherent integration and that for video integration, $L_i(n/k) = L_i(5) = 1.0$ dB, $L_i(20) = 3.2$ dB. With $k = 4 = 6$ dB, the scan distribution loss is $L_d(20,4) = 1.58 = 2.0$dB. *DKB*

Ref.: Barton (1988), pp. 31, 74.

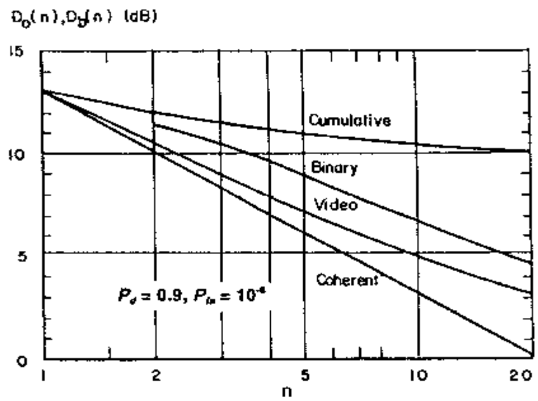


Figure L28 Comparison of detectability factors for different methods of integration.

Scanning loss is the result of scanning a beam at speeds such that the echo pulses arrive after the beam has moved significantly from its position at the time of transmission. The effect is to narrow the two-way beamwidth, as shown in Fig. L29. The resulting loss is shown in Fig. L30.

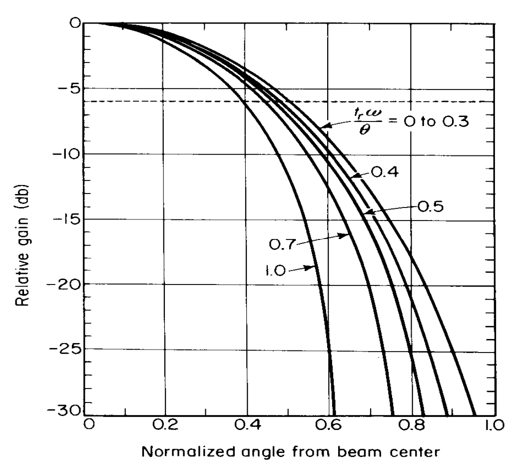


Figure L29 Effective beam patterns for rapid scan (two-way radar case) (from Barton, 1964, Fig. 5.6, p. 150).

The term *scanning (pattern) loss* has been used to denote beamshape loss (see Blake, 1980, p. 49), but that is an entirely separate issue, not to be confused with the effect of the receiving beam not being directed toward the echoes from the preceding transmission. *DKB*

Ref.: Barton (1964), p. 150.

signal-processing loss (see **miscellaneous signal-processing loss**).

The **squint loss** is the amount by which the gain is reduced when the mainlobe is radiated at an angle other than the normal to an array face, as in some frequency-sensitive waveguide feeds. The beam is also described as being squinted from the tracking axis in a conical-scanning radar, but in that case the loss is described as crossover loss). *DKB*

Ref.: Barton (1988), p. 150.

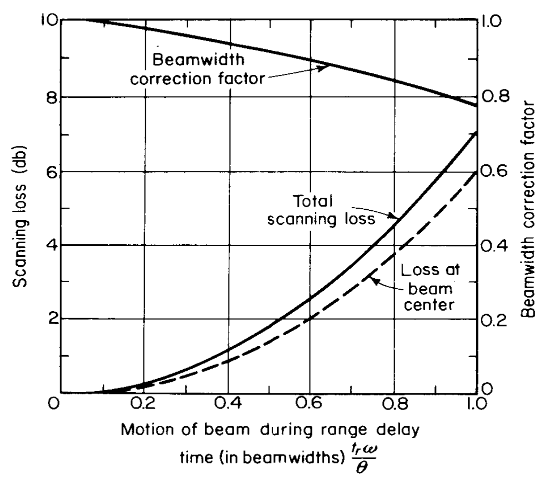


Figure L30 Scanning loss vs. scan speed (from Barton, 1964, Fig. 5.6, p. 150).

A **statistical loss** is one that depends on detection probability, such as beamshape loss or straddling loss. See Table L5 for a list of such losses.

Straddling loss refers to a loss resulting from targets' being distributed around points of maximum system response in range, angle, or doppler frequency. When this results from sampling in a digital system, it may be referred to as *sampling loss*. (See **angle straddling loss**, **filter straddling loss**, **range straddling loss**.) *DKB*

The **system loss** is the product of all losses applicable to a radar operating in a particular mode and environment. (See **loss budget**.)

Taper (illumination) loss refers to the reduction in antenna gain, relative to that obtained with a uniformly illuminated aperture. (See **APERTURE illumination**; **WEIGHTING**.) This loss normally appears as a reduction in antenna gain G_t or G_r used in the radar equation, by the illumination efficiency factor η_i that is one component of the aperture efficiency factor η_a in

$$G_t = G_0\eta_a = G_0(\eta_i\eta_f\eta_s\eta_b\eta_t)$$

where G_0 is the ideal gain for uniform illumination, and efficiency factors for the feed, spillover, blockage, and antenna tolerances are included along with illumination efficiency. If stated separately as a loss, the taper loss would be expressed as $L_{ti} = 1/\eta_i$. *DKB*

Ref.: Barton (1969), App. A.

Transient gating loss in a doppler radar system using batch-processing results from the need to delay processing until clutter with maximum observed range delay has entered the sample to be processed. Upon changing the beam position, carrier frequency, or waveform, a clutter transient occurs, which lasts until clutter from the maximum range has entered the receiver. If the coherent dwell time is t_{cd} and the maximum range of significant clutter is R_{mc}, corresponding to delay $t_g = 2R_{mc}/c$, the transient gating loss is

$$L_{eg} = \frac{t_{cd}}{t_{cd} - t_g}$$

DKB

Ref.: Barton (1988), p. 270.

Transmit [transmission-] line loss is the attenuation in the RF components connecting the point at which transmitter power is measured to the point at which antenna gain is measured. This loss may have components due to the waveguide or coaxial line, the duplexer, rotating joints, directional couplers for sampling the signal, and other RF devices through which the signal must pass. It is entered as a separate loss term in the standard radar equation and in the Blake chart. *DKB*

tropospheric loss (see **atmospheric loss**).

Velocity response loss is the result of some targets signals falling in the stopband of the MTI or pulsed doppler filter (velocity) response. It is defined as the increase in average signal energy required to achieve a given probability of detection with the doppler processor present, compared with that required with an all-pass filter. It depends on the detection probability as well as the velocity response of the filter. The loss may be reduced by use of PRF stagger or diversity, and when stagger is used by defining the velocities of interest as exceeding the width of the rejection notch. The moving target detector, using area MTI processing in the zero-doppler filter output, also reduces velocity response loss when considering targets spread over regions in which clutter gaps are present. (See **VISIBILITY, interclutter.**) *DKB*

Ref.: Barton (1988), p. 251, (1993), p. 134.

video integration loss (see **integration loss**).

Video mixing loss is a type of collapsing loss resulting from mixing of several receiver outputs (e.g., in a stacked-beam radar) before integrating the signals on a display or in a video integrator. (See **collapsing loss**.) *DKB*

Weighting loss is incurred in radar signal processing when a weighting function is applied to reduce time sidelobes in pulse compression or frequency sidelobes in a narrowband doppler filter. In the former case it is sometimes referred to as *pulse compression loss*, and is included in the range equation as the filter matching loss or matching factor $M = L_m$. In the latter case it is included as L_{mf}, a component of the miscellaneous signal-processing loss, L_x.

In a digital signal processor, the weighting function is a sequence of coefficients C_n applied to the input signal at times t_n. An example is the Hamming weighting function

$$C_n = a_0 - a_1 \cos\left(\frac{2\pi n}{N}\right),\ n = 0, \ldots, N-1$$

where $a_0 = 0.53836$, $a_1 = 0.46164$. For this weighting function the loss is

$$L_m \text{ (or } L_{mf}\text{)} = \frac{\left(\sum_{n=0}^{N-1} C_n\right)^2}{N \cdot \sum_{n=0}^{N-1} C_n^2}$$

In the case of pulse-compression weighting, the times t_n are spaced at intervals $1/B$ over the width τ of the uncompressed pulse, where B is the waveform bandwidth, while in pulsed doppler filter weighting they are spaced at intervals t_r over the coherent processing interval, where t_r is the pulse repetition frequency.

The loss is independent of the number of samples, and for Hamming weighting is equal to 0.73 dB. The loss for some other functions is given in Table L14. (See also **WEIGHTING.**) *SAL*

Ref.: Cook (1967), pp. 191–206, Barton (1993), pp. 98, 102.**)**

Table L14
Weighting Loss

Weighting function	Loss in dB
Uniform (rectangular)	0
Cosine	1.0
Cosine-squared	1.8
Taylor	$0.1 - 0.0041(G_s + 15)$
Dolph-Chebyshev	$0.01 - 0.05(G_s + 15)$
G_s is the design sidelobe level in dB relative to the main-lobe peak (from Barton (1993), Table 3.2, p. 98)	

M

MAGIC- [HYBRID-] TEE (see **BRIDGE, waveguide**).

A **MAGNETRON** is a crossed-field microwave tube (oscillator) characterized by the interaction of electrons with the electric field of circuit element in crossed steady electric and magnetic fields and converting to an RF power an output energy extracted from a constant electric field.

Historically, the magnetron is the device that made microwave pulsed radar practical, and it has been in used in different types of radars for over 50 years. It was invented by A. W. Hall in 1921 for use as a diode switch, and until the invention of the resonant cavity in 1939 it remained practically a laboratory device. Now it can obtain the megawatts of output power with efficiency up to 80%.

Early magnetrons were of the unstrapped resonator type. To improve stability, strapped-vane forms were introduced and now such magnetrons compose about 70% of these devices. The remaining 30% include primarily rising-sun magnetrons (considered obsolete) and coaxial magnetrons. The latter developments include a strapped-vane system and high-Q resonant cavity that provides greatly enhanced frequency stability. All existing magnetrons can be classified from the standpoint of the operation mode (pulsed or CW), from the standpoint of frequency variation (fixed-frequency or frequency-agile, tunable or tuned magnetrons), and from the standpoint of their structure (a variety of different types, such as conventional magnetrons, coaxial magnetrons, inverted coaxial magnetrons).

Although magnetrons are the oldest sources of microwave energy used in radar, they are still widely used today because of their "innate" positive qualities: small size, relatively light weight, reasonable operating voltages, excellent efficiency, and long life. The main assets of magnetrons are: wide range of operating frequencies (from meter to millimeter waves); high attainable power combining with relatively small size and weight (from the 15 cm^3, 1-kW peak power beacon magnetron to several megawatts peak power in air defence radars); very high efficiency inherent to crossed-field devices (up to 50 to 80%); low operating voltage (usually too low to generate dangerous x-rays); and low cost. The main disadvantages restricting the usage of these devices in some types of radar transmitters are the following: inability to ensure coherence from pulse to pulse (without using some special measures, such as frequency and phase-locking techniques, which are complicated and not generally attractive); insufficient range of pulse shaping (an order of few decibels); inherent frequency drift and frequency modulation by microphonics from ambient vibration; insufficient stability for generating very long pulses (more than 100 μs) and very short pulses (less than 50 ns); and high spurious power level that produces considerable electromagnetic interference across the bandwidth that is much wider than that of their signals. *SAL*

Ref.: Ewell (1981), pp. 22–37; Skolnik (1980), pp. 192–200, (1990), pp. 4.5–4.9; Currie (1987), p. 448; Brookner (1988), pp. 263, 317; Sivan (1994), Ch. 6.

A **coaxial magnetron** is one of the most common forms of magnetron, in which the stabilizing cavity surrounds the conventional resonators (Fig. M1). In such a configuration the straps are removed and the π-mode is controlled by coupling alternate resonators to a cavity surrounding the anode, giving an improvement in power, efficiency, frequency stability, and life over the conventional magnetron. *SAL*

Ref.: Skolnik (1980), p. 193.

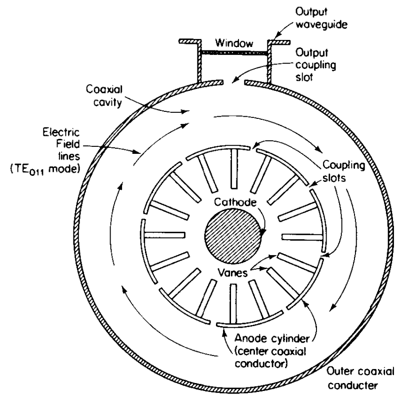


Figure M1 Cross-sectional sketch of the coaxial cavity magnetron (from Skolnik, 1980, Fig. 6.3, p. 194, reprinted by permission of McGraw-Hill).

A **continuous-wave magnetron** operates in the CW mode. The efficiency of such devices is about 30%, power levels are few hundred watts. The range of application is primarily in doppler radar and electronic countermeasure systems. *SAL*

Ref.: Ewell (1981), p. 33; Fink (1975), p. 9.50.

A **conventional magnetron** is the classical structure (Fig. M2) in which the anode is a large block of copper (1) with holes (2) and slots (3), the latter function as the resonant circuits. The holes correspond roughly to inductors, and the slots to capacitors. The cathode (4) is a flat cylinder of the

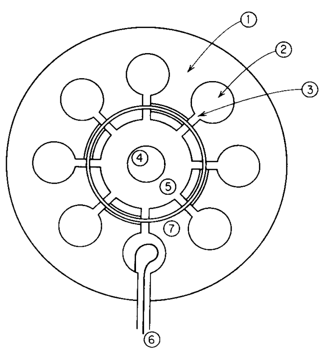


Figure M2 Cross-sectional sketch of classical cavity magnetron (from Skolnik, 1980, Fig. 6.1, p. 193, reprinted by permission of McGraw-Hill).

oxide-coated material. The process of the interaction of the electrons and dc electric and magnetic fields takes place in the interaction space (5). The RF power is extracted by placing a coupling loop (6) in one of the cavities, and the stability and efficiency of the tube is improved by the straps (7): metal rings connected to alternate segment of anode block. The preferred mode of magnetron operation (π-mode) corresponds to an RF field configuration in which the RF phase alternates 180° between adjacent cavities. The conventional magnetron can operate rather successfully through X- or K_u-band. Above this frequency, rising-sun or inverted coaxial magnetrons are typically used. Frequency stability was improved, compared with the conventional magnetron, by developing the coaxial magnetron. *SAL*

Ref.: Ewell (1981), p. 22; Skolnik (1980), p. 192.

A **dither-tuned magnetron** is a mechanically tuned magnetron with an integral motor and resolver to provide frequency-agile operation. A voltage output from a resolver, proportional to the magnetron frequency, is used to adjust the receiver local oscillator to track the rapidly tuned frequency of the magnetron. Such type of magnetron is capable of rapid tuning over a narrow band and also can be tuned to a frequency over a broad band in the normal manner using a geared drive. With servo-motor control it is possible to go from one frequency to another under 0.1 sec. Attainable tuning range and tuning rates are restricted by mechanical limitations imposed by acceleration forces.

Dither-tuning of coaxial magnetrons may also be obtained using an element termed a *ring tuner*, which consists of a narrow ring. This ring is installed in an annular groove cut into the outer wall of the cavity, and projects slightly into the cavity. By deforming the ring inward from mechanical motion applied to the ends of the ring, the frequency in the cavity is changed. *SAL*

Ref.: Skolnik (1980), p. 200; Fink (1975), p. 9.53.

A **frequency-agile magnetron** provides a variable output frequency by changing the resonant frequency of its cavity. In general there are two basic approaches to change the magnetron frequency: electronic tuning and mechanical tuning. Magnetrons using the first approach are termed *voltage-tuned magnetrons*, and those which use the second one are called *mechanically tuned magnetrons* (see **tunable magnetron**).

Ref.: Fink (1975), p. 9.44. *SAL*

A **gyro-tuned magnetron** is a coaxial magnetron providing frequency variation through rotation of several dielectric ceramic paddles in the stabilizing coaxial cavity. *SAL*

Ref.: Fink (1975), p. 9.54.

An **inverted-coaxial magnetron** is one in which the cathode surrounds the anode (Fig. M3). Such a configuration is preferable at the higher frequencies (typically above X-band) because at the high frequencies the cavity becomes very

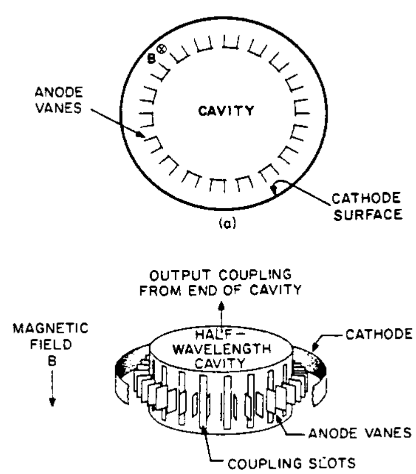


Figure M3 Inverted coaxial magnetron: (a) simplified cross-section; (b) simplified perspective (from Skolnik, 1990, Fig. 4.5, p. 4.7, reprinted by permission of McGraw-Hill).

small, so the usual construction would leave inadequate room for the anode and cathode structure. Sometimes this structure is also called an *inverted magnetron*. *SAL*

Ref.: Skolnik (1990), p. 4.7.

mechanically-tuned magnetron (see **tunable magnetron**).

A **multipactor-tuned magnetron** uses the effect of a multipactor discharge to vary the frequency of an auxiliary coupled resonator, which in turn changes the magnetron operation frequency. Each auxiliary cavity is coupled to a given anode and can be turned on or off independently, so in general $2n$ combinations of frequencies may be selected, where n is a number of cavities. *SAL*

Ref.: Ewell (1981), p. 28.

A **multiresonator magnetron** has an anode unit in the form of connected three-dimensional resonators. In the central opening of the anode unit is a cathode, and in one of the resonators is a connector loop by which the RF output is transmitted from the magnetron to the external circuit (Fig. M4). The anode unit, as a rule, is grounded, and a high negative potential is sent to the cathode.

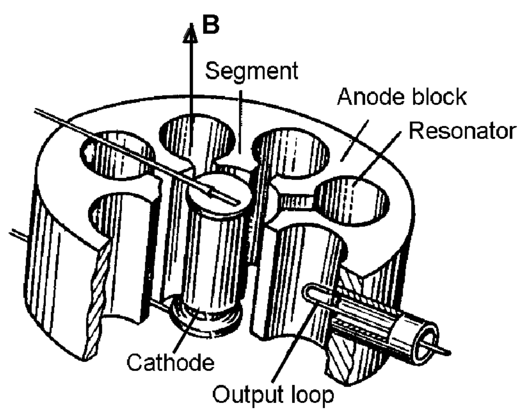


Figure M4 Multiresonator magnetron (from Leonov, 1988, Fig. 2.11a, p. 45).

Transmission of energy from the electron flow to the RF field occurs when the bunches of electrons pass near the slot of the resonator coincident with presence of the RF field in the necessary phase. The electron flows have a complex structure in the form of electron spokes between cathode and anode, rotating in the same space (interaction space). Electrons in the spokes move to the anode over loop-like trajectories as a result of interaction with the external constant magnetic and variable electrical fields.

Multiresonator magnetrons are the basic type used today. Magnetrons for different purposes cover the frequency range of 300 MHz to 300 GHz. The output power of continuous magnetrons is from fractions of one watt to several tens of kilowatts, and for pulsed magnetrons is from 10W to 10 MW. The electronic efficiency of high-power magnetrons can exceed 70%. The advantages of magnetrons include their high efficiency, high output power, the capability of frequency tuning and the use of frequencies over a wide band. However, they do not provide coherence from pulse to pulse, have low frequency stability and a comparatively high power of parasitic radiation.

This device is also called a *multicavity*, *multicircuit*, or *multisectional magnetron*. *IAM*

Ref.: Gilmour (1986), p. 352; Leonov (1988), pp. 45, 46; Andrushko (1981), p. 81.

A **pulsed magnetron** is one operating in a pulsed mode. Pulsed modulation typically is obtained by applying a negative rectangular voltage pulse to the cathode with the anode at ground potential. The devices have been developed covering frequency ranges from a few hundred megahertz to 100 GHz, the peak power up to several megawatts, the efficiency is about 30 to 40%. High-power pulsed magnetrons are primarily used in simple radar transmitters. Low-power pulsed magnetrons find applications in radar beacons.

Some commercially available high-power pulsed magnetrons are listed in Table M1. *SAL*

Ref.: Ewell (1981), p. 28; Fink (1975), p. 9.50.

Table M1
Some Commercially Available High-Power Pulsed Magnetrons

Tube type	Center freq. (GHz)	Peak power	Max duty cycle	Peak	
				kV	A
M 545	1.290	25 MW	0.0025	52	260
3M901	2.765	4.7 MW	0.001	75	135
SFD344	5.60	1.0 MW	0.001	37.5	65
VF 11	9.25	1.0 MW	0.0015	30	70
7208 B	16.5	100 kW	0.001	22	20
VF 20	16.5	400 kW	0.0015	26	40
SFD326	24	120 kW	0.0005	14	30
SFD327	34.86	150 kW	0.0005	23	22
BL235	52.5	10 kW	0.0012	14.5	9.5
DX221	69.75	10 kW	0.00055		
M 5613	95.5	2 kW	0.0002	10	9
DX252	120	2.5 kW	0.0002	10	11
(from Ewell, 1981, Table 2-1, p. 28).					

A **rising sun magnetron** is a magnetron in which large and small slots are alternated forming the “rising-sun” structure (Fig. M5). In this case stable oscillation occur in π-mode without using straps as in the conventional magnetron. Such a configuration is more suitable for the higher frequencies. *SAL*

Ref.: Skolnik (1980), p. 193.

A **spin-tuned magnetron** is a mechanically tuned magnetron in which frequency agility is provided by means of rotating

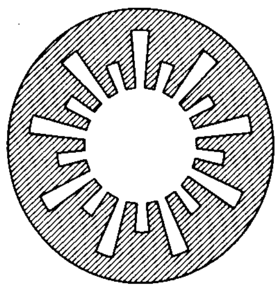


Figure M5 Rising-sun magnetron configuration (from Skolnik, 1980, Fig. 6.2b, p. 194, reprinted by permission of McGraw-Hill).

the slotted disk that is suspended above the anode resonators (Fig. M6). Rotation of this disk provides inductive or capacitive loading of the resonators, the frequency changing up or down, respectively. This technique was developed around 1960 was one of the first for achieving frequency agility in magnetrons. Very fast tuning rates are feasible, but when used for MTI radars stability is lower than with other tuners. *SAL*

Ref.: Skolnik (1980), p. 199; Skolnik (1990), p. 4.6; Fink (1975), p. 9.53.

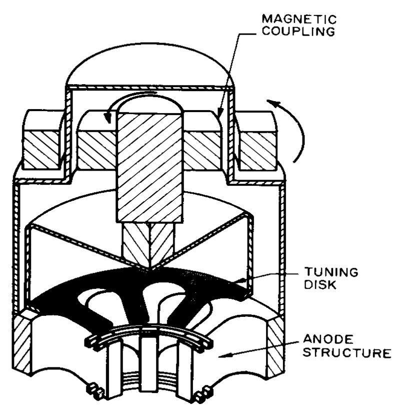


Figure M6 Magnetron rotary tuner (from Skolnik, 1990, Fig. 4.3, p. 4.6, reprinted by permission of McGraw-Hill).

A **stabilized magnetron** provides greater stability than the conventional magnetron. The most common types of stabilized magnetrons are coaxial and inverse-coaxial magnetrons. *SAL*

Ref.: Skolnik (1990), p. 4.7.

A **tunable magnetron** permits changing the output frequency by changing the resonant frequency of its cavity. There are two basic ways to realize change in frequency: electronic tuning and mechanical tuning. Magnetrons employing the first technique are called *voltage-tuned magnetrons* and magnetrons employing the second technique are called *mechanically-tuned magnetrons*.

Electronic tuning uses the electron beam to produce variable reactance in the resonant circuit. One of the techniques is to use ferrites or piezoelectric materials within the cavity to tune the magnetron, as when the voltage is applied across these crystals it can cause the change in cavity tuning. There are major difficulties in practical implementation of electronic tuning at high power levels, so satisfactory operation typically can be achieved only at modest power levels. The typical example of devices implementing electronic tuning are multipactor-tuned magnetrons.

Mechanical tuning uses tuning elements, such as rods or rings, which are inserted into the holes of resonators to change the inductance of the resonant circuit. These elements can move in a reciprocating or rotatory manner. Most of the readily available devices used in radar systems use the approach based on inserting some structures within the cavity and their motion inside it to tune the magnetron. The main techniques to implement mechanical tuning are rotatory (or spin) tuning (see spin-tuned magnetron), dither-tuning (see **dither-tuned magnetron**), and gyro-tuning (see **gyro-tuned magnetron**). Mechanical tuning over a 5 to 10% frequency range is typical (in some cases as much as 25% can be achieved).

The comparison of a number of different techniques to arrange the tuning in medium power K_u-band magnetrons are given in the Table M2. *SAL*

Ref.: Ewell (1981), pp. 26–33; Skolnik (1980), p. 199.

A v**oltage-tunable magnetron** is one using electronic tuning. An example is one using a circular-format, reentrant-stream injected beam that interacts with a standing wave on a low-Q resonant structure to achieve frequency agility. Low-power voltage-tunable magnetrons can find application as local oscillators or swept-frequency generators, while high-power ones are used in electronic countermeasure applications as a source of frequency-modulated noise. This type of magnetron has been designed to achieve CW power outputs at S-band of 500W over 10% tunable bandwidth, with efficiency of 65%. At X-band, power of 1 to 10W has been achieved at 25% efficiency, over tunable bandwidths of 5 to 10%. The structure and equivalent circuit of the device is shown in Fig. M7. *SAL*

Ref.: Ewell (1981), p. 26; Fink (1975), p. 9.54.

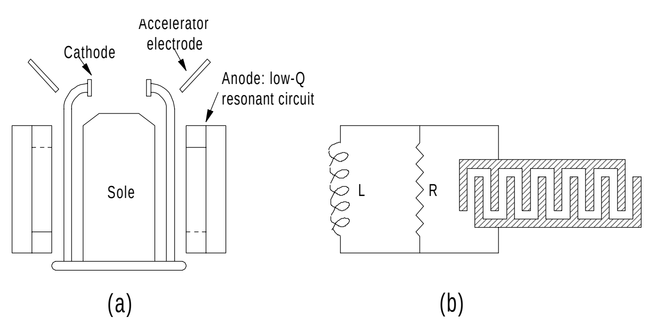


Figure M7 Schematic of voltage-tunable magnetron: (a) structure; (b) equivalent circuit (after Fink, 1975, Fig. 9-66, p. 9-54).

MAINTAINABILITY (see **SERVICE**).

Table M2
Comparison of Frequency-Agility Methods for Medium-Power, K_u-Band Magnetrons

Parameter	Method						
	Litton gyro	Litton E/M	Raytheon ring tuner	Varian dither I	Varian dither II	Amperex spin tune	EEV piezoelectric
Max rate (Hz)	600	400	200	200	200	200	Est 100
Max range (MHz)	300	Total band-width of tube	300	Total band-width of tube	300	1000	Est 50
Typical max rate - range combination	200 Hz 300 MHz	200 Hz 100 MHz	200 Hz 300 MHz	200 Hz 50 MHz	200 Hz 300 MHz	200 Hz 3000 MHz	Est 100 Hz 50 MHz
Rapid-tune mode	Sine only	Sine and/or random	Sine only	Sine	Sine only	Sine only	Sine and/or random
Broadband tunable	Yes	Built into E/M system	Yes	Yes	Yes	No	Yes
Readout accuracy (MHz, max rate range)	±5	±10	±5	±1	±5	No readout	No data available
Nominal operating life (h)	750 min	5,000	650 min	< 500	< 500	No data available	Est. 5,000
Flexibility	Good	Excellent	Est. fair	Poor	Poor	Good	Excellent
Complexity	High	Low, but requires servo amplifier	Moderate	Moderate	Moderate	Low	Low
Vibration	Good	Poor	Moderate	Good	Good	Moderate	Good off piezo resonance
Altitude	Good	Good	Moderate	Moderate	Moderate	Good	Good
Vacuum structure compromises	Complex vacuum sleeve	None	Mode-suppression *Q*s, additional vacuum seals	None	None	Vacuum bearings, *Q*s	Minor
Magnetron type	Coaxial	Conventional or coaxial	Coaxial	Coaxial	Coaxial	Conventional	Conventional

(from Ewell, 1981, Table 2-2, p. 34, reprinted by permission of McGraw-Hill).

MAP

clutter map (see **CLUTTER map**).

A **radar map** is an image of the earth's surface obtained using an terrain-observation radar. Like a terrain map, with imaging of small sections of the terrain including ice surfaces, it is used directly for economic, military, and other tasks. The usual scale of the map is 1:100,000 or 1:250,000. Radar maps are the starting material for the creation of special-purpose maps and topographic, geological, vegetation maps. (See **terrain mapping**.) *IAM*

Ref.: Mel'nik (1980), p. 169; Hovanessian (1984), Ch. 8.

sensitivity-time-control map (see **SENSITIVITY time control**).

A (radar) **weather map** is the representation of the weather conditions on the screen of the radar display. Typically, the different intensities of the signals reflected from the precipitation are digitized and encoded in a way so that each level of intensity is depicted by its own color. *SAL*

Ref.: Sauvageot (1992), p. 273.

MAPPING

Sky object mapping is performed by radar observation of extraterrestrial objects. It applies to the planets which provide a sufficiently stable distribution of surface reflection. Mapping is performed on the basis of doppler frequency shift or delay and doppler shift, with the required spatial resolution achieved using an interferometric method of observation. (See **INTERFEROMETER, radar.**) In radar mapping of the moon, maps were obtained without the use of interferometric methods through antenna angular resolution, which is possible due to the comparatively large angular dimensions of the moon. *IAM*

Ref.: Skolnik (1970), p. 39.25 - 39.28.

Terrain mapping is radar surveying to obtain different types of maps, including topographic. Side-looking radars mounted on aircraft or spacecraft and using either real or synthesized apertures, along with a precision navigational system, are used for terrain mapping. The onboard inertial navigation system of the aircraft provides a precision on the order of 1m, which makes it possible to create a quality radar image on a scale of 1:250,000 or 1:100,000. A specific feature of terrain mapping is the reference of the radar image to a topographic map. Referencing is done either from special marks whose coordinates are strictly associated with the coordinates of the platform, or by aligning the radar blips of the character reference points with the images on the topographic map. A radar mapping system with reproduction of terrain altitude uses radar interferometers, dual-beam stereoscopic sensors, or sensors which survey the terrain from different directions to obtain a stereoscopic effect.

The simplest method of mapping is to obtain a mosaic map. In terrain mapping, radar image correction and decoding are carried out, and joint processing of a large number of images using television-optical equipment and computers. (See also **RADAR, ground-mapping**.) *IAM*

Ref.: Mel'nik (1980), p. 169; Reutov (1970), p. 339.

MARK, calibration. Calibration marks are "the indications superimposed on a display to provide a numerical scale of the parameters displayed." An example is a range marker, a reference appearing on a display that is delayed by a known amount from the transmission, and used to determine the range of the target by reading visually the relative position of the received signal. *SAL*

Ref.: IEEE (1993), p. 150; Barton (1976), p. 38.

MASER (See **AMPLIFIER, paramagnetic quantum**).

MATRIX

beam-forming matrix (see **FEED**).

The **coherence matrix** comprises normalized reciprocal correlation functions of polarization components of a planar wave (components of the field intensity), $\dot{E}_1(t), \dot{E}_2(t)$, in a linear polarization base:

$$\rho = \frac{1}{\rho_{11}+\rho_{22}}\begin{bmatrix}\rho_{11} & \rho_{12}\\ \rho_{21} & \rho_{22}\end{bmatrix}, \rho_{ij} = \langle \dot{E}_i(t)\dot{E}_j\rangle_t,\ i, j = 1, 2$$

where $< x >_t$ is the averaged value of x over the argument t.

The coherence matrix completely determines the polarization state of an electromagnetic wave and can be measured in practice (see **POLARIZATION**).

To describe a field that is scattered by a long object, a more general matrix of space-time correlation functions is used:

$$\rho_{ij}(\Delta\vec{r}, \Delta t) = \langle \dot{E}_i(\vec{r}, t)\dot{E}_j^*(\vec{r}+\Delta\vec{r}, t+\Delta t)\rangle_{\vec{r}, t}$$

where $\vec{r}$ is the space coordinate vector. *IAM*

Ref.: Kanareikin (1966), p. 62; Kraszhuk (1988), p. 96.

A **conductance matrix** is a matrix *Y*, of 2*N* ports, that establishes the relation between *N* normalized voltages $\vec{u}_n$ acting on a microwave multiport and the reaction in the form of *N* normalized currents $\vec{i}_n$:

$$\vec{i}_n = Y_{N\times N}\vec{u}_n$$

The elements of the conductance matrix $\dot{y}_{mn}$ are determined with a short-circuit of the inputs, except for the excited one:

$$\dot{y}_{mn} = \left.\frac{\dot{i}_m}{\dot{u}_n}\right|_{\dot{u}_g = 0,\ q = \overline{1, N},\ q \neq n}$$

The nondiagonal elements are called reciprocal conductances, the diagonal ones natural conductances. The conductance matrix is the inverse of the impedance matrix *Z*: $Y = Z^{-1}$. *IAM*

Ref.: Sazonov (1988), p. 78.

The **covariance [correlation] error matrix** *M*, of measurements of the vector parameter $\hat{\alpha}$ is composed of values of the squares of the deviation of its components, α_i, from their estimates, $\hat{\alpha}_i$, obtained by one method or another:

$$M = \|M_{ij}\| = \left\|\langle(\alpha_i - \hat{\alpha}_i)(\alpha_j - \hat{\alpha}_j)\rangle\right\|, \quad i, j = \overline{1, N}$$

where *N* is the number of the components of $\vec{\alpha}$.

The correlation error matrix characterizes the quality of the estimate method and is used to determine the fundamental accuracy of radar measurement both of independent and of dependent parameters of signals. It is used in the theory of Kalman filtering.

The matrix that is the inverse to the correlation error matrix is called the *precision matrix*. *IAM*

Ref.: Shirman (1981), pp. 191, 207.

The **impedance matrix** is the matrix *Z*, of a 2*N* multiport that establishes the relation between the *N* normalized currents $\vec{i}_N$ acting on a microwave multiport and the reaction in the form of *N* normalized voltages $\vec{u}_N$:

$$\vec{u}_N = Z_{N\times N}\vec{i}_N$$

It is a generalization of the impedance in Ohm's law for a double-port. The element $\dot{z}_{mn}$ of the matrix is determined with excitation of the *n*th input by an ideal current source, i_n, when unloaded ($i_q = 0$, $q = \overline{1, N}$, $q \neq n$) at the other inputs:

$$z_{mn} = \left.\frac{\dot{u}_m}{\dot{i}_n}\right|_{i_q = 0,\ q = \overline{1, N},\ q \neq n}$$

Nondiagonal elements are called reciprocal impedances of inputs of the multiport, and diagonal ones are called natural impedances. *IAM*

Ref.: Montgomery (1947), p. 140; Sazonov (1988), p. 77.

A **multiport matrix** is a square matrix with dimensions $N \times N$, where *N* is the number of inputs of a microwave multiport, establishing the relation between the action matrix $\vec{r}_N$ on the multiport and the reaction vector $\vec{f}_N$. The action and reactions might be various electrical values: normalized (with dimensionality of the square root of watts) voltages, currents entering the multiport, normalized voltages of incident and reflected waves. This determines the many mutually related matrices of a multiport, each of which is a complete external characteristic, since it makes it possible to calculate the reaction to any external action. Three types of matrices have special uses: the conductance matrix, the impedance matrix, and the scattering matrix. The last is introduced analogous to the radar scattering matrix in the form of the relation of reflected and incident waves.

The matrix determines the behavior of a multiport only at a given frequency of oscillations. In the description of behavior of a multiport in the frequency band, the elements of any of its matrixes are converted to complex frequency functions. *IAM*

Ref.: Sazonov (1988), p. 78.

The **polarization scattering matrix** S, is the 2-by-2 matrix that relates the vector components of the scattered field E_s and the incident field E_r:

$$E_s = SE_r$$

The components of these field vectors are complex numbers and describe the components of the scattered and incident waves in some polarization basis. The elements of the scattering matrix are complex numbers which depend on the angle of incidence of the wave at the target, the angle of observation, and the distance between the antenna and the target. The scattering matrix is a complete characterization of the scattering properties of the target, inasmuch as it may be used to determine the amplitude, phase, and polarization of each spectral component of the scattered wave, for a given illuminating field. The matrix derived in this fashion is the bistatic scattering matrix. In the particular case of backscattering, the matrix is called the *monostatic scattering matrix*, and has equal diagonal elements. With a change in the polarization basis, the scattering matrix undergoes a linear transformation. The "relative" scattering matrix is obtained by taking outside the matrix the common amplitude-phase factor due to the target range R_0:

$$\frac{\exp\left(j2\pi\frac{R_0}{\lambda}\right)}{\sqrt{4\pi}R_0}$$

IAM

Ref.: Tuchkov (1985), p. 48; Kobak (1975), p. 48; Knott (1993), pp. 71–74.

The **scattering matrix** S, of a multiport ($2N$ ports), establishes the relation between N incident waves and the reaction vector in the form of N reflected waves. The simplest mode for determining matrix elements is to connect the source of the incident wave to each input of the multiport and the matched loads to all the other inputs. The element $\dot{s}_{mn}$ of the matrix S is equal to the ratio of the normalized voltages of the reflected waves scattered from the multiport $\dot{u}_{om}$, to the normalized voltage of the only incident wave $\dot{u}_{in}$:

$$\dot{s}_{mn} = \left.\frac{\dot{u}_{om}}{\dot{u}_{in}}\right|_{\dot{u}_{iq} = 0,\, q = \overline{1, N},\, q \neq n}$$

Elements of the scattering matrix are nondimensional: the diagonal elements are the reflection factors for each input of a multiport with matched loads at the inputs. *IAM*

Ref.: Gardiol (1984), p. 238; Sazonov (1980), p.72.

MAXWELL'S EQUATIONS (see **WAVE, electromagnetic**).

MEASUREMENT, radar. Radar measurement is the process of estimating target parameters, typically the angular coordinates θ (azimuth and elevation); range R; radial velocity v_r; and radar cross section σ. Information on these parameters is coded in the spatial and temporal structure of the echo signal $y(t)$, which can be represented as

$$y(\vec{p}, t) = y_s(\vec{p}, t) + n(t)$$

where $y_s(\vec{p}, t)$ and $n(t)$ are samples of the useful signal and noise, respectively, $\vec{p} = (\theta, t_d, f_d, A)$ is the vector of measured parameters, θ is the angular coordinate, t_d is time delay (proportional to R), f_d is doppler frequency (proportional to v_r), and A is signal amplitude (proportional to $\sqrt{\sigma}$). The formalized representation of the radar measurement task is then to find an estimate $\hat{p}$ of vector $\vec{p}$ with minimum error $\delta\vec{p}$, based on the received sample, $y(\vec{p}, t)$. Because the received signal is corrupted by random interference (noise, jamming, clutter, or their combinations), the radar measurement if a statistical process and is based on methods of statistical measurement theory. The best developed theory is for measurement in white, Gaussian noise (see **SIGNAL parameter estimation**).

The main parameter describing the quality of radar measurement is the accuracy, defined by the value of errors (see **ERROR, measurement**). In the radar channel there are two stages of measurement (Fig. M8), implemented to reduce the resulting error:

(1) At the first stage, the data on $\vec{p}$ is extracted from the radar echo of a single pulse (in Russian literature this is called *primary signal processing*);

(2) At the second stage, the information accumulated in a series of successive pulses or scans is used to reduce the measurement error (*secondary signal processing*).

Different methods of data filtering (e.g., the Kalman filter) are used at the second stage to improve the quality of radar measurement. (See **FILTER**.) *SAL*

Ref.: Barton (1969).

measurement, accuracy of (see **ERROR, measurement**).

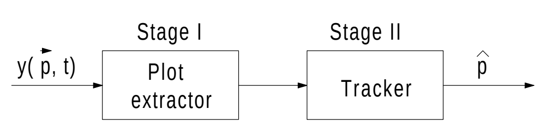


Figure M8 Two stages of radar measurement.

Angular measurement is the estimation of target angular coordinates based on analysis of the spatial structure of the radar signal. The coordinates of angular position are azimuth and elevation (in ground radar), yaw and pitch (in airborne radar), or direction cosines relative to an array face. (See **COORDINATES, radar.**) Angular measurements are based on directional properties of the radar antenna, using either of two signal parameters: amplitude, or phase. In the first, the amplitude of received signals is used to extract information about angle of arrival of the wave; while in the second the signal phase is used. (See **INTERFEROMETER, radar.**) In early radar the main technique for tracking radar was conical scanning, but the monopulse technique is now used to provide more accurate data. (See **MONOPULSE.**) *SAL*

Ref.:Barton (1964), p. 47; Leonov (1988), p. 13; (1986); Dulevich (1978), pp. 253–281.

antenna measurement (see **TESTING, antenna**).

Doppler (frequency or velocity) measurement is the estimation of target radial velocity by measurement of doppler frequency shift. The most common technique is to use a set of doppler filters distributed over an observable frequency interval (a filter bank) and to determine the radial velocity based on the filter in which the strongest signal appears. (See also **DOPPLER EFFECT**; **ERROR, doppler.**) *SAL*

Ref.: Barton (1969), Ch. 4.

measurement error (see **ERROR, measurement**).

Optimum measurement is the estimation of signal parameters using the optimum receiver structure. (See **SIGNAL parameter estimation.**) *SAL*

RCS measurement (see **RADAR CROSS SECTION**).

Range measurement is the estimation of target slant range by analysis of the temporal structure of the received signal. Any of three basic methods may be used: (a) time-related, (b) frequency-related, or (c) phase-related. The first, used in pulsed radars is based on the time delay t_d of the received signal relative to that transmitted. The target range R is computed as

$$R = \frac{ct_d}{2}$$

where $c = 2.9979 \times 10^8$ m/s is the velocity of light. The frequency-related method uses CW signals in which the carrier frequency changes as a function of time in a specific way (e.g., sawtooth modulation). When the modulation function is linear, a transmitted frequency $f_t = f_0 + at$ is mixed in the receiver with the echo frequency $f_r = f_0 + a(t - t_d)$ to obtain

$$\Delta f = f_t - f_r = at_d$$

The phase method uses the difference ϕ in phase between the transmitted and received waveforms that are modulated at a frequency f_m:

$$\phi = 2\pi f_m t_d$$

which gives a direct measure of time delay. This method is also used in CW radars.

The main advantage of the time-related method is the relative simplicity of hardware implementation, and the ability to measure range on many targets within the same beam. The main disadvantage is the inherent inability to measure very short ranges (those with time delay less than the transmitted pulse width), since the receiver cannot operate during the transmission. The main advantage of the frequency-related method is its ability to measure short ranges, and the lower peak power of the transmitted CW signal, as compared with the time-related method. Its main disadvantage is the complexity and size of the hardware (separate transmitting and receiving antennas are necessary for high-power CW radars), and the stringent requirements placed on linearity of modulation (when multiple targets are measured) and on the noise sidebands of the transmission. The phase-related method provides high measurement accuracy but produces many ambiguities, since phase can be measured unambiguously only in the range 0 - 2π. Multiple modulation frequencies may be imposed on the carrier to reduce ambiguity, but the practical implementation becomes even more complicated than with frequency ranging. *SAL*

Ref.: Barton (1964), p. 37; Leonov (1988), p. 13; Dulevich (1978), pp. 215–223.

METEOROLOGY, radar. Radar meteorology is the exploitation of meteorology and radio physics employing radar measurement of precipitation and atmospheric phenomena. Typically, radar applied to the observation of meteorological phenomena provides two types of information:

(1) The signal arriving from a small region (defined by the resolution volume) gives information on the properties of scattering medium (e.g., reflectivity).

(2) The signals arriving from extended targets make it possible to define the contour of the targets (e.g., clouds, thunderstorm areas), and the internal structures of different scale, their evolution and movement.

The first type of information typically is referred to as *quantitative*, and the second, because of the difficulty of incorporating it into quantitative models, is referred to as *qualitative*. Meteorological radar can be ground based or located on movable platforms: ships, aircraft, or spacecraft. Spaceborne meteorological radars make it possible to gather global meteorological data, but they have limitations because of the earth's surface background masking the signals from the hydrometeors. Frequency agility can be used to decrease the effect of the earth background because of the dependence of scattering properties of hydrometeors and surface from wavelength do not coincide. The main features determined by meteorological radars are the height of clouds, detection and classification of thunderstorm and hail clouds, intensity of precipitation, and air humidity. *IAM*

Ref.: Sauvageot (1992); Bogush (1989); Stepanenko (1973), p. 142; Belov (1976), pp. 5, 210; Mel'nik (1980), p. 223.

MICROPHONE [MICROPHONIC] EFFECT. The microphone effect is the changing of the electrical parameters of the components and circuits under the effect of mechanical jolts and oscillations. The microphone effect can create parasitic noise components, and sometimes also passive interference. The effect is reduced by increasing the strength and rigidity of microwave and other elements subjected to vibrations, as well as by employing compensation using a servo system in the event that the effect occurs at low frequencies. *IAM*

Ref.: Skolnik (1970), p. 16-6.

MICROWAVE ADAPTER (see **ADAPTER, microwave**).

MICROWAVE AMPLIFIER (see **AMPLIFIER**).

MICROWAVE ANTENNA (see **ANTENNA**).

MICROWAVE BRIDGE (see **BRIDGE, microwave**).

MICROWAVE CHOKE (see **CHOKE, microwave**).

MICROWAVE CIRCUIT (see **CIRCUIT**).

MICROWAVE CONNECTOR (see **CONNECTOR, microwave**).

MICROWAVE DEVICE (see **DEVICE, microwave**).

MICROWAVE DIODE (see **DIODE, microwave**).

MICROWAVE HYBRID (see **HYBRID, microwave**).

MICROWAVE INTEGRATED CIRCUIT (see **INTEGRATED CIRCUIT**).

MICROWAVE JOINT (see **JOINT, microwave**).

MICROWAVE MULTIPORT (see **MULTIPORT**).

MICROWAVE OSCILLATOR (see **OSCILLATOR**).

MICROWAVE RADIOMETER (see **RADIOMETER**).

MICROWAVE RESONATOR (see **RESONATOR**).

MICROWAVE SWITCH (see **SWITCH**).

MICROWAVE TETRODE (see **TETRODE**).

MICROWAVE TOMOGRAPHY (see **TOMOGRAPHY**).

MICROWAVE TRANSFORMER (see **TRANSFORMER**).

MICROWAVE TRANSISTOR (see **TRANSISTOR**).

MICROWAVE TRANSMISSION LINE (see **TRANSMISSION LINE**).

MICROWAVE TRIODE (see **TRIODE**).

MIRROR, antenna (see **ANTENNA, reflector**).

MISSILE, antiradiation. An antiradiation missile is one with a passive seeker, designed to destroy operating radars of the adversary and guided by their emissions. The seeker includes an antenna system within a radome, a receiver, and a control system. The antenna system as a rule contains one or several parabolic reflectors with horn elements, and also devices that scan the beams. Receivers in antiradiation missiles used against radars operating at constant carrier frequencies are narrowband. Antiradar missiles with wideband receivers are used against radars whose frequency is unknown or known only approximately. Such receivers have lower sensitivity and selectivity. The warhead is detonated by a proximity fuse at the moment that missile closes with the target, at a signal from the control system, or in case the radar goes silent by a contact fuse upon impact. *IAM*

Ref.: Popov (1980), p. 326; Skolnik (1990), p. 19.18.

MIXER. The mixer is "the stage in a heterodyne receiver in which the incoming signal is modulated with the signal from the local oscillator to produce the intermediate frequency signal." In some radars there may be two or three successive conversions, with mixers operating first at the RF and then at the first (and second) IFs. The main types of mixers are described below. *SAL*

Ref.: IEEE (1993), p. 812.

A **balanced mixer** is one providing local oscillator (LO) noise level reduction by means of phase balance at parallel switched diodes. The main structures of balanced mixers are double-diode mixers, double-balanced mixers, and circular and double-circular mixers. The main advantage of a balanced mixer in comparison with a conventional one is the reduction in the noise factor by 2 to 10 dB because of phase suppression of LO noise, improved decoupling of the signal and local oscillator circuits, permitting use of LOs with lower power, higher input power handling capability, and the suppression of the even IF harmonics. In modern receivers, balanced mixers are typically implemented with IC technology based on Schottky-barrier diodes. The noise factor in a 5- to 10-GHz band is about 7 to 10 dB, and conversion losses are about 5 to 8 dB. *IAM*

Ref.: Gassanov (1988), pp. 116, 118; Skolnik (1990), pp. 3.7–3.17.

A **circular mixer** is one using a balanced diode bridge. The signal and local oscillator (LO) voltages are applied to the orthogonal diagonals of this bridge by two matching transformers. The IF signal is filtered by means of microwave chokes. Circular mixers are more broadband than double-balanced mixers (several octaves bandwidth) and LO power is 5 to 10 mW and a typical IF is 70 MHz. The typical mixer based on Schottky-barrier diode performs signal conversion with a noise factor less than 10 dB in the frequency band 1 to 12 GHz.

The double circular mixer can be used to reduce conversion losses due to recovery of image frequency power. It operated analogously to the double-balanced mixer and also provides image frequency interference suppression. *IAM*

Ref.: Gassanov (1988), pp. 120, 122

A **crystal mixer** is one using the nonlinearity of the characteristic of a point-contact diode. The main shortcoming of this mixer is a big spread in parameters and low power handling capability. Conversion losses are about 3 to 10 dB. Typically, it is not used in modern radars.

Ref.:Ridenour (1947), p. 416; Fradkin (1969), p. 35.

A **diode mixer** is one using a microwave diode as the nonlinear component for frequency conversion. Most mixers use a nonlinear volt-ampere diode curve. The nonlinearity of the diode differential capacitance is used only in power varactor mixers for up conversion. Low-noise mixers, used in radar frequency converters and power mixers and to form transmitter frequency, are classified for application range. For operation principles, nonbalanced and balanced mixers are distinguished. The latter have wider range of application because of sidebands suppression and a lower noise level. *IAM*

Ref.: Tsui (1983), p. 439.

The **double-balanced mixer** has two mixing sections, designed to provide image channel reception suppression. The latter is provided due to the fact that the received signals are supplied to the two double-diode mixer in-phase and LO signals are supplied out-of-phase. Phase suppression is implemented through the addition of output IF signals. This mixer has 20 to 30% improvement in bandwidth compared with unbalanced mixer. *IAM*

Ref.:Tsui (1983), p. 443; Gassanov (1988), p. 119.

A **double-diode mixer** is a kind of balanced mixer suppressing even harmonics of RF and local oscillator signals. Typically it employs two single-ended mixers connected in parallel and 180° out of phase. Typically, the balanced mixer incorporates two diodes and coupling element with the signal source and LO. In a decimeter waves band, symmetric coupling loop inserted in coaxial volume resonator of preselector is typically used. In a centimeter waveband, T-bridges, circular bridges, and slot bridges are used. To improve the decoupling factor of the signal and LO inputs, diodes are connected to quadrature bridge (Fig. M9).

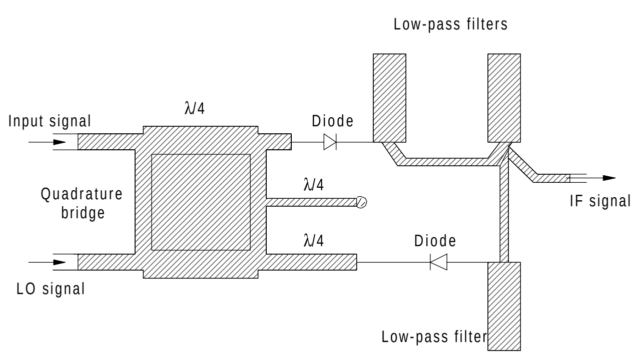


Figure M9 Microstrip double-diode balanced mixer (after Gassanov, 1988, Fig. 4.20, p. 117).

Because of partial circuit unbalance, LO noise is reduced only by 15 to 20 dB. *IAM*

Ref.: Gassanov (1988), p. 116; Skolnik (1980), p. 349.

A **Gunn diode mixer** is a high power mixer using the nonlinear resistance of a Gunn diode. This mixer has comparatively high efficiency (up to 9%) because of the descending part of the volt-ampere curve of the diode and the low level of nonlinear and frequency distortions. However, the level of the internal noise for this mixer is about 10 dB higher than for low-noise mixer based on Schottky-barrier diode. *IAM*

Ref.: Kukushkin, V. V., *Radiotekhnika*, no. 4, 1981, p. 9 (in Russian).

A **high-power mixer** is one used to generate transmitter carrier frequencies by conversion of IF oscillations to the microwave region, and is also used in local oscillator-exciters for the conversion of carrier or LO frequencies. The main requirements of high-power mixer performance are to ensure specified output power with adequate uniformity in operating band; to have sufficient suppression of spurious conversion frequencies; and to have low conversion loss and high stability. Typical high-power mixers use high-power varactors. These are comparatively narrowband and have rather complicated circuitry and tuning. *IAM*

Ref.: Gassanov (1988), p. 126.

An **image-recovery mixer** is one using a reactive image termination without narrowband components. Typically it uses two single-ended, balanced, or double-balanced mixers, and two hybrid junctions as shown in Fig. M10. The left hybrid-

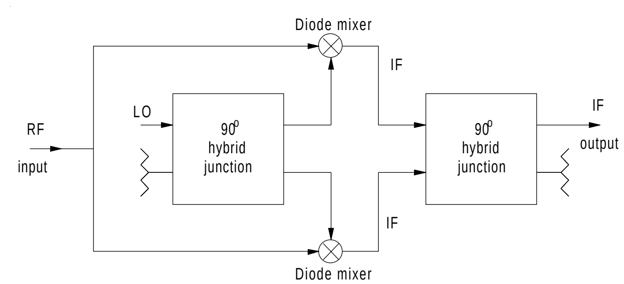


Figure M10 Image-recovery mixer (after Skolnik, 1980, Fig. 9.3, p. 349).

junction produces a 90° phase shift, and the next one adds another 90° phase differential. As a result, the IF signals from two mixers add in phase and the images cancel. This kind of mixers has high dynamic range, less susceptibility to burnout, and low intermodulation products. It is good as a receiver front-end. *SAL*

Ref.: Skolnik (1980), p. 349.

An **image-rejection mixer** is one having two conventional asymmetrical mixers fed with the signals from a local oscillator shifted by 90°, and received signals, and the bridge circuit to suppress the signal at image frequency. If both mixers have a balance configuration, the device is termed a *double-balanced mixer*. *IAM*

Skolnik (1970), p. 5.10.

A **Josephson-effect mixer** is one using a Josephson junction as the nonlinear component (see **JOSEPHSON effect**). The mixing component is a junction with a superconductor-insulator-superconductor structure implemented with thermal evaporation technology. The separate component or the grid of the components is typically mounted in a waveguide mixing structure analogous to those of Schottky diodes. Because of the extremely sharp volt-ampere characteristic, the power of the local oscillator can be less than 1 mW. This type of mixer is especially efficient at very high frequencies (e.g., at 452 GHz), and has a noise temperature of 350K and a 5-dB conversion loss.

IAM

Ref.: Reysanen, A., *Zarubezhnaya Radioelektronika*, no. 11, p. 67.

A **low-noise mixer** is one providing relatively low noise level at the frequency converter output. The main types of such mixers are transistor mixer, Schottky-barrier diode mixer, and Josephson effect mixer (Table M3). Tunnel diode and inverse diode mixers are also used as low-noise mixers. Typically, low-noise mixers employ balanced circuits (see **balanced mixer**).*IAM*

Ref.: Gassanov (1988), p. 134.

Table M3
Main Types of Low-Noise Mixers

Mixer	Operating frequency, GHz	Noise factor, dB
Transistor mixer	5 to 30	5 to 16
Schottky-barrier diode mixer	50 to 600	5 to 7 (100 GHz)
Josephson effect mixer	300 to 600	5 to 8

A **multidiode mixer** is an integrated-circuit-technology-based mixer employing double-diode balanced mixers and their combinations. The main advantage is that this mixer is based on the up-to-date integrated circuit technology. The high quality and identity of IC mixing diodes makes it possible to ensure decoupling of LO and signal sources without using bulky frequency-selective circuits, to suppress LO noise and sidebands, to reduce the conversion losses by returning to IF the power of the signal converted at one of the sidebands, and to ensure phase decoupling of both channels. Typically, multidiode mixers are based on monolithic ICs and are used in receiver frequency converters and active phased array transceivers in millimeter waveband. *IAM*

Ref.: Rozanov (1989), p. 74.

A **point-contact diode mixer** is one using the conventional point-contact diode as the nonlinear device providing frequency conversion. In comparison, the Schottky-barrier diode mixer it has better burnout properties, but higher noise figure than the point-contact mixer. *SAL*

Ref.: Skolnik (1980), p. 347.

quasioptical mixer (see **Josephson-effect mixer**).

A **Schottky-barrier-diode mixer** is one using a Schottky-barrier diode as an active component. Typically, these mixers are used in a millimeter wave and a submillimeter wave band. The best noise performance is achieved for the sharp volt-ampere curve. In a 3-mm band under ambient temperature, the noise temperature is about 400 to 700K, and losses of 4.5 to 7 dB. For cooling down to 20K, these parameters are 70 to 200K and 5 to 7 dB respectively. Cooled mixers require the power of the LO to be about 50 to 500 mW. A typical millimeter-wave mixer is a semiconductor IC, with more than 100 Schottky contacts which are inserted in the waveguide. (See **multidiode mixer**.) *IAM*

Ref.: Fink (1982), p. 14.61; Reysanen, A., *Zarubezhnaya Radioelektronika*, no. 19, 1984 (in Russian).

A **mixer spurious-effect chart** is a chart depicting the spurious components arising from the nonlinear nature of mixer operation. There are several forms of such data representation, one of them is shown in Fig. M11. Here the higher input frequency is designated by H and the lower one by L. The response caused by the first-order mixer product (H-L) originates mainly from the square-law term in the series describing the dependence of the current flowing in nonlinear resistance from the voltage across the resistor terminals (see **MIXER**), and the variation of normalized output frequency (H-L)/H with normalized input frequency L/H are shown with heavy lines. All other lines in the chart depict spurious effects originating from high-order terms in the series. This chart is convenient when one wants to see at a glance which combination of input frequencies and bandwidths are free of strong low-order spurious components. *SAL*

Ref.: Skolnik (1990), pp. 3.8–3.10.

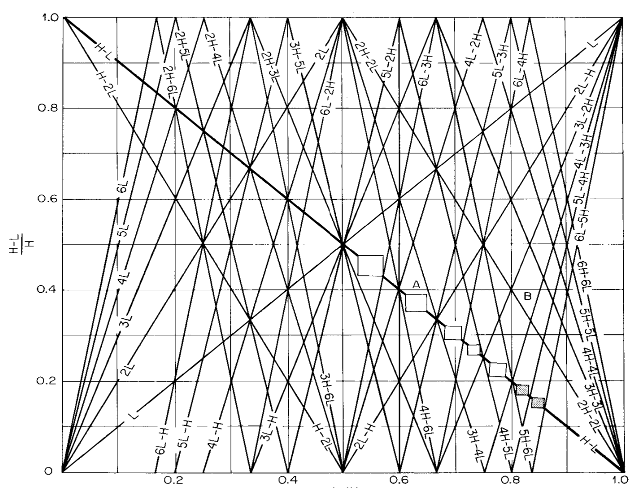


Figure M11 Downconverter spurious-effects chart. H = high input frequency, L = low input frequency (from Skolnik, 1990, Fig 3.2, p. 3.9, reprinted by permission of McGraw-Hill).

The **transistor mixer** uses as an active element a transistor, typically a bipolar transistor or a field-effect transistor. In a microwave band, Schottky-barrier field-effect transistors are typically used because they have a lower noise level and can operate more efficiently at frequencies higher than 10 GHz. Transistor balanced mixers and broadband mixers based on field-effect tetrodes are frequently used. The noise performance of field-effect transistor mixers is slightly worse than for diode mixers, but they perform the conversion with some gain (3 to 10 dB) reducing the IF amplifier noise figure requirements. The output power for transistor mixers is about an order more than for diode mixers, so the dynamic range is about 10 to 20 dB larger. *IAM*

Ref.: Fink (1982), p. 14.64; Gassanov (1988), p. 133.

The **tunnel diode mixer** is based on the tunnel diode as an active component, and has signal gain because of the negative resistance of the diode. Tunnel diode mixers are unstable while operating. Power handling capability and dynamic range are lower than for diode mixers, and noise performance is only slightly better. Consequently, they have no wide range of practical application. *IAM*

Ref.: Gassanov (1988), p. 112.

An **unbalanced mixer** incorporates a directional coupler, summing microwave signals and a single-band rectifier (multiplier) based on the mixing diode. The working point is located on the straight part of the volt-ampere curve, corresponding to the current of 0.5 to 1.0 mA produced by the rectified voltage of local oscillators. When the signal voltage is considerably less than the LO voltage, the mixer performs a linear transformation of the signal spectrum. Typically, unbalanced mixers with high sideband and those with low sideband are distinguished, depending on the location of the signal band relative to the LO frequency. Significant levels of sum ($\omega_s + \omega_{LO}$) and image ($2\omega_{LO} - \omega_{LO}$) frequencies are present at the mixer output besides the difference (intermediate) frequency. These oscillations are reflected from the diode and can be either absorbed by the matched load at mixer input, or reflected from the input by means of filters circuits. A mixer matched at the input has conversion losses of 8 to 10 dB and is broadband with linear frequency-phase response. Mixers with reflections have losses which are 1 to 2 dB less, but their bandwidth is lower.

Unbalanced mixers typically have waveguide, coaxial, or microstrip structure. The main disadvantages are the higher LO power compared to signal power and the conversion of LO noise to intermediate frequency. Consequently, they have a relatively high noise factor, 10 to 15 dB, (which can be reduced by cooling). *IAM*

Ref.: Gassanov (1988), p. 113.

MODEL

atmospheric model (see **ATMOSPHERE**).

clutter model (see **CLUTTER**).

detector model (see **DETECTOR**).

error model (see **ERROR model**).

The **radar model** describes a radar system operating against a specified class of targets in a specified environment. There are two basic approaches to modeling of radar operation:

(1) Modeling the target and interference environment, the propagation medium, and the major radar subsystems (antenna, transmitter, receiver, and signal processor), and then simulating radar operation with this model. Typically this approach is more flexible with respect to variation in the radar and environmental parameters, and it offers estimates of the entire set of radar characteristics in detection, tracking, and as applicable in the target discrimination mode. However, it requires the development of complicated and costly models, and is often based on Monte-Carlo simulation (statistical modeling) because analytical models (based on closed-form equations) cannot be applied. Algorithms for this approach are described in Leonov (1979).

(2) Development of a model for the particular radar characteristic of interest (e.g., probability of detection, measurement error, subclutter visibility, etc.) as a function of the factors affecting this characteristic (transmitter power, antenna gain, number of received pulses, type of signal processor, etc.). In this case the models can be less complicated, or combined (analytical-statistical) models can be developed (see **ERROR model**). This approach requires less development time, but is applicable only to estimation of the characteristic for which the model was developed, and the applicability of results is limited by the assumptions of the analytical model (e.g., operation in white, Gaussian noise, with a Rayleigh target, in Rayleigh clutter, etc.). *SAL*

Ref.: Leonov (1979).

(RCS) fluctuation model (see **RCS fluctuation**).

Swerling model (see **RCS fluctuation**).

MODULATION. Modulation is “the process by which some characteristic of the carrier is varied in accordance with a modulating wave” and the result of that process. In radar applications three basic types of modulation are used: (amplitude)-pulse modulation, frequency modulation, and phase modulation. *SAL*

Ref.: IEEE (1993), p. 816.

Amplitude modulation is “the process by which a continuous high-frequency wave (carrier) is caused to vary in amplitude by the action of another wave containing information.” The analytical formula for amplitude-modulated signal is

$$\begin{aligned} u(t) = {} & A_0 \sin(\omega_0 t + \phi_0) \\ & - \frac{M_0 A_0}{2} \cos[(\omega_0 + \Omega)t + \phi_0 + \psi] \\ & + \frac{M_0 A_0}{2} \cos[(\omega_0 - \Omega)t + \phi_0 - \psi] \end{aligned}$$

where A_0 is the amplitude of carrier oscillation, M_0 is the modulation coefficient, ω_0 and ϕ_0 are the frequency and initial phase of the carrier, and Ω and ψ are the frequency and the initial phase of the modulating signal.

The first term is the underlying unmodulated oscillation with carrier frequency ω_0; the second and third terms are the description of oscillations due to amplitude modulation process. The frequencies of these oscillations $\omega_0 + \Omega$ and $\omega_0 + \Omega$

are termed *upper* and *lower sidebands*. The spectrum of the amplitude modulated signal is shown in Fig. M12. The spectrum width is equal to 2Ω. *AIL*

Ref.: IEEE (1993), p. 34;Terman (1955), Ch. 15; Chistyakov (1986), p. 138.

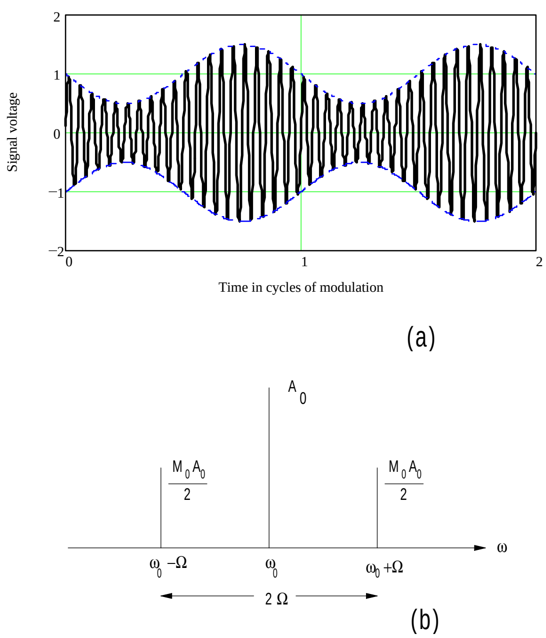


Figure M12 Amplitude modulation of the signal: (a) amplitude-modulated waveform (the dashed line indicates the envelope); (b) spectrum of resulting signal.

(Amplitude-)pulse modulation is amplitude modulation when the modulating signal is a sequence of pulses with parameters depending on the modulating signal. *AIL*

Ref.: Popov (1980), p. 28.

antenna-scanning modulation (see **MTI, limitation to performance**).

Bird-activity modulation is a term sometimes applied to denote waveforms obtained from birds. It should be mentioned that the spectral components of the bird-activity modulation pattern are rather stable and may be used for the purposes of identification. *SAL*

Ref.: Skolnik (1980), p. 510.

Coded modulation is modulation by information represented by codes. In radar applications, coded modulation is typically used to generate phase-coded waveforms. *AIL*

Ref.: Skolnik (1970), p. 16-27.

The **modulation coefficient** or *degree of modulation* is the amplitude modulation parameter equal to the ratio of maximum signal amplitude envelope variation ΔA to carrier amplitude A_0. The usual notation is M_0. *AIL*

Ref.: Terman (1955), p. 523; Popov (1980), p. 195.

Clipped noise modulation is noise modulation with a clipping action performed to broaden the bandwidth of the jamming signal. *DKB*

Ref.: Boyd (1961), p. 12.4.

Complex modulation is the factor $u(t)$ in the complex representation of a waveform:

$$u_c(t) = u(t)e^{j2\pi t f_0}$$

where

$$u(t) = a(t)e^{j\varphi(t)}$$

$a(t)$ is amplitude modulation, $\phi(t)$ is the phase modulation, and f_0 is the carrier frequency. The concept of complex modulation eliminates the necessity of manipulating the carrier frequency in the calculations associated with the complex representation of the waveform. *SAL*

Ref.: Brookner (1988), p. 125.

Cross-modulation is a type of intermodulation when the desired signal's carrier is modulated by an undesired signal wave. *SAL*

Ref.: Johnston (1979), p. 57.

digital modulation (see **coded modulation**).

Dual modulation is the combination of modulating modes employed in CW radar to improve the radar performance. Different variants of combination can be used: sawtooth plus sawtooth, triangle plus sine, triangle plus triangle, triangle plus noise, sine plus noise, and so forth. In general, one of the modulations enables to have a large deviation and another is chosen to have a perturbation type of behavior on the mixer spectrum. *SAL*

Ref.: Skolnik (1990), p. 14.30.

Frequency modulation changes the carrier frequency gradually in time in accordance with the modulating signal law. The frequency-modulated signal is shown in Fig. M13. If the

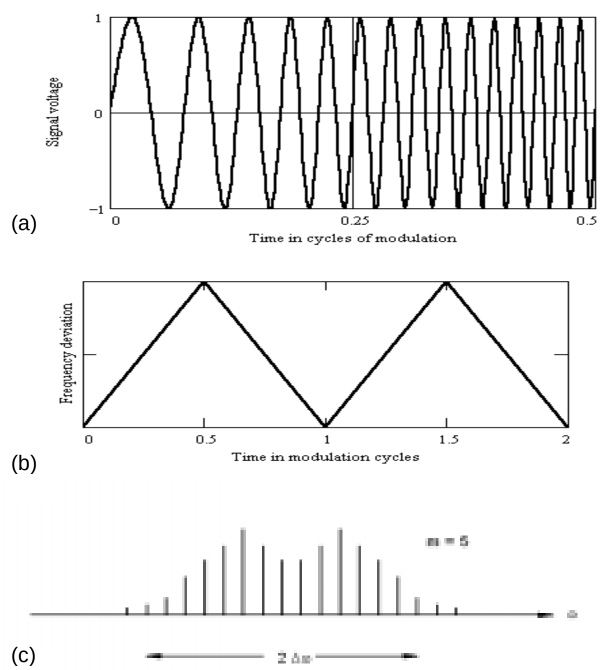


Figure M13 Frequency modulation of the signal: (a) FM waveform; (b) frequency vs. time; (c) FM spectrum.

initial phases are omitted, the signal can be written in the form

$$u(t) = A_0[(\sin\omega_0 t) + m_r(\sin\Omega t)]$$

where A_0 is the carrier amplitude, ω_0 is the carrier frequency; m is the index of modulation, and Ω is the modulation frequency.

The width of the frequency span, $\Delta\omega = 2\pi\Delta f$, is termed *frequency deviation*, is proportional to the amplitude of the modulating signal, and does not depend on Ω. Since it is impossible to implement continuous variation of the transmitter frequency, in practice the frequency can change periodically relative to a specified level, $\omega_0 = 2\pi f_0$. One of the simplest is the triangular frequency variation with a modulation period t_r (Fig. M13b). The spectrum of the FM signal consists of an infinite number of sidebands, which differ from carrier frequency by $n\Omega$, where n is an integer. The amplitude of the nth sideband is equal to $A_n = J_n(m)\cdot A_0$, where $J_n(m)$ is the Bessel function of the first kind of nth order with argument m (Fig. M13c). When $m >> 1$, the width of the spectrum is approximately equal to $\omega_e = 2\Delta\omega$. *AIL*

Ref.: Terman (1955), Ch. 17; Vinitskiy (1961), p. 184.

Frequency-modulation-by-noise is frequency modulation with random noise, as used in ECM to jam victim radars, especially those using AM and fixed-tuned FM receivers. *SAL*

Ref.: Johnston (1979), p. 60.

The **modulation index** is the frequency modulation parameter equal to the ratio of the frequency deviation Δf to modulation frequency f_m. The usual notation is m_r. *AIL*

Ref.: Terman (1955), p. 588; Popov (1980),

Intensity modulation is "a process used in certain displays whereby the luminance of the signal indication is the function of the received signal strength." *SAL*

Ref.: IEEE (1990), p. 19.

Intrapulse modulation is modulation of any pulse parameter (frequency, phase, or polarization) in accordance with a preset law. The most widely used types are frequency modulation and biphase modulation (shift keying) that widen the signal spectrum and provide for pulse compression. *SAL*

Phase modulation causes the phase and carrier frequency to vary in accordance with the modulating signal. If the carrier phase as the function of time is $\phi = \phi(t)$, then the instantaneous value of the frequency in the moment t is $\omega(t) = d\phi/dt$, and the complete variation of the phase during period t_r is

$$\phi = \int_0^{t_r} \omega(t)dt$$

There is a linkage between phase and frequency modulation, since time-dependent frequency variation $\omega(t)$ is equivalent to phase variation as the integral of $\omega(t)$, and time-dependent variation $\phi(t)$ is equivalent to frequency variation as the derivative of $\phi(t)$. *AIL*

Ref.: Terman (1955), p. 592; Popov (1980), p. 451.

Polarization modulation is intrapulse modulation whereby the waveform polarization changes in time in accordance with the preset law. Such a technique is more complex than frequency or phase modulation, but it is claimed that it can improve by a significant amount the radar jamming immunity. *SAL*

Ref.: Leonov (1988), p. 156.

(Modulation) sidebands are the components of the spectrum of a modulated signal, typically located symmetrically relative to the carrier frequency, and differing from the carrier frequency by the value equal to the modulation frequency, or a multiple thereof. For example, if the amplitude of the signal with the carrier frequency ω_0 is modulated by the sinusoidal oscillation with frequency Ω, ($\Omega << \omega_0$), then the output spectrum, besides the component of carrier frequency ω_0 will have sidebands components of $\omega_0 \pm \Omega$. *AIL*

Ref.: Terman (1955), p. 523; Popov (1980), p. 54.

MODULATOR. A modulator is the device that imposes modulation on the carrier. In radar applications it is used to modulate the transmitted waveform (see **pulse modulator**, **frequency modulator**) and in receiver circuits to transform the electrical signal into acoustical or optical form for signal processing. (See **acousto-optical modulator**, **space-time light modulator**.) *SAL*

An **acousto-optical modulator** is based on the effect of diffraction of light on acoustic waves. It consists of a light conductor made from lithium niobate or gallium phosphide in the microwave band, a piezo-converter of an electrical signal to an acoustic wave, an acoustic absorber, as well as an excitation circuit of the piezo-converter to ensure stability.

Acousto-optical modulators use various diffraction modes: Bragg diffraction (see **BRAGG effect**), Raman-Nath, and intermediate. Based on the type of acoustic and optical waves, we distinguish among modulators of three-dimensional acoustic, surface acoustic (for reflection and transmission) optical waves of infrared, visible-light, and ultraviolet bands.

In the microwave band, the use of acousto-optical modulators is limited to a frequency on the order of 3 GHz. The frequency band is 20 to 30%, and the value of the time aperture is equal to the ratio of the aperture of the modulator to the velocity of the acoustic wave, to units of microseconds. The basic advantage of acousto-optical modulators is the capability of inputting a radio signal into an optical system in real time, which establishes its basic use in acousto-optical devices for processing complex radar signals. The shortcomings of these modulators includes the short memory time, which limits the duration of the processed signals to around 100 μs, and the comparatively low number of channels (around 100). *IAM*

Ref.: Kulikov (1989), p. 5; Lukoshkin (1983), p. 261; Zmuda (1994), p. 215.

An **active-switch modulator** is one that uses active switches for controlling both the leading and trailing edges of a pulse.

They are subdivided into three basic types: cathode modulators, anode modulators, and grid modulators. The first control the full power of a beam of an oscillator tube either directly or through a connecting circuit. The anode pulse modulator provides a voltage equal to the full voltage of the tube beam, and the current is determined only by the charge of the capacitors of the circuit at the start and end of the pulse, since the modulating anode consumes a low current during the pulse. The output voltage of the pulse modulators for RF tubes with a controlling grid is much less, allowing the use of low-voltage components and circuits.

An anode pulse modulator makes it possible to produce an exceptionally flat pulse output. Grid modulators are marked by low parasitic capacitance effects and are especially suited for forming radar pulse trains.

A battery of capacitors with dimensions determined by the permissible droop of the pulse peak are used as the storage device of active modulators. Compared with linear modulators, modulators with active switches provide a better shape of pulses, due to the absence of special circuits containing elements with lumped parameters pulse shaping, which limit the rise time and create intrapulse ripple. *IAM*

Ref.: Skolnik (1970), pp. 7.78–7.87, (1990), p. 4.32.

A **complete-discharge modulator** is one based on pulse forming through complete discharge of energy from a storage device. Delay lines in a linear modulator, or a capacitor in a magnetic modulator, are used as the storage device.

Complete discharge in a delay line modulator occurs when the line matches the load resistance during the pulse, which is equal to twice the delay time of the line. In this case the shape of the pulse is rectangular. When there is a mismatch, the training edge of the pulse is distorted. This type of modulator is marked by a high stability of pulse length and, when hydrogen thyratrons are used as switches, high efficiency.

Complete discharge of a capacitive storage device in a magnetic modulator occurs through a small inductive resistance of the coil of a transformer with core in the saturation mode, and through the forming line.

The general drawbacks of modulators with complete storage discharge include the poor stability of timing of the pulse leading edge, the difficulty of forming interrogative code messages, and the impossibility of shaping pulses of different durations. *IAM*

Ref.: Glasoe (1948) Part II; Perevezentsev (1981), p. 136; Davydov (1984), p. 41.

A **frequency modulator** applies frequency modulation to a carrier. Frequency modulation is carried out either in the master oscillator of the transmitter, or in a low-power amplifier where it may take the form of phase modulation, converted into frequency modulation. The first method is the most common, and is realized by connection of a varicap to the resonator and the use of the nonlinearity of its barrier capacitance. The connection of the varicap with the resonator is usually capacitive. To reduce the level of harmonics of an even order, opposing connection of high-frequency varicaps is sometimes used.

A frequency modulator based on phase modulation is distinguished by its capability of using an independent high-stable local oscillator, but the frequency deviation of the signal at the output of the modulator does not exceed 1%. *IAM*

Ref.: Terman (1955), p. 601; Petrov (1989), p. 203; Shumilin (1981), p. 227.

An **ion-switch modulator** is a linear modulator using hydrogen thyratrons or ignitrons as switches. Modulators operate in a discharge mode using lines with capacitive reactance as the storage devices. Modulator circuits are classified based on the type of power source into modulators with dc or ac charging. The latter are used in those cases when the modulating pulse repetition frequency is synchronous with the frequency of the supply line, and the modulators themselves synchronize the radar. Modulators supplied from a dc source are more common.

In terms of storage circuits, we distinguish between modulators that use one (Fig. M14) or many delay lines. An increase in the number of delay lines makes it possible to reduce the supply voltage of the modulator by a factor of two or more in comparison with those using a single delay. *IAM*

Ref.: Skolnik (1990), p. 4.33; Rakov (1970). Vol 2, p. 86.

A **line-type [or linear] modulator** is a pulse modulator whose switching device uses a linear circuit for pulse formation (storage device) only for starting and maintaining discharge. The circuit determines the shape and duration of the pulses. A thyratron, ignitron, electron tube, thyristor, or a spark discharger may be used as the switch. A line-type modulator using partial discharge of the storage device is based on capacitors and has a power rating up to 1 kW, with an efficiency of 40 to 70%.

A line-type modulator with complete discharge of the storage devices based on one or two delay lines (Fig. M14).

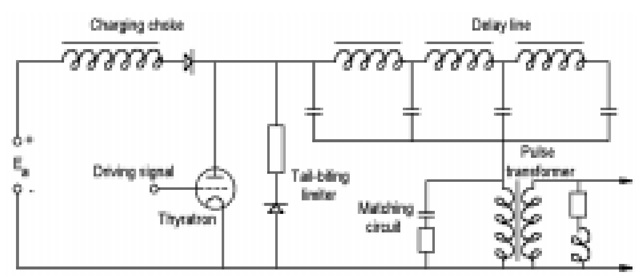


Figure M14 Line-type modulator (after Davydov, 1984, Fig. 2.6, p. 41).

At the moment that the thyratron is turned on, the voltage in the load decreases discontinuously and a wave is distributed along the line from left to right that is reflected from the end of the line, and moving in the opposite direction, continues to discharge the line capacitors. When the wave reaches the left end of the line, the process of formation of the pulse concludes, since the delay-line capacitors are completely discharged. As the wave is propagated along the line, the current in the line is equal to the load current, and the voltage at its input is close to the voltage of the supply source E_a. The duration of the formed pulse is equal to twice the delay time in the line. In this circuit a hydrogen thyratron is generally used, ensuring a high slope of the wave front of the pulse. *IAM*

Ref.: Davydov (1984), p. 40; Skolnik (1990), p. 4.33.

A **magnetic modulator** is a pulse modulator with a switch in the form of a nonlinear inductance coil. Nonlinear inductances used as coils are transformers with ferrite cores having a hysteresis curve of minimum area with the maximum possible saturation inductance. Operation of the modulator is based on the property of sharp increase in the inductance of the coils of such a transformer at magnetization current values corresponding to the top and bottom saturation regions.

Magnetic modulators are divided into two classes: ac modulators, whose pulse repetition frequency is equal to or a multiple of the frequency of the supply voltage, or dc modulators (thyristor-magnetic modulators) whose pulse repetition frequency is determined by the external triggering devices.

Modulators of the first type are most frequently used in radar equipment. The circuit of such a modulator consists of an input circuit, transforming stages, and terminal stage (Fig. M15).

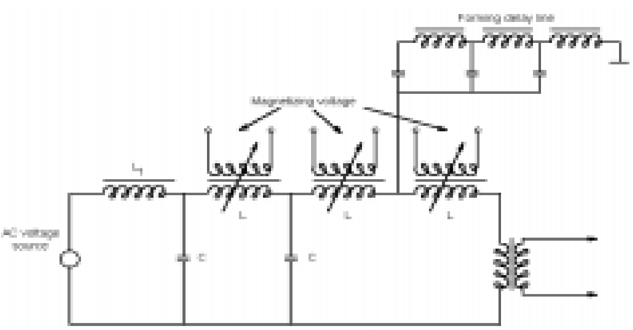


Figure M15 Magnetic modulator (after Shumilin, 1981, Fig. 15.22, p. 274).

The transformer stages are LC circuits, each of which compresses pulse of the charge current of the capacitor (up to 10 times). To obtain a typical pulse ratio on the order of 0.001, three to four stages are required. One output pulse corresponds to each period of supply-line voltage. The output stage is a single-section pulse-shaping circuit.

The advantages of magnetic modulators include immediate readiness for operation, unlimited service life, increased reliability, and low supply voltage. The drawbacks include the complexity of manufacture and tuning, the difficulty of changing the pulse repetition frequency, and insufficient stability of timing of the leading edge. *IAM*

Ref.: Skolnik (1970), pp. 7.76–7.78; Perevezentsev (1981), p. 147.

A **partial-discharge modulator** is a pulse modulator based on pulse formation through partial discharge of a storage capacitor. The discharge of the capacitor occurs during the starting pulse which opens the switch of the pulse modulator, through its small resistance. The recharging time constant is selected so that the capacitor will charge during the pulse repetition period to a voltage close to the constant voltage of the supply source. The discharge time constant is selected on the basis of requirements on the pulse shape and on the stability of the carrier frequency. Usually an acceptable change of voltage during the starting pulse is several percent. The shape of the trailing edge of the pulse is determined by the period of oscillations arising after closure of the commutator, in the circuit formed by the charge reactor the parasitic capacitances of the circuit. To compensate for undesirable oscillations, a damping circuit is used that contains a diode of opposite polarity of the source (see Fig. M14).

These modulators have good timing stability for both leading and trailing edges of the pulse. The basic drawbacks include their large dimensions and mass.

Modulators with partial discharge of the capacitive storage device are widely used in aircraft radars and secondary radars, because of their capability of generating pulses whose duration and PRF can be efficiently changed without switching to high-voltage circuits, and the capability of forming two- and three-pulse code messages. *IAM*

Ref.: Perevezentsev (1981), p. 131; Skolnik (1990), p. 4.35.

A **phase modulator** is a controlled phase-inverter. An oscillating circuit with nonlinear capacitance, controlled by a source of modulating oscillations, may serve as the simplest phase modulator. Often a varicap is used as the capacitor of the circuit, with the low-frequency modulating voltage applied to it. In such a device, it is possible to obtain a phase deviation within the limits of 30° with nonlinear distortions of 7 to 10%. To obtain a greater phase modulation, several weakly coupled circuits with varicaps are connected in series.

To produce phase-keyed signals, discrete phase-inverters are used that may be of the transmitting or reflective type. In the microwave range, the latter have become more common. Their operation is based on discrete switching of the length of the line between the input and output of the modulator. For a discrete change in the line length, switching elements are used; these are most often PIN diodes.

In modulators of the reflective type, devices are required for isolation of the incident wave and the reflective wave going to the load. Circulators or bridge power dividers or adders may be used as the isolating devices. A biphase modulator based on PIN diodes, used in millimeter-wave radars, has a switching time of less than 1 ns and insertion loss of less than 1 dB. *IAM*

Ref.: Terman (1955), p. 602; Petrov (1989), p. 208.

A **pulse modulator** is intended to control the output of an RF oscillator through connection during the pulse width to the power source. Starting pulses are used that are generated by a special circuit (submodulator) controlled by synchronization pulses.

The operating principle of the modulator is that in the interval between the synchronizing pulses, energy accumulates in a storage element (capacitor, line). After arrival of the starting pulse, the pulse modulator switch connects the oscillator (e.g., a magnetic oscillator) to the storage element, which is partially or fully discharged through it. Based on the type of storage element, we distinguish among modulators with partial discharge, modulators with complete discharge, linear modulators with pulse formation in the storage line circuit, and magnetic modulators that use a storage device with nonlinear elements. At the end of the pulse, the discharge circuit is opened and recharging of the storage element begins. The shape and width of the pulse are determined both by the circuit of the storage device and the linear modulator, and by the controlled switch in active-switch modulators.

Pulse modulator switches are electronic devices for starting and stopping discharge of an energy storage device. In linear modulators, the switches do not control the shape and duration of the pulse, but only initiate and maintain the discharge. Active modulators control timing of both the leading and trailing edges of the pulse. Hydrogen thyratrons, mercury ignitrons, thyristors as well as preignition dischargers in older types of modulators are used as switches for linear modulators. Special types of vacuum tubes and thyristors are used as active switches.

Hydrogen thyratrons have peak powers up to 100 MW, average powers of 200 kW, and operating lives of 500 to 5,000 hours. The tubes are kept supplied with hydrogen by a reservoir with an independent heater. Their power limitation is the heating of the anode. The ignition delay after delivery of the trigger pulse is in the order of tens of nanoseconds, and depends on various factors. For better timing stabilization the thyratron may contain a priming grid, which initiates a low-power discharge before the pulsing of the main control grid.

Ignitrons are marked by high peak and average powers but have limited use due to the complexity of the triggering and controlling devices, and the restrictions on the pulse repetition frequency to a value of 500 Hz due to the slow deionization of the mercury vapors.

Thyristors are similar in parameters to hydrogen thyratrons. They are characterized by their peak voltage (≈1,000V), speed of current buildup (up to 10^4 A/s). Their operating life with current operation is practically unlimited. To obtain high power, connection of several thyristors is widely used. The basic drawback of thyristors is their sensitivity to overloads, even instantaneous ones.

Vacuum-tubes of active switches are created for nominal voltages up to 260 kV and peak powers up to 30 MW. The drawbacks include high-voltage breakdowns and a high level of x-rays requiring protective measures for operating personnel.

A comparison of the main types of pulse modulators is given in Table M4. *IAM*

Ref.: Skolnik (1970), pp. 7.72–7.76; (1990), p. 4.32; Davidov (1984), p. 40.

A **space-time light modulator** performs space-time modulation of the phase (amplitude) of a coherent light wave in accordance with the change in parameters of the controlling signal. The operating principle is based on the change in the spatial distribution of the transmission factor in accordance with a time signal. Various physical effects are used for this, and determine the basic types of modulators.

Based on method of writing information, space-time light modulators are divided into three classes: electron-beam addressing, optical beam addressing, or electric signal addressing (see **acousto-optical modulator**). The first class includes modulators based on electro-optical crystals (lithium niobate, etc.), thermoplastic media, and dielectric oil film. Their action is based on deformation of the surface of the modulator through electrical interaction of the charges.

The second class includes modulators based on liquid crystals and photoconducting electro-optical media.

For description of the properties of modulators, the following characteristics are used; sensitivity, resolution, contrast, dynamic range, linearity, operating wavelength, presence of memory, speed (cycling type), and a number of others.

For processing of pulse-compression waveforms in real time, the acousto-optical modulator is the only one possible. Modulators of other classes are used in opto-electronic devices as references for processing of radar signals. *IAM*

Ref.: Voskresenskiy (1986), pp. 17, 24; Lukoshkin (1983), p. 259; Zmuda (1994), Ch. 6.

A **thyristor-magnetic modulator** is a synchronized magnetic modulator supplied by a dc voltage source with thyristors at the input circuit and first compression stage. The use of thyristors in the input circuit makes it possible to obtain output power stability with variations of input voltage, choice of pulse repetition frequency independent of the frequency of the supply line, and control of the repetition frequency with an external trigger. However the output pulse is delayed significantly from the trigger pulse, and this lag changes by several percent in the process of warm-up and with fluctuations in temperature. These changes are much less than in a purely magnetic modulator, but too great for use of these modulators for moving-target indicator radars.

Owing to the control of the pulse repetition frequency, and the capability of obtaining high (hundreds of microseconds) and low (units of microseconds) values of pulse duration, thyristor-magnetic modulators are more widely used than magnetic ones. *IAM*

Ref.: Skolnik (1970), p. 7.70; Rakov (1970), vol. 2, p. 104.

MONOPINCH is “a single axis monopulse used in search radars to provide effective beam narrowing by displacing the displayed azimuth by the target angle off axis.” It uses monopulse difference-channel information to narrow the angular width of the target echo on a display, giving the appearance of better angular resolution. A voltage taken from each target pulse in the difference channel is added to the display deflection voltage in such a way as to move the intensified blip toward the actual target angle, superimposing all target echoes onto a narrow angle cell. It has also been called an electrical correction system (ECS) and electronic beam sharpening (EBS). *DKB*

Ref.: IEEE (1990), p. 21; Rhodes (1959), p. 60.

Table M4
Comparison of Modulators

Modulator		Flexibility		Pulse-length capability		Pulse flatness	Crowbar required		Modulator voltage level
		Duty cycle	Mixed pulse lengths	Long	Short		Load arc	Switch arc	
Line type	Thyratron/ SCR	Limited by charging circuit	No	Large PFN	Good	Ripples	No		Medium/ low
Magnetic		Limited by reset and charging time	No	Large *C*s and PFN	Good	Ripples	No		Low
Hybrid SCR magnetic modulator		Limited by reset and charging time	No	Excellent; large capacitor bank	Good	Ripples	No		Low
Active switch	Series switch	No limit	Yes	Large coupling capacitor	Good	Good	Maybe	Yes	High
	Capacitor-coupled	Limited	Yes	Difficult; XF gets big; large capacitor bank	Good	Good	Maybe	Yes	High
	Trans-former-coupled	Limited	Yes	Excellent; large capacitor bank	Good	Fair	Maybe	Yes	Medium/ high
	Mod-anode	No limit	Yes	Excellent; large capacitor bank	OK, but efficiency low	Excellent	Yes	Yes	High
	Grid	No limit	Yes		Excellent	Excellent	Yes	-	Low

(from Skolnik, 1990, Table 4.3, p. 4.34, reprinted by permission of McGraw-Hill)

MONOPULSE is "a radar technique in which information concerning the angular location of a source or target is obtained by comparison of signals received in two or more simultaneous antenna beams, as distinguished from techniques such as lobe switching or conical scanning in which beams are generated sequentially. The simultaneity of the beams makes it possible to obtain a two-dimensional angle estimate from a single pulse (hence the term *monopulse*), although multiple pulses are usually employed to improve the accuracy of the estimate or to provide doppler resolution." The generic block diagram of a monopulse system is shown in Fig. M16.

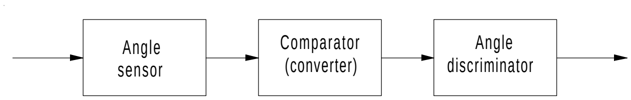


Figure M16 Block diagram of monopulse system (after Leonov, 1986, Fig. 1.2, p. 8).

It consists of an angle sensor that forms signals containing the target angle information, a comparator (or converter) which transforms the target angle information into some combinations of the amplitude and phase relation between the signals in two independent channels, and an angle discriminator having a single-valued relation to the angle of arrival of the signals. There are three basic types of techniques for angle sensors: amplitude comparison, phase-comparison, and combination (Fig. M17).

In the amplitude-comparison angle-sensing technique, two identical overlapping beams are formed with an offset of $\pm\theta_0$ from the equisignal direction or null axis (the axis along which the amplitudes of two patterns are equal). When the target is offset by an angle θ from the axis, the signal received through the lower beam has a greater amplitude then the signal received through the higher beam. The amplitude of this

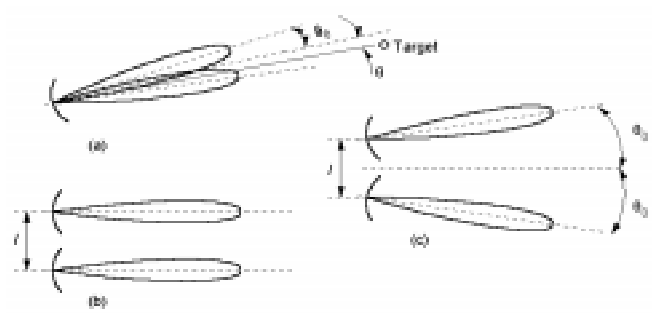


Figure M17 Elementary antenna patterns for monopulse radar: (a) amplitude comparison, (b) phase comparison, and (c) combination (after Leonov, 1986, Fig. 1.1, p. 3).

difference signal determines the magnitude of the angular offset from the null axis and the sign of the difference indicates the direction of the target.

In the phase-comparison system, the target angle is determined by the comparison of the signals received by a pair of antennas (Fig. M18). The difference between the paths from the antenna phase centers to the target, $\Delta R = l\sin\theta$, gives the phase difference $\Delta\phi = 2\pi R/\lambda = (2\pi l/\lambda)\sin\theta$, where λ is the wavelength, so the target angle is

$$\theta = \arcsin(2n\pi/k_\lambda)$$

where $n = 0, 1, \ldots$, and $k_\lambda = 2\pi/\lambda$ is the wave number. The presence of factor n causes measurement ambiguity that can be overcome by the proper choice of displacement between the phase centers of the antennas. In combination systems, both amplitudes and phases of the received signals are used to get the information about the target angle.

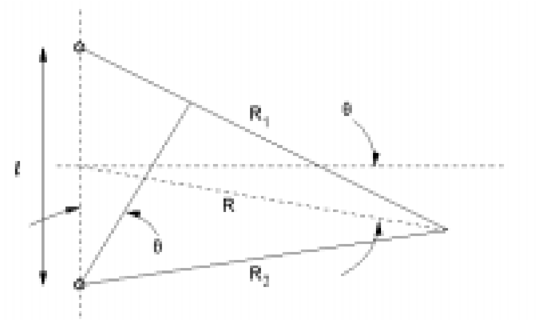


Figure M18 Geometry of phase-comparison monopulse (after Leonov, 1986, Fig. 1.1, p. 3).

There are only three possible methods of angle discriminators realization: amplitude, phase, and sum-and-difference. The block diagrams of these discriminators are shown in Figs. M19, M20, and M21. Since each of these may be applied in amplitude-comparison, phase-comparison, or combination systems, there are nine basic classes of monopulse systems as shown in Table M5.

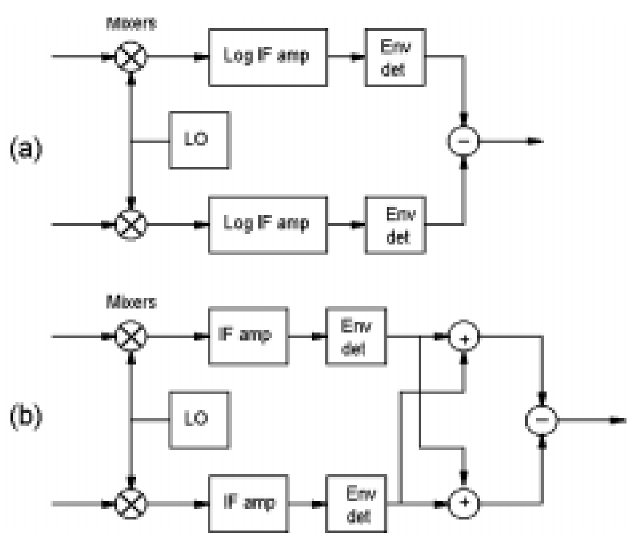


Figure M19 Amplitude angle discriminator: (a) with logarithmic IF amplifiers; (b) with normalization of the sum at video frequency (after Leonov, 1986, Fig. 1.3, p. 9).

Typically, in practice only four of them are used: amplitude-amplitude (AA), phase-phase (PP), amplitude sum-and-difference (ASD), and phase sum-and-difference (PSD).

The accuracy of monopulse angle-sensing is much better than for conical scanning systems and it is not sensitive to amplitude fluctuations (random variations in RCS). Although a monopulse system requires two independent channels for each coordinate and it is more complicated, it has become the basic angle sensing technique for precision tracking radars. *AIL*

Ref.: IEEE (1993), p. 821; Leonov (1986).

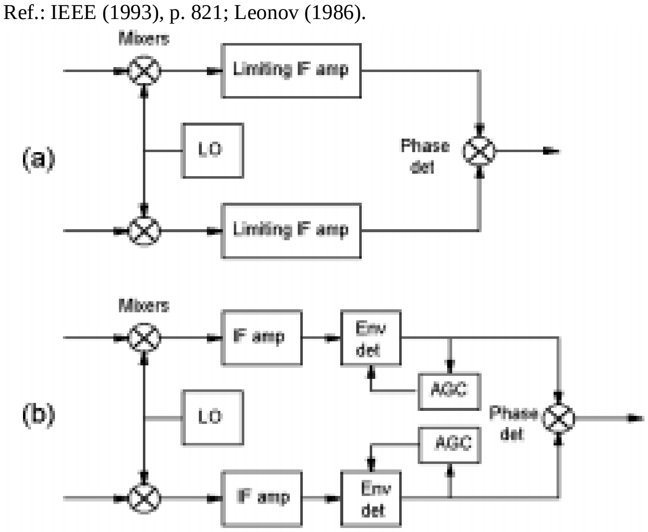


Figure M20 Phase angle discriminator: (a) with limiting IF amplifiers; (b) with AGC (after Leonov, 1986, Fig. 1.4, p. 10).

Table M5
Classes of Monopulse Systems

Method of measurement (type of angle discriminator)	Basic classes of monopulse sensing systems for three types of angle sensing		
	Amplitude (A)	Phase (P)	Combination (C)
Amplitude (A)	AA	PA	CA
Phase (P)	AP	PP	CP
Sum and difference (SD)	ASD	PSD	CSD
(from Leonov, 1986, Table 1.1, p. 7)			

Amplitude-amplitude monopulse is a method of determination of angular coordinates in monopulse systems using amplitude direction finding and an amplitude method of measurement of an angle (pulse-amplitude discriminator). Figure M22 shows the block diagram of an amplitude-amplitude monopulse system for target direction finding in one plane.

The difference in the amplitudes of received signals defines target deviation from the equisignal line. The sign of this difference characterizes direction and displacement of the equisignal direction relative to the target. When the equisignal direction coincides with the target, the amplitudes of the signals received by both patterns are equal, while their difference reverts to zero. The direction-finding characteristic of this system is

$$S(\theta) = 2\mu\theta$$

where μ is the slope of the antenna pattern.

A shortcoming of a system with a pulse-amplitude discriminator is the requirement to maintain exact amplitude match in the amplifier responses. *AIL*

Ref.:Sherman (1984), Ch. 5; Leonov (1984), p. 66.

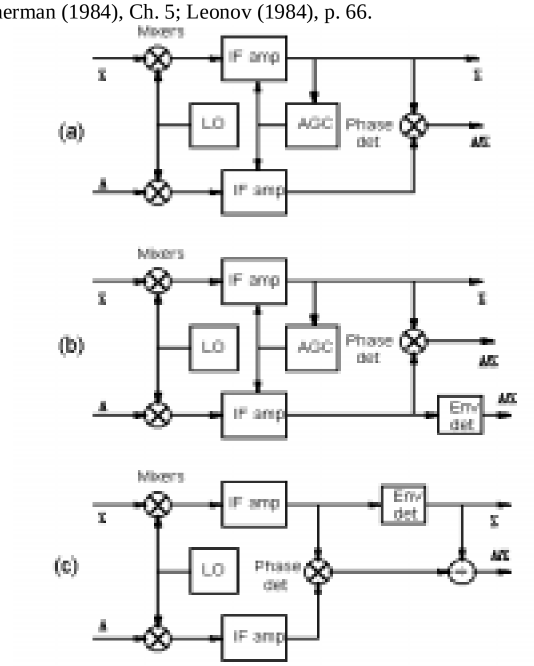


Figure M21 Sum-and-difference angle discriminators: with (a) combined, and (b) separate formation of the magnitude and sign of the error signal, and (c) with normalization by the sum signal at video frequency (after Leonov, 1986, Fig. 1.5, p. 11).

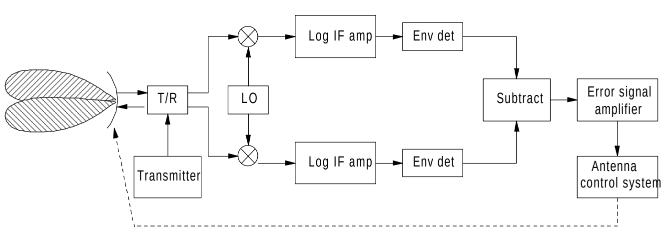


Figure M22 Radiation patterns and block diagram of an amplitude-amplitude monopulse system (after Leonov, 1986, Fig. 4.1, p. 76).

Combination monopulse is a method of determining angular coordinates in monopulse systems using both amplitude and phase direction finding with the sum-and-difference angle measurement method (sum-and-difference discriminator). Figure M23a is a block diagram of an integrated monopulse system for target measurement in two planes. Figure M23b shows the beam shaping principle. The antenna forms two beams in the vertical plane, inclined relative to one another by approximately one beamwidth, but parallel to each other in the horizontal plane, and separated by distance L, which provides target direction finding using the amplitude method in the vertical plane and the phase method in the horizontal plane.

Two phase detectors, one forming an elevation error signal and the other forming an azimuth error signal, are shown, to extract angular information relative to each coordinate at receiving channel input. A difference signal is supplied to the azimuth phase detector with a 90° phase shift relative to the overall signal, which is the reference signal. The direction-finding response is

$$S(\theta) = K_{pd}\tan\left(\frac{\pi L}{\lambda}\sin\theta\right)$$

where K_{pd} is the phase detector transmission factor θ is the deviation of the beam peak from the equisignal direction, and λ is the wavelength.

An advantage of monopulse systems with combined direction finding is that, for direction finding in two planes, it turned out to be possible to work with just two beams and two interconnected channels with one waveguide bridge at their input, something especially important for airborne radar equipment, where weight and size are of primary significance. *AIL*

Ref.: Barton (1974), pp. 269–275; Leonov (1986), p. 95.

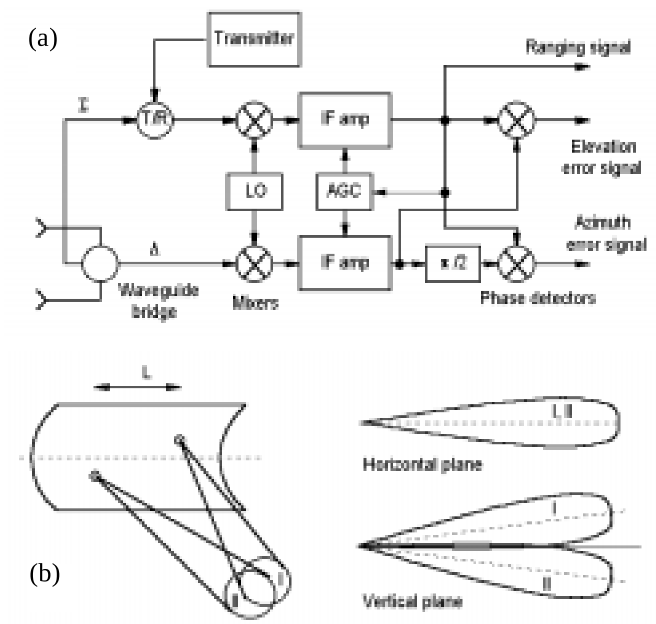


Figure M23 Combination monopulse system: (a) block diagram, (b) antenna patterns.

Phase-phase monopulse is a method of determination of angular coordinates in monopulse systems using phase direction finding and a phase method of angle measurement (phase detector). Figure M24 shows the elementary beams, signal comparison, and a block diagram of a phase-phase monopulse system for one plane. In the figure, θ is the deviation of the target from the equisignal direction, R_1 and R_2 are the distances from the target to the corresponding antennae, L is the distance between antennas, and ΔR is the difference in distances from target to the antennas.

The phase difference is

$$\Delta\phi = 2\pi\Delta R/\lambda = 2\pi L/\lambda \sin\theta,$$

where λ is the wavelength. This provides the capability to define angle θ relative to the measured magnitude of the phase shifts of signals reflected from a target.

The direction-finding characteristic is

$$S(\theta) = K_{pd}V_{lim}^2 \sin\left(\frac{2\pi L}{\lambda}\sin\theta\right)$$

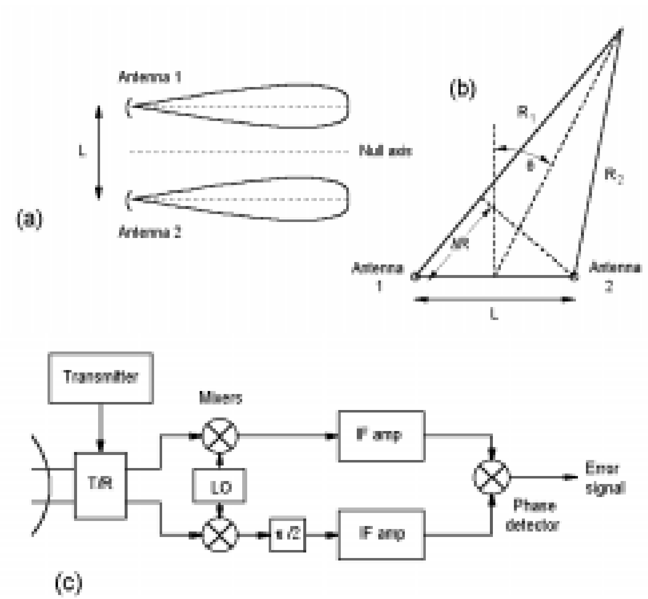


Figure M24 Phase-phase monopulse system: (a) elementary patterns, (b) signal paths, and (c) block diagram (after Leonov, 1986, Figs. 1.1 and 4.2, p. 68).

where K_{pd} is the phase detector transmission factor, and V_{lim} is the limiting threshold relative to amplitude of signals at phase detector input. A shortcoming of a system with a phase angular discriminator is the great dependence of direction finding accuracy on the degree of identity of the phase response of the receiving channels and their stability. *AIL*

Ref.:Sherman (1984), Ch. 5; Leonov (1984), p. 67.

Sum-and-difference monopulse is a method of determination of angular coordinates in monopulse systems using amplitude or phase direction-finding and a sum-and-difference angle measurement method (sum-and-difference discriminator). Most widely used are amplitude sum-and-difference monopulse systems, a block diagram of which for target direction finding in one plane is depicted in Fig. M25a. Figures M25b and M25c depict the elementary patterns, $F_1(\theta)$ and $F_2(\theta)$, the sum pattern $\Sigma(\theta)$, and difference pattern $\Delta(\theta)$, of such a system, with + and – signs designating the relative phase. It is evident from this figure that the phase of the difference signal changes depending on target direction relative to the equisignal direction and may be either at 0 or 180° relative to the phase of the sum signal. When the direction to the target coincides with the equisignal direction, signals in the receiving channel output have equal amplitude and the difference signal equals zero.

The difference signal is used to control the beam position during the tracking process. The sum signal is used both as a reference signal and for target acquisition and measurement of its range and velocity. The direction finding characteristic is

$$S(\theta) = K_{pd}\mu\theta\cos(\phi_1 - \phi_2)$$

where K_{pd} is the phase detector transmission factor, μ is the slope of the difference pattern over its operating region, and ϕ_1, ϕ_2 are the phase shifts in the channels. An advantage of such a direction finding system is the lax requirement for match of receiving channel responses, which explains its wide use in modern monopulse systems. *AIL*

Ref.: Sherman 91984), Ch. 5; Leonov (1986), p. 78.

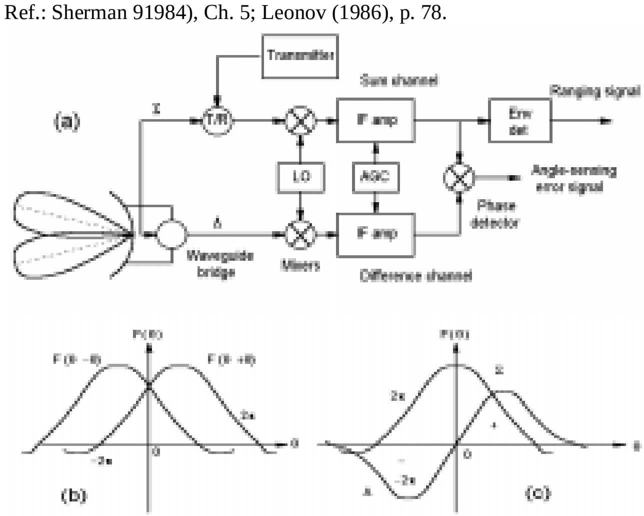


Figure M25 Sum-and-difference monopulse system: (a) block diagram, (b) elementary antenna patterns, (c) sum and difference patterns (after Leonov, 1986, Fig. 4.3, p. 70).

MOVING-TARGET DETECTOR (MTD). The MTD is an enhanced configuration of moving-target indicator (MTI) that combines a series of features to improve clutter rejection and target detection. The main features in an MTD are (a) the MTI precanceler, (b) the doppler filter bank, (c) use of burst-to-burst PRF diversity, (d) adaptive thresholding, and (e) the clutter map (Fig. M26).

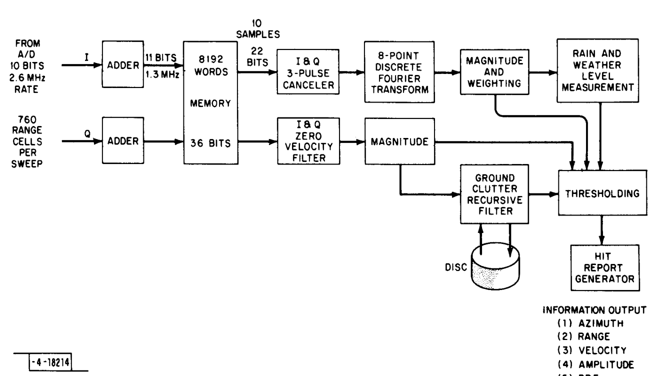


Figure M26 Block diagram of typical moving target detector (from Schleher, 1991, Fig. 1.29, p. 39).

The MTI precanceler eliminates most of the fixed clutter, reducing the dynamic range required in subsequent processing. The doppler filter bank is typically based on the FFT algorithm, providing the following advantages over a delay-line canceler: (1) the signal-to-noise ratio is improved by coherent integration within each of the n filters, whose bandwidth will be $1/n$ that of the canceler; (2) doppler frequency measurement is available, based on the filter number in which detection occurs; (3) the filter bandwidth can be adjusted by amplitude or frequency weighting (windowing), giving better range sidelobe reduction; (4) adaptive thresholding can be applied to each filter, permitting rejection of moving clutter (e.g., weather clutter).

The use of burst-to-burst PRF diversity is necessary to fill blind speeds that fall within the target velocity region, and it provides a means of rejecting multiple-time-around (range-ambiguous) target echoes. The adaptive thresholding applies separate thresholds to each filter: the nonzero velocity channels use a range-cell-averaging CFAR to adapt to moving clouds of precipitation, while the zero-velocity channel threshold is generated by the clutter map, which applies a separate threshold for each range cell. In the zero-velocity channel, which bypasses the MTI precanceler, targets at zero radial velocity and at the blind speeds can be detected if they exceed by a sufficient margin the clutter stored in the map.

The MTD is essentially a low-PRF pulsed doppler processor, and was originally developed in the mid-1970s at the MIT Lincoln Laboratory. The initial design used a three-pulse precanceler and an eight-pulse doppler filter bank. *DKB, SAL*

Ref.: Schleher (1991), pp. 37–52.

MOVING-TARGET INDICATION (MTI). MTI is the process of rejecting fixed or slowly moving clutter while passing echoes from targets moving at significant velocities. In most cases the MTI is sensitive only to radial components of velocity, but area MTI techniques can provide sensitivity to angular components as well.

In conventional MTI, discrimination between echoes from fixed and moving targets is based on the doppler effect. The filter technique can be realized either in the time or frequency domain. Implementation in the time domain is based on the fact that the phase of the fixed target echo does not change from pulse to pulse, while that of the moving target changes at a rate corresponding to the doppler frequency. This leads to the delay-line canceler implementation (Fig. M27), in which the phase-detected (bipolar) video signals (whose amplitudes depend on their phase angles) are fed into a delay equal to the pulse repetition interval and subtracted from the

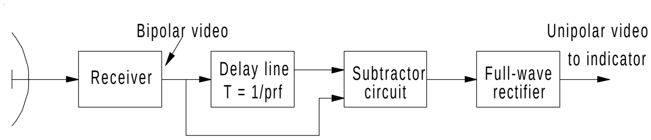


Figure M27 MTI receiver with delay-line canceler (after Skolnik, 1980, Fig. 4.4, p. 104).

signals of the next interval. The result is cancellation of fixed targets and passage of targets with pulse-to-pulse change in detected amplitudes.

The resultant video signal is

$$v(t) = k\sin(2\pi f_d t - \phi_0)$$

and that delayed from the previous transmission is

$$v(t-T) = k\sin[2\pi f_d(t-T) - \phi_0]$$

The resulting canceler output is

$$\Delta v = A_\Delta \cos\left[2\pi f_d\left(t - \frac{T}{2}\right) - \phi_0\right]$$

where k is the amplitude of the video signal, f_d is the doppler frequency, ϕ_0 is the initial phase shift, and T is the pulse repetition interval. The amplitude of the canceler output is

$$A_\Delta = 2k\sin(\pi f_d T)$$

and it is a periodic function of T and f_d, called the MTI (frequency) response (Fig. M28). This is the classical response of

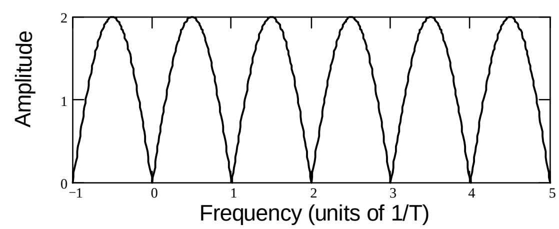


Figure M28 MTI frequency response.

a single-delay canceler. It shows that cancellation will occur not only for zero-velocity targets but for those with $f_{di} = i/T$. The speeds $v_i = \lambda f_{di}/2$ at which this undesired cancellation occurs are termed *blind speeds*.

To improve this simple response, more complex cancelers can be used: those of the three-pulse, or occasionally the four-pulse type; cancelers with feedback (infinite-impulse-response filters); or more complex filters operating on range-gated receiver outputs. The most common MTI systems pass video (baseband) pulses through the cancelers, using parallel channels for in-phase (I) and quadrature (Q) phase-detected components. Other approaches pass intermediate frequency (IF) signals through a vector canceler.

The voltage response of an m-pulse canceler to a target with velocity v is

$$H_m(v) = [2|\sin(\pi v / v_b)|]^m$$

shown in Fig. M29 for $m = 1$ and 2. Failure to center the rejection notch on clutter moving with velocity Δv leads to a clutter response $H_m(\Delta v)$, degrading the performance of the system.

The wide rejection notch formed for $m > 1$ causes serious loss in target detection (see **LOSS, velocity response**) unless staggered PRF or PRF diversity is used. A typical staggered PRF velocity response is shown in Fig. M30.

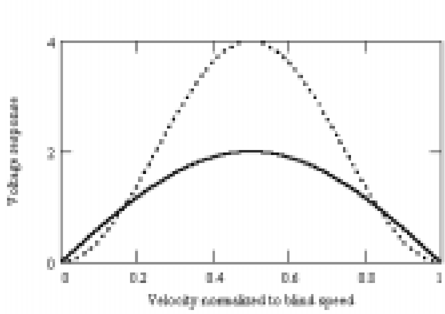


Figure M29 Velocity response of single- and double-delay MTI cancelers.

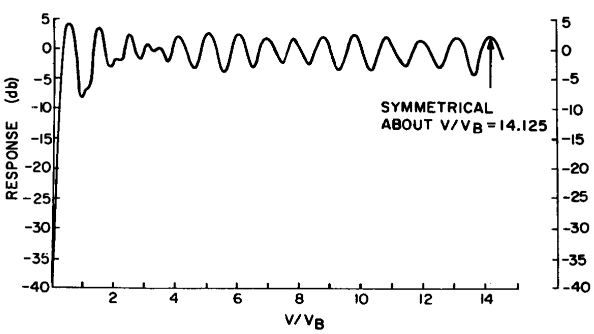


Figure M30 Velocity response of typical staggered PRF MTI canceler (from Skolnik, 1970, Fig. 42, p. 17.40, reprinted by permission of McGraw-Hill).

Moving-target processing in the frequency domain is based on doppler filter implementation to separate signals from targets with different velocities. This approach is more common in modern radars because it provides better performance, but it is more complex in comparison with delay-line cancelers where a single network operates at all frequencies and range delays and does not require separate filters for each velocity.

There are two basic approaches to implementation of MTI systems: (a) coherent MTI and (b) noncoherent MTI. The first provides better performance, but the second is simpler, using clutter to perform the same function as does the reference signal in coherent MTI. This is termed *clutter-referenced* or *externally coherent MTI*.

In general, the more pulses that are used to shape the MTI frequency response the better will be the performance. Using more than one PRF (staggered PRF or PRF diversity) provides additional flexibility in response shaping and reduces the effect of blind speeds. Use of coherent signal processing, digital filter banks, PRF staggering, clutter maps, and adaptive thresholds for rejection of undesired echoes results in the moving target detector configuration (see previous article on this technique).

The main factors used to describe MTI performance are MTI improvement factor, subclutter visibility, MTI gain, MTI response, clutter attenuation, clutter visibility factor, and cancellation ratio. (See also **MTI improvement factor.**) *DKB, SAL*

Ref.: IEEE (1993), p. 824; Skolnik (1970), Ch. 17, (1980), pp. 101–150; Bakulev (1986); Schleher (1991).

Adaptive MTI uses a variable rejection notch that changes its center velocity (and sometimes its width) in response to the velocity spectrum of the average clutter in resolution cells surrounding the target detection cell. Adaptation may be provided by controlling the frequency of a coherent oscillator (COHO) to minimize the total output of a fixed canceler, by varying the complex weights applied to the m pulses used in the canceler, or by selecting from the output of several parallel canceler circuits the one with minimum total output. *DKB*

Ref.: Skolnik (1990), pp. 15.61–15.65.

Airborne MTI (AMTI) refers to a system used to cancel clutter observed by a moving radar. Although the “A” originally stood for “airborne,” the term is now used for MTI radar on any moving platform. The techniques used to support AMTI include adaptive MTI, to remove the mean radar-to-clutter velocity and broaden the rejection notch; displaced phase center antennas, to reduce the clutter velocity spread cause by radar platform motion; and combinations of MTI cancellation with pulsed doppler filtering. *DKB*

Ref.: Skolnik (1980), p. 140.

Area MTI compares the envelope-detected outputs of successive scans to select targets that move in range or azimuth between scans, and in this sense differs from other MTI techniques in not using doppler frequency information directly. The requirement for detection is that the target echo must move by a significant fraction of the resolution cell width in either coordinate. Where the conventional MTI responds to radial motion of a fraction of a wavelength between pulses (e.g., 0.001m in 1 ms, or 1 m/s), the typical area MTI requires target motion of tens of meters between scans (e.g., 20 m in 5 sec, or 4 m/s). The clutter cancellation of area MTI is limited by clutter amplitude fluctuation between scans. The short-pulse area MTI has no blind speeds and is more attractive at higher carrier frequencies where conventional MTI suffers from excessive blind speeds, and where available bandwidth is greater.

The operation of the zero-velocity channel in the moving target detector is a form of area MTI, in which the clutter map stores the reference picture of fixed clutter and targets are detected if they lie above the mapped level. *DKB*

Ref.: Skolnik (1980), p. 147.

Batch-processed MTI processes a group of m pulses as a given PRF, obtaining a single output before changing the PRF (and sometimes the carrier frequency). The technique is used in place of pulse-to-pulse stagger when multiple-time-around clutter is present, as the m pulses may contain n_f fill pulses to permit cancellation of this clutter, at the expense of using only $m - n_f$ pulses in the MTI canceler. *DKB*

Clutter referenced MTI is “a type of noncoherent MTI that uses clutter as a reference.” Usually it is an adaptive MTI in which the average velocity of clutter surrounding the target cell is used to control the center velocity of the rejection notch. *DKB*

Ref.: IEEE (1990), p. 8.

Coherent MTI is "a form of MTI in which moving a target is detected as a result of pulse-to-pulse change in echo phase relative to the phase of a coherent reference oscillator." Typically it is a system in which a coherent oscillator (COHO) at intermediate frequency provides the reference for phase detection of the signals passed to the canceler or filter. The COHO may be continuously running, providing the offset of the transmitter frequency from the receiving local oscillator, or in coherent-on-receive MTI it may be locked in phase to a downconverted sample of each transmitter pulse. Coherent MTI is distinguished from area MTI and from noncoherent MTI, in which clutter at or near the target provides a phase reference. *DKB*

Ref.: IEEE (1993), p. 206; Skolnik (1990), p. 101.

Digital MTI uses an A/D converter to process phase-detected signals for coherent MTI (or envelope detected, for noncoherent MTI) before their passage to a digitally implemented filter circuit (usually a delay-line canceler or filter bank). *DKB*

Ref.: Skolnik (1980), p. 119.

externally coherent MTI (see **noncoherent MTI**).

MTI gain is the average response of the MTI to targets, defined as the ratio of signal power at the MTI output S_o to that at the input S_i averaged over all target radial velocities of interest:

$$G_{mti} = \overline{\left(\frac{S_o}{S_i}\right)}$$

Because the MTI improvement factor is defined as

$$I_m = \frac{\overline{(S_o/C_o)}}{\overline{(S_i/C_i)}}$$

it can also be written as

$$I_m = \frac{\overline{(S_o/S_i)}}{\overline{(C_i/C_o)}} = G_{mti}\mathrm{CA}$$

where CA is the clutter attenuation.

The effective MTI gain for targets distributed uniformly over the response is the same as the gain for white noise, and hence there is no gain (or loss) in signal-to-noise ratio in passage through an MTI canceler. There may be, however, a loss in detection capability caused by introduction of correlation between successive noise samples, loss of quadrature components of both signal and noise, and nonuniformity in target response. (See **LOSS, MTI processing.**) *SAL*

Ref.: Blake (1980), p. 63.

The **MTI improvement factor** I_m is a measure of MTI performance, defined as "the signal-to-clutter ratio of the output of the clutter filter divided by the signal-to-clutter ratio at the input of the clutter filter, averaged uniformly over all target radial velocities of interest." Equations giving I_m for single-, double-, and triple-delay cancelers as a function of the normalized clutter velocity spread, $z = 2\pi\sigma_v/v_b$, are

$$I_{m1} = \frac{1}{1-\exp(-z^2/2)} \approx \frac{2}{z^2}$$

$$I_{m2} = \frac{1}{1-(4/3)\exp(-z^2/2)+(1/3)\exp(-2z^2)} \approx \frac{2}{z^4}$$

$$I_{m3} \approx \frac{4}{3z^6}$$

where σ_v is the standard deviation of the clutter spectrum, v_b is the blind speed of the waveform, and the canceler notch is assumed centered on the clutter spectrum.

There is a dependence between improvement factor and other basic parameters describing MTI performance:

$$I_m = G_{mti}\ \mathrm{CA}$$

$$I_m = D_{xc}\ \mathrm{SCV}$$

$$I_m^{-1}{}_{\mathrm{actual}} = I_m^{-1}{}_{\mathrm{ideal}}\ \mathrm{CR}^{-1}$$

where G_{mti} is the MTI gain, CA is the clutter attenuation, SCV is the subclutter visibility, D_{xc} is the clutter detectability factor, and CR is the cancellation ratio determined by system instabilities. The ideal improvement factor is computed on the assumption that instabilities have no effect on the clutter returns: CR $\to \infty$.

The ratio of MTI improvement factor at a specific doppler frequency to that averaged over all frequencies is often termed the *MTI velocity response*. *DKB, SAL*

Ref.: IEEE (1993), p. 825; Schleher (1991), p. 108; Skolnik (1980), p. 129.

Limitations to MTI performance are defined in terms of the maximum attainable improvement factor, I_m in the presence of a specific source of limitation. There are two basic types of limitation:

(1) Those determined by the characteristics of input clutter.

(2) Those determined by the parameters of the radar channel itself.

The limit set by the first type is expressed in terms of the spectral spread of the clutter, $\sigma_{fc} = 2\sigma_v/\lambda$, where σ_v is the clutter velocity spread defined by the clutter model. (See **CLUTTER spectrum.**) The limitations due to radar parameters can be divided further into (1) antenna scanning modulation (or scanning fluctuation, the causing a spectrum of nonzero width at the radar input because of the finite time on target as the antenna beam scans over it); (2) use of PRF stagger; and (3) radar equipment instabilities.

The effect of antenna scanning modulation for an *m*-delay canceler is to limit the improvement factor to

$$I_{ma} = \frac{2^m}{m!}\left(\frac{f_r}{2\pi\sigma_f}\right)^{2m} \qquad (1)$$

where f_r is the PRF, $\sigma_f = \sqrt{\sigma_{fc}^2 + \sigma_{fa}^2}$, and

$$\sigma_{fa} = \frac{\sqrt{\ln 2}}{\pi} \cdot \frac{f_r}{n}$$

is the scanning modulation spectral spread of clutter, and n is the number of hits per scan. Equation (1) shows the maximum attainable improvement factor against clutter with a Gaussian spectrum centered at zero frequency.

Staggering of the PRF results in a limit to improvement factor due to unequal time spacing between the clutter samples:

$$I_{ms} = \left(\frac{2.5n}{\gamma - 1}\right)^2$$

where γ is the ratio of maximum to minimum PRF.

The sources of equipment instability are given in Table M6, with the corresponding limits to improvement factor in decibels. To find the attainable improvement factor in the presence of N different limiting factors, the improvement factors are converted to power ratios and combined according to

Table M6

Instability Limitations in MTI

Pulse-to-pulse instability	Limit on improvement factor
Transmitter frequency	$I = 20 \log [1/(\pi\ \Delta f\ \tau)]$
Stalo or coho frequency	$I = 20 \log [1/(2\pi\ \Delta f\ T)]$
Transmitter phase shift	$I = 20 \log (1/\Delta\phi)$
Coho locking	$I = 20 \log (1/\Delta\phi)$
Pulse timing	$I = 20 \log [\tau/(\sqrt{2}\Delta t\sqrt{B\tau})]$
Pulse width	$I = 20 \log[\tau/(\Delta PW\sqrt{B\tau})]$
Pulse amplitude	$I = 20 \log (A/\Delta A)$
A/D jitter	$I = 20 \log [\tau/(J\ \sqrt{B\tau})]$
A/D jitter with pulse compression following A/D	$I = 20 \log [\tau/(JB\tau)]$

where
- Δf = interpulse frequency change
- τ = transmitted pulse length
- T = transmission time to and from target
- $\Delta\phi$ = interpulse phase change
- Δt = time jitter
- J = A/D sampling time jitter
- $B\tau$ = time-bandwidth product of pulse compression system ($B\tau$ = unity for uncoded pulses)
- ΔPW = pulse-width jitter
- A = pulse amplitude, V
- ΔA = interpulse amplitude change

(from Skolnik, 1990, Table 15.4, p. 15.42, reprinted by permission of McGraw-Hill)

$$I_m^{-1} = \sum_{i=1}^{N} \frac{1}{I_{mi}}$$

where the individual factors include those shown in Table M6 plus I_{ma} and I_{ms} from inherent clutter spread, antenna scanning, and PRF stagger. *SAL*

Ref.: Skolnik (1980), p. 129, (1990), p. 15.41.

Noncoherent MTI uses clutter echoes from the target region as the phase (velocity) reference for detection before the canceler or filter. One common form of noncoherent MTI uses an envelope detector at the receiver output, the output amplitude of which varies at the target doppler shift as the signal phasor rotates around end of the clutter phasor (Fig. M31). This amplitude change component passes the canceler.

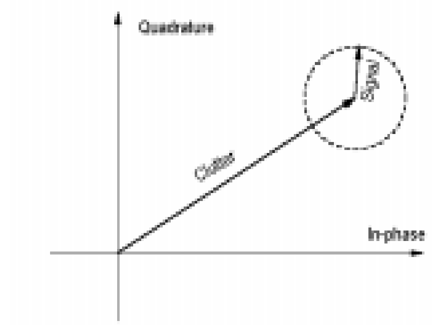


Figure M31 Phasor diagram of noncoherent MTI detector.

The improvement factors for single- and double-delay noncoherent MTI using a square-law detector are

$$I_{m1} = \frac{2}{1 - \exp(-z^2)} \approx \frac{2}{z^2}$$

$$I_{m2} = \frac{2(1)}{1 - (4/3)\exp(-2z^2) + (1/3)\exp(-4z^2)} \approx \frac{1}{z^4}$$

In the conventional noncoherent MTI system, targets in range cells containing no clutter are lost for want of a phase reference. Various clutter gating procedures are used to enable a normal video channel in such cells, bypassing the canceler to avoid loss of targets. A alternative noncoherent MTI detector uses the hard-limited output of adjacent cells as the reference to a phase detector, such that moving targets are detected in the absence of clutter through the random phase of noise in the reference cells (Fig. M32).

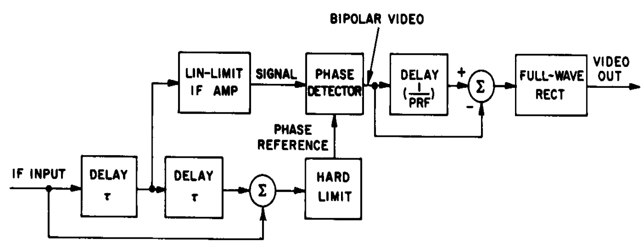


Figure M32 Phase-detection circuit for noncoherent MTI (from Skolnik, 1970, Fig. 59, p. 17.54, reprinted by permission of McGraw-Hill).

Advantages of noncoherent MTI are simplicity and its inherently adaptation to moving clutter. Disadvantages are reduced improvement factor and inability of most types to operate in the absence of clutter. *DKB*

Ref.: Skolnik (1980), p. 138.

The **optimum MTI** uses an Urkowitz (inverse) filter response to whiten the output interference spectrum. *DKB*

Ref.: Barton (1988), p. 236.

PRF-diversity MTI produces an effect similar to stagger-PRF MTI, except that sequential bursts of pulses at each PRF are transmitted, received, and batch-processed. Addition of p fill pulses to each batch supports cancellation of clutter from ambiguous range intervals, $R_u < R_c < (p + 1)R_u$, where

$R_u = c/2f_r$ is the unambiguous range. The disadvantage of batch processing is that only a single output is obtained from each batch, decreasing the number of integrated pulses by the factor $m + p$ for an m-pulse canceler with p fill pulses. *DKB*

Ref.: Barton (1988), p. 239.

Staggered-PRF MTI uses a train of pulses with pulse-to-pulse stagger (change of PRI) to reduce target loss in blind regions of the velocity response. The clutter cancellation notch remains at zero velocity and is usually repeated at some multiple of the average blind speed (see Fig. M30). Cancelation of multiple-time-around clutter is impossible when using pulse-to-pulse stagger. *DKB*

Two-frequency MTI uses two carrier frequencies that are beat together to generate a difference frequency as the IF. When this is processed in an MTI canceler, the resulting blind speed is that of a carrier equal to the difference frequency. The spectral spread of the clutter is increased by the beating of the two signals, resulting in decreased improvement factor. It has been found that the two-frequency MTI gives no net advantage over a properly designed single-frequency system, in most applications, and it has not found wide use. *SAL*

Ref.: Skolnik (1980), p. 147.

MULTIPATH (see **ERROR, radar**; **PROPAGATION**).

MULTIPLEXING is the combining of two or more signals into a single wave (the multiplex wave) from which the signals can be individually recovered.

MULTIPORT, microwave. A microwave multiport is any combination of conductors, dielectrics, and other microwave components that has several inputs in the form of transverse sections of transmission lines with type types of waves in each line.

A certain fictive pair of poles in the supply transmission line is assigned to each input of a microwave multiport. For this reason the term "2N-pole" or "2N-port" means N supply transmission lines with N types of waves.

Multiports are subdivided into passive (without amplification or microwave power generation inside the multiport) and active (with the opposite properties). The parameters of a multiport are written in the form of multiport matrices, which mathematically describe the properties of reciprocal, nondissipative, symmetrical, and other types of multiports.

In a nondissipative multiport there are no internal losses of electromagnetic energy (losses are vanishingly small against the background of the total power applied to the inputs). A condition of nondissipation is this equation for the resistance matrix Z:

$$Z + Z^{T*} = 0$$

where T is the transposition sign, and * is the complex conjugation sign. There is an analogous condition for the conductance matrix Y.

A nondissipative and reciprocal multiport often has purely imaginary matrices Z and Y and for this reason are called reactive. A reactive 2N multiport is characterized by $N(N + 1)/2$ real parameters.

Examples of nondissipative multiports include ideal directional couplers and ideal circulators.

A reciprocal multiport is a microwave multiport that has a symmetrical, nonnormalized conductance matrix. For reciprocal multiports, the theorem, familiar from electrodynamics, regarding the reciprocity of any two inputs applies (see also **RECIPROCITY**). Normalized matrices of conductivity, resistance, and scattering are also symmetrical.

A reciprocal 2N-pole is described by $N(N + 1)$ complex parameters.

The property of reciprocity is ensured by the absence of anisotropic electromagnetic media inside the multiport (e.g. magnetized ferrites or plasma).

A symmetrical multiport is one for which renumeration of inputs is possible, without leading to a change in the multiport matrices. We distinguish between geometrical and electrical symmetry. The latter is achieved through special selection of the nominal characteristics of the elements of the multiport. Geometrical symmetry is manifested in the fact that the multiport remains like itself during symmetrical conversions (rotations of the multiport about its axis, mirror images relative to planes of symmetry). Geometric symmetry is easily established *a priori* and entails electrical symmetry. Mathematically, symmetry of a multiple is described by the introduction of a special matrix of symmetry G, which is commutative with the matrices of the multiple parameters Z:

$$GZ \equiv ZG$$

These equations for the resistance matrix, conductance matrix, and scattering matrix are used to reduce the number of independent parameters of a multiport. A double waveguide T-bridge is an example of a symmetrical multiport. *IAM*

Ref.: Sazonov (1988), pp. 86–92.

A **MULTIVIBRATOR** is a two-stage oscillator with positive feedback between the stages, which generates video pulses with the shape close to a rectangular one. They can operate in two modes: a self-oscillations mode and a waiting mode. The first class device is a two-stage amplifier with resistors, the output of each amplifier being connected with input of another one via a separating capacitor (Fig. M33). The process of generation consists of opening one transistor and closing another one in succession. As a result, the pulses of different polarity and duration appear at transistors output.

Pulse duration τ_p, pulse repetition period T_p, and pulse amplitude A_p are defined as

$$\tau_{p1} = R_{b1}C_2 \ln\left[\frac{2 + (I_{c0}R_{b1}/E_c)}{1 + (I_{c0}R_{b1}/E_c)}\right]$$

$$\tau_{p2} = R_{b2}C_1 \ln\left[\frac{2 + (I_{c0}R_{b2}/E_c)}{1 + (I_{c0}R_{b2}/E_c)}\right]$$

$$T_p = \tau_{p1} + \tau_{p2}; \qquad A_{p1} = A_{p2} = E_c$$

where I_{c0} = initial current at collector.

A waiting multivibrator possesses a single stable state and generates a single pulse when external voltage is applied.

There is big variety of waiting multivibrators block diagrams. The circuit with an emitter coupling and voltage versus time diagrams for it are shown in Fig. 1. *AIL*

Ref.: Terman (1945), p. 625; Druzhinin (1967), pp. 232–236.

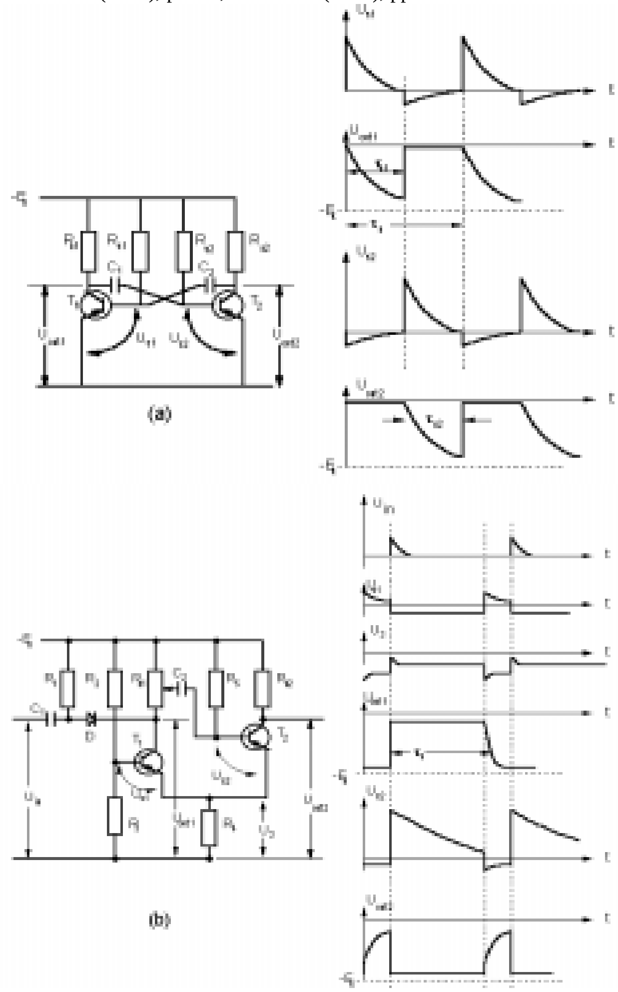


Figure M33 Multivibrators: (a) in self-oscillating mode, (b) in triggered mode.

N

NAVIGATION, radio. Radio navigation is the branch of science and technique covering radio engineering methods and aids for navigation of ships, aircraft, spacecraft, and other vehicles. Typically, radio navigation systems are classified according to the following features:

(1) The mode of object location: positional (angle finding, angle-range finding) or nonpositional systems.

(2) The type of the measured parameter: amplitude, time, frequency, or phase measurement systems.

(3) The band used: decameter to optics band systems.

(4) The range of operation: space, global, short- or long-range navigation systems.

(5) The place of reference station location: space-based or ground-based systems.

Positional methods of radar navigation are typically classified as angle-finding, range-finding, angle-range-finding, or difference-range-finding methods.

Ref.: Smith (1978); Yarlykov (1985); Sonnenberg (1978).

Angle-finding navigation is based on measurement of at least two angular directions. There are two basic configurations of angle-finding systems: in the first, the receiver antenna is directional and the transmitter antenna is omnidirectional, and vice versa for the second. When the transmitter and the receiver are located in the same plane (e.g., at the surface of the Earth), the direction to the beacon (transmitter) is its bearing (or azimuth). If it is referenced to the geographical latitude, it is termed the true bearing or the true azimuth. It the two bearings α_1 and α_2 (Fig. N1) are determined, the location of the receiver can be determined as the point where two position lines intersect. If the system is located above the ground surface, the third beacon is required to determine the location of the receiver. The examples of angle-finding systems are VOR, DVOR, and PDVOR.

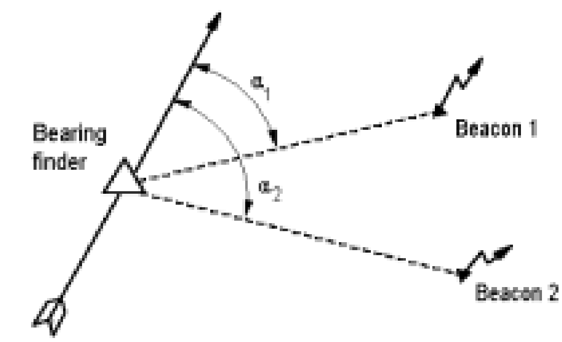


Figure N1 Angle-finding navigation method.

Range-finding navigation systems are based on the measurement of range in a secondary radar system (Fig. N2). If the time of interrogation and responding signals propagation is the same $\tau_i = \tau_r$, and the transponder delay can be neglected, the range measured on the transponder is

$$R = c(\tau_i + \tau_r)/2,$$

where c is the velocity of light. The the position surface is the surface of a sphere with a radius equal to R, and lines of position are circumferences. The position of the object is the

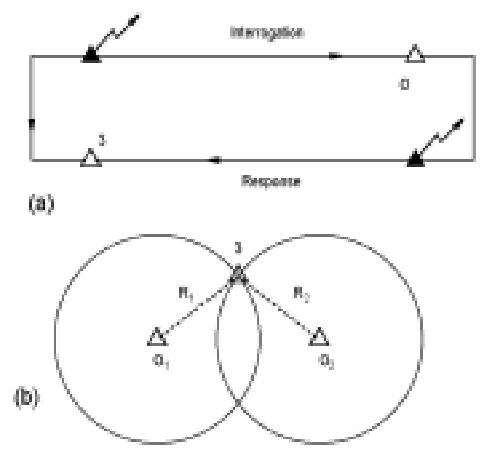


Figure N2 Range-finding navigation method.

intersection of two lines of position. To eliminate the ambiguity, additional data are used with the crude accuracy sufficient to determine one of two intersection points as the true location. As the achievable accuracy of range measurement can be very high, the coordinates of the object can be determined with high accuracy. The DME system is an example of range-finding navigation system.

Range-difference navigation systems measure the difference in time between the reception of signals from two reference stations: A and B_1 (Fig. N3). The distance between stations is called a *baseline.* For two stations, a set of hyperbolas with focuses in points A and B_1 can be determined. The two stations can determine only the position line. To determine the exact point of object location on a position line another couple of stations is required; the baseline d_2 of these stations must be positioned at an angle α to the baseline d_1 of the first couple.

The accuracy of range-difference systems is higher than the accuracy of angle-finding system and approaches the accuracy of range-finding system. But its throughput capability is much higher, because it is not necessary to have an onboard transmitter as for the range-finding system. The examples of difference-range finding systems are “Loran-A,” “Loran-C,” and “Omega.” A description of the major positional navigation systems is given in Table N1.

Nonposition methods of navigation are based on an autonomous approach to determine the location of the object. They can be based on the measurement of the object velocity vector and integration to obtain the distance traveled, or on measurement of any physical field and comparison of the measurement data with the reference data. The advantage of the systems employing nonposition methods is autonomous operation, while the disadvantage is complexity and necessity to process all data with an onboard computer. *AIL*

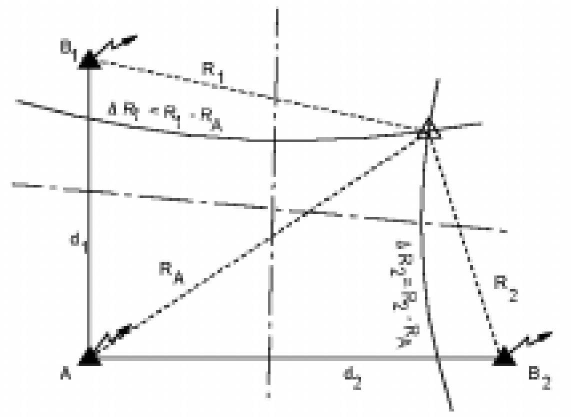


Figure N3 Range-difference navigation method.

Ref.: Skolnik (1970), p. 16.33; Kazarinov (1990), pp. 182–194; Belavin (1967), pp. 5–62; Sosulin (1992), pp. 11–15

NAVIGATOR, doppler. A doppler navigator is a CW radar used for navigation, relying on the nonpositional method. (See **NAVIGATION, radio.**) *SAL*

Ref.: Skolnik (1990), p. 14.37.

Table N1
Major Positional Navigation Systems

System	Description
VOR	VHF Omnidirectional Range. A ground transmitter radiates CW signals on one of the even frequencies between 108 and 118 MHz. A cardioid antenna pattern rotates at 30 Hz, producing an AM signal at 30 Hz in the airborne receiver. A 30-Hz reference frequency is radiated as FM, ±480 Hz on a 9,960-Hz subcarrier. The bearing is measured as a function of the phase between the FM and the AM modulations as read by the airborne receiver.
DVOR	Doppler VOR. A wide-aperture VOR to reduce site error. It uses a 15m diameter circular array of antennas to generate the variable phase, and the reference phase is transmitted by a single, central antenna. The format is compatible with standard VOR receivers.
TACAN	TACtical Air Navigation. Constant duty cycle DME beacon to which a rotating cardiod antenna pattern plus a 9-lobed pattern is added. It generates a 15-Hz coarse bearing and 135-Hz fine bearing in the aircraft receiver.
DME	Distance-Measuring Equipment. The airborne interrogator transmits about 30 pulse pairs per second on one of 126 channels between 1,025 and 1,150 MHz. The ground transponder replies on one of 126 channels between 962 and 1,027 or 1,151 and 1,213 MHz. The airborne indicator reads transmit-to-receive time delay on a meter calibrated in nautical miles.
LORAN A, C	LOng RAnge Navigation. Pulse of 45 ms duration are transmitted by a master station on a carrier of 20 to 34 MHz, and repeated by a slave station some 450 km away. An aircraft, using an oscilloscope display, measures the difference in time of arrival and computes its position by using two or more slave stations. In comparison with LORAN-A, the LORAN-C system uses the same basic principle, but with a carrier at 100 kHz where ground-wave propagation is possible to long ranges.
Omega	A range-difference system using high-power transmitters with time-shared CW signals at about 10 kHz. The baselines and operational ranges are about 9,000 km.

NETWORK

all-pass time-delay network (see **DELAY LINE**).

array (antenna) feed network (see **FEED**).

beam-forming network (see **FEED, beam-forming network**).

digital beam-forming network (see **FEED, beam-forming network**).

An **electrical delay network** is one with dispersive characteristics used in analog pulse compression technology. Typically, it is used as an electrical-network class of linear FM waveform generators and has linear delay-versus-frequency characteristics. The most common representatives of delay network are all-pass network (a four terminal lattice network with constant gain at all frequencies and a phase shift varying with the square of frequency to ensure the constant delay slope; this the low-frequency device using lumped elements), a folded-type meander line, which is the microwave analog of all-pass network, and waveguides operating near its cutoff frequency. *SAL*

Ref.: Skolnik (1990), p. 10.13.

NOISE is unwanted electromagnetic energy with random parameters that interferes with the ability of radar to perform in an ideal, error-free mode. Typically, the noise is represented as a random process, and its presence at radar input has the fundamental importance for estimation of radar performance in any mode of operation: detection, measurement, and discrimination. Classification of noise can be based on different factors:

(1) From the standpoint of its origin, noise is classified as target noise (caused by radar target, its shape and motion peculiarities), external (environmental) noise (caused by the environment in which radar signal propagates) or internal noise (caused by radar hardware, primarily by the antenna and receiver).

(2) From the standpoint of its spectral characteristics, as white or colored.

(3) From the standpoint of its distribution as Gaussian or non-Gaussian (often Rayleigh).

The main characteristics of noise are noise power and power density. When noise is represented as a random process it may be described with the parameters typical of this kind of random function representation: mathematical expectation, variance and correlation function. (See **FUNCTION, random.**) Pure noise typically has zero mathematical expectation and a delta correlation function (uncorrelated). *SAL*

Ref.: Bendat (1958).

Ambient noise (see **environmental noise**).

Amplitude noise is the fluctuation of echo signal amplitude caused by the complex shape of the target and its changing aspect angle, excluding the effects of changing target range. It is the result of echo signal amplitude modulation by interference between scatterers on the target, and can be represented as the fluctuating sum of many small vectors changing randomly in relative phase. Amplitude noise can include periodic components and typically is divided into low-frequency and high-frequency amplitude noise, depending on its spectral characteristics. Amplitude noise, which is the reason for scintillation error, is termed amplitude fluctuation. (See **FLUCTUATION**; **RCS fluctuation.**) *SAL*

Ref.: IEEE (1993), p. 34; Skolnik (1990), p. 18.34.

Angle noise is "the noise-like variation in the apparent angle of arrival of a signal received from a target, caused by a change in phase and amplitude of target-scattering sources and including angular components of both glint and scintillation error." Random variation in observed angular location of the target results from a change with time in the apparent location of the target with respect to a reference point, which is usually a center of gravity of the reflectivity distribution along the target. In some cases this location can fall at a point completely outside the extremities of the target. Angle noise typically has Gaussian distribution and the errors caused by angular noise are inversely proportional to range. The angle-noise phenomenon affects all types of radars but it is mainly of concern for precision tracking radars operating at medium and short range. The main measures for angle noise reduction are proper choice of servo bandwidth and AGC characteristics. (See also **ERROR**.) *SAL*

Ref.: IEEE (1993), p. 39; Skolnik (1990), p. 18.37.

Antenna noise is the result of the noise received by antennas from external sources (primarily, because of atmospheric radiation) and the internal noise generated inside antenna and microwave feed structures (primarily generated in the ohmic components: resistive conductors and imperfect insulators). Typically, it is described by the antenna noise temperature. (See **TEMPERATURE.**) *SAL*

Ref.: Skolnik (1990), p. 2.28.

Atmospheric noise [atmospherics] is a noise component produced by radiation from lightning strokes (not to be confused with atmospheric absorption noise). The spectrum of atmospheric noise falls off rapidly with increasing frequency, and for frequencies above 50 MHz has no significant effect on radar. (See also **ATMOSPHERICS**.) *SAL*

Ref.: Skolnik (1980), p. 463.

Atmospheric absorption noise is produced by the atmosphere by the same phenomena that result in attenuation of the electromagnetic waves. The resulting noise has at the radar input a spectral density $\Delta N_0 = kT_\alpha(1 - 1/L_{\alpha 1})$, where T_α is the ambient temperature of the atmospheric gases contributing the loss ($T_\alpha \approx 290$K) and $L_{\alpha 1}$ is the one-way loss through the atmosphere. *SAL*

Ref.: Skolnik (1980), p. 461.

Automatic video noise leveling is a "CFAR technique in which the receiver gain is readjusted to maintain a constant video noise level." In this case the noise level is sampled at the receiver output at the end of each range sweep prior to the next transmission and the resulting gain is fixed throughout the next sweep. (See **CONSTANT-FALSE-ALARM RATE.**) *SAL*

Ref.: IEEE (1990), p. 6.

Noise bandwidth is the bandwidth of an equivalent rectangular *filter* whose noise-power output is the same as the filter with frequency-response characteristic $H(f)$. The usual notation is B_n, given by the formula:

$$B_n = \frac{\int_{-\infty}^{\infty} |H(f)|^2 df}{|H(f_0)|^2}$$

where f_0 is the frequency of maximum response (usually at midband). The measurement of noise bandwidth involves a complete knowledge of the response characteristic $H(f)$, so in practice the 3-dB bandwidth is used as an approximation. *SAL*

Ref. Skolnik (1980), p. 18.

Colored noise is a noise component having nonuniform spectral density, as opposed to *white noise*. *SAL*

cosmic noise (see **solar noise**).

Direct amplified noise is noise directly amplified without a carrier frequency, used in jamming to saturate the radar receiver noise level. In this case the radar operator may not realize the radar is jammed if AGC or automatic noise leveling are employed. It is used as barrage noise jamming with a bandwidth of more than 50 MHz. This method is not widely used, as linear wideband amplification is not a very efficient process. This technique is also termed direct noise amplification (DINA). *SAL*

Ref.: Johnston (1979), p. 58.

Noise duty cycle is the parameter sometimes used in description of receivers analogously to signal duty cycle and equal to the product of average receiver range-gate-on-interval and pulse repetition frequency in the assumption that the receiver noise generated after the range-gate switch is negligible. *SAL*

Ref.: Currie (1987), p. 794.

Environmental noise is ambient noise arising from the natural environment. For much of the radar band this noise is relatively low in comparison with internal receiver noise and has to be taken into consideration usually only for radars employing receivers with very-low-noise input stages. The main sources of environmental noise are atmospheric noise, urban or industrial noise, and cosmic noise in which solar noise sometimes is separately distinguished. Sometimes this noise is called *ambient* or *external noise*. *SAL*

Ref.: Skolnik (1990), p. 463.

Noise factor [figure] is the ratio of the noise at the output of a practical linear network (e.g., a receiver) to the noise at the output of an ideal network at standard temperature. The typical notation is F_n and the evaluation equation is

$$F_n = N_o/(kT_0B_nG_a) = \frac{S_i/N_i}{S_o/N_o} = 1 + \frac{T_e}{T_0}$$

where k is Boltzmann's constant, N_o is the output from the network, T_0 is the standard temperature (290K), T_e is the effective noise temperature of the network, B_n is the noise bandwidth, G_a is available gain, and S_i/N_i and S_o/N_o are signal-to-noise ratios at the input and the output of the network. The noise factor may serve as a measure of signal-to-noise degradation when the signal passes the network. Typically, it is measured in decibels (10 log F_n). If the receiver incorporates several cascaded stages the resultant noise figure may be shown as

$$F_n = F_{n1} + \frac{F_{n2}-1}{G_1} + \frac{F_{n3}-1}{G_1G_2} + \dots$$

where i = 1, 2, ... is the number of the stage and F_{ni} and G_{ni} are noise figure and gain of corresponding stage. So if the gain is rather high, it is the first stage (usually RF amplifier) that determines the noise properties of the whole receiver. *SAL*

Ref.: IEEE (1993), p. 849; Skolnik (1980), pp. 19, 344; Morchin (1993), p. 411; Leonov (1988), p. 56.

Gaussian noise is described by the random process with the Gaussian distribution of probability density function:

$$W(u) = \frac{1}{\sqrt{2\pi}\sigma_n}\exp\left[-\frac{(U_n-\overline{U_n})^2}{2\sigma_n^2}\right]$$

where u is noise voltage, $\overline{U_n}$ is the constant component of noise voltage, and $\sigma_n{}^2$ is noise variance.

If $\overline{U_n} = 0$, Gaussian noise from k different independent sources are summed as the powers:

$$U_{n\Sigma}^2 = U_{n1}^2 + U_{n2}^2 + \dots + U_{nk}^2$$

AIL

Ref.: Barton (1964), p. 11; Svistov (1977), p. 85

industrial noise (see **environmental noise**).

Johnson noise (see **thermal noise**).

Noise power is the power of the noise at the output of a network. The usual notation is N. Noise power is given by the formula:

$$N = kT_sB_n$$

where k is Boltzmann's constant, T_s is the system noise temperature, and B_n is noise bandwidth. *SAL*

Ref.: Morchin (1993), p. 407.

Phase noise is random phase modulation of the carrier produced by signal sources such as oscillators. This kind of noise often referred to as *single-sideband noise*, it causes carrier sidebands whose amplitude vary as a function of frequency offset from the carrier frequency, and it is the predominant factor of short-term stability. *SAL*

Ref.: Morchin (1993), p. 414.

Noise power (spectral) density is the noise power per unit bandwidth, n/B_n.

Quantization noise is noise arising from the conversion of sampled voltages into the integer numbers in analog-to-digital converters. This noise is the source of random error additional to the errors caused by the noise in analogous networks of radar receiver. *SAL*

Ref.: Skolnik (1990), p. 3.40.

Range noise is "the noise-like variation in the apparent distance of the target, caused by changes in amplitude and phase of the target-scattering sources, and including radial components of glint and scintillation error." Range noise is similar to angle noise. It can also result in an error greater than the wander of the target center of gravity, can fall outside the target span, can be a major limitation to the accuracy of velocity obtained from range rate, and can cause angle-tracking errors in multilateration systems that use high-precision range measurements. A beacon installed on a target can eliminate an error arising from range noise, provided it has very stable circuitry avoiding pulse jitter and drift. (See also **ERROR**.) *SAL*

Ref.: Skolnik (1990), p. 18.45.

Rayleigh noise is the noise described by the random process with the Rayleigh distribution of probability density function:

$$W\left(\frac{u}{\sigma_n}\right) = \frac{u}{\sigma_n}\exp\left(-\frac{u^2}{2\sigma^2}\right),\ u \ge 0$$

$$= 0,\ u < 0$$

where u is instantaneous noise voltage and σ_n^2 is noise variance. The envelope of receiver noise, consisting of two quadrature components of Gaussian noise, has a Rayleigh distribution. *AIL*

Ref.: Barton (1964), p. 11; Svistov (1977), p. 866.

Servo noise is the noise arising from different mechanical instabilities of servomechanisms in radars employing mechanical scanning. The sources of this noise, which limits angular accuracy in some tracking radars, are backlash and compliance in the gears, shafts, and structures of the mount. *SAL*

Ref.: Skolnik (1980), p. 170.

Shot noise is the fluctuation of electronic current or ionic emission caused by the continuous changing of the number electrons or ions being emitted when the structure of the emitting surface is unchanged. It is the main source of internal noise of microwave amplifiers due to current fluctuations in the process of electron emission. *AIL*

Ref.: Chistyakov (1986), p. 23; Fink (1982), p. 4.4.

Solar noise is the environmental noise arising from electromagnetic radiation emitted by sun. Its power, unlike most other noise sources, increases with frequency, and in some cases it can affect radars with low-noise receivers. *SAL*

Ref.: Skolnik (1980), p. 463.

A **noise source** is a device employed for generation of noise in ECM technique. Usually, these are different types of microwave tubes and solid-state devices. *SAL*

Ref.: Johnston (1979), p. 64.

speckle noise (see **SPECKLE**).

Target noise is the fluctuation in amplitude, location, and doppler caused by the target, excluding atmospheric effects. Typically, target noise includes amplitude noise, angle noise, range noise, and doppler scintillation. (See also **FLUCTUATION, GLINT.**) *SAL*

Ref.: IEEE (1990), p. 29; Skolnik (1990), p. 18.34.

noise temperature (see **TEMPERATURE**).

Thermal noise is noise generated by the thermal motion of the conduction electrons in tubes or solid-state components of radar receivers. The power of this noise is directly proportional to the temperature of the ohmic portions of the circuit and the receiver bandwidth (see **noise power**). Thermal noise is the main source of uncertainty resulting in fundamental inability of error-free radar operation in real conditions, as it results in arising of errors when radar operates in any of its modes, performing detection, measurement, or discrimination function. Thermal noise sometimes is called *Johnson noise*. *SAL*

Ref.: Skolnik (1980), p. 18.

tracking noise (see **ERROR, measurement**).

Urban noise is noise arising from automobile ignition, power tools, and other man-made sources of unintentional interference. It can affect radars operating at low frequencies. *SAL*

Ref.: Skolnik (1980), p. 463.

White noise is a random process having the uniform spectrum density at all frequencies, and correspondingly, a delta-function-type correlation function. This means that any two values of white noise in any two moments of time regardless of how close they are to each other, are always independent (uncorrelated). Theoretically, white noise has an infinite spectrum but in practical applications any noise in the passband of a real radar circuit can be presented as white because the passband is always finite and the out-of-the-band components do not affect the circuit. White noise is termed so by analogy with the white light that has an approximately uniform spectrum within its visible range. *AIL*

Ref.: Barton (1964), p. 323; Kazarinov (1990), p. 19.

NOMENCLATURE, radar. The radar nomenclature system used in U.S. military equipment was developed during World War II, and is formally entitled the Joint Electronics Types Designation System (JETDS). It is described by Military Standard MIL-STD-196D and was formerly known as the Joint Army-Navy Nomenclature System or AN System, Table N2. The designation begins with the letters AN, followed by a slash and three additional letters from Table N2 specifying where the equipment is installed, its type, and purpose. The final numerical value indicates the order in which the equipment was placed into development or entered into the nomenclature system. Other systems of designation are used for civil radars and those of non-U.S. origin. *SAL*

Ref.: Skolnik (1990), p. 1.19.

Table N2
JETDS Equipment Designators

Installation (first letter)		Type of equipment (second letter)		Purpose (third letter)	
A	Piloted aircraft	A	Invisible light, heat radiation	A	Auxiliary assembly
B	Underwater mobile, submarine	C	Carrier	B	Bombing
D	Pilotless carrier	D	Radiac (radioactive detection, indication, and computation devices)	C	Communications (receiving and transmitting)
F	Fixed ground	E	Laser	D	Direction finder, reconnaissance and/or surveillance
G	General ground use	G	Telegraph or teletype	E	Ejection and/or release
K	Amphibious	I	Interphone and public address	G	Fire control or searchlight directing
M	Ground, mobile	J	Electromechanical or inertial wire-covered	H	Recording and/or reproducing (graphic meteorological)
P	Portable	K	Telemetering	K	Computing
S	Water	L	Countermeasures	M	Maintenance and/or test assemblies (including tools)
T	Ground, transportable	M	Meteorological	N	Navigational aids (including altimeters, beacons, compasses, racons, depth sounding, approach and landing)
U	General utility	N	Sound in air	Q	Special or combination of purposes
V	Ground, vehicular	P	Radar	R	Receiving, passive detecting
W	Water surface and underwater combination	Q	Sonar and underwater sound	S	Detecting and/or range and bearing, search
Z	Piloted and pilotless airborne vehicle combination	R	Radio	T	Transmitting
		S	Special types, magnetic, etc., or combinations of types	W	Automatic flight or remote control
		T	Telephone (wire)	X	Identification and recognition
		V	Visual and visible light	Y	Surveillance (search, detect, and multiple-target tracking) and control (both fire control and air control)
		W	Armament		
		X	Facsimile or television		
		Y	Data processing		

(From Military Standard Joint Electronics Type Designation System, MIL-STD-196D, Jan. 19, 1985)

NOTCHER

A **mainlobe notcher** is a canceler used to suppress high-duty cycle interference entering the mainlobe of radar antenna pattern from a single direction in space. It is implemented in a way analogous to coherent sidelobe cancelers, but the auxiliary antenna for the mainlobe notcher has a gain much higher than that of auxiliary antenna in sidelobe canceler because the mainlobe has much higher gain than sidelobes. *SAL*

Ref.: Lewis (1986), p. 127.

sidelobe notcher (see **ALGORITHM, sidelobe cancellation**).

NUCLEAR EFFECTS. The effect of nuclear blasts on radar operation is to produce “blackout” from an increased electron density in the ionosphere (for high-altitude nuclear blasts). For limited periods of time electron densities can go up to 10^{13} per cc or greater. Densities 100 times greater than those of the normal F-layer (10^8 as compared to the normal 10^6 per cc) may prevail over paths of 300 or 500 km. This ionization decays over a period of about an hour. The attenuation varies with the square of the wavelength, so radars operating at shorter wavelengths will be affected for much shorter times than radars operating in the long wave portion of radar band. *SAL*

Ref.: Barton (1976), p. 471.

NULL AXIS (see **AXIS, antenna**).

NULLING

Feed-through nulling is a technique used in CW radars to null out the transmitted signal leaking into the receiver. It uses negative feedback at the receiver input to cancel whatever transmitter signal may appear there. It is arranged to cancel only the signal carrier, as signals offset from the carrier include the desired signals with doppler shift. *SAL*

Ref.: Skolnik (1990), p. 4.3; Brookner (1988), p. 58.

Nulling within the main lobe (see **NOTCHER, mainlobe**).

Nulling within the sidelobes (see **CANCELLATION, sidelobe**).

NYQUIST

The **Nyquist formula [theorem]** describes the dependence of the mean-square value of the thermal noise voltage v_n in an active resistance R, on the value of the absolute temperature T and the frequency bandwidth B, as

$$v_n = \sqrt{4kTRB} \quad \text{(V)}$$

where k is Boltzmann's constant. The available noise power output from the resistor is 1/4 this value. *IAM*

Ref.: Rakov (1970), vol. 2, p. 110; Skolnik (1990), p. 2.26.

Nyquist rate (see **SAMPLING**).

Nyquist sampling theorem (see **SAMPLING**).

O

OPERATOR, radar. A radar operator is the person who performs two basic functions in the process of radar service: to ensure proper operation of the radar through its maintenance and to ensure that the radar performs its main task such as detection, tracking, target recognition, and so forth. if the performed task requires operator involvement. The degree of operator involvement depends on the concept of radar operation. In automatic radar, an operator does not interfere in the process of operation, but only controls proper operation through the radar monitoring system and eliminates the failure if it had occurred. In automated radars, an operator is the important link in the process of operation (Fig. O1). Working with radar display, an operator detects the targets and selects which of them are important for subsequent processing. There are some specific requirements depending on the tasks performed by radar, the features of its structure, and conditions of work, which require special knowledge and training. (See **TRAINER, radar.**) *AIL*

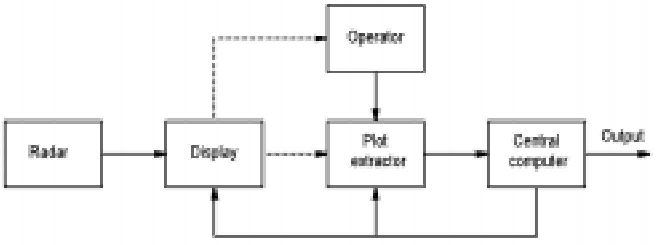


Figure O1 Block diagram of automated radar data processing.

Ref.: Skolnik (1962), pp. 439, 566; Romanov (1980), pp. 5-24.

OSCILLATION. Oscillation is "the variation, usually with time, of the magnitude of a quantity with respect to a specified reference when a magnitude is alternately greater and smaller than the reference." In radar applications, this term is typically applied to the variation of an electromagnetic wave. *SAL*

Ref.: IEEE (1993), p. 896.

Coherent oscillation is when two (or more) oscillations have a phase difference between them which does not change over time in interval *T.* (See **COHERENCE.**) *AIL*

Ref.: Popov (1980), p. 178.

A **damped oscillation** is a natural oscillation with amplitude continuously decreasing based on an exponential law, due to loss of energy in an oscillatory system. *AIL*

Ref.: Popov (1980), p. 131.

Forced oscillation arises in an isolated system driven externally. During forced oscillations, the oscillatory process depends on both the nature of the external force and the properties of the system. *AIL*

Ref.: Popov (1980), p. 75.

harmonic oscillation (see **sinusoidal oscillation**).

Modulated oscillation results from the interaction of two oscillations, *modulated* and *modulating*. A modulated oscillation is an oscillation serving as a carrier for a modulating oscillation. A sinusoidal oscillation or a periodic square pulse train is used as the modulated oscillation. A modulating oscillation is one containing transmitted information and used to change a certain parameter (depending upon modulation type) of a modulated oscillation. (See **MODULATION.**) *AIL*

Ref.: Popov (1980), p. 235.

Monochromatic oscillation has amplitude, frequency, and phase that do not depend on time. *AIL*

Ref.: Popov (1980), p. 238.

Natural oscillation arises in an isolated system as a result of an initial external effect delivering energy for the entire process. Such an effect might be connection of voltage or opening (closing) a section of a circuit. Natural oscillations arise at the resonant frequency of a system, the value of which is determined by system parameters. *AIL*

Ref.: Popov (1980), p. 402.

Parametric oscillation arises under the effect of an external periodic force that transmits energy to a system through a change in a certain system parameter (usually capacitance or inductance). *AIL*

Ref.: Popov (1980), p. 276.

Pi-mode oscillation appears at combination frequencies arising in an oscillator such as a magnetron where phases of adjacent resonators differ by π radians. *AIL*

Ref.: Popov (1980), p. 180.

Self-sustaining oscillation is maintained in a circuit due to an external energy source, through automatic regulation of the input of energy to the circuit. Losses of the energy of the oscillations in the circuit are compensated for automatically by energy from a source. Self-sustaining oscillations are continuous oscillations. *AIL*

Ref.: Popov (1980), p. 10.

Sinusoidal oscillations are the simplest periodical oscillations described as a function of time t by

$$a(t) = A \sin(\omega t + \phi)$$

where A is the amplitude, ω is frequency, and ϕ_0 is initial phase. In Russian literature this is termed *harmonic oscillation*. *AIL*

Ref.: Popov (1980), p. 80.

OSCILLATOR, microwave. A microwave oscillator is a device that converts the energy of a power source into the energy of microwave oscillation. Along with a self-excited oscillator (master oscillator), the oscillator often includes a frequency multiplier and power amplifiers.

The basic electrical parameters of an oscillator are its power level, efficiency, operating frequency and range of tuning, stability of frequency and output power, noise level close to the carrier, and frequency spectrum.

Oscillators are classified according to purpose, type of active components, value of parameters (power, frequency range, etc.), types of stabilization of oscillation.

In terms of type of active components, oscillators are divided into two large classes: semiconductor and tube oscillators (Table O1). Low-power oscillators (units of milliwatts) include klystron oscillators, tunnel-diode oscillators, and low-power transistor oscillators. Other semiconductor oscillators are used in more powerful devices (hundreds of milliwatts, tens of watts).

Table O1
Classification of Microwave Oscillators Based on Types of Active Components

Tube	Semiconductor
klystron	Gunn diode
magnetron	avalanche transit-time diode
amplitron	bipolar transistor
backward-wave tube	field-effect transistor
traveling-wave tube	
gyrotron	
grid-control tube	

In radar applications, high-power oscillators (up to megawatts) use magnetron tubes, backward wave tubes, amplitrons, and gyrotrons, which are also distinguished by their high frequencies (up to 100 GHz).

Microwave oscillators are used in transmitters and local oscillators of radars, in measurement apparatus, in pumping oscillators of parametric amplifiers, exciters, and so forth. *IAM*

Ref.: Gassanov (1988), p. 174; Rakov (1970), vol. 2, p. 5

A **amplitron oscillator** is one that uses an amplitron for generation or amplification of the frequency of the master oscillator. In the first case, the oscillator is also called a *stabilitron*. Depending on the operating mode, the amplitron is used in pre-amplification stages with a gain coefficient of 12 to 15 dB and low efficiency, and in terminal stages of radar transmitters with a gain coefficient of 6 to 8 dB and efficiency of 55 to 80%. An amplitron provides high phase stability (less than 3° to 5° per 1% change in operating voltage), bandwidth (7 to 20%), and output power (2 to 5 MW). To eliminate the excitation inherent in a amplitron at low levels of output power, ferrite duplexers are used at the input and output of instruments in multi-stage amplitron-based transmitters.

Amplitron oscillators are used in certain air-traffic control radars. *IAM*

Ref.: Perevezentsev (1981), pp. 167–172.

An **avalanche transit-time diode oscillator** is an oscillator that uses an avalanche transit-time diode possessing a negative dynamic conductivity in avalanche breakdown mode as its active component.

The basic oscillation mode is the IMPATT mode, which allows production of hundreds (tens) of milliwatts of power at frequencies on the order of units (tens) of gigahertz, with an efficiency of 15 to 20%. The frequency limit of operation of these oscillators is 200 GHz. With retuning of the oscillating circuit, the range of oscillation may reach an octave. Several times more power and efficiency up to 50 to 70% are obtained in the TRAPATT mode. However, owing to the great amount of noise, criticality with respect to tuning, and limitation of operating frequency to 10 GHz, at which the transistors operate effectively, oscillators using avalanche transit-time diodes are not widely used in the TRAPATT mode.

Oscillators using avalanche transit-time diodes can have a waveguide, coaxial, or microstrip design. For stabilization of the operating mode of the avalanche transit-time diodes, stabilization of the bias current is required. Oscillations of the current of the diodes are rich in higher harmonics, which may be separated by connecting an additional cavity to the diode. When a diode is connected to a low-quality oscillating system, a wideband noise oscillator is produced.

The basis use of the avalanche transit-time oscillators is in phased-array antennas and in the output stages of the transmitters of radio relay stations. *IAM*

Ref.: Gassanov (1988), p. 183; Voskresenskiy (1981), p. 317.

A **backward-wave-tube oscillator** is one whose basic element is a backward-wave tube. The generation mode in a backward-wave tube arises when there is equality of the transfer rate of electrons and the phase rate of the reverse spatial harmonic of the wave (condition of synchronization) and the current is higher than the starting current. This is assured through selection of the slow-wave structure and the electrical mode.

For backward-wave tube oscillators, the dependence of the frequency of oscillations on the accelerating voltage is characteristic. Backward-wave tube oscillators usually are used in continuous mode. The parameters of an oscillator depend on the type of backward-wave tube. In oscillators of the backward-wave tube type M, the dependence of the frequency on the voltage is more linear. For electronic frequency tuning, less of a change in voltage is required in than in type O backward-wave tube oscillators. Backward-wave tube oscillators of type M have these characteristics: range 1 to 90 GHz, output power in the CW mode up to several tens of kilowatts for decimeter waves, hundreds of kilowatts for centimeter waves, voltage rating of 3 to 4 kV, and efficiency 50 to 60%. Oscillators of the O type operate in all microwave and millimeter-wave bands with a power from tens of milliwatts to several watts in the CW mode, voltage rating 2 to 10 kV, and efficiency of several percent.

At present, type M backward-wave tube oscillators are the most powerful oscillators with electronic frequency tuning. *IAM*

Ref.: Andrushko (1981), p. 77; Dubin (1972), p. 71.

A **blocking oscillator** is a relaxation oscillator of short pulses, which constitutes a single-stage amplifier with transformer feedback. Owning to the strong inductive feedback, the blocking generator forms nonsinusoidal, practically rectangular pulses with a duration from units of nanoseconds to several tens of microseconds, with a wide range of change of their repetition frequency. Blocking oscillators can operate in the triggered mode, the auto-oscillation mode, and the synchronization mode. They are used as oscillators for modulating, synchronizing, blanking and strobing pulses, and sawtooth voltages and currents. *IAM*

Ref.: Grigor'yants (1981), p. 133.

A **crystal oscillator** can be of the tube or solid-state type and uses a piezoelectric crystal (usually quartz) as the frequency-determining circuit. The frequency of operation can be from 1 kHz to about 200 MHz when using bulk-wave crystal resonators, and up to 1 GHz for SAW devices. The *Q*-factor of a typical crystal is in the order of 10^6 to 10^7, and stabilities in the order of one part in 10^{10} per day can be obtained. Temperature-controlled ovens are commonly used to maintain constant frequency within a few parts per million, and control to one part on 10^7 can be obtained by compensation for measured temperature changes. *SAL*

Ref.: Fink (1982), pp. 7.21–7.24.

A **digital oscillator** is a digital device that carries out the function of an analog oscillator in digital form. A digital oscillator at frequency ω_0 generates, at clock moment *kt*, a pair of outputs $c_k = \cos\omega_0 kT$ and $s_k = \sin\omega_0 kT$, where T is the clock interval. We distinguish between table and recursive oscillators. The former are realized using a permanent storage device with a very large memory capacity, which performs the conversion of $\omega_0 kT$ into c_k and s_k. The latter calculate the readings of c_k and s_k using recursive correlations. Recursive oscillators use special methods of stabilization for prevention of increasing computation error as k grows. Digital frequency synthesizers are used in digital signal processing devices for transfer and inversion of the spectrum. (See also **GENERATOR, waveform.**) *IAM*

Ref.: Beskin, L. N., *Radiotekhnika*, no. 4, 1984, pp. 63–65, in Russian.

An **extended interaction oscillator** is usually a klystron oscillator in which the cavity of the multiresonance klystron is replaced by a system of two or several associated cavities. A multigap cavity is produced, with whose pole the electron flux interactions. Self-excitation arises in the forward spatial harmonic with positive feedback due to reflections from the ends of the slowing system. The oscillators have electronic and mechanical tuning of the generation frequency, an output power up to 1 kW at 30 GHz, 20W at 140 GHz in the continuous mode, and 10 kW at 95 GHz, 10 W at 280 GHz in the pulse mode. Frequency stability is much greater than in reflex klystrons. They are used in millimeter-band radars. *IAM*

Ref.: Andrushko (1981), p. 60.

A **field-effect tetrode oscillator** is one that uses a field-effect transistor with two gates as its active component. In power generators they are inferior to field-effect transistors, since they have less power output per unit of gate width. Their basic use is as special circuits for oscillators with higher frequency stability, for example for Doppler radars. In master oscillators based on tetrodes, the first half of the instrument is used directly for oscillation, and the second half is a pulse modulator with a high switching speed. The frequency deviation of such an auto-oscillator at a frequency of 8.6 GHz does not exceed 300 Hz. Industrial types of microwave tetrodes are used with frequencies up to 12 GHz. The circuit of a combined oscillator-multiplier, in which the first part of the instrument operates as an oscillator, and the second as a multiplier, with a frequency of output signal of up to 22 GHz, is in use. *IAM*

Ref.: Joshi, J. S., and Pengelly, R. S., *Proc. Int. Microwave Symp.*, Washington, D.C., 1980.

A **grid-controlled (tube) oscillator** is an oscillator that uses oscillator triodes or tetrodes in bands up to 2 GHz. (See **triode oscillator.**) At present its use is comparatively rare due to restrictions in frequency that lead to a sharp reduction in output power with an increase in frequency (inversely proportional to the square of the frequency).

Typical parameters: peak power 2 MW at a frequency of 425 MHz, high efficiency of 40 to 60%, bandwidth of 1 to 2% with stage amplification of 10 to 20 dB. *IAM*

Ref.: Perevezentsev (1981), p. 156.

A **Gunn-diode oscillator** is one that uses a Gunn-effect diode as its active component. Microwave oscillations in the cavity of the oscillator arise as a result of formation and movement of domains that give up their energy in the braking half-periods of the variable field (see **GUNN effect**). The basic operating mode of the diode in the oscillators is the mode of limited accumulation of volume charge (LAVC). To produce this mode, a constant voltage greater than critical is applied to the diode, which is contained in a high-quality cavity, and the oscillating system is tuned to a frequency is many times greater than the transit frequency (up to several gigahertz). Frequency retuning is possible in a band higher than an octave. The power of the oscillators amounts to units of Watts in continuous mode, with an efficiency up to 15 to 20%, and units of kilowatts in pulse mode. In practice, hybrid modes combining the LAVC mode with other domain modes, and possessing less sensitivity to changes in parameters and loads are widely used. Stabilization of the supply voltage is required for stabilization of the operating mode.

Compared with the avalanche transit-time diode, Gunn-effect diode oscillators have less power, less noise level, and lower supply voltage (up to 10V). In the decimeter, centimeter, and millimeter bands, Gunn diodes put out hundreds, tens, and units of milliwatts, respectively.

The fields of application are roughly the same as in avalanche transit-time diodes. Gunn-diode oscillators are used as local oscillators in the millimeter band, and harmonic oscillation modes are used in frequencies above 100 GHz. *IAM*

Ref.: Gassanov (1988), p. 186.

A **gyrotron oscillator** is one based on a gyrotron. It is the most powerful and efficient source of power in the millimeter-wave band, capable of operating in long-pulse modes (up to 75 ms) or the CW mode. The efficiency of gyrotron oscillators in the short-pulse mode reaches 65%, and the output power in the CW mode is more than 200 kW at a frequency up to 70 GHz. Powerful oscillators (up to 1 MW) are made in the higher frequency range of 100 to 200 GHz.

Gyrotron oscillators are used in developmental radars and communications systems in the millimeter-wave band and also for heating plasma in large-scale tokamacs. *IAM*

Ref.: Temkin, R. J. et al., *Int. J. Infrared and Millimeter Waves*, 1982, vol. 3, no. 4; Carmel, Y. et al., *Phys. Rev. Let.t* ***50***, 1983, no. 2.

A **klystron oscillator** uses a klystron as its active component. They may be based either on a reflex \klystron or on a transit klystron. In the transit klystron, self-excitation is provided by feedback between the output and input cavities, connecting them by a coaxial transmission line. The length of the line provides the necessary phase of the oscillations applied to the input. Excitation occurs when the current is above a certain threshold (starting current).

The maximal output power and the efficiency of klystron oscillators are the same as for transit klystron amplifiers. Transit-klystron oscillators have comparatively limited uses, basically in the range of 5.5 to 44 GHz as highly stable medium-power oscillators (0.2 to 200W) of doppler radars, radio beacons, and also for pumping of parametric amplifiers.

Reflex klystrons are widely used as low-power (0.1 to 1W oscillators, thanks to their design simplicity, and the capacity for electronic frequency tuning (usually amounting to ±10 to 15%).

They are used in low-power transmitters of radar navigation apparatus and as microwave local oscillators. *IAM*

Ref.: Andrushko (1981), pp. 30, 43; Perevezentsev (1981), p. 158.

local oscillator (see **LOCAL OSCILLATOR**).

magnetron oscillator (see **MAGNETRON**).

A **master oscillator** is a source of microwave (RF) oscillations, operating in a given frequency range with the necessary frequency stability. A master oscillator is usually a low-power, high-stable (transistor, tube, etc.) auto-oscillator and is used for excitation of oscillations, which are then amplified in subsequent stages.

One distinguishes between single-circuit auto-oscillators, which are composed in an inductive or capacitive circuit; double-circuit, which are implemented with series connection of an additional circuit to an anode circuit of any single-circuit oscillator; and auto-oscillators with transformer feedback. (See **blocking oscillator.**) The frequency of the generated oscillations in master oscillators is usually close to the natural frequency of the circuit and may be changed over wide limits by the tuning elements. *IAM*

Ref.: Popov (1980), p. 128.

A **monotron oscillator** is one in which the continuous electron beam transfers energy to the electromagnetic field in the individual resonant cavity. The use of high-quality superconducting cavities makes it possible to design powerful monotron microwave oscillators whose frequency stability is as high as that of crystal oscillators. *IAM*

Ref.: Popov (1980), p. 238.

A **nonresonant oscillator** is an electronic device with one or multiple interactions in which the oscillating system in the operating frequency band does not have resonant properties and is excited in the traveling electromagnetic wave mode. Examples of such devices include the traveling-wave tube, the backward-wave tube, and the stabilitron. (See also **traveling wave tube oscillator** and **backward-wave tube oscillator**.) *IAM*

Ref.: Popov (1980), p. 382.

A **radio-frequency (RF) oscillator** is a generator of electromagnetic oscillations in the bands 100 kHz to 100 MHz. The RF oscillator is used as the basic element of radio transmitters in this range, and low-power RF oscillators as local oscillators of superheterodyne receivers. *IAM*

Ref.: Popov (1980), p. 81.

A **resonant oscillator** is an electronic device that creates discrete types of oscillations. Such oscillators are based on the short-term interaction of electrons with the electronic field in the gap of a hollow cavity. These include devices with long and multiple interactions, whose slow-wave system operates in the standing-wave mode and forms an oscillating system with discrete types of oscillations. Examples of such oscillators include microwave triode and tetrode oscillators (see

grid-control tube oscillator), **klystron oscillator** and **magnetron oscillator**. *IAM*

Ref.: Popov (1980), p. 382.

A **semiconductor oscillator** uses semiconductors as its active components. Tunnel diodes and low-power transistors are used in semiconductor low-power microwave oscillators (units of milliwatts). The more powerful devices (tens and hundreds of milliwatts or more) use avalanche transit-time diodes, Gunn diodes, and powerful transistors. (See **transistor oscillator**; **tunnel-diode oscillator**; **DIODE, avalanche-transit-time**; and **DIODE, Gunn.**) Production of large output powers is restricted due to the difficulty of removing the heat from the active zone of semiconductor instruments. To increase the output power, heat sinks and forced cooling are used, and semiconductor devices are being developed that combine several diode transistor cells in one housing. *IAM*

Ref.: Gassanov (1988), p. 174.

A **shock-excited oscillator** is based on electronic tubes (or transistors) in which generation of oscillations occurs as a result of application of a pulsed signal to its input, exciting the circuit with the current increment. The most common is a circuit with positive feedback, in which the input pulse closes the input electronic device (tube), shunting the circuit, and causing sinusoidal oscillations to arise, supported through the feedback circuit with a cathode follower. This circuit is used as a mark generator for producing range-scale marks.

A shock-excited oscillator is used for producing short, peaked pulses. In this case, a critical mode is used for which the loop in the anode circuit of the tube is shunted by the resistor $R = (\sqrt{L/C})/2$. The driving input is a negative pulse with an amplitude exceeding the voltage of the source.

Shock-excited oscillators are also used in circuits for producing rectangular pulses of various duration, for expansion of pulses, and in other circuits. *IAM*

Ref.: Druzhinin (1967), p. 218.

A **solid-state oscillator** uses solid-state passive components: oscillating devices, matching transformer circuits, adders, etc. Circuits based on discrete L and C components are used as the solid-state components of oscillating systems: cavities of traveling or standing waves in sections of transmission lines, including radial, spiral, printed, and dielectric; spheres of yttrium-iron garnet [YIG].

To reduce the power consumption, mass and cost of the devices, solid-state oscillators are made in the form of hybrid integrated circuits. Thanks to their high reliability and small size, solid-state oscillators are widely used in multifunction radar modules and other radar apparatus. *IAM*

Ref.: Gassanov (1988), p. 24.

A **surface-acoustic-wave (SAW) oscillator** is an amplifier to which a positive feedback circuit is applied, and which includes a stabilizing (frequency-selector) component. A delay line or SAW cavity is used as the latter.

The range of the basic surface acoustic wave frequencies reaches 3 GHz, with frequency deviation up to 1%, which significantly exceeds the corresponding parameters of the crystal cavity in crystal oscillators. The frequency stability with the simplest thermostats (±4°C) is no worse than 10^{-6} per degree, which is comparable with a crystal oscillator. The relative change in frequency due to aging is an average of $\pm 10^{-6}$ per year.

These oscillators can be produced in integrated form with accommodation on one substrate of both a stabilizing surface acoustic-wave component, and the electronic components of the oscillator (amplifiers, transistors, and diode switches). Thanks to this, and the high values of the characteristics, SAW oscillators are used widely as tunable-frequency oscillators. The tuning range in a programmable frequency synthesizer based on periodic connection of various channels of a delay line is 30%. Because of the capacity of operation in the mode for generation of pulses with a high degree of coherence, SAW oscillators are used in communications systems for formation and processing of a coherent pseudonoise signal with frequency hopping. *IAM*

Ref.: Gratze, S. C., and Barton, R. K., *Electronic Engineering*, March 1975, pp. 49–51; Hartemann, P., *Electronics Letters* ***11***, 1975, no. 5, pp. 119–120.

A **transistor oscillator** uses a bipolar or field-effect transistor as the active component and constitutes an amplifier with positive feedback in the self-excitation mode. Oscillators based on bipolar transistors are capable of operating at frequencies up to 10 GHz, while those based on field-effect transistors can operate up to 100 GHz). The latter are inferior to oscillators based on bipolar transistors in terms of noise properties, but they have a lower supply voltage and lower value of internal feedback and therefore are becoming more and more common, especially in semiconductor integrated circuits

Transistor oscillators supply output power up to 1W. To obtain higher values, parallel transistor amplifiers are used. Thanks to the use of a frequency stabilization with a cavity in the form of a sphere of iron-yttrium garnet, it is possible to obtain a high nonlinearity of tuning (to ±0.05%) over a significant band. *IAM*

A **triode oscillator** is one in the meter or decimeter band based on a microwave triode. The tube is usually connected in a circuit with coaxial anode-grid and grid-cathode tuned cavities. The feedback between them is provided through the interelectrode capacitance of the tube, and if necessary through additional inductance and capacitive components.

The efficiency of triode oscillators is up to 50% for waves longer than 30 cm, and lower at shorter waves. Pulse power in the meter band reaches several megawatts, and frequency stability is high. *IAM*

Ref.: Druzhinin (1967), p. 279.

A **traveling-wave-tube (TWT) oscillator** is a generator of microwave oscillations in which a traveling-wave tube is used as the active component. When an O-type traveling-wave tube is used, the generation occurs due to the positive feedback between its input and output. This connection can be made either over a special channel or through internal reflections of the electromagnetic wave at discontinuities of the coil, chiefly at points of the transmission from the spiral waveguide structure to the waveguide or coaxial transmission line, through which the signal is sent to the device or taken from it to the load.

Traveling-wave-tube oscillators of the O type with low power are widely used as wideband voltage-controlled oscillators (VCOs) and also for local oscillators and RF frequency-agility oscillators. Variants of such higher power tubes for radar transmitters have not been developed due to the requirement for precise stabilization of the high voltage and the lack of advantages over tubes of other types. Traveling-wave tubes are more widely used in transmitters as amplifiers of oscillations of the master oscillator. (See **AMPLIFER, traveling-wave-tube.**) *IAM*

Ref.: Popov (1980), p. 82; Perevezentsev (1981), p. 165.

A **tube oscillator** is one based on an electronic tube. The main uses of various types of tube oscillators are shown in Table O2. *IAM*

Ref.: Popov (1980), p. 206.

A **tunnel diode oscillator** consists of an active component (the diode), and LC-circuit, formed by the capacitive reactance of the diode and an external inductance coil, for example in the form of a shortened section of coaxial line with length less than a quarter wave length. Oscillation arises with compensation of circuit losses by the negative resistance of the diode. To increase frequency stability, a high-quality coaxial cavity is selected with a low temperature coefficient of expansion. In the microstrip version of the design, a dielectric cavity is used as the resonator.

Table O2
Types of Tube Oscillators

Type of Oscillators	Basic Use
klystron	local oscillator for radar in cm or mm bands
magnetron	power generator for radar in dm, cm, or mm bands
amplitron (stabilitron)	generator for medium- and high-power radar in dm, cm, or mm bands with high frequency stability
backward-wave-tube	generators for low- and medium-power radars in cm and mm bands with electronic frequency tuning
traveling-wave-tube	wideband generators for medium power in cm and mm bands
gyrotron	powerful radar oscillators in mm band
grid-control tube	generator for high-power radar in dm and cm bands

The low level of generated power (around 1 mW), is a significant drawback of tunnel-diode oscillators, so they have not become widely used. These oscillators may operate at frequencies up to several tens of GHz and can be used as local oscillators, for example. *IAM*

Ref.: Gassanov (1988), p. 180.

A **voltage-controlled oscillator (VCO)** is one in which the frequency varies with an applied voltage. This device may be employed to form frequency-modulated waveforms. The characteristics of some common VCOs are given in Table O3.

Table O3
Characteristics of VCO Devices

Device	Center-frequency range (MHz)	Maximum frequency deviation as percent of center frequency	Maximum linearity as percent of deviation	Maximum center-frequency stability	Comments
LC oscillator	Up to 50	±15%	±0.5%	±10 to ±100 ppm	
Crystal oscillator	0.1 to 300	±0.25%	±1%	±1 to ±10 ppm	
Three-terminal gallium arsenide oscillator	60 to 2500	±2%	±2%	±1%	
Voltage-tunable magnetron	100 to 10,000	±50%	±1%	±2%	Requires anode-voltage-control range of 750 to 3000V
Backward-wave oscillator	2,000 to 18,000	±20%	±0.3%*	±0.2%	Requires helix-voltage-control range of 400 to 1500V

* Deviation from an exponential frequency-versus-voltage curve.

(from Skolnik, 1990, Table 10.3, p. 10.16, reprinted by permission of McGraw-Hill).

The backward-wave oscillator has an exponential frequency-versus-voltage characteristic, and all others have a linear characteristic.*SAL*

Ref.: IEEE (1990), p. 30; Skolnik (1990), p. 10.17.

P

PARAMETRIC ECHO EFFECT. The parametric echo effect is the phenomenon whereby the amplitude of a radio wave is stored in ferromagnetic samples by the accumulation of energy in resonant elements of the magnetized sample, which are formed on the basis of the magnetic moments of the nuclear magnetization. The resonant frequencies of the elements of the magnetized sample, which differ as a result of the heterogeneity of the magnetic field, cover a specific band. Within the limits of this band it is possible to process nonsinusoidal (linearly frequency modulated) oscillations owing to the stored information of the signal oscillations and subsequent reading of the pulses with the same deviation, but with a duration half as long. Using this same process, frequency-modulated pulses can be compressed.

The effect can be used in analog pulse compression, where its merit is the simplicity of its application to random modulation of the signal. *IAM*

Ref.: Shirman (1974) p. 162.

PATTERN, antenna (radiation). The antenna pattern is "a graphical representation of the radiation properties of the antenna as a function of space coordinates." It is usually a plot of power gain (a dimensionless quantity) versus angular coordinates (the power gain pattern), measured at sufficient distance from the antenna that the gain is independent of range (i.e., in the far field, see **ANTENNA radiation regions**). The same plot with a change of scale can represent radiation intensity or power flux density produced at a given range by a particular transmitter radiating through the antenna (power per unit solid angle), or directivity, which differs from power gain only by omission of the radiation efficiency. Alternatively, the voltage gain (dimensionless) or the field strength (V/m) produced by a particular transmitter radiating through the antenna may be plotted, these also having the same curve with a change of scale. The complete antenna pattern also includes the phase, relative to a given reference (e.g., at the beam axis), and the polarization. Since most antennas are reciprocal, the "radiation properties" are identical to the corresponding "receiving properties" of the antenna.

For convenience in plotting, the pattern is often expressed as two separate gain versus angle plots, measured in the principal planes (e.g., azimuth and elevation) cutting through the beam axis (Fig. P1). Other methods of plotting are contour plots (Fig. P2) and three-dimensional plots (Fig. 3).

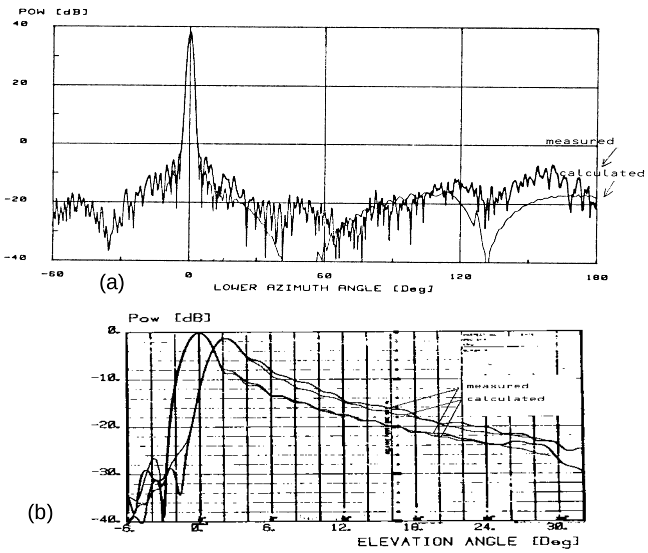


Figure P1 Example of search radar antenna patterns in principal planes: (a) azimuth pattern; (b) elevation pattern.

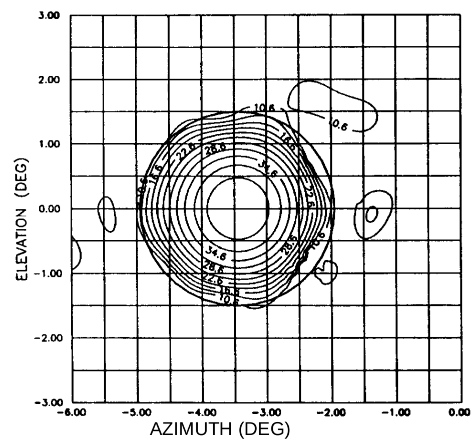


Figure P2 Example of antenna pattern contour plot.

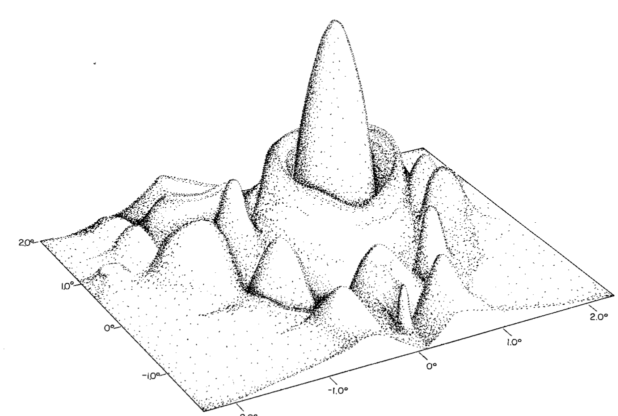


Figure P3 Example of three-dimensional antenna pattern plot (from Barton, 1988, Fig. 4.1.4, p. 147).

Either power gain or voltage gain can be used in plotting antenna patterns, and decibel scales in gain are often used to provide visible data over a dynamic range of 60 dB or greater. A power pattern does not indicate phase (or polarity) of the actual radiation patterns, but a plot of the voltage pattern shows the alternating phase of the principal sidelobes, 180° and 0° relative to the mainlobe. A more complete characterization of the radiated field requires plots of amplitude and phase separately, as functions of angle. *DKB*

Ref.: IEEE (1993), p. 1057; Johnson (1984), p. 1.5.

An **antenna pattern approximation** is a mathematical model approximating the actual pattern with some analytical expression to be used for analysis or synthesis. The most common approximations are the Gaussian approximation and the (sin*x*)/*x* functions, and the pattern of a cosine-illuminated aperture is also used to match a class of practical antennas. The first is an idealized pattern of an antenna having a smooth mainlobe with no sidelobes. The voltage gain pattern is

$$f(\theta) = \exp\left[-1.3863\frac{\theta^2}{\theta_3^2}\right]$$

where θ_3 is the half-power beamwidth. The power gain pattern, $G(\theta) = f^2(\theta)$, has the same form but with the constant 2.7725 in the exponential argument. The Gaussian approximation is used when only the mainlobe of the actual antenna needs to be considered (e.g., in calculation of beamshape loss or spectral spreading of clutter echoes). A true Gaussian pattern would require an infinite aperture with Gaussian illumination function, but approximations may be produced by truncating the aperture. For example, truncation at the –9-dB level of illumination produces a pattern with –21-dB sidelobes.

The (sin*x*)/*x* pattern describes the voltage gain of a uniformly illuminated linear or rectangular aperture:

$$f(\theta) = \frac{\sin\left[\left(\pi\frac{w}{\lambda}\right)\sin\theta\right]}{\left[\left(\pi\frac{w}{\lambda}\right)\sin\theta\right]} = \frac{\sin\left(2.7831\frac{\theta}{\theta_3}\right)}{\left(2.7831\frac{\theta}{\theta_3}\right)}$$

where w is the aperture width, and λ is the wavelength. This model has higher sidelobes than are normally found in radar antennas, but gives the narrowest beamwidth and is sometimes an adequate representation for analysis.

When sidelobes are of concern, but are not as great as those of the (sin*x*)/*x* pattern, the pattern of a cosine-illuminated aperture can be used:

$$f(\theta) = \frac{\cos\left[3.7352\left(\frac{\theta}{\theta_3}\right)\right]}{1 - 5.6544\left(\frac{\theta}{\theta_3}\right)^2}$$

This is the sum pattern illustrated in Fig. P6.

Another type of pattern approximation arises when discrete measured values of the pattern are available, and continuous values need to be found. A cubic approximation is commonly used. For a three-point cubic approximation, the gain $G(\theta)$ within a specified region $\theta_1 < \theta_2 < \theta_3$, for which gains G_1, G_2, G_3 have been measured, can be found as

$$G(\theta) = k_1 G_1 + k_2 G_2 + k_3 G_3$$

where

$$k_1 = \frac{(\theta - \theta_2)(\theta - \theta_3)}{(\theta_1 - \theta_2)(\theta_1 - \theta_3)}$$

$$k_2 = \frac{(\theta - \theta_1)(\theta - \theta_3)}{(\theta_2 - \theta_1)(\theta_2 - \theta_3)}$$

$$k_3 = \frac{(\theta - \theta_2)(\theta - \theta_2)}{(\theta_3 - \theta_1)(\theta_3 - \theta_2)}$$

DKB, SAL

Ref.: Barton (1969), pp. 254, 280–285; (1991), App. F.

The **array (antenna) pattern**, $f_{arr}(\theta,\phi)$, is the product of two factors: the array element pattern, $f_e(\theta,\phi)$, and the array factor, $f_a(\theta,\phi)$. The first is the pattern produced when an individual element in the array is excited, while the second expresses the effect of controlled excitation of all elements of the array. Thus:

$$f_{arr}(\theta, \phi) = f_e(\theta, \phi) \cdot f_a(\theta, \phi)$$

The element pattern is relatively broad (resulting, for example, from a single dipole), and it is the array factor that determines the width and direction of the radiated (or received) beam. For a uniformly spaced linear array it can be shown that, in the single plane θ,

$$f_a(\theta) = \sum_{n=0}^{N-1} A(x_n)\exp\{j[\phi(x_n) + knd\sin\theta]\} \qquad (1)$$

where $A(x_n)\exp(j\phi x_n)$ is a complex amplitude-phase distribution of the illuminating (exciting) field over the aperture (A is the amplitude, ϕ is the phase, $x_n = nd$ is the coordinate of the nth element of the array, Fig. P4), d is the element spacing, and $k = 2\pi/\lambda$ is the wave number.

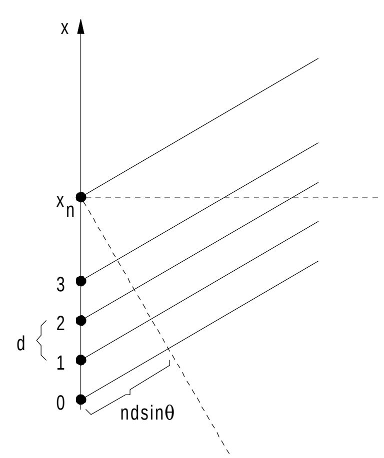


Figure P4 Array elements and coordinates.

Thus the array pattern in angle θ depends on the aperture illumination $A_n\exp(j\phi_n)$, the parameters of the array, n and d, the wavelength λ and the element pattern $f_e(\theta)$. Setting $\psi_0(\theta) = kd\sin(\theta) - \xi$ for the case in which the phase distribution is described by a constant phase shift ξ between adjacent elements, $\phi(x_n) = -n\xi$, (1) becomes

$$f_a(\theta) = \sum_{n=0}^{N-1} A(x_n) \exp[jn\psi_0(\theta)]$$

giving maximum directivity for $\exp(j\psi_0(\theta)] = 1$ at an angle:

$$\theta_m = \operatorname{asin}\left[\frac{\lambda}{d}\left(\frac{\xi}{2\pi} + m\right)\right] \quad , \quad m = 0, 1, 2, \ldots \qquad (2)$$

Equation (2) is the basic formula showing the properties of the array pattern:

(1) The pattern does not have a single maximum, but rather a set of peaks corresponding to $m = 1, 2, \ldots$ (called grating lobes), which must be suppressed by proper design (choice of the element pattern and the ratio d/λ).

(2) There are two means of scanning the array pattern electronically: to change λ (frequency scanning), or the element-to-element phase shift ξ (phase scanning).

For a two-dimensional planar array of $n \times m$ elements having aperture illumination $A_{nm}\exp(j\phi_{nm})$, the array factor is

$$f_a(u, v) = \sum_{n=0}^{N-1} \sum_{m=0}^{M-1} A_{nm} \cdot \exp[j(\phi_{nm} + kx_n u + ky_m v)]$$

where u, v are sine-space coordinates and x_n, y_m are linear coordinates of the nmth element in the array plane. *SAL*

Ref.: Johnson (1984), pp. 20.3–20.17.

The **array element pattern** is "the actual radiation pattern of an element in the array taken in the presence of all other elements and taking into account all mutual coupling effects and mismatches." (See **array (antenna) pattern.**) *SAL*

Ref.: Johnson (1984), p. 20.27.

The **Bickmore-Spellmire pattern**, produced by the Bickmore-Spellmire distribution, is given by

$$E(u) = \Lambda p(\sqrt{u^2 - A^2})$$

where Λ is the lambda function, p and A are the constants of the Bickmore-Spellmire distribution, and

$$u = \frac{\pi D \sin\theta}{\lambda}$$

Here, λ is the wavelength, θ is the angle from the normal to aperture, and D is antenna diameter.

The Bickmore-Spellmire pattern reduces to the Chebyshev pattern when $p = -1/2$. (See **WEIGHTING.**) *SAL*

Ref.: Currie (1987), p. 533.

The **cosecant-squared pattern** is an elevation pattern designed to obtain constant signal level, as a function of range, on target at constant elevation relative to the radar. The mainlobe pattern is conventional, but above the half-power point, at angle θ_1, the mainlobe skirt and sidelobes are replaced by a function (Fig. P5).

$$G(\theta) = G(\theta_1)\left(\frac{\csc\theta}{\csc\theta_1}\right)^2$$

where G is antenna power gain and θ is elevation angle.

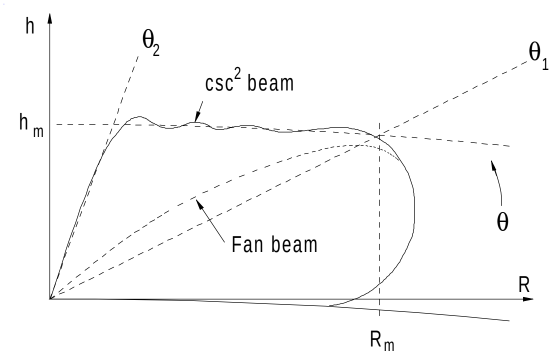


Figure P5 Example of radar coverage using $\csc^2$ antenna (after Barton, 1988, Fig. 4.2.9, p. 163).

The same pattern, inverted, is used in airborne surface-search radar (see **COVERAGE, radar**). *DKB*

Ref.: Skolnik (1980), p. 258; Barton (1988), pp. 26–27.

The **difference pattern** of a monopulse antenna is the error-sensing pattern formed, in a simple feed, by subtracting the signal voltage of the lower horns from that of the upper. In more complex feed systems, the difference pattern is formed by coupling of horns or array elements to produce an illumination function with odd symmetry and with a shape designed to create the steepest possible slope about the center null, for given sidelobe levels.

Figure P6 shows the antenna sum pattern formed using a $\cos x$ illumination function on a rectangular aperture, and the difference pattern representing the first derivative of the sum pattern, formed with an illumination function $x\cos x$. (See also **MONOPULSE.**) *DKB*

Ref.: Barton (1988), pp. 198–205.

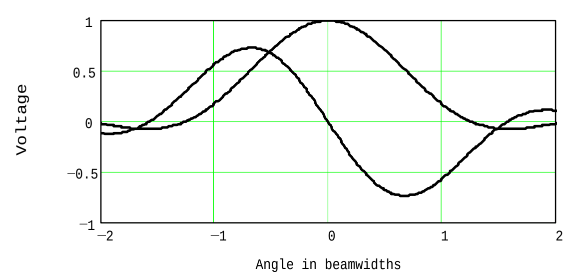


Figure P6 Sum and difference patterns for cosine illumination function.

A **Dolph-Chebyshev pattern** is the optimal antenna pattern created by an aperture illumination of the Dolph-Chebyshev family, having minimum beamwidth for a given sidelobe level. (See **WEIGHTING.**) *DKB*

A **multibeam pattern** is the pattern of an antenna that radiates more than one beam. In stacked-beam, three-dimensional search radar multiple beams are generated in the elevation coordinate. In monopulse radar, multiple beams are generated in both azimuth and elevation, difference patterns being formed in both coordinates to supplement the on-axis sum pattern. *DKB*

An **omnidirectional pattern** is one with constant value of gain over a sphere, or in a single angular coordinate (e.g., azimuth). *SAL*

Ref.: Johnson (1984), p. 1.13.

pattern-propagation factor (see **PROPAGATION**).

The **spillover pattern** is a set of sidelobes, typically at angles 90° to 120° from the mainlobe of a reflector antenna, that result from horn radiation that misses the periphery of the reflector and radiates directly into space. *DKB*

The **sum pattern** of a monopulse antenna is the pattern of the on-axis beam formed, in a simple four-horn feed, by summing all four horns in-phase. In more complex feed systems, the sum beam is formed by coupling the horns or array elements to produce an illumination function with even symmetry and with a shape designed to create the highest possible gain at the beam axis, for given sidelobe levels. (See **difference pattern**; **MONOPULSE**.) *DKB*

Ref.: Barton (1988), pp. 198–205, 399–408.

The **two-way pattern** of an antenna is the product of the transmit pattern and the receive pattern. In general, for transmit beamwidth θ_{3t} and receive beamwidth θ_{3r} at the –3-dB points, the beamwidth of the two-way pattern at the –6-dB level is given by

$$\theta_6 = \frac{\sqrt{2}\theta_{3t}\theta_{3r}}{\sqrt{\theta_{3t}^2 + \theta_{3r}^2}}$$

which can be considered an effective one-way, –3-dB beamwidth for the pattern pair. *DKB*

Ref.: Barton (1993), p. 88.

A **V-beam antenna pattern** is a pattern comprising two fan-shaped beams. One beam is located in the vertical plane, while the other is inclined at an angle of 45° relative to the vertical beam (Fig. P7). Antennas with a V-shaped radiation pattern are used as three-dimensional search radars. Both beams rotate jointly and are fed simultaneously from one or from different transmitters. Meanwhile, a separate receiver is used with each beam. The vertical beam accomplishes the azimuth and range search, with data from both beams used to determine target altitude. Antenna rotation is chosen in such a way that the target will first be traversed by the vertical beam and then by the inclined beam. Target altitude is given by

$$h = \frac{R\sin\theta}{\sqrt{1 + \sin^2\theta}}$$

where R is the slant range to the target and θ is the antenna rotation angle between the center of the target blips in the channels of the vertical and inclined beams. *AIL*

Ref.: Skolnik (1970), p. 22.3.

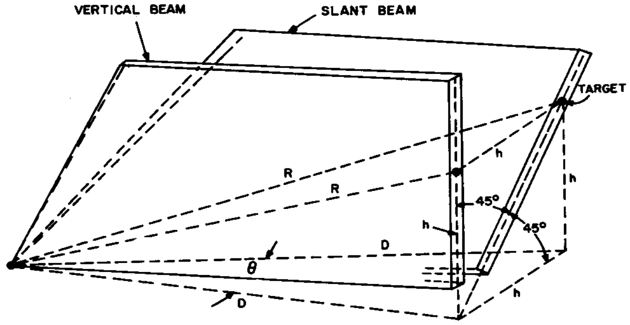


Figure P7 V-shaped radiation pattern (from Skolnik, 1970, Fig. 2, p. 22.3, reprinted by permission of McGraw-Hill).

PERFORMANCE, radar. Radar performance is the set of characteristics defining the quality of radar operation. In Russian literature, radar performance figures are usually referred to as *tactical-technical characteristics*. Tactical characteristics describe top-level performance and typically include (1) maximum and minimum operational ranges (detection range, for search radar); (2) coverage angles; (3) resolution; (4) measurement accuracy (or errors); (5) throughput capacity; (6) interference immunity; and (7) availability. The technical characteristics include lower level parameters: operating frequency, transmitter power, pulse repetition frequency, antenna gain (directivity), receiver sensitivity, and so forth. The radar performance can be evaluated through a combination of analysis, simulation, and subsystem and system testing. The flow of the performance evaluation in the process of design and testing is shown in Fig P8. *SAL*

Ref.: Leonov (1988), p. 23; Barton (1991) p. 13.1.

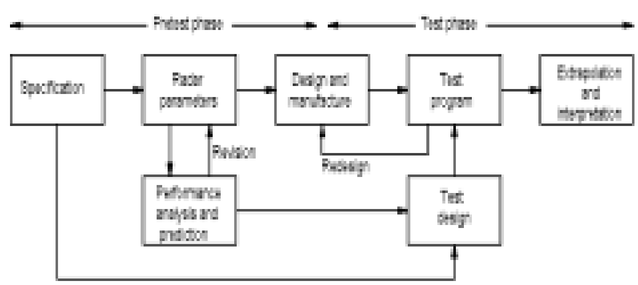


Figure P8 Flow of radar system design and testing program (after Barton, 1991, Fig. 13.1, p. 13.2).

The **radar performance figure** is "the ratio of the pulse power of the radar transmitter to the power of the minimum signal detectable by the receiver." *SAL*

Ref.: IEEE (1993), p. 1,052.

PERMEABILITY. Permeability is "a general term used to express various relationships between magnetic induction and magnetizing force. These relationships are either (1) absolute permeability, that in general is the quotient of a change in magnetic induction divided by the corresponding change in magnetic force; or (2) specific (relative) permeability, which is the ratio of the absolute permeability to the magnetic constant."

The permeability of free space is $m_0 = 1.257 \times 10^{-6}$ H/m, and the relative permeability of a material with permeability μ is $\mu_r = \mu/\mu_0$. *DKB*

Ref.: IEEE (1993), p. 935.

PERMITTIVITY. Permittivity is "the incremental change in electric displacement per unit electric field when the magnitude of the measuring filed is very small compared to the coercive electric field. The small signal relative permittivity, κ is equal to the ratio of the absolute permittivity ε to the permittivity of free space, that is $\kappa = \varepsilon/\varepsilon_0$." The relative permittivity is also called the *relative dielectric constant*, denoted by ε_r. The permittivity of fee space is $\varepsilon_0 = 10^{-9}/36\pi = 8.854 \times 10^{-12}$ F/m. *DKB*

Ref.: IEEE (1993), p. 936.

PHASED ARRAY (see **ARRAY ANTENNA**).

PHASE. Phase is a property of a periodic phenomenon $f(t)$ with a period T, defined at time t as "the fractional part t/T of the period T through which t has advanced relative to an arbitrary origin." In radar applications, the concept of phase is usually applied to oscillation of electromagnetic waves. (See **OSCILLATION, sinusoidal.**) *SAL*

Ref.: IEEE (1993), p. 939.

The **blind phase** is a phenomenon observed in MTI systems when "the echo of interest is in quadrature to the reference signal." It occurs in systems that have only the in-phase channel, and can be eliminated by using in-phase/quadrature (I/Q) channels in the signal processor. *SAL*

Ref.: Schleher (1978), p. 8.

phase front (see **WAVE front**).

PHASE SHIFTER. A phase shifter is "a device in which output voltage (or current) may be adjusted, in use or in its design, to have some desired phase relation with the input voltage (or current)." The operation of phase shifters is usually based on the variation of the electrical length of a transmission line, connection of a reactive element to a transmission line, or vector addition of several signals. Phase shifters can be classified as

(1) Electrical or mechanical, depending on the physical principle of phase control;

(2) Analog (continuous) or digital (discrete), depending on the granularity of control;

(3) Feedthrough or reflex, based on the mode of connection to external circuits; and

(4) Active or passive, depending on whether amplification of the signal is included.

The main types of microwave phase shifters are shown in Table P1

The main requirements on modern phase shifters are to provide the required phase shift with high accuracy (several degrees), to provide high speed (several nanoseconds), to have minimum insertion loss, to provide good matching to the microwave circuits, to have adequate power-handling capacity, and to have stable and repeatable parameters, small size, and light weight. These requirements are met by both diode and ferrite phase shifters. The advantage of diode types is lower weight and size, higher speed, and compatibility with integrated circuit technology and packaging. Ferrite phase shifters provide higher power handling capacity and have lower loss at frequencies above about 2 GHz. Integrated circuit phase shifters and those based on new physical principles (e.g., plasma phase shifters) are promising in certain applications.

Phase shifters are widely used in phased-array radar, and other applications include generation of phase-coded waveforms, monopulse tracking circuits, and others. *SAL*

Ref.: IEEE (1993), p. 944; Skolnik (1970), Ch. 12; Gassanov (1988), p. 146; Lavrov (1974), p. 340; Koul (1991, 1992).

Table P1
Classification of Phase Shifters

Electrical phase shifters	
Ferrite phase shifters:	
(a) Nonreciprocal	Toroidal [twin-slab], strip-line, coaxial- waveguide, helical
(b) Reciprocal	Reggia-Spencer, Faraday rotator
Semiconductor phase shifters	PN diode, PIN diode, integrated circuit
Huggins phase shifter	
Mechanical phase shifters	Fox phase shifter, helical line, waveguide trombone

An **active phase shifter** is one that includes amplification along with phase shifting. Examples are field-effect tetrode and field-effect transistor phase shifters. If the phase shift is provided without amplification, the phase shifter is termed passive. *SAL*

Ref.: Gassanov (1988), p. 146.

A **capacitance phase shifter** is the phase shifter inverting the phase because of variation in capacitance. The most widely used are the semiconductor phase shifters employing the variation of the PN junction capacitance. *IAM*

Ref.: Popov (1980), p. 125; Skolnik (1970), p. 12.55; Sazonov (1988), p. 178.

A **[combined] coaxial-line-waveguide phase shifter** is a nonreciprocal ferrite phase shifter that combines the electrical performance of the toroidal (twin-slab) digital waveguide phase shifter with the compactness of a coaxial transmission line (Fig. P9). *SAL*

Ref.: Skolnik (1970), p. 12-21.

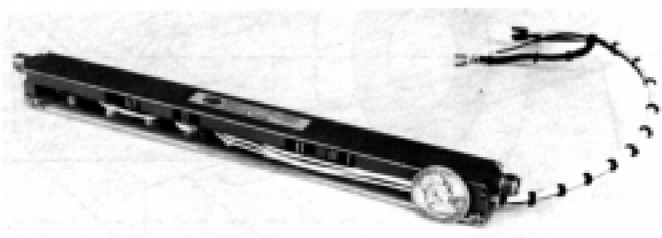


Figure P9 Combined coaxial-line-waveguide phase shifter (from Skolnik, 1970, Fig. 23, p. 12.25, reprinted by permission of McGraw-Hill).

A **continuous phase shifter** is one with a gradual (continuous) phase shift within the required limits when the external exposure is applied. In semiconductor phase shifters, such an exposure is the control voltage, in ferrite phase shifters, it is external magnetic field, in some types of phase shifters it is mechanical force (see **mechanical phase shifter**). Continuous phase shifters can be feed-through or reflex phase shifters, typically based on varactors and field-effect tetrodes. They enable the receptions of the continuous phase inversion from 0 to 360° with a high degree of phase linearity. *IAM*

Ref.: Skolnik (1970), p. 12.41; Gassanov (1983), pp. 146, 153; Voskresenskiy (1981), p. 355

A **diode phase shifter** uses as the control component a semiconductor PN diode (varactor) or PIN diode. Phase shifters based on PN diodes provide a gradual variation of the reactive component of diode resistance in the mode when the PN junction is closed. Typically they are used in low-power continuous phase shifters of the reflex type. A reflex phase shifter based on a single varactor provides phase shift up to 180° with a nonlinearity not worse than 5% in 5-band. PIN diode phase shifters are based on switching of diode states with forward and reverse biases and are used in high-power (up to 50 kW) discrete phase shifters. The switch time in high-voltage phase shifters is about several microseconds, in low-voltage phase shifters 10 to 100 ns. The most widely used discrete feed-through phase shifters provide the phase shift accuracy of 5° to 8° in a 5 to 10% bandwidth with an insertion attenuation of 1.0 to 1.5 dB.

Diode phase shifters have small weight and size and can be done with hybrid IC technology. *IAM*

Ref.: Skolnik (1970), pp. 12.45–12.63; Sazonov (1988), p. 162; Kaganov (1981), p. 75.

A **discrete phase shifter** produces fixed phase shifts that differ from each other by the specified value. Discrete phase shifters are based primarily on PIN diodes, field-effect transistors, and field-effect tetrodes in feed-through or reflex circuits. There are three main approaches to designing these phase shifters: switched line sections, lines loaded with LC-filters, and reflex types with switch elements. The first are typically used in phased arrays, but the band is comparatively narrow. These phase shifters have several sections for different phase shifts (e.g. 22.5, 45, 90, 180° sections). The phase adjustment error up to 16° for a 4-bit phase shifter reduces the possibilities for using programmable control for phase-frequency response correction. Phase shifters based on LC filters use the capabilities of low-pass and high-pass filters for phase inverting. The use of these phase shifters enables the reduction of the losses (up to 0.3 to 0.8 dB) and the production of discrete phase control from 0 to 180°. Field-effect transistors phase shifters have a more stable phase shift in broader band than PIN diode phase shifters and they have a switch time of about an order less. *IAM*

Ref.: Skolnik (1970), p. 12.41;Gassanov (1988), pp. 148, 151; Voskresenskiy (1981), p. 360.

Electrical phase shifters implement phase control by means of semiconductor device electrical control, or as the result of dielectric and magnetic permeability variation when external electrical and magnetic fields are applied to dielectric and magnetic plates. (See **semiconductor phase shifter**, **ferrite phase shifter**, **Fox phase shifter.**). In opposition to mechanical phase shifters, they give the possibility of reaching high phase shift speed: hundreds of megahertz instead of hundreds of hertz for mechanical phase shifters. *IAM*

Ref.: Lavrov (1974), p. 340.

A **Faraday-rotation phase shifter** is a reciprocal phase shifter based on the Faraday effect, sometimes implemented in a phase shifting unit (section of waveguide) with $\lambda/4$ ferrite plate. With such a plate it works with both linear and circular polarized waveforms, the direction of circular polarization does not change and the plane of linear-polarized signal shifts by 90°. To eliminate this effect, two plates with the opposite direction of magnetic fields can be used. *IAM*

Ref.: Skolnik (1970), p. 12-31; Johnson (1984), pp. 20–36, 47.

A **feed-through phase shifter** changes the phase of the signal propagating along a transmission line. Phase shifters of this type are used both for continuous (see **continuous phase shifter**) and discrete (see **discrete phase shifter**) phase controls. *IAM*

Ref.: Sazonov (1988), p. 165, Kaganov (1981), p. 74.

Ferrite phase shifters use the interaction of the electromagnetic field with ferromagnetic materials. (See **FERRITE.)** They are divided into reciprocal and nonreciprocal devices. The former include Reggia-Spencer phase shifters, stripline analog latching phase shifters, and Faraday rotation phase shifters. The latter include twin-slab phase shifters and strip line phase shifters. Nonreciprocal phase shifters have smaller size, insertion loss, and drive power then reciprocal types, but they must be switched between transmit and receive modes in the systems using a single antenna for transmitting and reception. The main advantage of ferrite phase shifters is high power handling capability; the main disadvantages are the dependence of ferrite materials performance on temperature and difficulties in obtaining a high switching rate because the control magnetic systems have large inductances. Typically, ferrite phase shifters produce a phase shift of 0 to 360°, they have phase errors less than ±5° in a 10% frequency band, and insertion loss is about 1 dB. *IAM*

Ref.: Skolnik (1970), pp. 12.1–12.55; Johnson (1984), p. 20.32.

A **field-effect tetrode phase shifter** is a semiconductor phase shifter with active components based on a field-effect tetrode. There are two configurations: analog (continuous) and discrete phase shifters, based on the dependence of phase shift on bias voltage in a gate electrode or on vector summation of several signals. The first have a phase shift up to 180° in a narrow frequency band and gain of more than 8 dB. The latter types realize discrete control in a 0 to 360° range and have a bandwidth of about an octave. The advantages of field-effect tetrodes are low losses, a high speed of phase variation that is beneficial for the formation of microwave phase-coded waveforms and using in broadband phased arrays. *IAM*

Ref.: Gassanov (1988), p. 151.

A **field-effect transistor (FET) phase shifter** is a semiconductor phase shifter based on a field-effect transistor as the main active component. Typically, it is used as a discrete phase shifter, having a configuration analogous to the diode phase shifter, and incorporates additional matching circuits because of the large shunting reactive resistance of the interelectrode area. The loss at X-band is about 5 to 11 dB. To compensate for manufacturing differences, frequency and temperature dependence of phase shifter performance, combined discrete-analog phase shifters with programmable control are used. The control is implemented by the gradual variation of the transistor bias voltage (0 to 3V). Field-effect transistor phase shifters are characterized by a low switch time and a wide bandwidth and are good for use in broadband phased arrays. *IAM*

Ref.: Ayasli, Y., et al., *IEEE Trans **MTT-32**,* No. 12, 1984.

A **fixed phase shifter** provides a constant phase shift. Typically, this type of phase shifter is based on a delay line. It is possible to realize any desired phase shift and there are practically no limitations for input signal dynamic range (more than 100 dB). For frequencies of 0.3 to 40 GHz, striplines, microstrip lines, and coaxial lines are used, while for frequencies more than 40 GHz, waveguides are used. The accuracy of phase shifting in the range of 0 to 360° is several percent and less. *IAM*

Ref.: Sokolov (1984), p. 122.

A **Fox phase shifter** is a mechanical waveguide phase shifter with mechanical phase shift control by rotation of one section of transmission line relative to another one. It is used primarily in measurement equipment. *IAM*

Ref.: Lavrov (1974), p. 344; Gardiol (1984), p. 253.

A **helical phase shifter** is a nonreciprocal ferrite phase shifter consisting of shielded helix wound on a ferrite toroid for compactness at frequencies below the S-band. The unit can be either analog or digital. A typical 4-bit digital latching phase shifter at S-band has an insertion loss of 1.2 dB, peak-power handling capability about 10 kW and provides 360° of phase shift. *IAM*

Ref.: Skolnik (1970), p. 12.21.

A **Huggins phase shifter** is one using a frequency change technique to introduce the required phase shift, with the subsequent conversion back to the input frequency. It consists of two mixers and a delay line (Fig. P10). The signal at frequency f_0 is mixed with a signal from the variable frequency oscillator f_c in the first mixer. The phase-shifted signal from the output of delay line and the signal from the first mixer are mixed in the following stage to obtain the resultant signal with the same frequency f_0 and phase shift $\phi = 2\pi f_c \tau$. *SAL*

Ref.: Skolnik (1980), p. 303.

An **integrated(-circuit) phase shifter** is a semiconductor phase shifter in the form of an integrated circuit. Structurally,

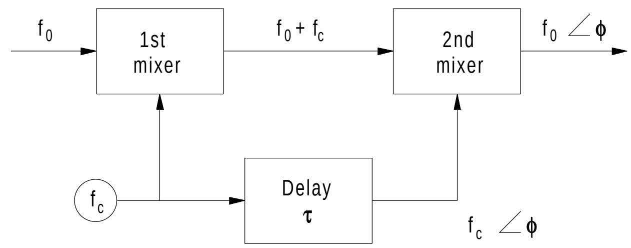


Figure P10 Schematic diagram of the Huggins phase shifter (after Skolnik, 1980, Fig. 8.19, p. 304).

integrated phase shifters are made with one type of microstrip line, most often asymmetrical strip, but in modern devices with different types of lines, slotted, coplanar, and so forth, that form a phase shifter in the form of a three-dimensional integrated circuit.

Most common are circuits of reflex phase shifters with parallel connection to an asymmetrical stripline of serial diode circuits: varactor diode and correcting inductance coil in the form of a loop. A 360° phase shifter contains two identical serial diode circuits connected to a line at a distance of a quarter-wavelength from one another.

A phase shifter based on a combination of different lines, for example an asymmetrical strip and symmetrical slot, has diodes connected, respectively, to an aperture of the strip connector and a short-circuited slot line located orthogonally to it on the opposite side of the substrate.

These reflex phase shifters in combination with quadrature directional couplers create the component base of integrated transfer phase shifters.

Integrated phase shifters are marked by their small dimensions, and are used in radar control devices, and in signal processing devices where high power-handling capacity is not required. *IAM*

Ref.: Gvozdev, V. I., *Radiotekhnika*, no. 4, 1991, p. 33, in Russian.

A **latching phase shifter** is a reciprocal ferrite phase shifter in which the magnetic state of the ferrite is set with a current pulse, leaving the remanent magnetization constant until reset by another current pulse. This mode of operation eliminates the constant drive dissipation between phase settings, and the possibility of undesired phase modulation resulting from current fluctuation during transmission and signal reception. *DKB*

Ref.: Skolnik (1970), p. 12.11.

A **mechanical phase shifter** is based on the variation of electrical length of transmission line by changing its geometrical length or cross section or by inserting in a waveguide dielectric or ferromagnetic plates (see **waveguide phase shifter**). An example of a mechanical phase shifter is the Fox phase shifter. Mechanical phase shifters have low phase shifting speed (hundreds of hertz) and are not used in modern radars. *IAM*

Ref.: Skolnik (1962), p. 308; Sazonov (1988), p. 152.

A **multielement phase shifter** is a discrete phase shifter providing phase variation in a specified range with the specified discrete $\Delta\phi$. Actually, it is multidigit phase shifter, each digit of which can be in either of two states: the phase shift of the ith digit $\Delta\phi_i$ is present or absent. The number of digits n is defined via specified $\Delta\phi$ as

$$n = 1 + \log_2(\pi/\Delta\phi)$$

Multielement phase shifters are realized through diode phase shifters, field-effect transistor phase shifters, and other electrical phase shifters. They are widely used in phased arrays (number of digits is typically 5 or 6). *IAM*

Ref.: Voskresenskiy (1981), p. 369.

A **piezoceramic phase shifter** is an electromechanical waveguide phase shifter based on the reverse piezoelectric effect. It consists of a waveguide section, a phase shifting element made as a springy metal plate close to the narrow wall of the waveguide connected to the piezoceramic element, and a short circuit element. The phase shifter operation is based on the metal plate motion as the piezoceramic element bends under the exposure of control voltage. This is a continuous phase shifter, with a phase shift up to 315°, active losses less than 0.5 dB, and a switch time of 20 μs. The advantages of this phase shifter is the linear dependence of the phase shift, the capability for initial phase setting mechanical adjustment, and the capability of incorporating in the structure of phased array. *IAM*

Ref.: Skalzhakin, A. I., *Radiotekhnika*, no. 6, 1991, p. 76 (in Russian).

A **plasma phase shifter** is one using plasma in the waveguide so the electromagnetic waves going through this section get a phase shift depending on plasma density. This density can be controlled by static electrical and magnetic fields. *IAM*

Ref.: Popov (1980), p. 290.

A **reflex phase shifter** operates on the principle of the phase variation of a reflected signal when the length of a transmission line is switched. To separate coming and reflected waves, the microwave bridges and circulators are used. The reflex section of circulators of a diode phase shifter incorporates a varactor, the short circuit section of transmission line shorter than $\lambda/4$ providing bias to the diode and the compensation of its spurious reactance parameters, and the elements for decoupling of microwave and supply circuits. The main advantage of the reflex phase shifter is that the voltage standing wave ratio is less than for the transmission phase shifter. *IAM*

Ref.: Gassanov (1988), p. 146.

A **Reggia-Spencer phase shifter** is a reciprocal ferrite phase shifter consisting of a bar of ferromagnetic material located axially within a section of waveguide (Fig. P11). The longitudinal magnetic field produced by a solenoid wound around waveguide causes variation in the permeability of material and variation in propagation constant of the energy that results in a phase shift that can be controlled by a driving current. Operating frequencies are 8 to 70 GHz, the advantage is simplicity and phase control implemented up to 360° with comparatively weak control field intensities, the disadvantage is worse switching time and switching power in comparison to latching ferrite phase shifters. *IAM*

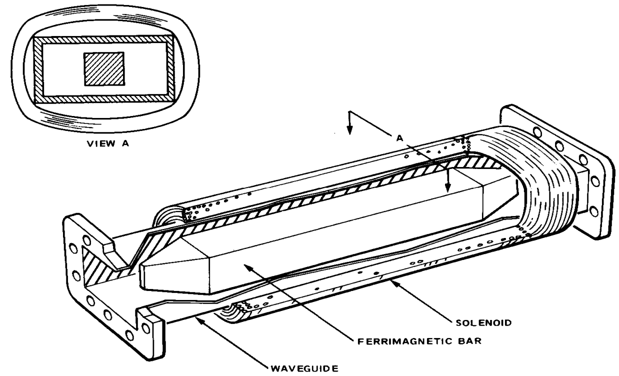


Figure P11 Typical Reggia-Spencer phase-shifter configuration (from Skolnik, 1970, Fig. 27, p. 12.27, reprinted by permission of McGraw-Hill).

Ref.: Skolnik (1970), pp. 12.23–12.29.

Semiconductor phase shifters use semiconductor devices as control components. Typically, they are divided into diode phase shifters, field-effect transistor phase shifters, and field-effect tetrode phase shifters. When the phase shift is large, diode phase shifters have considerable insertion loss, nonlinear dependence of phase shift on control voltage, and a comparatively large switch time. Transistor and tetrode-based phase shifters have an order better speed of phase shifting (a few nanoseconds or less) and lower losses. Semiconductor phase shifters are the main type of phase shifters used in high-speed phase control units in modern radars. *IAM*

Ref.: Skolnik (1970) pp. 12.45–12.63; Sazonov (1988) p. 178.

A **strip-(transmission-)line phase shifter** is based on variation of the electrical length of a strip transmission line. Typically, these phase shifters use ferrite dielectric inserted into the transmission line. Toroid cores from ferrite or garnet materials are used. To diminish the size of the phase shifter, a slow-wave structure (e.g., a meander line located in the center slot of a ferromagnetic toroid) can be used. Strip-transmission-line phase shifters operate in frequency bands of 0.18 to 7 GHz, they have a bandwidth of 8 to 10%, and an insertion loss less than 2 dB. In comparison to waveguide phase shifters, they have lower size and weight, but lower power handling capacity. *IAM*

Ref.: Skolnik (1970), p. 12.20; Sokolov (1984), p. 124.

A **switched-line phase shifter** is a discrete phase shifter of the feed-through typed based on variation of microwave path by diode or transistor switches. (See **diode phase shifter**, **field-effect transistor phase shifter.**) This type of phase shifter is simple and the insertion loss has a weak dependence on phase shift that is beneficial for large values of phase shift. However, the upper frequency of these phase shifters is limited by the resonant phenomena due to the capacitance of switching diode or transistor. *IAM*

Ref.: Voskresenskiy (1981), p. 363; Kaganov (1981), p. 79; Fink (1982), p. 25.64.

A **toroidal phase shifter** is a nonreciprocal ferrite phase shifter with the transverse magnetic field, using one or several sections of ferrite material having toroidal geometry. It can be implemented as an analog or digital unit, typically are based on a waveguide or strip transmission line. Switching time is about 0.5 to 2 ms, frequency band is 10%, insertion losses are 0.8 to 1.2 dB, and average power handling capability is 0.2 to 0.4 kW. *AIL*

Ref.: Skolnik (1970), p. 12.15; Sazonov (1988), p. 178.

A **transverse magnetic field phase shifter** is a nonreciprocal ferrite phase shifter providing phase shifting as a result of magnetic permeability variation in ferrite with a transverse magnetic field created by electromagnets. This phase shifter is of the nonreciprocal type. When it uses rectangular waveguide, two ferrite plates are typically inserted to increase the phase shift. A coaxial variant with an area between ferrite plates partially filled by dielectric is also used. Continuous phase shifters of this type require a permanent current supply to the control winding. Discrete phase shifters do not have this disadvantage. (See **toroidal phase shifter.**) *IAM*

Ref.: Sazonov (1988), p. 177.

twin-slab phase shifter (see **toroidal phase shifter**).

A **waveguide phase shifter** is based on the variation of electrical length of a waveguide. According to the type of control used they are divided into mechanical and electrical phase shifters. Mechanical waveguide phase shifters have different modes of electrical length variation: the variation of geometrical length, the use of a dielectric plate inserted parallel to the electrical field through a nonradiating slot, or the use of an elastic section of waveguide made from a springy material. Typically, waveguide phase shifters are used in the band of 6 to 35 GHz. *IAM*

Ref.: Skolnik (1962), p. 308; Sazonov (1988), p. 152; Lavrov (1974), p. 340.

PHASOR. A phasor “a complex number expressing the magnitude and phase of a time-varying quantity.” It is used to describe a signal vector, relative to some arbitrary reference phase (e.g., the phase of the reference oscillator, or COHO, in an MTI system). *DKB*

Ref.: IEEE (1993), p. 946.

PLATFORM. In radar terminology, the platform refers to the vehicle on which the radar is mounted.

Platform motion effects are the effects introduced in radar performance by a movable platform, which may also be unstable and deformed (such as ship, aircraft, spacecraft, or etc.). This effect results in additional measurement errors (see **ERROR, platform dependent**) and other effects such as velocity spread across the antenna beamwidth in airborne MTI), and others. (See **SLANT-RANGE EFFECT.**) *SAL*

Ref.: Skolnik (1990), pp. 16.3-16.8; Leonov (1990), pp. 174-203.

A **PLATINOTRON** is a microwave tube that combines the properties of a magnetron and a backward-wave tube with a transverse magnetic field (M-type). In contrast to the magnetron, the platinotron, like the backward-wave tube, has an open structure and operates with the inverse spatial harmonic. An amplifying or oscillating platinotron is called an amplitron or a stabilitron, respectively.

A platinotron is used in decimeter and centimeter wavebands basically in output stages of pulsed-radar transmitters. *IAM*

Ref.: Skolnik (1962), p. 231; Popov (1980), p. 291.

PLUMBING. “A colloquial expression for pipelike waveguide circuit elements and transmission lines.”

Ref.: IEEE (1993), p. 963.

POINCARÉ SPHERE. “A sphere whose points are associated in a one-to-one fashion to all possible polarization states of a plane wave [field vector] according to the following rules: The longitude equals twice the tilt angle, and the latitude is twice the angle whose cotangent is the negative of the axial ratio of the polarization ellipse.” The north pole represents left-hand circular polarization, the south pole right-handed circular, and diametrically opposite points represent orthogonal polarizations (Fig. P12). If the angular distance between two points is 2ξ, then the polarization efficiency (the fraction of power transferred between antennas having polarizations represented by the points) is

$$\eta_p = \cos 2\xi$$

SAL

Ref.: IEEE (1993), p. 963; Johnson (1984), p. 1.9.

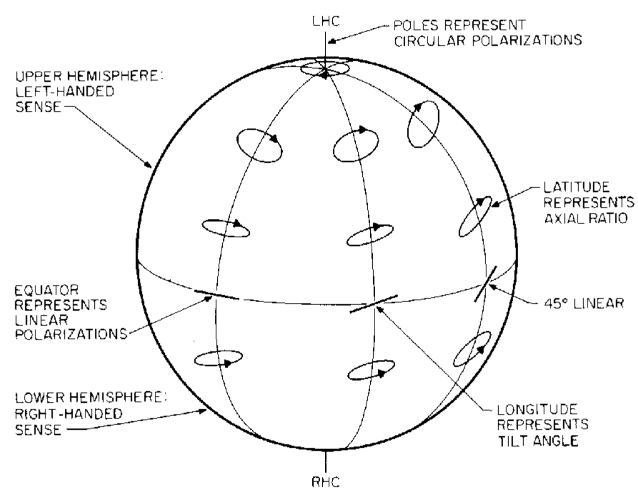


Figure P12 Polarization states on the Poincaré sphere (from Johnson, 1984, Fig. 1.4, p. 1.9, reprinted by permission of McGraw-Hill).

POLARIZATION. In the context of radio-wave propagation, pointing is “the locus of the tip of the electric field vector observed in a plane orthogonal to the wave normal.” All polarizations are special cases of elliptical polarization, which may be divided into right-hand and left-hand. Special cases are circular and linear polarization, and special cases of this last are horizontal and vertical polarizations. Complete information about a backscattered electromagnetic signal (radar return) is represented by polarization scattering matrix. (See **MATRIX.**) Techniques using all the information content of the polarization scattering matrix are called *matrix tech-*

niques; those requiring dual-polarization or polarization agility and utilizing only part of scattering matrix content are called *vector techniques*; and under the assumption of a constant polarization state the techniques are called *scalar.* The effect of polarization may be used in target detection, discrimination (or classification), and suppression of interference (e.g. echoes from weather disturbances). A convenient method of representing polarization states is the Poincaré sphere. *SAL*

Ref.: IEEE (1993), p. 967; Fink (1975), p. 18.14; Johnson (1984), pp. 1.7, 23.1; Skolnik (1980), p. 227; Currie (1987), p. 158.

Polarization agility is the technique using variation of polarization, in which the transmitted polarization is switched on a pulse-to-pulse or batch-to-batch basis. *SAL*

Ref.: Currie (1987), p. 158; Leonov (1988), p. 156; Stutzman (1992).

The **polarization basis** is the set of two orthogonal unit vectors into which the vector of electromagnetic field intensity can be expanded. *IAM*

Ref.: Kanareikin (1966), p. 30.

polarization calibration (see **CALIBRATION, polarization**).

A **polarization chart** is the chart graphically representing the ratio of the magnitudes of the vertically and horizontally polarized waves and the phase angle between them. The phase angle is shown as the relative phase of the vertical element when the relative phase of the horizontal element is zero (Fig. P13). For example, if an ellipticity (a figure reciprocal to axial ratio, see **elliptical polarization**) is 3 dB and polarization orientation is 37°, one draws a line from the center of the chart to the periphery at the 37° point and lays off a distance from the chart center to the value V/H = 3 dB, measured along the horizontal. One can read V/H = 0.8 dB and a phase angle of 71° for the right-hand component and 289° for the left-hand component. *SAL*

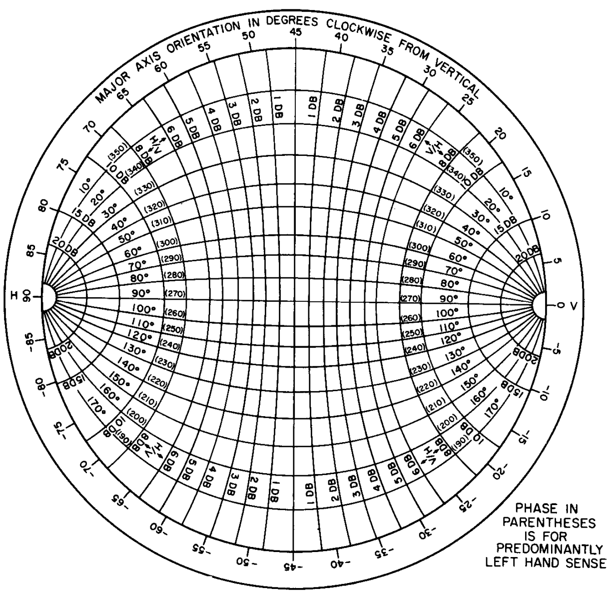


Figure P13 Polarization chart (from Johnson, 1984, Fig. 23-5, p. 23-6, reprinted by permission of McGraw-Hill).

Ref.: Johnson (1984), p. 23.6.

Circular polarization is a special case of elliptical polarization corresponding to the case when electric field vector traces out a circle (Fig. P14). Depending on the direction of rotation of the vector circular polarization is right-hand or left-hand polarization. Circular polarization is more difficult to generate than linear polarization, but it is often more desirable for radars operating in jamming or clutter environment (for example, in radars which must see through weather disturbances). *SAL*

Ref.: Skolnik (1980), p. 227; Lothes (1990), p. 122.

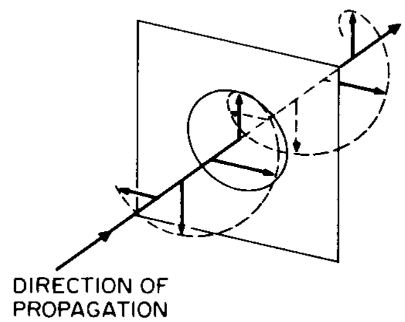


Figure P14 Circular polarization (right-hand) (from Johnson, 1984, Fig. 23-1, p. 23.1, reprinted by permission of McGraw-Hill).

Polarization coding is a method of pulse compression in which subpulses are transmitted and received with either of two orthogonal polarizations, rather than with 180° phase shifts. This technique is more complex than conventional phase coding, but it offers the possibility of improving radar jamming immunity (an example is the so-called Intrapulse Polarization Agile Radar). *SAL*

Ref. Eaves (1987), pp. 490–492.

copolarization (see **desired polarization**).

Cross-polarization is the undesired component, orthogonal to the desired polarization for which an antenna has been designed. *SAL*

Ref.: Lothes (1990), p. 123.

The **desired polarization** is "the polarization of radio wave for which an antenna system is designed." This is also called copolarization. *SAL*

Ref.: IEEE (1990), p. 23.

The **polarization distortion matrix** is a four-by-four complex matrix that contains all of the distortion terms injected into a measured target polarization matrix by a measurement system. The polarization scattering matrix actually measured by radar is the product of the polarization distortion matrix and actual target matrix. *SAL*

Ref.: Currie (1987), p. 768; Currie (1989), p. 129.

Polarization efficiency is "the ratio of the power received by an antenna from a given plane wave of arbitrary polarization to the power that would be received by the same antenna from a plane wave of the same power density and direction of propagation, whose state of polarization has been adjusted for a maximum received power." *SAL*

Ref.: IEEE (1993), p. 967.

The **polarization ellipse** is an elliptical path along which the electrical field vector of a polarized wave moves in the plane orthogonal to the wave normal. *SAL*

Ref.: Johnson (1984), p. 1.7.

Elliptical polarization is the general case in which the electric field vector traces out an ellipse (Fig. P15). Elliptical polarization is characterized by an axial ratio, which is the ratio of the major axis to the minor axis of the polarization ellipse (sometimes the term ellipticity, which is reciprocal to axial ratio is used), the tilt of the axis of the polarization ellipse, polarization orientation (the direction in which the major axis lies), and the direction of rotation of the E-vector. Linear and circular polarizations are special cases of elliptical polarization. *SAL*

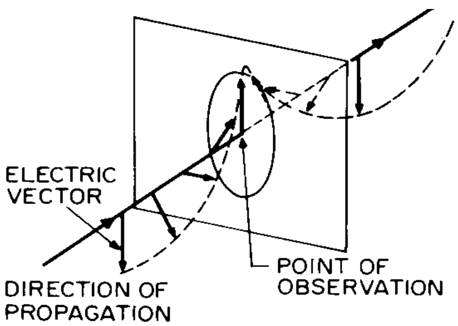


Figure P15 Elliptical polarization (from Johnson, 1984, Fig. 23-1, p. 23-1, reprinted by permission of McGraw-Hill).

Ref.: IEEE (1993), p. 432; Johnson (1984), p. 23.2; Skolnik (1980), p. 227; Lothes (1990), p. 122.

polarization factor (see **LOSS, polarization**).

Polarization isolation is the measure of separation between two channels in a dual-polarized radar. Contamination of the receiver channel designed to receive one polarization (e.g., vertical) by another polarization (e.g., horizontal) can occur in the antenna, transmitter or receiver. The main reasons for contamination are: imperfections in setting of the desired polarization in the transmitting antenna; imperfection in the reference reflector returning the signal; improper operation of the orthomode transducer separating co- and cross-polarizations; and presence of a leakage path between the receiver channels. Responses of different reference targets that can be used for measuring polarization isolation to various illumination polarizations are given in the Table P2. *SAL*

Ref.: Currie (1989), p. 124

Linear polarization is the special case of elliptical polarization in which the electrical field vector at all times lies along a fixed line (Fig. P16). Linear polarization can have any orientation angle, special cases being horizontal and vertical polarizations. *SAL*

Ref.: Skolnik (1980), p. 227; Johnson (1984), p. 23.1.

Table P2
Reflected Polarizations for Various Reflectors

Reflector	Illuminated Polarization					
	V	H	+45°	–45°	Circular RH	Circular LH
Flat plate	V	H	+45°	–45°	LH	RH
Sphere	V	H	+45°	–45°	LH	RH
Trihedral	V	H	+45°	–45°	LH	RH
Dihedral (V)	V	H	–45°	+45°	RH	LH
Dihedral (45°)	H	V	+45°	–45°	RH	LH
Dihedral (22.5°)	+45°	–45°	V	H	RH	LH

(after Currie, 1984, Table 4-1, p. 128)

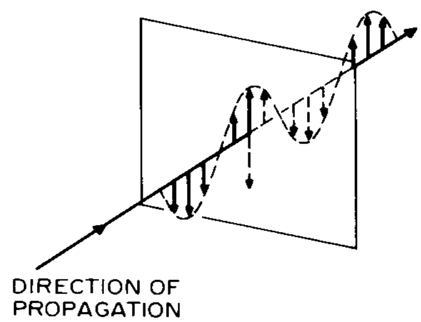


Figure P16 Linear polarization (from Johnson, 1984, Fig. 23-1, p. 23-1, reprinted by permission of McGraw-Hill).

A **polarization meter** is the device to measure the polarization parameters of an electromagnetic wave. The structure of a polarization meter is defined by the methods of polarization parameter measurement. The simplest polarization meter is a linear polarized antenna rotating around its axis, a receiver and output signal indicator (e.g. oscilloscope). Waveguide sections with ferrite, $\lambda/4$ or $\lambda/2$ dielectric plates can be used As polarization analyzers to extract a wave with a specified polarization. *IAM*

Ref.: Kanareikin (1966), p. 107.

Null polarizations are any pair of orthogonal polarization states of the target polarization scattering matrix such that when this matrix operates on these polarizations the target echo is nulled. *SAL*

Ref.: Currie (1987), p. 303.

Polarization parameter measurement is the method of measuring different sets of polarization parameters (the geometrical parameters of polarization ellipse, the modulus and argument of the phasor, Stoke's parameters). The main methods of measurement are shown in Table P3. The first two are used to measure the parameters of full-polarized waves, while the last three can be also applied also to partially-polarized waves (having a polarization component of random origin).*IAM*

Ref.: Kanareikin (1966), p. 107.

Table P3
Methods of Polarization Parameter Measurement

Name	Essence	Measured parameters	Range of application
Polarization diagram method	The dependence of field intensity at the detector output as a function of angle of rotation around the axis (wave propagation direction) for linear polarized wave is measured	Polarization ellipse parameters	Fully polarized waves, the direction of field vector rotation is not measured
Compensation method	The compensating differences in phases and amplitudes of orthogonally linear polarized components are measured	Phasor modulus and argument, polarization ellipse parameters	Fully polarized waves, the measurements are done in succession
Method of expanding the wave into orthogonally polarized components	The amplitudes of orthogonal components and phase shift between them are measured	Parameters of polarization ellipse	Fully and partially polarized waves
Method of several antennas	The voltages at the matched loads of four antennae with known polarization are measured	Stoke's parameters	Fully and partially polarized waves
Modulation method	The phase lag is introduced in one plane of propagation and the parameters of polarization are measured simultaneously	Stoke's parameters	Fully and partially polarized waves

The **polarization plane** is the plane orthogonal to the direction of propagation and containing the electric and magnetic fields. *SAL*

Ref.: Johnson (1984), p. 1.7.

Polarization purity is the ratio of the principal polarization response to the cross-polarization response of an antenna, often expressed in decibels. *SAL*

Ref.: Lothes (1990), p. 123.

The **polarization ratio** is ratio of the power in the received signal with copolarization to the power of the received signal with cross-polarization. *SAL*

Ref.: Currie (1987), p. 810.

The **polarization ratio discriminant** is a technique using frequency-agile waveforms to distinguish the desired target from clutter or other targets on the basis of its polarization ratio. Its operation is based design of the receiving antenna such that the polarization ratio is larger for the echo from a cell containing a desired target, and smaller when the return arrives from cells containing undesired targets or clutter only. The polarization ratio discriminant is a vector discrimination technique. *SAL*

Ref.: Currie (1987), p. 285.

The **polarization response** is the polarization received by antenna in response to a signal transmitted with preset polarization. The desired response is termed *desired polarization*, *copolarization*, *principal polarization*, or *antenna's design polarization*. The undesired orthogonal response is called *cross-polarization*. *SAL*

Ref.: Lothes (1990), p. 123.

polarization rotation (see **FARADAY rotation effect**).

polarization scattering matrix (see **MATRIX**).

POLARIZER. A polarizer is a structure that forms or responds to the polarization of an electromagnetic wave. Typical polarizer configurations are of transmission or reflection types. The transmission type uses an anisotropic medium with the anisotropy adjusted to achieve phase quadrature for two waves whose (linear) polarization vectors are mutually orthogonal. To produce such anisotropy the structure can use (1) parallel metal plates, (2) parallel dielectric plates, or (3) lattices of strips or rods. These structures are bulky and inconvenient to adjust but are inexpensive and have high power-handling capacity. The reflection type polarizer is a half-length transmission type polarizer mounted against a conducting sheet. Polarizers are typically used to convert a linear polarization to circular or vice versa. *SAL*

Ref.: Johnson (1984), pp. 23–25.

POLYPLEXER. "Equipment combining the functions of duplexing and lobe switching." *SAL*

Ref.: IEEE (1993), p. 974.

POTENTIAL, radar. Radar potential is the generic radar parameter describing the maximum detection range R_0 of a radar operating against the target with specified RCS σ_t. To combine all radar-related terms in a radar range equation and call it radar potential Q_r, the formula for signal-to-noise ratio is a very simple representation:

$$\frac{E_1}{N_0} = \frac{Q_r \sigma_t}{R_0^4} = D_x$$

Thus, maximum detection range R_0 depends on RCS σ_t, the detectability factor D_x (the required energy ratio, specified through probability of detection and false alarm), and radar potential incorporating all radar-related parameters defining the quality of detection (peak power, antenna gain, losses, etc.). The term is widely used in Russian radar literature. *SAL*

Ref.: Leonov (1988), p. 28.

POWER. Power, in the broad, physical sense, is the rate of doing work, dW/dt. The international standard (SI), or mks unit of power for electrical systems is the watt (W), equivalent to the expenditure of one joule of energy per second. One joule = 10^7 ergs in Gaussian units. The term power is used in several different ways in radar; this entry will deal with those most common. *PCH*

Power-aperture product is the product of a radar's average transmitter power and the effective aperture area (i.e., power-aperture product = $P_{av}A_r$). The antenna effective aperture area is equal to

$$A_r = A\eta_a = \frac{G_r\lambda^2}{4\pi}$$

where A is the physical area, η_a is the aperture efficiency, G_r is the antenna gain, and λ is the radar wavelength. That the detection range of a radar is directly proportional to the power-aperture product can be seen in the search radar equation:

$$R^4{}_m = \frac{P_{av}A_r t_s \sigma F^4}{4\pi k T_s D_0(1) L_s}$$

where t_s is the radar frame time, σ is the target's RCS, k is Boltzmann's constant, D_0 (1) is the detectability factor (or required SNR) for a single pulse, and L_s is the total search loss. (See **APERTURE, antenna**; **RANGE EQUATION**.) *PCH*

Ref.: Barton (1988), pp. 12–26.

Available power is (1) the maximum output power at the transmitting antenna terminals, after all losses between the output of the transmitter and the antenna input terminals have been incurred. Typical losses include ohmic or dissipative losses such as line loss, which attenuates the signal as it travels through the waveguide, and insertion loss when the transmitted signal passes through a circulator or T/R switch designed to protect the radar receiver from excess power leakage during transmission. In some cases the actual power at the antenna input terminals may not be equal to the maximum available power (e.g., when a three-dimensional radar uses power management to allocate transmitter power as a function elevation beam position). (See also **LOSS, transmission-line.**)

(2) Available power may also refer to the average level of electrical prime power, in watts available at the output of an electrical battery, generator/inverter, or line source used to supply the original source of electrical power. (See **primary power.**)

(3) From a signal generator (e.g., as used to determine the properties of a receiver), or from any linear network, the available power is the maximum power that can be taken from the network by a suitably adjusted load. *PCH*

Average power in radar is the average transmitted power P_{av} over the pulse repetition interval (PRI) or some longer period. If P_t is the peak pulse power, τ the pulse width, and f_r the pulse repetition frequency (PRF), then $P_{av} = P_t \tau f_r$. Since the PRF is the reciprocal of the PRI t_r, $P_{av} = P_t(\tau/t_r)$. The quantity τ/t_r is called the *radar duty factor* D_u. For a CW radar, the peak and average power are identical. *PCH*

Center-line power in the frequency spectrum of a waveform is the power in the central line, or carrier frequency, f_0. If the signal amplitude at frequency f_0 is a, the center-line power is equal to $a^2/2$. *PCH*

Power combining refers to adding the outputs of separate power sources, especially as related to radar radiated power. RF power combining of two or more radar transmitter tubes to meet radar power requirements beyond the capability of any single tube has been practiced since the 1950s. With the development of solid-state transmitters, power combining has become widely practiced and necessary, since the individual transmitters are, relative to RF tubes, low-power devices. In addition to achieving the total radar power required, combining the outputs of several RF transmitters has the advantage of increasing total radar system reliability. When large numbers of solid-state transmitters are combined, as in an active array radar, the additional benefit of "graceful degradation" may be claimed, in that the random failure of a few transmitter modules will have little effect on the performance of the overall array or of the radar in general.

In an alternate approach to the combination of complete transmitter modules in an active array, the outputs of individual solid-state power generating devices, such as gallium-arsenide (GaAs) IMPATT diodes, may be combined within a single resonant cavity called the combiner, which then serves as the power module of a central transmitter for a radar system using a conventional (passive) antenna. To date, consideration of this approach to solid-state power combining has been restricted to airborne applications having severe power-aperture limitations, such as radar missile seekers. (See **TRANSMITTER, solid-state.**) *PCH*

Ref.: Ostroff (1985), Ch. 5.

Power density is the distribution of power over a surface area, W/m^2. For an isotropic radiator with power P_t, the power density at range R from the source will be $I_s = P_t/4\pi R^2$. If a transmitting antenna with power gain G_t is used in place of the isotropic antenna, the power density along the axis of the beam now will be $P_tG_t/4\pi R^2$. A spherical target of radius a will have a projected area $\sigma = \pi a^2$. If the target is on the radar beam axis at range R, the target will intercept a fraction of the radiated radar power given by $P_i = I_s\sigma$. This power will be reradiated isotropically by the spherical target and the power density of the return signal at the radar I_r will be $P_tG_t\sigma/(4\pi R^2)^2$. The signal after capture by a radar antenna with effective aperture A_r will have the power:

$$S = I_r A_r = \frac{P_t G_t A_r \sigma}{(4\pi R)^2}$$

PCH

Ref.: Johnson (1984), p. 1.6; Barton (1988), p. 10.

Effective radiated power (ERP) is the product of radar transmitted power and antenna gain, P_tG_t, in watts. Here, radar power is generally expressed in terms of peak pulse power; for a CW radar, the values of peak and average power are identical. Similarly, the ERP of a jammer is P_jG_j and the jammer effective radiated power density (ERPD) is P_jG_j/B_j, where B_j is the bandwidth over which the jammer signal is transmitted. *PCH*

Ref.: Schleher (1986), p. 422; Eaves (1987), p. 691.

Power efficiency may be calculated by dividing the average power output of a device or system, by the average power required at the input. For a complete radar system, the total power efficiency would be measured by dividing the average transmitted RF power, P_{av}, by the primary power supply input (watts/watt). *Power-added efficiency* is a term sometimes used to describe the efficiency of a particular subsystem or component (e.g., a solid-state transmitter amplifier or a transceiver module). *PCH*

Power factor is the average real power drawn, in watts, by a electrical load, divided by the rms voltage-amperes drawn by the same load.

Ref.: IEEE (1993), p. 987.

Power fluctuations, as related to the primary power supply of a radar, are variations in line voltage with time due to transients (voltage spikes), ripple effects, or load changes. Control of power fluctuations is especially important to maintain proper transmitter operation. In tube-type transmitters, regulators are used to ensure that the output of the radar high-voltage power supply (HVPS) is constant within a few percent; the type of regulator used depends on the requirements of the particular transmitter power tube. Solid-state transmitters also require control of power fluctuations, and special electronic conditioning circuits are usually designed into the transceiver module to regulate the power, which is at much lower voltage than that used in tube-type transmitters. *PCH*

Ref.: Skolnik (1990), p. 5.16; McQuiddy, David N., et al., "Transmit/Receive Module Technology for X-Band Active Array Radar," *Proc. IEEE* **79**, no. 3, Mar. 1991.

Power gain is "in a given direction and at a given point in the far-field, the ratio of the power flux per unit area from an antenna to the power flux per unit area from an isotropic radiator at a specified location with the same power input as the specified antenna." It is a measure of the ability of an antenna to concentrate energy in a preferred direction. Numerically, power gain G (generally referred to simply as the *antenna gain*), may be calculated as 4π times the maximum power radiated/unit solid angle divided by the net power accepted by the antenna. Equivalently, power gain is equal to the maximum radiation intensity from the subject antenna divided by the radiation intensity from a lossless isotropic source with the same power input. Gain is related to the effective aperture A_r by $G = 4\pi A_r/\lambda^2$. The power gain of an antenna is always less than the directive gain, since it includes antenna dissipative losses. It is the power gain G that should be used in the radar range equation. (See **ANTENNA gain**; **APERTURE, effective**; **RANGE EQUATION.**) *PCH*

Ref.: IEEE (1993), p. 988; Skolnik (1980), p. 225.

Power handling capability [capacity] describes the maximum power that can be carried by a transmission line or other RF component in the transmitter path. This value is determined by the dimensions of the component, spacing of components subject to high field strengths, and presence of atmospheric pressure or pressurized gases.

Instantaneous power, when a voltage $E = E_m \sin \omega_t$ is applied to a resistance R, is equal to EI, where $I = E_m \sin \omega t/R = I_m \sin \omega t$. E is the instantaneous voltage, E_m is the maximum amplitude of a sine wave of frequency ω, and I is the instantaneous current. Since $EI = E_mI_m \sin^2 \omega t$, and by substitution,

$$(\sin \omega t)^2 = \frac{1}{2} - \frac{1}{2}\cos 2\omega t$$

the instantaneous power is also equal to

$$\frac{E_mI_m}{2} - \frac{E_mI_m}{2}\cos 2\omega t$$

The average value of this power can be seen to be $E_mI_m/2$, and in radar engineering this value, equal to one-half of the maximum instantaneous power, is called the peak power P_t of a radar transmitted pulse. *PCH*

Ref.: Skolnik (1980), p. 52; Hovanessian (1984).

noise power (see **NOISE**).

peak power (see **instantaneous power**).

Primary [prime] power is the source of electrical power used to operate a radar. For land-based civil radars and some fixed-site military radar systems, prime power is provided by the commercial power grid (land lines); in airborne radar systems, prime power is usually provided by electrical generators driven by the aircraft engines. Field army radars are generally powered by portable diesel generators. Radar missile seekers have short operational lives, and in this case, electrical batteries serve as the source of prime power. *PCH*

Power programming is the variation of radar transmitted power in accordance with different radar mode requirements. For example, an agile-beam phased-array radar may use a "burnthrough" waveform against noise jammers that uses higher power (and longer dwell times) than the normal volume search waveform; it may also program a lower average power waveform on targets already under track than to targets being transitioned from search to track mode. Some types of transmitter tubes, such as the crossed-field amplifier (CFA), are more amenable to stepped programming than others, and solid-state active array radars have an inherent capability for transmitted power programming via computer control of individual transceiver modules. Power programming can be implemented via a software template, wherein the radar resources are allocated by radar operational cycle, or it may be adaptively implemented using more sophisticated algorithms that respond to the sensed environment. *PCH*

power spectrum (see **SPECTRUM**).

POYNTING('S) VECTOR. The Poynting vector of an electromagnetic wave is "the vector product of the instantaneous electric and magnetic field vectors":

$$\vec{P} = \vec{E} \cdot \vec{H}$$

It gives the power density of antenna radiation in the far-field region and is the density of power flow at a specified point in W/m^2. In the single-frequency case, the complex Poynting vector is used:

$$\overrightarrow{P_c} = \frac{1}{2}\vec{E} \cdot \vec{H}^*$$

where $\vec{H}^*$ is the conjugate to the complex vector $\vec{H}$. *SAL*

Ref.: ITT (1978), p. 45; Fink (1982), p. 1.41.

PRECISION is the measure of ability of a system to operate with a preset level of errors. In radar the term *precision* primarily refers to the random spread of measurement errors. (See **ERROR, measurement.**) *SAL*

Ref.:Barton (1964), p. 235; Leonov (1988), p. 25.

PRECLASSIFICATION is the stage of target classification in which targets of insufficient interest for further analysis by classification algorithms are excluded from further consideration. This process often is based on clustering techniques. In this case the proximity of an incoming signature to all signatures that are under consideration is measured, and if the incoming signature is sufficiently close to this cluster (in accordance with preset threshold) it passes for the further processing; if not, it is excluded from further processing. An example is deletion of the light decoys accompanying a ballistic missile warhead at the stage of preclassification, when only heavy decoys and warheads pass for ultimate classification with identification algorithms. Sometimes this technique is called alien separation. (See **TARGET RECOGNITION AND IDENTIFICATION.**) *SAL*

Ref.: Long (1992), p. 461.

PRINCIPLE

The **principle of uncertainty in radar** is defined by the fact that the volume covered by ambiguity functions is constant and does not depend on the shape of the radar waveform. By varying the type of the waveform, one can change only the shape of the ambiguity function diagram. For a fixed time-bandwidth product, it is possible to increase the resolution and accuracy of the range measurement only at the expense of decreasing the resolution and accuracy of the doppler velocity measurement, and vice versa. But by increasing the time-bandwidth product, it is possible to improve the quality of measurement both in range (time delay) and velocity (frequency). The term has been introduced by analogy with the uncertainty principle in quantum mechanics that states that the position and the velocity of atomic particles cannot be simultaneously determined to any degree of accuracy desired. *IAM*

Ref.: Skolnik (1980), p. 408; Vakman (1965), p. 5.

PROBABILITY (see **DETECTION probability**, **FALSE-ALARM probability**).

PROCESS, random (see **FUNCTION, RANDOM**).

PROPAGATION, wave. Wave propagation is "the travel of waves through or along the medium." In radar applications, the most important media are the atmosphere and ferrites. (See **propagation medium.**) Atmospheric propagation is the primary consideration for radar and is divided into free-space, tropospheric, and ionospheric propagation, with special consideration to propagation over Earth. In this last case the characteristics of free-space propagation are modified by proximity to the surface, introducing multipath (reflection) and diffraction effects, accounted for, in the range equation, by the pattern-propagation factor. The basic effects associated with wave propagation are interference, scattering, (atmospheric) attenuation, refraction, and diffraction, all of which are discussed below.

Wave propagation is dependent on frequency. There are four different types of waves, depending on the propagation mode:

(1) Direct or freely propagating waves, traveling along straight (or almost straight) lines, characteristic of bands from UHF to millimeter waves.

(2) Tropospheric waves that propagate within a waveguide-like tropospheric structure (ducting), characteristic of VHF and higher frequencies.

(3) Ground waves that follow the curvature of the Earth over some limited range due to the effect of diffraction (VHF band).

(4) Ionospheric waves, propagating as a result of single or multiple reflections from ionospheric layers (HF band). The latter two waves constitute the basis for over-the-horizon radar. *SAL*

Ref.: IEEE (1993), p. 1017; Burroughs (1979); Skolnik (1980), Ch. 12; Blake (1980), Ch. 6; Barton (1988), Ch. 6; Rohan (1991).

Anomalous [abnormal, nonstandard] propagation occurs when the state of the atmosphere departs from its normal or standard condition. The main cause of such irregularities is a change in the index of refraction profile, leading to superrefraction, ducting [trapping], or subrefraction. (See **ATMOSPHERIC ducting.**) *SAL*

Ref.: Skolnik (1980), p. 450; Blake (1980), p. 224.

Free-space propagation occurs in a vacuum in the absence of proximity to a surface or other medium that can modify propagation conditions. Departure from free-space conditions can take the form of (1) refraction, (2) atmospheric attenuation, or (3) scattering of waves from the underlying surface. (See **pattern-propagation factor**, **propagation over the earth.**) *SAL*

Ref.: Skolnik (1980), p. 441.

Ionospheric propagation is characterized by the strong dependence of propagation effects on radar frequency. Decimeter, centimeter, and millimeter waves are practically not scattered by the ionosphere, but some attenuation is introduced. When the wavelength is 5 to 10m or less, the ionosphere serves as a reflecting medium that makes it possible to implement over-the-horizon detection of targets. (See **RADAR, over-the-horizon.**) The main effects of the ionosphere are inversely proportional to the square of the frequency. For a frequency of 300 MHz, the maximum variation in range is about 300m, and in elevation (due to refraction) 0.5 mrad, that is much less than for propagation in the troposphere. The maximum polarization rotation for quasilongitudal propagation (Faraday rotation effect) is about 13 rad (the variation in polarization structure due to quasitransversal propagation are negligible). For wideband waveforms there is an effect of signal distortion (pulse dispersion, or "blurring") that can be neglected for relative bandwidths of $\Delta f \leq 1.5\ f_0^2$ where f_0 is the carrier frequency in gigahertz. For 300 MHz, the attenuation in the ionosphere is about 1.3×10^{-5} dB/km. In a general case for wavelength less than 5 to 7 cm, ionospheric effects are negligible. *IAM*

Ref.: Davies (1965); Kravtsov (1983), pp. 65, 110.

The **propagation medium** is the portion of space through which the propagating wave passes. Typically, a propagation medium is characterized by electric and magnetic permeability (ε, μ), $(\vec{P})$ and $(\vec{M})$. Propagation media are classified as:

(1) Conductors or dielectrics depending on how the current density depends upon the external electrical field.

(2) Isotropic, anisotropic, gyrotropic, linear, or nonlinear), depending on how the parameters, ε, μ, $(\vec{M})$, $(\vec{P})$ depend upon the characteristics of external field.

(3) Homogenous or nonhomogeneous, depending on variability of μ, ε and specific conductivity σ.

The classification of propagation media is given in Table P4.

Table P4
Media Classification

Type	Characteristic	Comments
Homogeneous	Constant ε, μ, σ	
Nonhomogeneous	At least one of the parameters ε, μ, σ is not constant	Stratified media, most of the natural media
Isotopic	Scalar ε, μ, σ	
Anisotropic	Tensor ε, μ, σ: The medium properties depend on the direction of external field	Crystal materials, gyrotropic media
Gyrotropic	Resonant dependence of ε, μ on the frequency of external field	Magnetized plasma, ferrites
Linear	ε, μ, σ do not depend on the intensity of external field	
Nonlinear	ε, μ, σ depend on the intensity of the external field	Ferromagnets, segnetoelectrics
Conductors	$\sigma/(\omega\varepsilon_0\varepsilon) >> 1$	
Dielectrics	$\sigma/(\omega\varepsilon_0\varepsilon) << 1$	

For radar applications, the most important considerations are wave propagation in the atmosphere (see **propagation in the troposphere**, **ionospheric propagation**) and propagation in a ferrite medium whose properties are used in different microwave devices: phase shifters, attenuators, and so forth. Typically, the propagation medium is characterized by the losses it inserts in the propagating wave, that can lie from fractions of a decibel to hundreds of decibels (e.g., for interplanetary medium in radar astronomy). *IAM*

Ref.: Nikol'skiy (1969), p. 35.

Propagation over the earth results in scattering and diffraction of waves by the surface. This is described as multipath propagation within the radar line of sight, and diffraction propagation beyond line of sight, the latter producing significant loss in radar signal strength for most frequencies of interest. Multipath propagation results from the interference between the direct wave and one reflected from the surface, leading to a lobed structure in the radar coverage (Fig. P17).

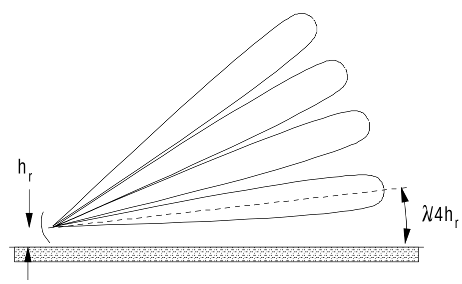


Figure P17 Vertical lobing caused by surface reflec-

Two levels of approximation are used to evaluate multipath propagation effects: flat-Earth and spherical-Earth. In the first (Fig. P18), the reflected ray can be assumed to originate in an image antenna or image target, located below the real surface. This geometry then yields values for the target elevation angle, the depression angle to the reflected ray (equal to the grazing angle at the surface), and the pathlength difference between the two rays. These are used, along with the elevation pattern of the antenna and the reflection coefficient of the surface, to calculate the pattern-propagation factor.

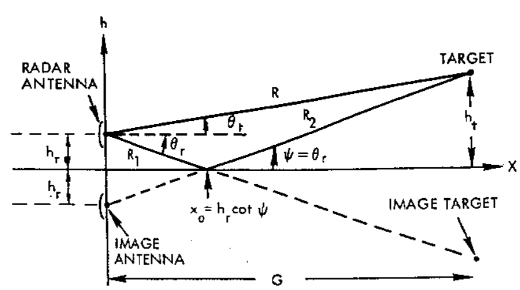


Figure P18 Multipath geometry over flat Earth.

The angles and pathlength difference may be corrected as needed for longer paths, using spherical-earth geometry. In that case, three propagation regions may be distinguished (Fig. P19):

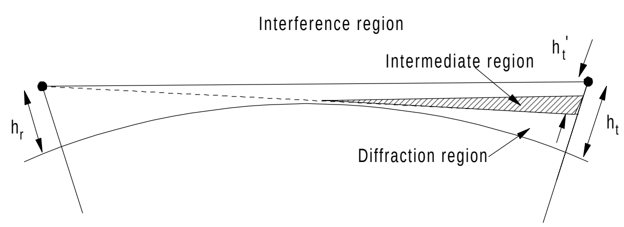


Figure P19 Three regions for propagation over the Earth.

(1) The interference (optical) region, for which $h_t' > 0$ and pathlength difference $\delta_0 > \lambda/8$.

(2) The intermediate (or transition) region, $h_t' > 0$ and $\delta_0 < \lambda/8$.

(3) The diffraction region, $h_t' < 0$ and $\delta_0 < \lambda/8$.

In region (1) the principles of ray optics are applicable and coverage has a lobing structure; in (2) and (3) ray optics cannot be applied. In region (2) the lower edge of the first reflection lobe must be modified by a diffraction propagation factor, while in (3) a pure diffraction field exists. *SAL*

Ref.: Skolnik (1980), p. 442; Barton (1988), pp. 288–302.

Over-[beyond-]the-horizon (OTH) propagation carries the electromagnetic wave into the diffraction region (beyond the direct line of sight). Either of two phenomena may be exploited:

(1) Ground (surface) waves, typically in the VHF band.

(2) Ionospheric bounce, in the HF band.

VHF ground waves follow Earth's curvature, but the ranges are not great (usually less than a few hundred km). Radars relying on this phenomenon are called ground-wave OTH radars. Ionospheric bounce provides ranges of thousands of kilometers, and under some circumstances the waves circle the entire Earth. However, this propagation mode suffers from several effects that reduce the accuracy and reliability of the resulting radar data:

(1) Ionospheric electron density varies from day to night, seasonally, and with sunspot activity, causing the ionospheric layers to vary in height and introducing physical and geometrical limitations on reflected paths.

(2) Because the reflected rays reach the earth from above, clutter reflections from both land and sea surfaces are large.

(3) Calculation of the pattern-propagation factor is complicated by the ionospheric phenomena involved, in addition to the usual surface effects, and this factor is often well below unity (unless some focusing effect is present in the ionosphere).

(4) Significant sources of atmospheric interference exist, such as aurora borealis, which can distort the propagation medium. *SAL*

Ref.: Blake (1980), p. 425; Kolosov (1987), Ch. 2.

The **pattern-propagation factor** is "the ratio of the signal strength that is actually present at a point in space to that which would have been present if free-space propagation had occurred with the antenna beam directed toward the point in question." The factor denotes a one-way voltage ratio, and hence appears in the radar equation as F^4, with range in direct proportion:

$$R = R_0 F$$

where R_0 is the free-space range. (See **CHART, Blake**; **RANGE EQUATION**.) The usual notation is F, sometimes modified to F_t or F_r to distinguish the transmit path from the receive path ($F^4 = F_t^2 F_r^2$). Atmospheric attenuation is excluded from F and appears as a separate loss in the range equation.

It is sometimes convenient to separate the effects of the antenna from those of the propagation path, and a propagation factor is then defined as the value of the pattern-propagation factor that would have been obtained with a broad-beam antenna, such that the underlying surface and the target are both fully illuminated. In the transition region (see **propagation over the earth**), F can be calculated as the weighted sum of a reflection factor F_R and a diffraction factor F_D.

The main effects of radar coverage modification by F are the introduction of lobing in the interference region, descending through the intermediate region into a deep propagation null in diffraction region. In the interference region the pattern-propagation factor can be written as a function of target elevation angle $F(\theta_t)$:

$$F_R(\theta_t) = f(\theta_t)\left[1 + \frac{f(\theta_r)}{f(\theta_t)} D\rho\exp(-j\alpha)\right]$$

where $f(\theta_t)$ is the antenna voltage pattern at the target angle, $f(\theta_r)$ is the voltage pattern at the angle of the reflected ray, D is the divergence factor of the surface, ρ is the magnitude of the reflection coefficient, α is the phase of the reflected ray, and the term in brackets is the propagation factor. More detailed formulas and analytical approximations used to compute the pattern-propagation factor can be found in Barton (1993), and more rigorous procedures are discussed in Meeks (1982). *SAL*

Ref.: IEEE (1993), p. 923; Blake (1980), p. 239; Meeks (1982); Barton (1988), pp. 288–302, (1993), p. 145.

Propagation in the troposphere is determined by the dielectric constant of the air (average value near the Earth surface $\varepsilon = 1.00075$), which decreases with height and by the presence of random irregularities with different values of ε. The regular component of ε defines the effects of the wave path variation introducing the errors in range measurement about 90m maximum, angle measurement due to refraction 30 arc-min maximum. The regular effects in the troposphere do not depend on radar frequency. The random troposphere irregularities cause the fluctuation of regular correction in a path variation with root-mean-square error about 0.15m, the fluctuation of wave front causing errors in angle measurement, the signal amplitude fluctuation (fading) and refraction fluctuations including the appearance of abnormally large angles (the effect of far troposphere propagation). In the bands of centimeter waves and especially millimeter waves there are considerable effects of wave attenuation by hydrometeors and molecules of gases O_2 and H_2O, which has a strong frequency-dependent character. (See **ATTENUATION by clear air.**) *IAM*

Ref.:Bean (1968); Kravtsov (1983), pp. 83, 118, 185.

PULSE. A pulse is "a wave that departs from a first nominal state, attains a second nominal state, and ultimately returns to the first nominal state." It is the basic waveform used in radar applications, where the first and final state has zero amplitude and the second state has a nominally constant amplitude. The ideal rectangular pulse is shown as a dashed line in Fig. P20, with a typical real pulse as a solid line, characterized by a leading edge with a rise time t_r, a trailing edge of duration t_f, a pulse droop from its initial value, and a ripple superimposed on the top of the pulse.

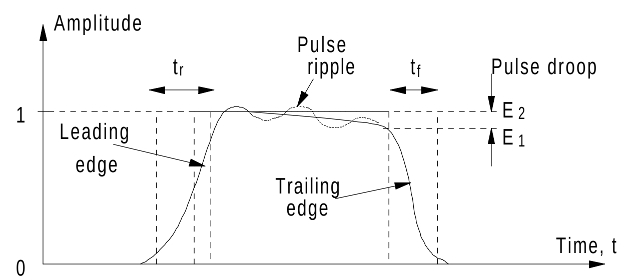


Figure P20 Actual and ideal pulse shapes.

The pulse rise time is "the interval between the instants at which the instantaneous value reaches specified lower and upper limits, namely, 10 and 90% of the peak pulse value unless otherwise specified." The pulse fall time t_f is defined similarly. Pulse rise and falls times are important parameters of pulse shape. A pulse with a fast rise time and a near-ideal rectangular shape is necessary to realize the inherent range resolution of the pulse. A slow rise time makes accurate measurement of time-of-arrival difficult and also wastes transmitter energy. Pulse rise time requirements also depend on the type of transmitter tube used.

The slow decay of the top of the pulse from its initial value is called pulse droop, *D*:

$$D = \frac{E_1 - E_2}{E_1}$$

There may also be oscillations on the top of the pulse, identified in Fig. P20 as *pulse ripple.*

The duration of the pulse is commonly called *pulse width,* normally defined between the half-power points of the leading and trailing edges. Pulse width is also called *pulse length* or *pulse duration.*

Other than rectangular pulse shapes are possible, but these normally are produced by pulse compression or other filtering in the receiver, rather than transmitted, because radar transmitters operate most efficiently in the saturated mode, producing constant output between the leading and trailing edges. The major pulse shapes and their corresponding spectra are shown in Table P5.

A single pulse is seldom used in radar. Typically, a sequence of pulses, termed a *pulse train* or *pulse burst,* is transmitted into each beam position (Fig. P21).

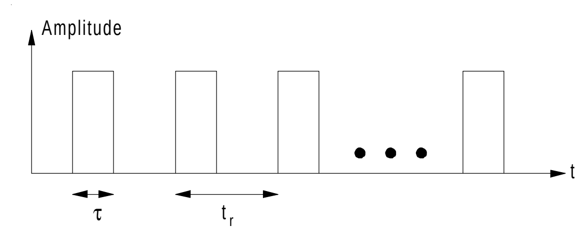


Figure P21 Pulse train.

The pulse train is characterized by the pulse width τ and the pulse repetition interval (PRI) t_r, the latter being the reciprocal of the pulse repetition rate (PRF), f_r. Since any discrete approximation of a continuous time-domain function (sampling of a single pulse or transmission of a pulse train) results in a periodic spectrum, the spectrum of a coherent pulse train is periodic at $f_r = 1/t_r$, with an envelope determined by the spectrum of the individual pulse in the train (Fig. P22). The width of the spectral lines, is the reciprocal of the duration of the train t_o and Fig. P22 is drawn for essentially infinite duration. *PCH, SAL*

Ref.: IEEE (1993), pp. 1033, 1035; Skolnik (1980), p. 432; Brookner (1988), p. 136.

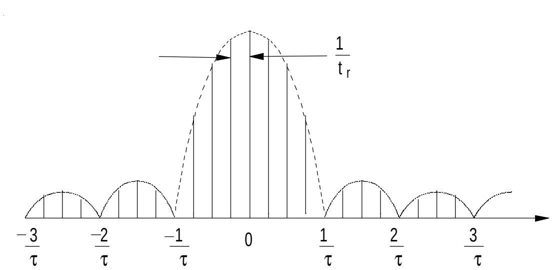


Figure P22 Spectrum of coherent pulse train.

A **coded pulse** is a pulse employing intrapulse modulation used for pulse compression or identification. (See **WAVEFORM.**) The corresponding device "for varying one or more of the characteristics of a pulse or of a pulse train so as to transmit information" is called a pulse coder, and the processing device in the receiver is a pulse decoder. *SAL*

Ref.: IEEE (1993), p. 202.

A **pulse-forming line (network)** is "a passive electric network in a radar modulator whose propagation delay determines the length of the modulation pulse." *SAL*

Ref.: IEEE (1990), p. 23.

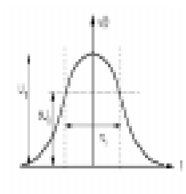
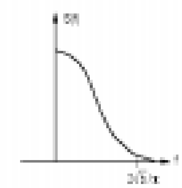
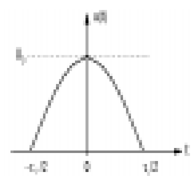
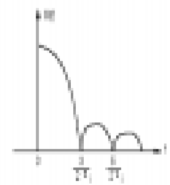
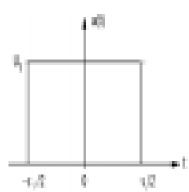
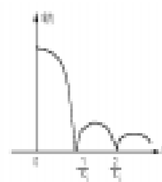
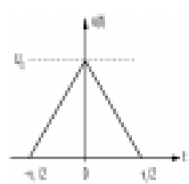
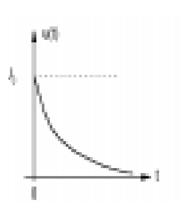
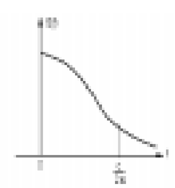
pulse integration (see **INTEGRATION**).

pulse length (see **PULSE**).

A **pulse packet** is "the volume of space occupied by a single radar pulse. The dimensions of this volume are determined by the angular width of the beam, the duration of the pulse, and the distance from antenna." In Russian literature, the term pulse packet refers to a train or burst of pulses. *SAL*

Ref.: IEEE (1993), p. 1,031.

Table P5
Pulse Shapes and Spectra

Pulse description	Pulse shape	Spectrum shape	Spectral width	Time-bandwidth product
Gaussian: $u(t) = U_0 \exp(-a_1 t^2)$, $-\infty < t < \infty$ $\sqrt{a_1} = \frac{4 \ln(1/b)}{\tau_i}$			$0.26\sqrt{\beta_1}$	0.426
Cosine: $u(t) = U_0 \cos(\pi t/\tau_i)$, $\lvert t\rvert < \tau_i/2$, $u(t) = 0$, $\lvert t\rvert > \tau_i/2$			$\frac{0.73}{\tau_i}$	0.43
Rectangular: $u(t) = U_0$, $\lvert t\rvert < \tau_i/2$, $u(t) = 0$, $\lvert t\rvert > \tau_i/2$			$\frac{0.81}{\tau_i}$	0.73
Triangular: $u(t) = \frac{2U_0}{\tau_i}\left(\frac{\tau_i}{2} - \lvert t\rvert\right)$, $\lvert t\rvert < \tau_i/2$, $u(t) = 0$, $\lvert t\rvert > \tau_i/2$			$\frac{0.84}{\tau_i}$	0.46
Exponential: $u(t) = U_0 \exp(-ct)$, $\lvert t\rvert < \tau_i/2$, $u(t) = 0$, $\lvert t\rvert > \tau_i/2$			$0.98c$	1.13

Predetected pulse is the term sometimes used for video-detected pulses employed to make a decision in target detection. *SAL*

Ref.: Wehner (1987), p. 69.

pulse repetition frequency (see **PULSE REPETITION FREQUENCY**).

An **uncoded [simple] pulse** is an amplitude modulated waveform with carrier frequency f_0 without intrapulse modulation. Typically, this waveform has a rectangular envelope (Fig. P23). The time-bandwidth product for a simple pulse is equal to unity. Usually, long and short pulses are distinguished: the first corresponds to the narrowband waveform (approaching CW) and has good resolution in doppler and poor resolution in range; the second is a wideband waveform having good range but poor doppler resolution. Simple pulses were widely used in old-generation radars as magnetrons primarily used in those facilities are not easily compatible with pulse-compression waveforms. Now it is used primarily in

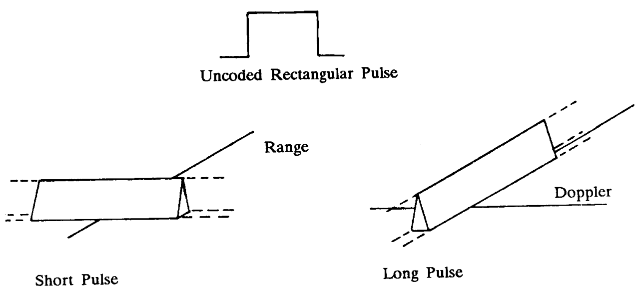


Figure P23 Simple rectangular pulse (from Bogush, 1989, Fig. 3.54, p. 193).

inexpensive radars when signal generation and signal processing cost must be minimized, and when the requirements to have sufficient energy for detection and tolerable range accuracy can be met simultaneously. Sometimes simple pulses are called uncoded pulses or Class A waveforms. *SAL*

Ref.: Bogush (1989), p. 192.

PULSE COMPRESSION is "the processing of a wideband, coded signal pulse, of initially long time duration and low-range resolution, to result in an output pulse of time duration corresponding to the reciprocal of the bandwidth and, hence, higher range resolution, and with approximately the same pulse energy." In principle, the process of pulse compression is the by-effect when the signal with intrapulse modulation is processed in the matched filter to maximize signal-to-noise ratio. Since the matched filter does not preserve the initial shape of the waveform at its output, but on the contrary, distorts it to obtain the benefit of superposition of the maxima of different harmonics (to get the highest possible signal-to-noise ratio), the resultant output waveform compresses in time in comparison with the input waveform by value of the pulse compression ratio (Fig. P24). This useful property of matched filter processing gives the benefit of radiating long pulses on transmit (and, hence obtaining efficient use of power capability), and simultaneously obtaining short pulses on receive (and, hence obtaining good range resolution) when employing pulse-compression waveforms. The side effect of pulse compression is appearance of range or time sidelobes that can mask nearby echoes, requiring the use of special measures for their suppression.

There are two basic ways to implement intrapulse modulation: coding either the frequency or the phase of the transmitted pulses, resulting in frequency-coded or phase-coded waveforms, which are the basic types of waveforms used in modern radars. (See **WAVEFORM, pulse-compression**.) The basic methods of implementation of pulse compression are analog and digital pulse-compression techniques. The main advantages of pulse compression, leading to wide usage of this technique in modern radar are the following: increased detection capability inherent in long, high-energy pulses is combined with increased resolving capability inherent in short pulses; generation of high peak power common in short-pulse systems can be avoided and more efficient use of average power can be obtained without increasing pulse repetition frequency, which would result in decreasing the unambiguous range; since the long pulse is used on transmit, increased doppler resolution is possible; when coded waveforms are used, a radar is less vulnerable to interference. The cost to be paid for these advantages is greater complexity relative to simple pulse transmissions, complexity of pulse-compression waveform generation and processing, all increasing the cost of the radar system. *SAL*

Ref.: IEEE (1990), p. 23; Barton (1988), pp. 220–230; Barton (1991), pp. 7.2–7.31; Skolnik (1980), pp. 420–434; Skolnik (1990), pp. 10.1–10.39; Brookner (1988), pp. 143–148; Lewis (1986), pp. 7–116; Leonov (1988), p. 62.

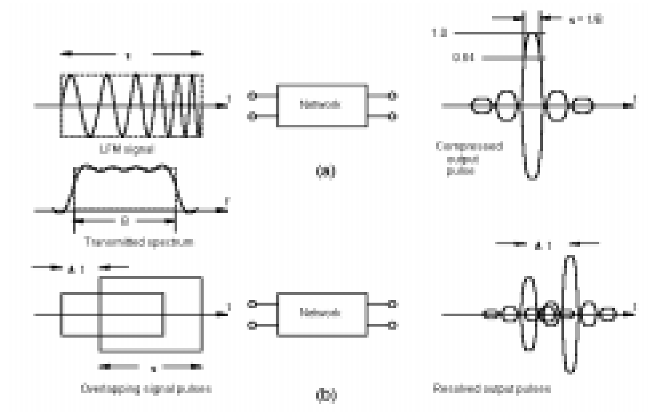


Figure P24 Pulse compression (a), and resolution (b) of pulse-compression waveforms after processing in matched filter (after Leonov, 1988, Fig. 2.21, p. 62).

Analog pulse compression involves the use of analog methods to generate and process pulse-compression waveforms. The main techniques incorporate active devices, primarily oscillators (see **OSCILLATOR, voltage controlled**) and passive devices, primarily different delay lines for linear FM waveforms. Passive devices can be divided into two general classes: ultrasonic devices (bulk-wave or surface-acoustic-wave (SAW) types) and electrical devices using the dispersive characteristics of an electrical network. In ultrasonic devices the input electrical signal is transformed into an acoustic wave propagating through the medium at sonic speed, and at the output the signal is converted back to an electrical waveform. Longer delays may be achieved than with purely electrical devices of comparable size, as the wave travels at sonic speed. The most popular technique for linear FM waveforms is the SAW delay line. In SAW technology the energy is concentrated in a surface wave, making it much more efficient than bulk-wave devices where the wave propagates through the crystal. The main limitation of bulk devices is the necessity to arrange the coupling between acoustic medium and electrical signal, typically with transducers inserting high losses (currently interdigital transducers are considered to transform an electrical signal most efficiently to acoustic energy and vice versa). Electrical networks with dispersive characteristics are typically electrical delay networks having a linear delay-versus-frequency characteristics. The main characteristics of analog pulse compression devices for linear FM waveforms are given in Table P6.

Table P6
Characteristics of Passive Linear-FM Devices

	B, MHz	T, μs	BT	f_0, MHz	Typical loss, dB	Typical spurious, dB
Aluminum strip delay line	1	500	200	5	15	−60
Steel strip delay line	20	350	500	45	70	−55
All-pass network	40	1000	300	25	25	−40
Perpendicular diffraction delay line	40	75	1000	100	30	−45
Surface-wave delay line	40	50	1000	100	70	−50
Wedge-type delay line	250	65	1000	500	50	−50
Folded-tape meander line	1000	1.5	1000	2000	25	−40
Waveguide operated near cutoff	1000	3	1000	5000	60	−25
YIG crystal	1000	10	2000	2000	70	−20

(from Skolnik, 1990, Table 10.2, p. 10.13, reprinted by permission of McGraw-Hill)

The main disadvantages of analog pulse compression are the instability of network parameters, resulting in poor repeatability of compressed waveforms, and larger size than digital circuits of comparable performance. *SAL*

Ref.: Skolnik (1990), pp. 10.10–10.15.

Digital pulse compression involves use of digital technology for generating and processing pulse-compression waveforms. It has some distinct advantages over analog pulse-compression technique. It is more flexible: it allows implementation of different types of pulse-compression waveforms (binary- and polyphase-code waveforms, linear and nonlinear FM waveforms or even other exotic waveforms at a designer's discretion); it can operate with signals of different durations; the same implementation can be used to handle multiple types of waveforms; and digital waveform generators are very stable devices

A system for digital implementation of a binary-coded waveform compression employing in-phase and quadrature channels (often called homodyne or zero IF) is shown in Fig. P25. In this system the phase of each transmitted binary element is defined by the code generator and is equal to 0 or 180° with respect to the local oscillator signal. The phase of the received signal is shifted in respect to the LO by an amount depending on the range and velocity of the target. Each correlator may consist of several correlators depending on the number of quantization bits in the digitized signal and only one channel instead of two channels can be used that will insert an average loss about 3 dB. A typical scheme that may be used to generate digital frequency-coded or polyphase- coded waveforms is shown in the Fig. P26.

The development of the digital pulse-compression approach is closely tied with the general development of the digital technology and will develop with the further increase of memory units capability and computational speed. Nowadays the main limitation is the upper limit of frequency available for digital devices, so digital approach is restricted in bandwidth under about 100 MHz. One of the ways to increase the bandwidth limitation for linear FM waveforms is to use

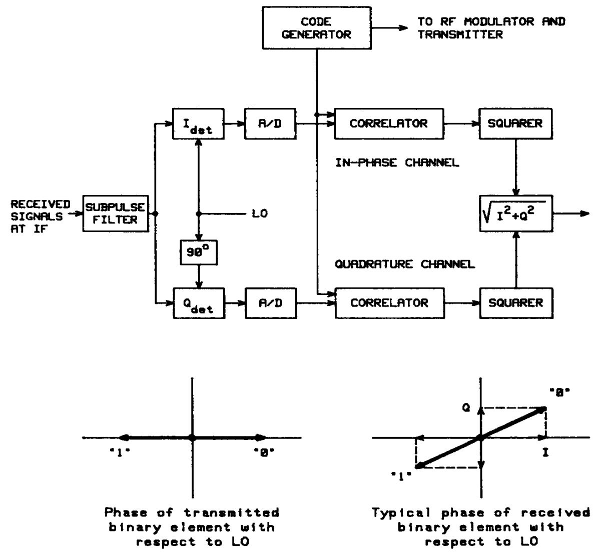


Figure P25 Digital pulse compression for binary-coded waveform (from Skolnik, 1990, Fig. 10.12, p. 10.23, reprinted by permission of McGraw-Hill).

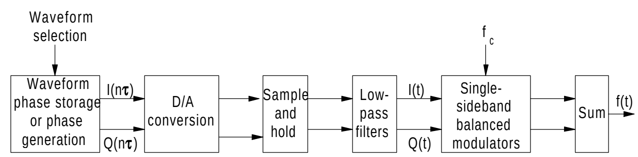


Figure P26 Digital pulse compression for FM and polyphase-coded waveforms (after Skolnik, 1990, Fig. 10.3, p. 10.7).

frequency multiplication combined with stretch processing. The popular digital technique for pulse compression is the fast Fourier transform. *SAL*

Ref.: Skolnik (1990), pp. 10.7–10.10; Wehner (1987), p. 138; Leonov (1988), p. 68.

Frequency-modulated [FM] pulse compression is one of the basic techniques for improving the range resolution of a radar while maintaining adequate duty factor and average power. The simplest approach in this technique is to use linear FM waveforms. As shown in Fig. P27, the carrier frequency is varied linearly during the pulse, broadening the transmitted spectrum. Upon reception, a matched filter produces the output of Fig. P27(e), with a narrow pulse surrounded by time sidelobes.

When using a complex signal with linear frequency modulation (chirp), the carrier frequency change during the pulse τ_i is linear at rate $\Delta f/\tau_i$, for a frequency deviation Δf (Fig. P27c). The instantaneous value of signal frequency is determined by

$$f(t) = f_0 + \Delta f\frac{t}{\tau_i},\ |t| \le \frac{\tau_i}{2}$$

where f_0 is the central frequency of the signal.

If the product $\tau_i\Delta f$ is large, the increase of the carrier frequency after the linear sweep (Fig. P27c) will yield an almost rectangular spectrum (Fig. P27b). The dependency of delay t_e

in time on filter frequency is a linearly decreasing function of frequency (Fig. P27d). The linear characteristic of delay in time provides large delay of low-frequency components at the beginning of the pulse in comparison with high-frequency components at its end.

All spectral components are delayed in an optimum filter as well as during digital processing by the amount of time necessary to arrive simultaneously at its output. Having the same zero phase, they add to form a peak signal blip (Fig. P27e). This also explains the increase of signal amplitude after passing through the filter and the compression in width. The output pulse at a level –4 dB below the maximum has width $\tau_o = 1/\Delta f$. The ratio of pulse width at the input τ_i to width at the output τ_o for an optimum filter is called the *pulse compression ratio*, $k_{pc} = \tau_i/\tau_o = \tau_i\Delta$.

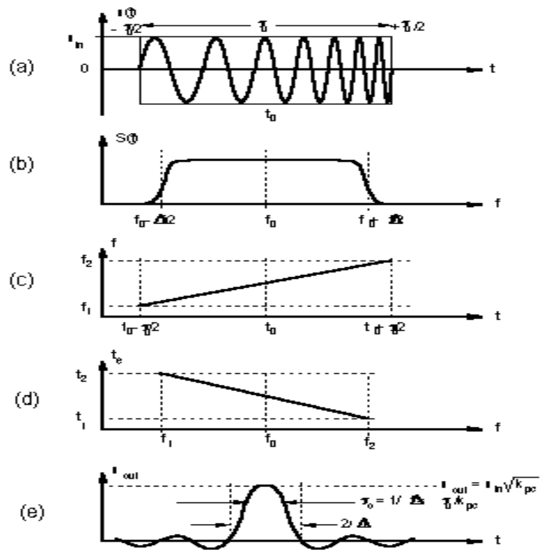


Figure P27 Compression of a linear-FM (chirp) signal (after Fal'kovich, 1989, Fig. 6.13 and 6.14, page 245).

A disadvantage of pulse compression is the appearance of sidelobes in the compressed signal that can lower the radar resolution. To control this, special measures are taken. To reduce sidelobes one can use a filter with smooth but sharply descending amplitude-frequency characteristic (for e.g., a Gaussian response) in place of an ideal filter. This allows one to reduce the sidelobe level considerably with only slight widening of the mainlobe.

Time sidelobes may be reduced by applying weighting in the compression filter, causing a weighting loss but greatly reducing the sidelobe levels. The characteristics of various amplitude weighting functions are shown in Table P7 (see also **WEIGHTING'**).

Greater attenuation of the sidelobes can be obtained by using special weighting processing. Its essence consists of selecting the pattern of amplitude variation (weighting function) of the amplitude-frequency characteristic of the filter, through which the signal passes. As a result one can distort the signal spectrum in such a manner it approaches a signal spectrum that has low sidelobe level. Another approach is to use a nonlinear FM waveform, in which the transmitted spectrum is weighted, providing a low-sidelobe response with a matched filter. *AIL*

Ref.: Cook (1967), Chs. 6, 7; Rihaczek (1969), Ch. 7.

Phase-coded pulse compression uses phase-coded waveforms. With pseudorandom phase-coding having *n* subpulses, the ratio of the mainlobe power over the average sidelobes is approximately *n*. Therefore, for large *n*, on the order of a thousand, sidelobes are small and measures to correct them are not called for.

Frequency signal compression is carried out in a correlator. To compress a complex signal in frequency it is necessary to compensate for signal phase. To do this one must supply a copy of the signal to the correlator synchronously and in phase. One thereby removes intrapulse modulation and the spectral width of a signal with duration t_i will equal $f_s = 1/t_i$ following compression. Thus, the correlator performs spectrum compression.

To the same degree that time compression improves range resolution, frequency compression improves velocity resolution. By using the combined correlation-filter method of complex signal processing, one can accomplish compression in both time and frequency.

The compression ratio of a phase-keyed signal $k_{pc} = t_i/t_o$ is equal to the number of increments *n* (in the example of Fig. P28, $n = 7$).

Thus, compression of phase-keyed signals in time substantially improves radar range resolution for a given average transmitted power. Range resolution is determined by duration of the compressed pulse τ_o and duration of the compressed pulse will depend on spectral width of the complex signal. For a phase-coded signal the spectral width is determined by the equation $f_s = 1/\tau_0$.

Radar sets using signals that are phase keyed radiate a long pulse τ_i, which consists of several short pulses τ_o with identical duration and frequency, but which differ in phase, by values 0 or 180° (Fig. P28a). In this figure we show the typical pattern of a phase-keyed signal, the phase-keying of which is accomplished in accordance with a 7-element ($n = 7$) Barker code.

Table P7

Characteristics of Pulse Compression Weighting Functions

Weighting function	Max. sidelobe level, dB	Relative width of mainlobe	Change of *S*/*N* ratio, dB
Without correction	−13.2	1.0	0
Hamming	−42.8	1.5	−1.34
Cosine-square	−31.4	1.59	−1.76
Cosine-cube	−39.1	1.41	−2.38
Dolph-Chebyshev	−40	1.41	-

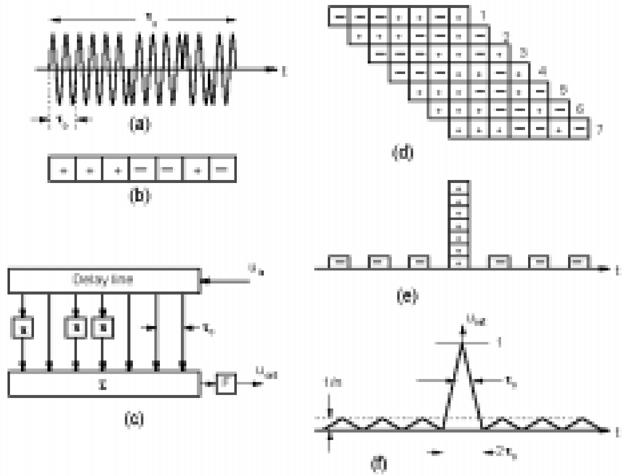


Figure P28 Compression of a phase-coded signal.

Biphase waveforms (phase-coded by 180°) using binary pseudorandom sequences have been most popular. As with a chirp pulse, a phase-coded signal is compressed by using a matched filter (Fig. P28c). It consists of a delay line with taps, phase inverters (π), adder (Σ), and filter (F). In Fig. P28d we have shown pulses which come from taps of the delay line to the adder. The result of summation is shown in Fig. P28e, and in Fig. P28f we obtain an envelope of a signal τ_o at the filter output F, that is compressed relative to τ_i by the factor *n*. *AIL*

Ref.: Cook (1967), Ch. 8; Popov (1980), p. 105; Sosulin (1992), p. 40.

The **pulse-compression ratio** is the ratio of transmitted pulsewidth to the pulsewidth after the process of pulse compression (at the output of the matched filter). An alternative term used is *time-bandwidth product*, although this term is also applicable to uncompressed waveforms (e.g., pulse trains). *SAL*

Ref.: Barton (1988), p. 221.

PULSE REPETITION FREQUENCY (PRF). The pulse repetition frequency "is the number of pulses per unit of time, usually per second." The choice of proper PRF is very important in radar design. There are three classic cases of PRFs are distinguished in radar: Low PRF (LPRF) gives unambiguous range measurements for all the targets of interest for the given radar; medium PRF (MPRF) is too large for unambiguous measurement of all targets of interest and too small for unambiguous doppler measurement of these targets, so the expected targets are ambiguous both in range and in doppler. High PRF (HPRF) is high enough to obtain unambiguous doppler measurement (and, in airborne radar, to obtain a clear doppler region for detection of approaching targets). LPRF gives unambiguous range measurement with a single PRF and permits use sensitivity time control (STC) to reduce the dynamic range requirements for the receiver. However, the average power may be low unless pulse compression is used to meet the combined needs of good range resolution and high average power (with long pulses). For HPRF the measurements are ambiguous in range, so typically several different PRFs are required to eliminate range ambiguity. Eclipsing losses are appreciable, and sensitivity time control cannot be used, resulting in requirements for the receiver to handle a wide range of amplitudes. The advantages are that doppler measurement is unambiguous (so blind speeds do not occur from ambiguous target doppler within the main-beam clutter notch), and long transmitter pulses are not required to obtain high average power. The MPRF to large extent is a compromise mode between LPRF and HPRF, so it tends to share both advantages and disadvantages of these modes.

Since the requirements to the proper choice of optimal PRF are contradictory, a useful mode of radar operation is staggered PRF (i.e., the mode when several different PRFs are used in a definite sequence). Sometimes this mode is called *multiple PRFs* and when the interpulse interval varies in a random manner it is termed *PRF jitter*. Pulse trains employing staggered PRF are termed *staggered pulse trains*. The PRF may be switched on a pulse-to-pulse basis, every other scan, or every time the antenna is scanned a half beamwidth. Such a mode may be used to eliminate range ambiguity (Fig. P29), to improve the characteristics of blind speed cancellation in MTI radars, or to provide enhanced ECCM capability against jamming. *SAL*

Ref.: IEEE (1990), pp. 23, 20, 21,18; Johnston (1979), pp. 64, 67; Barton (1991), p. 7.44; Skolnik (1980), p. 114; Long (1992), p. 259; Chrzanowski (1990), p. 60; Neri (1991), p. 421; Nathanson (1990), p. 330.

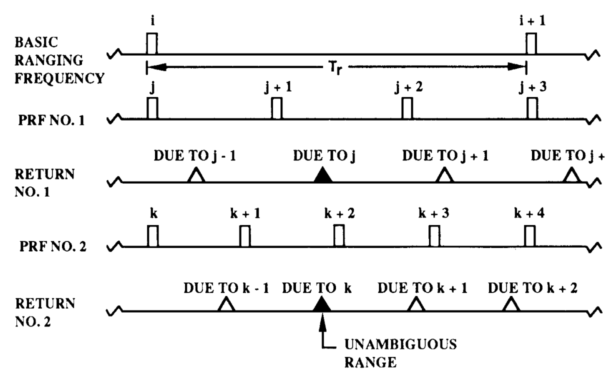


Figure P29 Removal of range ambiguity with staggered PRF (from Long, 1992, Fig. 6.15, p. 254).

High PRF (HPRF) has a special meaning for pulsed doppler radar: a pulsed doppler radar with a HPRF waveform is unambiguous in doppler but totally ambiguous in range. This is highly desirable for those cases in which targets can be distinguished from clutter on the basis of radial velocity. For this reason HPRF waveforms have found extensive application in airborne radar, where the strongest clutter return lies at a range no closer than the aircraft altitude, and all mainlobe and sidelobe clutter is located in the doppler frequency region $\pm 2v_r/\lambda$, where v_r is the velocity of the radar-equipped aircraft. The PRF is selected so as to create a large clutter-free doppler region that encompasses the range of expected closing velocities. Range ambiguities occur every $c/2f_r$ meters, where c is the speed of light and f_r is the waveform PRF, but target range, if required, can be resolved through the use of staggered PRFs. HPRF pulsed doppler radars at X-band typically operate at PRFs from 100 kHz to over 300 kHz, at duty cycles up to 50%.

For nonclosing or slowly closing target conditions, the target signal must compete with sidelobe and backlobe clutter, and if sufficient clutter attenuation (CA) cannot be achieved in the radar system, an alternate, medium PRF (MPRF) waveform may have to be used. *PCH*

Ref.: Schleher (1991), pp. 65,66.

PRF jitter (see **staggered PRF**).

Low PRF (LPRF) describes a waveform whose unambiguous range is greater than the maximum target range. Such waveforms are particularly suited to long-range airborne surveillance and to surface-based radars that must detect low-altitude targets in land or sea clutter. Most of these radars employ moving-target detection (MTD) techniques, such as MTI, pulsed doppler, or both, to detect the moving target against a background of stationary clutter, but use of the LPRF waveform allows only an ambiguous measurement of target radial velocity. The LPRF radar must detect targets having radial velocities from near zero to several times the blind speed of the radar. To avoid loss of detection of targets near the radar blind speed and its multiples, a pulse-to-pulse stagger can be introduced in the pulse repetition interval. Clutter cancellation is retained for clutter lying within the unambiguous range. If clutter exists beyond the first range ambiguity, another procedure, referred to as *PRF-diversity* can be used. With PRF-diversity MTI, the PRF is held constant until echoes from the longest-range clutter are received and processed.

If target radial velocity information is required in addition to target range (e.g., for target designation to a separate tracking radar), PRF-stagger or PRF-diversity, when used in a pulsed doppler radar, can be used to resolve the doppler ambiguities. (See **MTI**; **RADAR, pulsed doppler.**) *PCH*

Ref.: Barton (1988), pp. 234,239.

Medium PRF (MPRF) describes a waveform that is ambiguous in both range and doppler. Simply stated, use of MPRF is a compromise to ameliorate the clutter effects posed by the range-ambiguous HPRF waveform on the one hand, and those created by the velocity-ambiguous LPRF waveform on the other.

The MPRF waveform is especially important in airborne intercept (AI) radar, where detection of slow-moving and receding targets is required in an often-severe look-down clutter environment. With a LPRF waveform, in the presence of a broad clutter spectrum such as that produced with wind-blown rain or chaff, it may not be possible to realize a blind speed that is high enough to reject clutter without rejecting a significant fraction of targets as well. At the PRFs typically employed by modern high-speed interceptor aircraft HPRF radar, sidelobe clutter from virtually all ranges folds into the range interval corresponding to the interpulse period. If the sidelobe clutter amplitude in the target doppler resolution cell is sufficiently strong (e.g., from short-range clutter) and the target may be unresolvable in doppler. Such conditions can regularly occur for slowly moving and receding targets.

Figure P30 is a range-doppler "map" for a typical MPRF radar, showing the blind regions caused by the range and doppler ambiguities. With limited clutter attenuation, the first several range cells after the transmitted pulse will contain strong clutter, which may eclipse the target, adding to eclipsing by the transmitter itself. This creates blind range bands of width Δ_r, which cross the blind doppler bands of width Δ_v produced by blind velocities. To obtain a high probability of detection in each observation t_o, bursts at different PRFs must be used. The number of bursts needed depends on the fraction of clear area F_a in range-velocity space:

$$F_a = \left(1 - \frac{\Delta_r}{R_u}\right)\left(1 - \frac{\Delta_v}{v_b}\right)$$

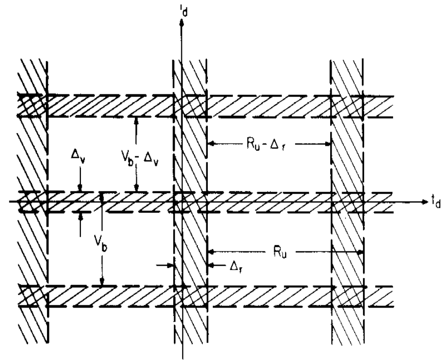


Figure P30 Blind regions in MPRF PD radar (from Barton, 1988).

Since sidelobe clutter poses a major problem in MPRF radar, ensuring low antenna sidelobes is a priority. Use of pulse-compression techniques to narrow the width of the effective range resolution cell is another step that can be taken to reduce the relative contribution of ambiguous range clutter. *PCH*

Ref.: Barton (1988), pp. 260–262. Schleher (1991), pp. 68–71.

Staggered PRF is a technique employed in MTI and pulsed doppler radars in which the radar's interpulse period is altered on a pulse-to-pulse or burst-to-burst basis to allow detection of targets whose velocities fall within the blind regions of the radar's velocity response that occur at multiples of the radar's PRF. In burst-to-burst stagger, also called "block" stagger, the PRF is changed after a number of radar pulses have been transmitted at a constant PRF. Pulse-to-pulse stagger generally provides a better velocity response than burst-to-burst stagger, but imposes more severe stability requirements on the system, especially the transmitter. Burst-to-burst stagger allows multiple-time-around clutter responses (from clutter beyond the first range ambiguity) to be canceled in a coherent MTI system. *PCH*

Ref.: Schleher (1991), pp. 9, 390–392.

PULSER. A pulser is a modulating device used in transmitters of pulsed radars to shape the transmitted pulse. The basic types are the cathode pulser, modulating-anode (or mod-anode) pulser, and grid pulser. The first is typically an active switch that is either driven hard enough to bring its voltage drop as low as possible, minimizing dissipation and maximizing efficiency, or operating as a constant-current device by limiting its drive. The basic circuit implementations are direct-coupled, capacitor coupled, transformer-coupled, or capacitor-and-transformer-coupled. A basic type of mod-anode pulser is floating-deck modulator used, for example, with a klystron. It can provide very good pulse flatness due to the usage of capacitor bank and sufficient pulse lengths as there is no limit on maximum pulsewidth except for capacitor bank size. The grid pulser is the smallest, easiest, and least expensive type of pulser, but it can be used only in RF tubes with the grids (see also **MODULATOR**). *SAL*

Ref.: Skolnik (1990), p. 4.37.

Q

Q(-FACTOR). The Q factor characterizes the resonant properties of an oscillatory system. Series and parallel tuned circuits are widely used in radars as oscillatory systems. For them, resonant frequency:

$$\omega_r = \frac{1}{\sqrt{LC}}$$

where L is the inductance and C is the capacitance of the circuit.

For a series tuned circuit, the Q-factor is

$$Q_1 = \frac{U_{Lr}}{U_i} = \frac{U_{Cr}}{U_i} = \frac{X}{R}$$

where U_{Lr} and U_{Cr} are the voltages across the inductor and capacitor, U_i is the input voltage, $X = \omega_r L = 1/\omega_r C$ is the impedance at the resonant frequency, and R is the resistance of the circuit.

For a parallel circuit:

$$Q_2 = \frac{I_k}{I}$$

where I_k is the current in the circuit and I is the current in a common conductor at circuit input.

Thus, for a series circuit, Q-factor illustrates the extent to which voltage at the output is greater than that at circuit input. For a parallel circuit, Q-factor illustrates the extent to which current at circuit output is greater than that at the input. *AIL*

Ref.:Terman (1955), p. 45; Gonorovskiy (1986), p. 120.

The **Q-FUNCTION** describes the dependence of the cross-sectional area of the square of the modulus of the ambiguity function $\chi(\cdot)$, drawn parallel to the axis of the time delays, on the doppler frequency shift f_d:

$$Q_u(f_d) = \int_{-\infty}^{\infty} |\chi_u(\tau, 0)|^2 e^{j2\pi f_d \tau} d\tau$$

It is used in determining the ability of a signal to suppress interference from local objects uniformly distributed in range with a given doppler shift, which differs from the doppler shift of the target by the value f_d. *IAM*

Ref.: Skolnik (1970), p. 3.40.

QUANTIZATION. Quantization is "a process in which the range of values of the wave is divided into a finite number of smaller subranges, each of which is represented by an assigned (or quantized) value within the subrange." In radar applications it is performed in the process of analog-to-digital conversion, consisting of two functions: waveform sampling (discretization in time), and quantization (discretization in amplitude). (See also **CONVERTER, analog-to-digital**; **ERROR, quantization.**) *SAL*

Ref.: IEEE (1993), p. 1,046; Barton (1969), p. 187.

R

"(RUNNING) RABBITS" are asynchronous pulses interfering with target visibility on a display. When the interfering pulse train differs from the radar pulse repetition interval by Δt, the rabbits move in range by $\Delta R = c\Delta t/2$ per PRI, corresponding to a velocity $c\Delta t/2t_r$. Thus, two radars having the same nominal PRI but differing by a timing oscillator offset of 0.001% will create rabbits moving at $10^{-5}c/2 = 1500$ m/s, potentially interpreted by an untrained operator as high-speed targets. Even if the rabbits are recognized as interference, their presence may detract from target detection. *DKB*

Ref.: Johnston (1979), p. 65.

RADAR is "a device for transmitting electromagnetic signals and receiving echoes from objects of interest (targets) within its volume of coverage. Presence of a target is revealed by detection of its echo or its transponder reply. Additional information about a target provided by a radar includes one or more of the following: distance (range), by the elapsed time between transmission of the signal and reception of the return signal; direction, by use of directive antenna patterns; rate of change of range, by measurement of doppler shift; description or classification of target, by analysis of echoes and their variation with time. The term *radar* was originally an acronym for *ra*dio *d*etection *a*nd *r*anging. Some radars can also operate in a passive mode in which the transmitter is turned off and information about targets is derived by receiving radiation emanating from the targets themselves or reflected by targets from external sources."

Radar is also recognized as the field of science and technology that includes the methods and equipment to perform the following basic operations against the targets of interest:

(1) Radar detection (see **DETECTION**).
(2) Radar measurement (see **MEASUREMENT**).
(3) Radar recognition, discrimination, and identification (see **TARGET RECOGNITION AND IDENTIFICATION**).

The quantitative and qualitative description of radar operation provides radar performance figures, the most basic of which are coverage, probabilities of detection and false alarm, resolution, measurement errors, and throughput capacity. (See **PERFORMANCE.**) Major subsystems affecting radar performance are the antenna, transmitter, receiver, and signal processor. (See **SUBSYSTEM.**) Most radars operate in the radio frequency bands. (See **FREQUENCY.**) Active sensors operating in the optical band are called laser radars, even though the first letter of the radar acronym stands for radio, indicating that lower band of frequencies.

There are many specific features that can be used in classification of radars:

(1) Depending on the nature and location of the source of electromagnetic radiation, which supplies information on the radar target and is called a radar signal, one can distinguish active (or primary) radar, radar with active response (secondary radar), or passive radar.

(2) Depending on maximum effective range, one can distinguish short-range, horizon (direct visibility), or over-the-horizon radar.

(3) Depending on the number of channels used to determine angular coordinates in a given coordinate plane, one can include monopulse radar.

(4) Depending on the position of receiving and transmitting antennas, one can distinguish monostatic or multistatic radar.

(5) The term radar is applied to applications for solving scientific and practical problems in different branches of human activity. For example, one refers to subsurface radar, radar astronomy, geographic research radar. *AIL*

Ref.: IEEE (1993), p. 1051; Skolnik (1980, 1988, 1990).

RADAR APPLICATIONS AND TYPES

An **above-the-horizon radar** is one in which the target is observed over line-of-sight propagation paths that do not extend below the horizon. The term applies to the vast majority of radars, and hence is seldom used except to emphasize the distinction with over-the-horizon or subsurface radars. *AIL*

Ref.: Dulevich (1978), p. 11.

An **acquisition radar** in air defense systems is associated with an antiaircraft artillery (AAA) or surface-to-air missile (SAM) battery or shipborne defense system that acquires targets for hand-off to a separate target tracking radar. A 360° azimuth scan usually is used as aerial target acquisition radars. They may be two-dimensional (measuring range and azimuth) or three-dimensional (measuring range, azimuth, and elevation). Height-finding radars may operate in association with 2D radars. Target parameters furnished by the acquisition radar usually include range, azimuth, elevation or altitude, and target radial velocity (doppler). These target parameters are used by the tracking radar to initiate its own local target acquisition mode. A air defense battery or shipborne system may have more than one target acquisition radar, such as one radar dedicated to low-altitude targets (horizon search), and a second for medium-to-high- altitude targets. The target acquisition radar in an air defense battery may operate autonomously or it may be cued by a long-range search radar located at or near a higher headquarters or, in the case of naval systems, aboard a ship near the fleet center.

Aerial target acquisition radars may be fixed or mobile. The main requirements levied on them include long detection range and the capability to acquire aerial targets at all possible operating altitudes. As a rule, modern aerial target acquisition radars use automatic detection circuits, and some also initiate tracks automatically. Acquisition radars are sometimes used also for air traffic control. The AN/FPS-7 and AN/TPS-59 are examples of three-dimensional radars, while the AN/FPS-89 is an example of a height-finder. (See also **air surveillance radar.**) *AIL, PCH*

Ref.: Skolnik (1980), p. 153; Leonov. (1988), pp. 72–103.

active radar (see **primary radar**).

An **active radar missile seeker** is a small, self-contained radar installed in a missile for acquiring a target and performing homing guidance on the energy returned from the missile-borne radar's illumination of the target. The difference between an active radar seeker and a semiactive radar seeker is that the former provides its own source of radar energy (via the transmitter), while the latter relies on passive reception of target-reflected energy provided by a separate, off-board, target illumination and tracking radar.

Active radar seekers are used in air-to-air (AAM) and surface-to-air (SAM) defense missiles, as well as in tactical air-to-surface (ASM) and surface-to-surface (SSM) weapon systems. Because of the physical limitations of weight, volume, and power attendant to installation in a tactical missile, active radar seekers are generally used in the missile's terminal guidance phase, acquiring the target during the last few seconds of flight after having been guided to the active seeker acquisition "basket" by one or more separate, midcourse guidance modes. Active radar seekers give an AAM-equipped aircraft the advantage of "fire-and-forget" soon after launch, leaving the aircraft free to perform maneuvers that decrease its own vulnerability. (See also **GUIDANCE, radar** and **HOMING, radar.**) *PCH*

Ref.: Skolnik (1990), p. 19.15.

Adaptive radar can be adapted to an unknown or changing radar situation. As a rule, adaptive antenna arrays are used in such radars. The fact that the characteristics of the targets and interference sources, their coordinates, number, and mutual positioning change during the process of acquisition and determination of target coordinates, explains the requirement to create adaptive radars. A processing algorithm optimum for one situation produces a significant loss in others. Therefore, the requirement arises to continually correct the processing algorithm so it remains optimum in spite of changes in the radar situation. This problem is solved in adaptive radars through use of special algorithms realized on a computer. *AIL*

Ref.: Skolnik (1980), p. 332; (1990), p. 9.14; Lukoshkin (1983), p. 35.

An **aerostat radar** (usually a search radar) is installed on a tethered balloon, increasing the detection range for low-altitude or surface targets. Power for the radar can be supplied, and data returned to the surface station, through the tethering cable. The operational difficulties of maintaining the aerostat in position under harsh weather conditions have precluded most military exploitations of aerostat radars, but some success has been achieved in application to border surveillance, as for drug enforcement and immigration control. Figure R1 shows an radar surveillance aerostat designed for battlefield surveillance. *DKB*

Ref.: Skolnik (1988), pp. 262–263; Long (1992), Ch. 11.

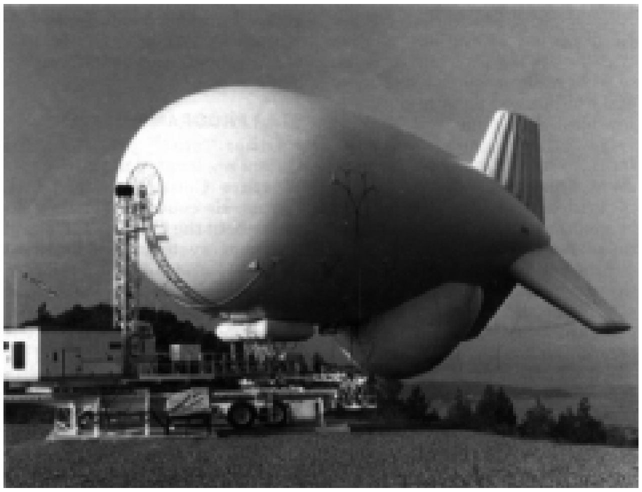


Figure R1 Small aerostat surveillance system, designed by Westinghouse for battle field surveillance, using a modified AN/APG-66 AI radar.

Airborne radar is any radar installed aboard a platform designed to operate within the earth's sensible atmosphere. Thus an active radar missile seeker is technically an airborne radar, as is an aerostat radar or one installed in a remotely piloted vehicle (RPV). However, most airborne radars are found in manned aircraft. Commercial airliners, transport aircraft, and many private aircraft are equipped with airborne weather radar, to allow the pilot to detect and avoid severe weather cells or storms. The air forces of many nations employ airborne early warning (AEW) radar at UHF to S-band frequencies for surveillance of their national air, land, and sea boundaries, and air defense fighter aircraft are equipped with airborne intercept (AI) X-band radars for fire control and missile guidance. Other types of military airborne radar include terrain-following radar to permit all-weather, low-altitude penetration of enemy airspace, and tail-warning radar to alert a bomber aircraft of rear attack. Both military and civilian organizations utilize synthetic aperture radar (SAR) techniques at various operating frequencies for high-resolution air-to-surface mapping.

Because of the nature of the platform, size and weight are critical issues in airborne radar, and the radar's power-aperture product is limited also by vehicle configuration and by the primary power supply, which is usually supplied by the aircraft propulsion system (engines). (See also **airborne early warning radar**; **airborne intercept radar**; **airborne mapping radar**; **airborne weather avoidance radar**; **aircraft tail-protection radar**; **doppler radar**; **synthetic aperture radar.**) *PCH*

Ref.: Povejsil (1961); Stimson (1983); Morris (1988); Morchin (1990); Neri (1991), p. 175; Long (1992).

An **airborne early warning (AEW) radar** is carried by an airborne or spaceborne platform whose function is the detection of airborne intruders into the air space under surveillance. For aircraft-based AEW radar, the primary targets are usually aircraft and cruise missiles, but the mission of modern AEW can extend to the detection and track of several additional types of targets, including ships, land- and sea-launched ballistic missiles, and various kinds of land-based targets.

AEW radar characteristics vary depending on mission requirements and the characteristics and limitations of the host platform. Two U.S. systems stand out as prime examples of military AEW radar applications. The first, the E-3A Airborne Early Warning and Control System (AWACS), shown in Fig. R2, is a Boeing 707-320 jet aircraft modified to carry the Westinghouse-developed, S-band, pulsed doppler AN/APY-1 radar and other command and control equipment and crew. The AWACS radar has several operational modes, including long range air search, maritime search, and detection of moving targets over land and sea clutter. The AWACS radar has recently been adapted, for Japan, to the more modern Boeing 767 aircraft.

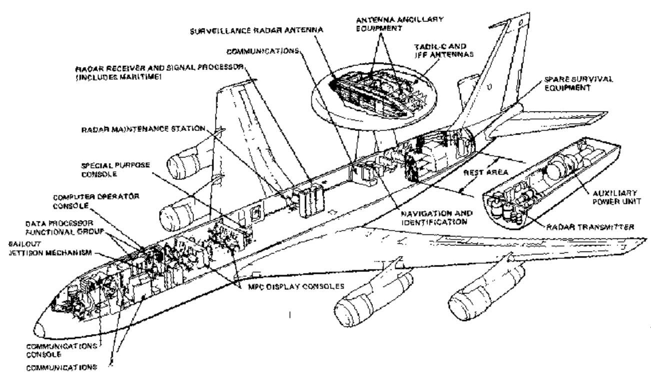


Figure R2 The Boeing E3A AWACS (from Morchin, 1990, Fig. 1.2, p. 11).

The E-2C Hawkeye, shown in Fig. R3, was developed by Grumman Aerospace for the U.S. Navy, specifically for carrier-based operations. Designed to serve as "the eyes of the fleet," the UHF, LPRF radar, developed by General Electric, has undergone several evolutions since its introduction in 1960.

Major improvements have been made in the radar's moving-target detection, multiple-target tracking, and electronic counter-countermeasure (ECCM) capabilities.

Both the AWACS and the Hawkeye AEW systems were designed to operate as airborne command, control, and communications (C^3) centers, with capability to direct multiple interceptor aircraft engagements of hostile airborne intruders.

Figure R3 E-2C AEW aircraft.

As such, these systems represent long-lived and expensive solutions to the military AEW problem. However, not all airborne surveillance missions require high cost and technologically sophisticated solutions, and many low-cost AEW radars have been installed on many types of airborne platforms, particularly for maritime surveillance, including commercial aircraft, helicopters, and airships (blimps).

An important problem for airborne radar is motion compensation is applied to moving-target indicator (MTI) systems on fast-moving platforms. Two effects must be overcome: the shifting mean of the clutter spectrum, resulting from projection of the platform velocity vector v_p on the beam axis, and the spreading of the spectrum caused by differing projected vectors across the beam and in the sidelobes. The shift in the mean is readily accomplished through control of an offset oscillator (usually at the radar IF), in accordance with the azimuth of a scanning beam and the elevation to the surface (as a function of range). Spreading is more difficult to counter, requiring use of so-called displaced phase center antenna (DPCA) techniques. In the true DPCA, the first pulse of a pair is transmitted and received by an antenna whose phase center is nearer the nose of the platform, and the second by an antenna whose phase center is displaced toward the tail by $t_r v_p$, the amount of platform motion during the pulse repetition interval t_r. The two resulting echoes from fixed clutter will be identical, appearing to have been produced by a fixed radar platform. In practice, it may prove desirable to transmit from a single antenna, feeding the delayed pulse train of an MTI canceler from a receiver connected to the leading phase center and the undelayed pulse train from the trailing phase center. In this case the two receiving phase centers are separated by $2t_r v_p$, and the virtual two-way phase centers formed by the transmit antenna and the two displaced receive antennas will have the desired separation $t_r v_p$.

When the antenna width w significantly exceeds the platform motion $t_r v_p$, a switched array (Fig. R4) may be used to shift the phase center between pulses. Another approach uses monopulse sum and difference patterns to synthesize DPCA. The transmitter is connected to the sum port of the antenna, and the two receive channels are formed as $\Sigma + jkv_p\Delta$ and

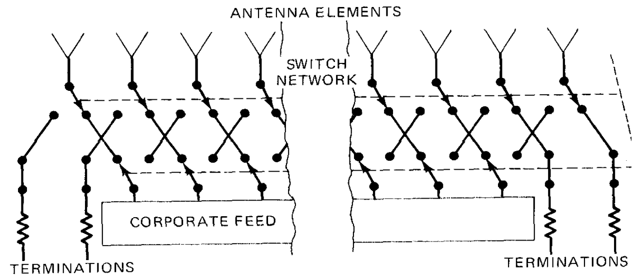


Figure R4 Switching network to synthesize displaced subarrays within an antenna array (from Skolnik, 1990, Fig. 16.10, p. 16.10, reprinted by permission of McGraw-Hill).

$\Sigma - jkv_p\Delta$. By setting $k \approx 2t_r v_p/w$, the delayed and undelayed signal trains may be made identical. *PCH, DKB*

Ref.: Stimson (1983); Hirst (1983); Morchin (1990); Long (1992); Skolnik (1988), pp. 251–263, (1990), Ch. 16; Cantafio (1989), pp. 425–440.

An **airborne intercept (AI) radar**, usually mounted on fighter or interceptor aircraft, is used for detecting, tracking, and supporting the engagement of hostile aircraft. In addition to target acquisition, AI radar supports the fire control function of the interceptor aircraft (i.e., it provides the human pilot with information required to vector the aircraft into a favorable weapon-firing position). The AI radar and fire control computer combination tells the pilot when to fire the guns or missiles. If the interceptor aircraft is equipped with semi-active radar air-to-air missiles, the AI radar illuminates the target or targets with the RF energy required for homing guidance.

Most AI radars operate in the X-band region of the radar spectrum, because it is within this region, nominally 10 GHz, that the best compromise can be reached between the radar's all-weather performance requirements and the radar size and power constraints. The Hughes AN/APG-63, a pulsed doppler, multimode radar installed in the F-15 aircraft, is an example of a currently deployed, modern AI radar. With a fully coherent radar, PRF-flexible waveforms, and programmable signal processor, the AN/APG-63, and its successor the AN/APG-70, are capable of detecting low-RCS targets at any aspect in a look-down clutter environment. Air-to-air modes include track-while-scan (TWS), range-while-search (RWS), single-target track (SST), super-search in a head-up display reference, as well as vertical acquisition modes with either automatic or manual lock-on. The radar's air-to-ground modes include real-beam mapping, doppler beam-sharpening ground-mapping, air-to-surface ranging, fixed and moving-ground-target track. Figure R5 shows the AN/APG-66 AI radar during final testing on the F-16 production line.

Trends in AI radar design appear to be focused on the use of wideband, active, conformal arrays instead of gimballed, narrowband antennas, and wider instantaneous bandwidths for better immunity to ECM. *PCH*

Ref.: Stimson (1983); Brookner (1988), p. 215.

An **airborne mapping radar** can be one dedicated to map-making, or a multimode airborne radar aboard a tactical air-

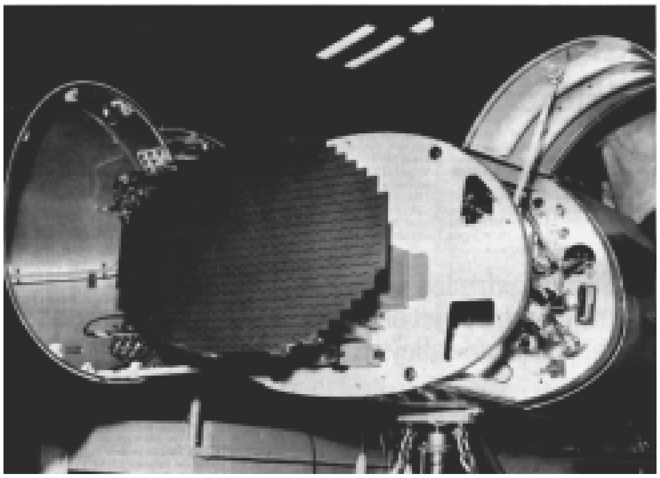


Figure R5 AN/APG-66 AI radar mounted in nose of F-16 fighter.

craft, which includes a surface mapping mode. Since in either case the surface of the earth (or some other astronomical body) is the target, the problem is not signal-to-noise ratio, which is generally a very large value, but resolution. There are two basic radar techniques used in aerial mapping of the surface: (1) real-beam mapping and (2) synthetic array mapping. In real-beam mapping, the azimuth resolution is limited by the actual radar beamwidth θ (i.e., at a range R the azimuth or cross-range resolution will be $R\theta$). Because antenna azimuth beamwidth is proportional to the radar wavelength λ divided by the width of the antenna, high resolution can be achieved with long antennas and short wavelengths. Weather conditions usually dictate that airborne mapping radar operate at X-band frequencies or lower, and maximum antenna dimensions are determined by the airborne platform's physical length and configuration. As an example, an X-band, side-looking, real-beam array antenna 5m long, can achieve azimuth resolutions sufficiently fine so as to resolve small ships at a range of approximately 10 km.

Ground-mapping using the synthetic array radar (SAR) technique can provide much higher azimuth resolution than possible using real-beam techniques, because the antenna "array" is formed by the linear motion of the radar-equipped aircraft. Thus the length L of a synthetic aperture array may be hundreds of meters long, and the two-way beamwidth of the array at a position 4 dB down from peak antenna gain is equal to $\lambda/2L$, or half the one-way beamwidth of a real-beam array of equivalent length. Practical SAR arrays are in use that yield azimuth resolutions in terms of a few meters.

Both real-beam and SAR mapping radars can take advantage of pulse compression techniques to enhance resolution in the downrange dimension. (See also **sidelooking radar**, **synthetic aperture radar**.) *PCH*

Ref.: Stimson (1983), pp. 521, 527–548; Currie (1987), Ch. 15.

An **airborne weather-avoidance radar** is a radar installed aboard an aircraft for detecting and avoiding severe weather. Virtually all large commercial aircraft, military aircraft not already equipped with tactical airborne radar, and many private aircraft carry weather-avoidance radar. Commercial aircraft weather-avoidance radars typically operate at C-band. The limited space and payload characteristic make X-band the frequency of choice for most private aircraft and small commercial "feeder" airlines. A simple and effective digital X-band weather-avoidance radar can weigh less than 50 lb, including displays and controls, and require only 4A of 28V dc power. Figure R6 shows the antenna and cockpit units of a weather-avoidance radar designed for small commercial aircraft.

Figure R6 AVQ-55 weather-avoidance radar components: (1) transmitter-receiver unit, (2) antenna, and (3) display (RCA photograph).

Because the radar return signal intensity is a function of precipitation (rain) drop size, by sensing the rate of change of signal strength with range, most modern weather-avoidance radars provide the pilot with a color-coded display of contours where the rain intensity changes abruptly. A typical weather avoidance radar uses a cone-shaped antenna beam which is scanned in azimuth (e.g., $\pm 60°$) and covers 5° to 8° in elevation. Simple pulsed radars, tilted in elevation to avoid excessive ground return, are adequate for most weather-avoidance radar applications. More sophisticated MTI and pulsed doppler techniques can be used to accurately measure wind shear within a rainstorm and reject ground return, but these features are generally not found on the inexpensive variety of radar common to private aviation. Figure R7 shows the cockpit display of data from a typical weather radar. *PCH*

Ref.: Stimson (1983), pp. 41, 127; *Allied Signal Technologies Pilot's Guide*, RDS 86 Digital Weather Radar; Hovanessian (1984), pp. 334–338; Davydov (1988), pp. 176–182.

aircraft control radar (see **air traffic control radar**; **ground-controlled approach radar**).

An **aircraft tail-protection radar,** also referred to as *tail-warning radar*, is an acquisition radar installed in the tail section of an aircraft to warn the crew of the presence of an enemy fighter or missile in the aircraft's rear hemisphere. The radars are used to warn the aircraft crew of danger, ensuring a timely aircraft maneuver and weapon employment. To the

Figure R7 Cockpit display of WXR-300 weather-avoidance radar data (Collins photograph).

extent that such radar may also activate a defense weapon system, a tail-protection radar may be technically indistinguishable from an airborne fire control or airborne intercept (AI) radar. Although radar-directed, tail-mounted weapon (gun) systems have been used on older generations of bomber aircraft, modern aircraft appear to rely on passive warning receivers combined with stealth and maneuverability for rear-area defense. Tail-warning radar can be used to augment these techniques in some cases, but cost and complexity of an additional radar often outweigh the perceived benefits. As the conformal array radar concept matures, tail-warning, or all-aspect radar protection, may become more practical. *PCH*

Ref.: Popov and Grigor'yants (1980), pp. 341–342.

Air-defense radars, depending on the types of observed targets, are classified as the radars operating against airborne targets (aircraft, helicopters, low-flying tactical missiles) and the radars operating against space targets (ballistic missiles and satellites). The first type of radars typically are classified as 3D or 2D surveillance radars (depending on the number of measured target coordinates), height-finding radars, or fire control radars. Radars operating against space targets are typically classified as early warning systems, antiballistic missile radars, spacecraft tracking radars, and instrumented radars located on test ranges. The radars used in early warning systems were developed as both above-the-horizon and over-the-horizon (OTH) systems.

Space target radars are the most sophisticated radar facilities, employing the latest achievements in radar technology, as they have to operate in an environment where a large target with a very small RCS (often less than 0.001 m^2) must be detected at long distances in very complicated jamming and false-target environment. *AIL*

Ref.: Leonov (1988).

An **airport surface detection equipment (ASDE) [radar]** is "a ground-based radar for observation of the positions of aircraft and other vehicles on the surface of an airport." This type of radar usually operates in K_u-, K-, or K_a-band with a narrow pulse (20 to 50 ns), narrow beam (0.25° to 0.4°), and rapid scan rate (60 rpm or greater), with display ranges from one to 5 km. The resolution is intended to be adequate to provide recognition of different classes of aircraft, as well as resolving one aircraft from another as they stand in taxiways at the end of an active runway. The radar must also detect and display automobiles and trucks operating on or near taxiways. Rain clutter presents a problem at the high microwave frequencies used, and this is minimized by using a circularly polarized antenna.

The newer generation of ASDEs used in the United States is represented by the ASDE-3, the parameters of which are listed in Table R1. *DKB*

IEEE (1993), p. 23; Skolnik (1988), pp. 116–126.

Table R1
Parameters of ASDE-3 Radar

Parameter	Units	Value
Frequency	GHz	16
Peak power	kW	10
Pulse width	ns	50
Pulse repetition frequency	kHz	20
Average power	W	10
Antenna size	m	5 × 1.3
Beamwidths	deg	0.25° × 1.6°, csc^2
Antenna gain	dB	46
Scan rate	rpm	60
Receiver bandwidth	MHz	38
Clutter rejection methods		CP, FA

An **airport surveillance radar (ASR)** is "a medium-range (for example, 60 nautical miles) surveillance radar used to control aircraft in the vicinity of an airport." The typical ASR has a detection range of 70 to 100 km on small aircraft, a scan period of 4 to 5 sec, operates at S-band, and provides two-coordinate data (range and azimuth). A secondary surveillance radar (SSR) antenna is usually mounted on the rotating ASR reflector.

The ASR-9, designed in the United States by Westinghouse, is typical of ASRs developed during the 1980s (Fig. R8). The major parameters of this radar are listed in Table R2. Important performance characteristics of an ASR, other than detection range, are resolution, data rate, ability to reject both land and weather clutter, ability to process and output data on tens of targets within the scanned volume, and high reliability. In many cases, duplicate transmit and receive chains are diplexed to a common antenna, providing improved detection performance through frequency diversity

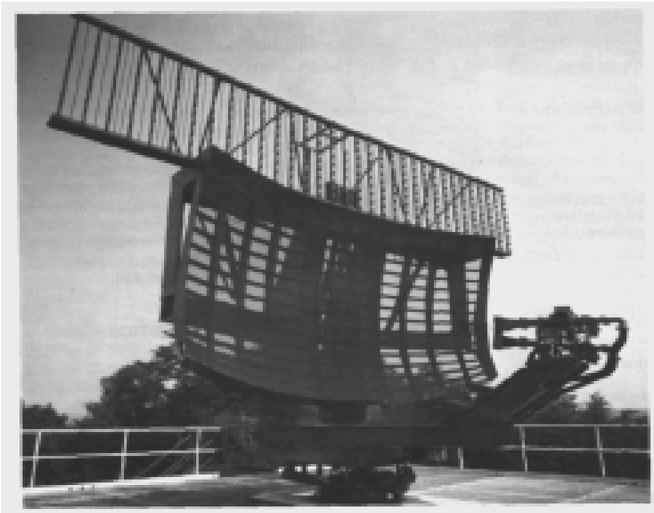


Figure R8 ASR-9 radar antenna (from Skolnik, 1990, Fig. 6.14, p. 6.19).

when both are operating and graceful degradation of performance in the event of failure in one chain.

The latest generation of ASRs is represented by the Raytheon solid-state ASR-10SS (Fig. R9). The all-solid-state design simplifies radar servicing, gives high reliability and maintainability, and permits unattended operation at remote sites. It also provides operational and cost flexibility, as the number of amplifier modules used in the transmitter can easily be changed.

Table R2
Major Parameters of ASR-9 and ASR-10SS

Parameter	Units	ASR-9	ASR-10SS
Frequency band		S	S
Peak power	kW	1,300	15[1)], 30[2)]
Pulse width	μs	1.0	1.0, 100
Pulse repetition frequency	Hz	1,200	825[3)]
Average power	kW	1.6	1.3[1)], 2.6[2)]
Antenna gain	dB	33.5	34.5
Azimuth beamwidth	deg	1.3°	1.4°
Elevation beamwidth	deg	4.8° $\csc^2$ to 30°	5° $\csc^2$ to 40°
Scan rate	rpm	12.5	15[3)]
Receiver noise factor	dB	4.5	3.3
Clutter rejection methods		MTD, STC	MTD, pulse compression, STC
Detection range (1 m^2, 80%)	km	100	125[1)], 150[2)]

Notes: [1)] 8-module transmitter; [2)] 16-module transmitter, [3)] average for 100-km instrumented range

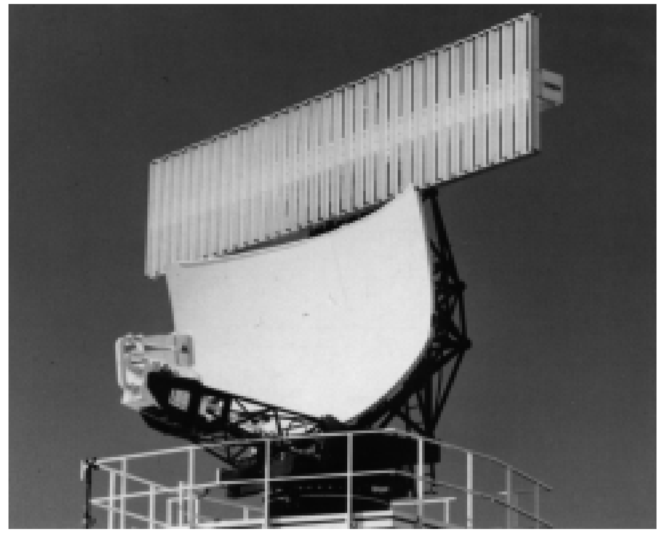


Figure R9 ASR-10SS radar antenna (Raytheon Company photo).

The main features of the ASR-10SS are

(1) Dual-beam antenna with sharp lower edge cutoff.

(2) Selectable circular or linear polarization.

(3) Frequency diversity, frequency agility, and PRF stagger.

(4) Dedicated weather and target channels.

(5) Highly reliable, multichannel, fully solid-state transmitter with fail-soft operation.

(6) Dual receiver-processor channels with automatic reconfiguration in event of failure.

(7) Digital signal processing with I- and Q-channels and doppler MTD. *DKB, SAL*

Ref.: IEEE (1993), p. 23; Skolnik (1988), pp. 89–94

An **air-route-surveillance radar (ARSR)** is "a long-range (for example, 200 nautical miles) surveillance radar used to control aircraft on airways beyond the coverage of airport surveillance radar (ASR)." A typical ARSR has a scan period of 10 to 12s, operates at L-band, and provides two-coordinate data (range and azimuth). A secondary surveillance radar (SSR) antenna is usually mounted on the rotating radar antenna. Important performance characteristics of an ARSR are long detection range, ability to reject both land and weather clutter, ability to process and output data on many tens of targets, and reliability.

The Alenia ATCR-22 is typical of ARSRs developed during the 1970s, and its major parameters are shown in Table R3. The newest U.S. ARSR is the ARSR-4 (Fig. R10), a joint development for the FAA and the U.S. Air Force, which provides long-range two-dimensional data for air traffic control and also three-dimensional data for the continental air defense network. The latest generation of ARSR is represented by the Raytheon ASR-23SS (Fig. R11), which is a fully solid-state radar, whose parameters are also shown in Table R3. The basic features of the ASR-23SS are identical to those of ASR10-SS (see **airport surveillance radar**). *DKB*

Ref.: IEEE (1993), p. 23; Skolnik (1988), pp. 70–88.

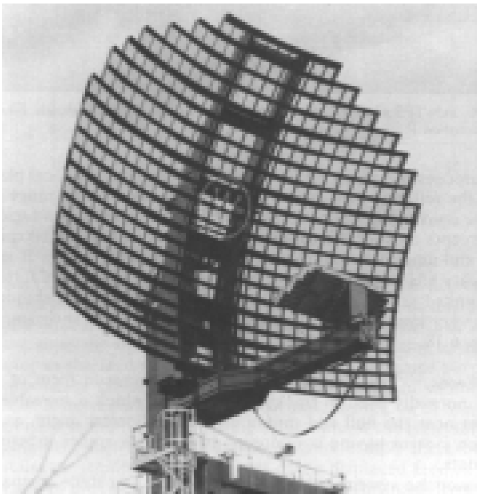


Figure R10 ARSR-4 radar antenna (from Skolnik, 1990, Fig. 6.18, p. 6.22).

Table R3
Major Parameters of ATCR-22 and ASR-23SS

Parameter	Units	ATCR-22	ASR-23SS
Frequency band		L	L
Peak power	kW	2,000	21[1], 40[2]
Pulse width	μs	3	1.0, 100
Pulse repetition frequency	Hz	420 average	302[3]
Average power	kW	2.5	0.7[1], 1.2[2]
Antenna gain	dB	36.5	36.0
Azimuth beamwidth	deg	1.2°	1.25°
Elevation beamwidth	deg	4°, $\csc^2$ to 45°	4°, $\csc^2$ to 45°
Scan rate	rpm	6	5[3]
Receiving noise temperature	K	649	
Detection range on $2\ m^2$, 80%	km	385	290[1], 340[2]
Notes: [1] 8-module transmitter; [2] 16-module transmitter, [3] average for 370-km instrumented range			

An **air(craft) search radar** is one "used primarily for detection of [airborne] targets in a particular volume of interest." Note that ability to measure target position and to form track files is not a necessary feature of search radar, while it is assumed for a surveillance radar. Targets for air search radar are understood to include conventional aircraft, helicopters,

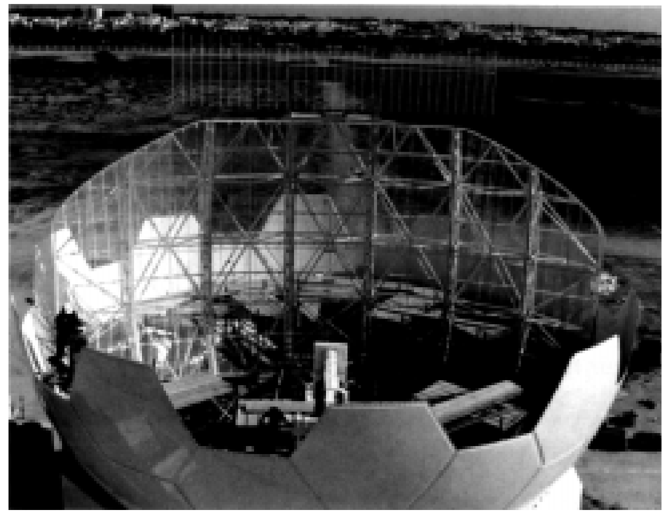


Figure R11 ASR-23SS radar antenna (Raytheon Company photo).

and missiles operating within the troposphere and stratosphere. Civilian uses are primarily for air traffic control (ATC), while military uses include early warning, target acquisition, and ground-controlled intercept (GCI) as well as ATC for military airfields, aircraft carriers, and their approaches. In some cases, joint civil and military needs are served by the same ASR or ARSR. *DKB*

Ref.: IEEE (1993), p. 1,176.

An **air surveillance radar** is "a search radar used to maintain cognizance of selected [air] traffic within a selected area, such as an airport terminal or en route area." It is normally understood that maintaining cognizance includes the function of forming and maintaining track files on detected targets. *DKB*

Ref.: IEEE (1993), p. 1315; Skolnik (1980), pp. 536–541.

An **air traffic control (ATC) radar** is a surveillance radar used to provide aircraft position data to controllers at airports or regional control centers. Both primary and secondary surveillance radars (SSRs) are used for ATC, with the SSR serving as the primary source of data. (See **airport surveillance radar**; **air route surveillance radar**; **secondary surveillance radar**.) *DKB*

All-around-looking radar is the translation of a term used in Russian literature, referring to a search radar having 360° azimuth scan. *DKB*

An **artillery fire control radar** is one designed for determination of target coordinates and projectile impact points with the target. They have high range and angular coordinate resolution and may be used for fire control against aerial, ground-based, or sea targets. As a rule, these are automatic radars. (See also **gunfire control radar**.) *AIL*

Ref.: Popov (1980), p. 343; Friedman (1981), pp. 54–58, 172–184; Macfadzean (1992), Ch. 4.

An **artillery and mortar location radar** is a ground-based, tactical battlefield radar used for locating the source (origin) of artillery, mortar, and rocket fire. The requirements for rapid, all-weather detection of incoming projectiles in an environment that may include natural clutter as well as chaff, has led to radar designs incorporating phased array and frequency scanning techniques combined with fully coherent waveforms and signal processing. A prime example of such a system is found in the Hughes-developed AN/TPQ-36 and -37 "Firefinder" radars operational with U.S. Army and many foreign nations. These X-band, three-dimensional, pulsed doppler, range-gated radars are designed to operate cooperatively. The older AN/TPQ-36 phase scans in azimuth and frequency scans in elevation, while the AN/TPQ-37 employs phased array techniques in both dimensions. Figure R12 shows the AN/TPQ-36 system.

Figure R12 AN/TPQ-36 mortar location radar (Hughes Aircraft Co. photo).

Both radars are capable of fully automatic operation in the detection, verification, tracking, and trajectory analysis of projectiles within their field of view. Backward extrapolation of the track data identifies the location from which the projectile was fired, which is then transmitted by voice or digital communication links to the command center for counterfire action. Typically, a single radar provides coverage over a 90° sector in azimuth and is capable of nearly simultaneous location of approximately 40 separate firing points. The AN/TPQ-36 is positioned within 4 km or less of the front, where mortar, short-range artillery, and rocket fire are most prevalent, with the AN/TPQ-37 located further to the rear to cover the longer range artillery and rocket fire. *PCH*

Ref.: Jane's (1995), p. 167; Brookner (1977), p. 51, (1988), p. 191.

An **automobile radar** is installed in an automobile and intended to prevent accidents and rear-end collisions, to maintain constant speed (distance between vehicles) in a transportation flow, to set up a passive system for protection of the driver and passengers, and to control an interlock protection device. Automobile radars are built with continuous emission and pulse emission. In continuous radars, FM signals are used to determine the range, speed, and approach signal. Pulse vehicle radars emit nanosecond duration pulses and allow one to detect and resolve closely located objects.

Due to the high cost and absence of operating experience, automobile radars have not been widely used. *AIL*

Ref.: Skolnik (1988), p. 445; *IEEE Int. Conf. Radar-95*, Washington, DC, 8–11 May 1995, papers by: Woll, J. D., "Vorad Collision Warning Radar," pp. 369–372; Rohling, H., and Lisel, E., "77 GHz Radar Sensor for Car Application," pp. 373–379; Eriksson, L. H., and Ås, B.-O., "A High Performance Automotive Radar for Automatic AICC," pp. 380–385.

Automated radar is radar in which an operator accomplishes the search for and acquisition of targets, while tracking of acquired targets, measurement of their coordinates, and determination of their motion parameters are done automatically by computers. *AIL*

Ref.: Popov (1980), p. 10.

Automatic radar is radar in which the search for and acquisition of targets, as well as radar information processing, are accomplished automatically using computers. The functions of a human operator in such radars mainly are restricted to observation of radar operation and its servicing and maintenance. Information processing is understood to mean evaluation of the coordinates and the number of targets; acquisition of target trajectories; calculation of target movement parameters (velocity, course, etc.); and, on that basis, coordinates smoothed and predicted in some segment of time. Most modern radars are automatic. *AIL*

Ref.: Kuz'min (1967), p. 8.

A **bistatic radar**, by definition, is one in which the transmitter and receiver sites are separated by a significant distance. The received signal power S for a bistatic radar system is expressed as

$$S = (P_t G_t)\left(\frac{\sigma_b}{4\pi R_1^2}\right)\left(\frac{A_r}{4\pi R_2^2}\right)$$

where P_t is the peak transmitted power, G_t is the transmit antenna gain, σ_b is the bistatic value of the target radar cross section, expressing the ability of the real target to scatter energy incident from the direction of the transmitter into the direction of the receiver, R_1 is the range from the target to the transmitter, R_2 is the range from the target to the receiver, and A_r is the effective aperture area of the receiving antenna.

Two bistatic radar types may be distinguished:

(1) Narrow-angle bistatic geometry, in which the transmit and receive paths intersect at the target with angles less than about 45°.

(2) Wide-angle bistatic geometry, where the paths intersect at angles from 45° to 180°.

In the first case, the radar modes and target characteristics are similar to those of monostatic radar. The observed range delay sum for the two path segments and the target doppler shift vary in a manner similar to that for monostatic geometry. Synchronization is required between the two radar sites, both in times of transmission and reception for delay measurement and in beam position or scanning. The target position is measured by a combination of range delay sum and angle of arrival of the target echo at the receiving site. The target RCS is comparable to that of monostatic radar, except for the absence of corner-reflector enhancement.

In the wide-angle bistatic system, target RCS statistics may differ significantly from those of monostatic radar, especially for bistatic angles approaching 180°. (See **RADAR CROSS SECTION, bistatic.**) Also, for this case, the relationships between range delay sum and observed target doppler shift and the corresponding target position and velocity may differ greatly from those in monostatic and narrow-angle bistatic radars. In the limit, as the bistatic angle approaches 180°, the observed doppler drops to zero and the range sum delay becomes equal to the site separation, independent of the target velocity and position along the path connecting the two sites.

Bistatic radar has been developed for use in long-range, over-the-horizon (OTH) surveillance. The AN/FPS-118, operating in the 5 to 28 MHz region of the HF band, and consists of very large (1,100m long) transmit and receive antenna arrays separated by approximately 200 km. The transmitter array forms a 7.5° steerable beam in azimuth while the receiving array divides the radar's 60° azimuth coverage sector into 24 separate beamwidth each 2.5° wide. In normal operation, the AN/FPS-118 establishes surveillance over a sector 60° in azimuth by 800 km in range. A sequential scan technique is used to illuminate four contiguous, 7.5° by 800-km sectors. Five parallel receive beams, each with 2.5° center spacings, are formed to be coincident with each range-azimuth sector. The AN/FPS-118 system, developed to allow long-range detection of low-flying cruise missiles, reportedly demonstrated detection of targets at ranges as large as 3,000 km.

Bistatic radar has also been studied as a possible answer to the problem of detecting very-low-RCS or "stealth" targets. Most practical airbreathing vehicle exhibit enhanced cross sections at broadside aspects, and bistatic radar has the potential for exploiting this characteristic. While the use of bistatic radar in a ground-based air defense system provides some important technical advantages, its use would not necessarily "solve" the problem of defense against stealth targets. From a practical standpoint, the total life-cycle cost of the number of bistatic sites required would have to be weighed against the cost of deploying fewer, but more powerful monostatic radars, or a mix of the two types of radar. (See also **over-the-horizon radar.**) *PCH*

Ref.: Skolnik (1990), Ch. 25; Willis (1991); Jane's (1995), p. 191; Currie (1989), Ch. 5; Cantafio (1989), Chap 5.

A **carrier-controlled approach (CCA) radar** is "an aircraft-carrier radar system providing information by which aircraft approaches may be directed by way of radio communication." It is the equivalent of a ground-controlled approach (GCA) radar except that it is mounted on an aircraft or helicopter carrier whose flight deck serves as the runway for aircraft landing. The stringent requirements for accuracy and control-loop bandwidth in CCA systems, along with the limited types of aircraft to be handled, have led to inclusion of precision tracking pencil-beam radars at K_u- or K_a-band as elements in the final approach stages of modern systems. The "airport surveillance radar (ASR)" function in CCA systems is performed by radars similar to land-based ASRs.

Examples of CCA radars in the U.S. Navy are the AN/SPN-6 and AN/SPN-43 S-band surveillance radars and the AN/SPN-35 (X-band), AN/SPN-41 (K_u-band), and AN/SPN-42 (K_a-band) precision approach radars. Figure R13 shows the azimuth antenna assembly of the AN/SPN-41 CCA system. The AN/SPN-44 speed-detecting radar is an X-band CW doppler radar used to provide approach velocity data to the landing signal officer. *DKB*

Ref.: IEEE (1993), p. 161; Friedman (1981), p. 61; Law (1986), pp. 284–297, 373-382.

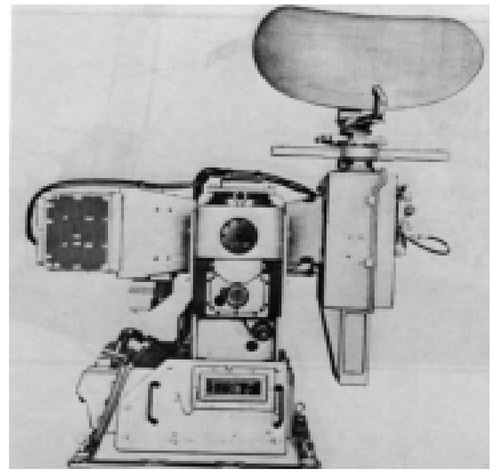


Figure R13 Azimuth scanning antenna of AN/SPN-41 carrier-controlled-approach radar system (from Law, 1983, Fig. 4.46, p. 377).

carrier-frequency-agile radar (see **frequency-agile radar**).

Chain radar refers to a series of instrumentation radars located at stations along a test range and capable of passing tracks from one to the next station as the test mission progresses. Such chains were established at the White Sands Missile Range and the Atlantic Missile Range during the 1950s, and are features of most ranges at which long-range vehicles are tested. The term should not be confused with the *radar chain* installed in Britain before World War II, known as Chain Home. *DKB*

Ref.: Barton (1964), p. 505; Scavullo (1965), pp. 230–236.

chirp radar (see **PULSE COMPRESSION**).

Close-in radar is radar operating at short ranges, from fractions of units to a dozen or hundreds of meters. Range in such radars is commensurate with the geometric dimensions of mutually acting objects and with errors in issuance of execution commands. Close-in radars are built on the same principles and bases as other radars. They can be active and passive. Close-in radars can be categorized as pulse, continuous, and doppler depending upon the type of signals used. Compared with other radars, differences are linked with emission and receipt of electromagnetic oscillations by antennas at short ranges, i.e., they mainly operate in the close-in radiation zone (in the Fresnel region), which imposes special features on shaping the antenna radiation pattern. Moreover, differences involve the power of the reflected signal, both regarding output and its duration, nature of output commands, and signal processor operating speed. Close-in radars are used in different closure, docking, and landing systems, to ensure safety in transport systems, low-altitude flights, collision-warning equipment, and the like. *AIL*

Ref.: Shelukhin (1989), pp. 5–7, 178–226.

Coastal radar is radar designed for use in a system of coastal observation of the ocean near the shore. Coastal radars may be fixed and mobile. They are categorized based on tactical assignment as mobile and aerial target acquisition, fire control, guidance, identification, and other radars. *AIL*

Ref.: Popov (1980), p. 48; Jane's (1995), pp. 173–230.

Coherent radar uses coherent signals and signal processing. (See also **COHERENCE**; **DETECTION, coherent**; **MTI, coherent.**) Coherence provides the ability to maximize signal-to-noise ratio, measure target radial velocity, and provide MTI and other doppler-based clutter rejection techniques. Coherent pulsed radars are the most widely used types, although coherent CW radars appear in specialized applications such as missile guidance and police speed-control radar.

Coherent pulsed radars are classified as truly coherent (Fig. R14a), or pseudocoherent, also known as coherent-on-receive (Fig. R14b). Both types are known as internally

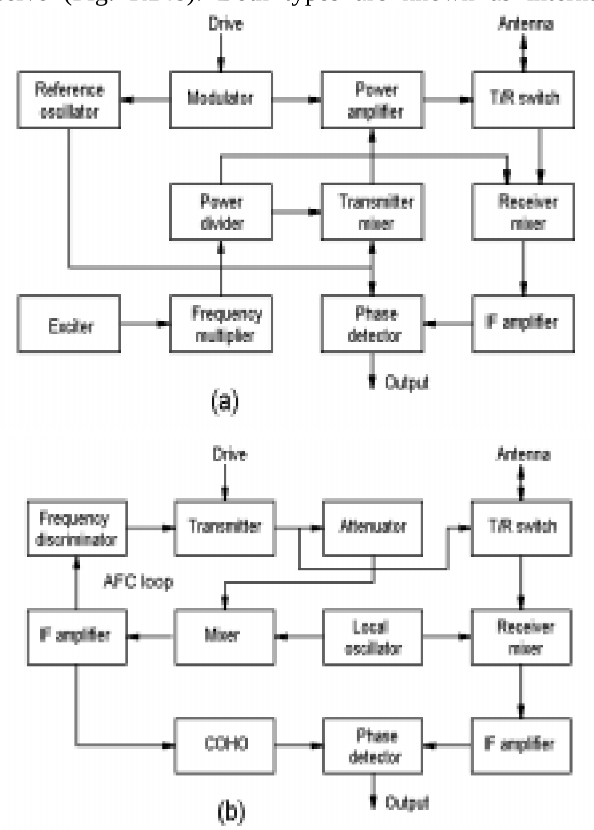


Figure R14 Coherent pulse radar: (a) coherent; (b) pseudocoherent.

coherent, to distinguish them from externally coherent types, which use the clutter itself as a phase reference. (See **MTI, noncoherent.**) In coherent radars, the IF phase reference is provided by a coherent oscillator (COHO), which in the truly coherent system is also used to offset the transmitted carrier from the stable local oscillator (STALO).

In the system shown, the master quartz oscillators generate highly stable oscillations which are supplied to the frequency multipliers. At the output of the frequency multiplier, continuous oscillations are generated at the frequency equal to the STALO frequency f_s, from which oscillations at the transmitter frequency are generated $f_t = f_s + f_c$. To ensure pulse operation of the transmitter and reference oscillation generator one uses a modulator. At the output of the phase detector video pulses are formed, the amplitude of which will depend on the phase shift between the received and referenced signals. This configuration is also termed a coherent radar with power-amplifier transmitter, or a master-oscillator-power-amplifier (MOPA) system

In a pseudocoherent radar (Fig. R14b) one uses a COHO which is phased by each transmitted pulse as it comes from the IF amplifier of the AFC loop. The transmitter generates a noncoherent train of pulses with constant carrier frequency but with random initial phases. Part of the transmitter power is sent to the input of the AFC system and then to the IF amplifier of this system to phase the coherent heterodyne. Signals from the COHO and receiver IF amplifier are applied to the input of the phase detector. At the output of the phase detector video pulses are formed, the amplitude of which depends on the phase shift between the received signal and the COHO reference. This radar approach is used when the transmitter is inherently noncoherent (e.g., a magnetron), and is termed a power-oscillator transmitter. Advantages of a radar with internal coherence lie in their high sensitivity and the possibility of measuring the doppler frequency shift with good accuracy. Disadvantages are the relative complexity and need to ensure high stability of all the oscillators used in the system.

In a radar with external coherence, the COHO is phased not by the transmitted signal but by a clutter signal, or the clutter itself is used as the reference voltage. Advantages with a radar with external coherence lie in the possibility of suppressing extended interference. A disadvantage is that the detection of moving targets in them is possible only if there is extended clutter present. If the clutter does not enter the antenna pattern of the radar, then phasing of the COHO is controlled only by target signals, which leads to suppression of those signals. *AIL, SAL*

Ref.: Skolnik (1962), p. 118, (1980), Ch. 4; Barton (1964), pp. 192–193, (1988), Ch. 5; Schleher (1991), p. 80.

coherent-on-receive radar (see **coherent radar**; **MOVING-TARGET INDICATION**).

Collision-warning [-avoidance] radar is used on a land, sea, or air vehicle to detect nearby vehicles posing threat of collision. For land vehicles, the application is relatively recent, and the predominant technology is that of short-range millimeter-wave radar. (See **automobile radar.**) Radar warning of collisions at sea has long been provided by the conventional navigation and surface search radars, supplemented by special computers or interpretation of display data by adequately skilled operators. (See **navigation radar.**) Air collision warning is dependent on proper operation of the air traffic control system to prevent conflicting flight paths, supplemented by exploitation of transponder signals triggered by air traffic control secondary surveillance radars (SSRs) or by onboard collision-avoidance systems (CAS). *DKB*

Ref.: Kayton (1990), pp. 236, 285

A **conical-scan radar** is a pencil-beam tracking radar using rapid scan of the beam axis in a small cone centered on the tracking axis (see **SCAN, conical**). *DKB*

A **conical-scan-on-receive-only radar** is a pencil-beam tracking radar in which the transmitting beam remains centered on the tracking axis while the receiving beam performs conical scan. The principal advantage of this technique is that it deprives the target of information on the frequency and phase of the scan, preventing a self-defense jammer from exploiting efficient AM jamming techniques such as inverse gain and spot scan-frequency jamming. Unless the intercept receiver can detect unintended modulation at the scan frequency, or can select a fixed, known scan rate for the tracker type, the AM jamming must cover the band of potential scan frequencies, reducing its effectiveness and generally precluding breaking of the track. *DKB*

Ref.: Schleher (1986), pp. 151, 248; Skolnik (1990), p. 9.27; Chrzanowski (1990), p. 141.

A **continuous-wave (CW) radar** is one that transmits a continuous wave, as opposed to a pulsed signal. A CW waveform may be described by the expression

$$E_t = E_0 \cos \omega_0 t$$

By definition, a CW radar has a duty factor of 100%, and a pulse repetition frequency (PRF) of zero. CW radar has an inherent ability to detect targets on the basis of their doppler frequency shift f_d:

$$f_d = \frac{2v_c}{\lambda} = \frac{2v_c f_0}{c}$$

where v_c is the target's radial or closing velocity with respect to the radar, λ is the radar wavelength, f_0 is the radar carrier frequency, and c is the speed of light.

Figure R15(a) shows a simplified block diagram of an unmodulated CW radar. The transmitter signal consists of a continuous oscillation at frequency f_0. If the target is moving with respect to the radar with velocity v_r, the radar's received signal will be shifted in frequency from the transmitted frequency by an amount of the target's doppler frequency $\pm f_d$. By convention, if the target is approaching the radar, the sign of the doppler frequency is positive, and if it is receding, the sign is negative. By a heterodyning process, the radar receiver isolates and detects the target doppler signal at frequency f_d. The doppler amplifier amplifies the target signal so that it can be displayed or heard. The frequency response characteristic of the amplifier (Fig. R15(b)) is designed to eliminate return from stationary targets (clutter) while passing through signals within the range of expected target doppler frequencies.

CW radars are required to detect signals that may be on the order of 200 dB smaller in amplitude than the transmitted

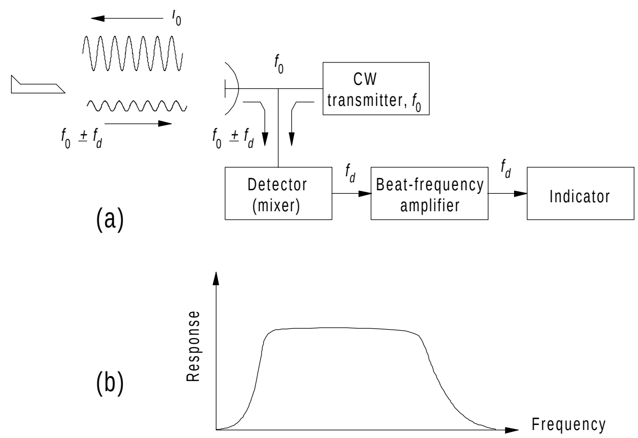


Figure R15 Simplified CW radar block diagram and frequency response (after Skolnik, 1980, p. 70).

signal, and unlike a pulsed radar, to do so while the transmitter is operating. To realize the required transmit-to-receive signal isolation dictates the use of separate transmitter and receiver antennas in CW radar, as well as careful control of transmitter noise and microphonics that may enter the receiver through transmitter leakage. These formidable engineering problems notwithstanding, successful CW radars have been in use for nearly four decades, primarily as low-altitude target acquisition, tracking, and semiactive radar homing guidance illuminators. Figure R16 shows the dual antenna of the Sea Sparrow tracking illuminator system, designed by Raytheon for use on NATO and Canadian ships.

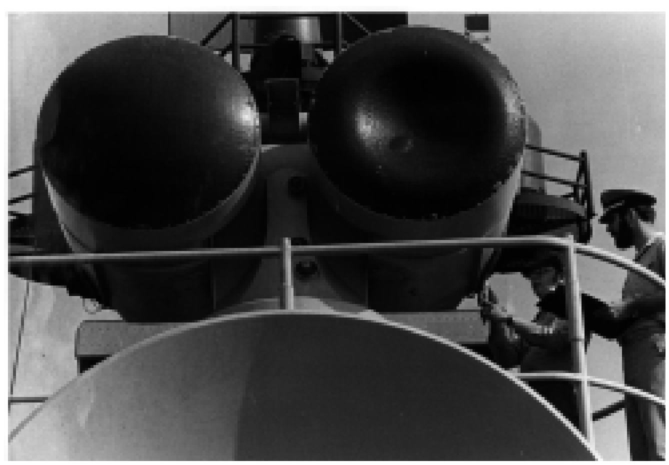


Figure R16 Antenna system for Sea Sparrow CW tracking illuminator radar (Raytheon photo).

A frequency-modulated CW radar is one in which a ranging capability is provided by frequency modulation of the transmitted signal. An unmodulated CW radar, unlike a pulsed radar, cannot determine target range based on the time delay between transmit and reception of a radar pulse reflected from the target. By imposing a frequency modulation (FM) on the CW carrier however, a timing mark can be created that will provide a reference from which target range can be derived. Figure R17 illustrates the principle of FM-CW radar ranging.

Figure R17(a) represents a linear FM process. The solid line represents the transmitted frequency as function of time, while the dashed line represents an echo signal from a fixed target. If the target is at range *R*, its echo signal will be received after time $t_r = 2R/c$. When the echo signal is heterodyned with a portion of the transmitted signal, a beat frequency f_b will be produced, which, in the absence of a doppler shift, will be due only to the target's range, that is:

$$f_b = \left(\frac{df_0}{dt}\right)t_r = \frac{2R}{c}\left(\frac{df_0}{dt}\right)$$

where df_0/dt is the rate of change of the CW carrier frequency.

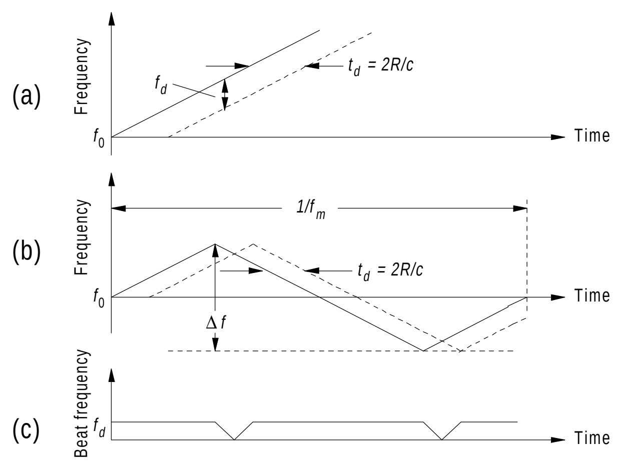


Figure R17 FM-CW ranging technique (after Skolnik, 1980, Fig. 3.10, p. 82).

Figure R17(b) shows a triangular frequency modulation, illustrating one method of implementing FM-CW in a practical radar for which the modulating frequency excursion Δf is limited. Figure R17(c) shows that the beat frequency for this fixed target example is constant except at the frequency turn-around points. If the frequency is modulated at rate f_m over the frequency excursion Δf, the beat frequency is

$$f_b = \frac{2R}{c}2f_m\Delta f$$

and thus the measurement of the beat frequency determines target range:

$$R = \frac{f_b c}{4f_m\Delta f}$$

The principle of FM-ranging has been described based on a stationary target. Moving targets will, of course, superimpose their own doppler shift on the beat frequency, yielding an erroneous measure of target range. This situation can be corrected, within limits, by determining the average beat frequency over many periods of the FM waveform.

Interrupted CW radar is a term sometimes used to describe coherent pulsed doppler radar using a high duty factor. The feature distinguishing this mode of operation from other pulsed doppler radar is the use of a single range gate for reception and processing of echo signals.

A multifrequency CW radar is designed to measure range by comparing the phase of signals transmitted and received on different carrier frequencies and hence differing in phase shift as a function of target range. The method is practical only when echoes from a single target can be resolved in angle or doppler. Hence the technique is seen primarily in instrumentation applications, where a transponder can be used. A variant of the technique applies two or more sinusoidal modulations to the carrier, creating sidebands that meet the criteria for differential phase shift.

Spillover in CW radar refers to the unintended coupling of transmitter power to the receiver (also known as *feedthrough*), and constitutes a major design problem in this type of radar. For example, for a receiver noise figure of 4 dB, the noise level in a 1-kHz filter may be –140 dBm. In a system transmitting 1 kW (+60 dBm), the noise level is 200 dB below the transmission, and signals near noise level must be detectable to take full advantage of the radar resources. If the receiver dynamic range is 100 dB, saturation will occur with an input of –40 dBm, and spillover must be at least 100 dB below the transmitted power. Systems of this power level must use separate antennas for transmitting and receiving, with adequate isolation provided by a septum between them or by using a tunnel structure around one or both antennas (Fig. R18). In addition, echoes from short-range clutter must be held below the saturation level, often requiring special circuit such as the feedthrough nulling loop used in the Hawk high-power illuminator. *DKB, PCH*

Ref.: Barton (1978), pp. 169–175, 177-185; Nathanson (1969), pp. 360–369; Skolnik (1980), pp. 68–84, 95, (1990), p. 14.19–14.28; Hovanessian, (1984), p. 84.

Figure R18 AN/MPQ-46 tracking illuminator for the Hawk missile system (Raytheon photo).

A **counterbattery radar** is designed to acquire and determine the coordinates of artillery systems, missiles, and mortars at several points of the trajectory of their flight and extrapolation of their trajectory, with subsequent determination of firing position location. The doppler effect and moving-target indicator systems may be used to eliminate ground clutter. (See **artillery and mortar location radar.**) *AIL*

Ref.: Popov (1980), p. 340; Brookner (1977), p. 51, (1988), p. 191.

A **doppler radar** is one "which utilizes the doppler effect to determine the radial component of relative target velocity or to select targets having particular radial velocities." (See **coherent radar**; **pulsed-doppler radar.**) *DKB*

Ref.: IEEE (1993), p. 382.

A **doppler-navigation radar** is an airborne radar with several (usually three) beams observing the underlying surface. The three frequency shifts are processed to determine the aircraft velocity vector with respect to the surface, and this vector is integrated to find the aircraft displacement from an initial position fix. *DKB*

Ref.: Skolnik (1980), pp. 92–95; Barton (1980), pp. 125–129.

A **doppler weather radar** is a meteorological observation radar in which the doppler spectrum of the observed precipitation or air mass is analyzed to yield wind-field data. Both surface-based and airborne meteorological radars use doppler observations to identify conditions threatening safety of personnel or property. The U.S. Nexrad is the primary doppler weather radar used in nationwide weather forecasting and storm warning. For air safety, most modern aircraft carry weather avoidance radars, some models of which display doppler data on turbulence. The FAA has obtained Terminal Doppler Weather Radars to monitor wind shear conditions in the vicinity of major airports. The parameters of this radar are shown in Table R4. *DKB*

Ref.: Doviak (1984); Skolnik (1988), pp. 395–426, (1990), Ch. 23; Heiss, W. H., McGrew, D. L., and Sirmans, D., "Nexrad: Next Generation Weather Radar (WSR-88D)," *Microw. J.*, Jan 1990, pp. 79–98; Michelson, M., Shrader, W. W., and Wieler, J. G., "Terminal Doppler Weather radar," *Microw. J.*, Feb 1990, pp. 139–148; Schleher (1991), pp. 537–560.

Table R4
Parameters of Terminal Doppler Weather Radar

Parameter	Units	Value
Frequency	GHz	5.6–5.65
Peak power	kW	250
Pulse width	μs	1.1
Pulse repetition frequency	Hz	2,000
Average power	W	550
Antenna diameter	m	7.6
Antenna gain	dB	50
Beamwidth	deg	0.55°
System noise factor	dB	2.3

dual-frequency radar (see **frequency diversity radar**).

An **early-warning radar** is a long-range search radar intended to give initial warning of the approach of hostile aircraft. This function is often included with air surveillance, ground- or air-controlled intercept, and target acquisition radars. (See **air surveillance radar.**) *DKB*

A **radar with electronic scanning** is one in which scanning (i.e., movement of the antenna beam in space) is accomplished by an electronic method. One can distinguish amplitude, frequency, and phase electronic scanning. As a rule such radars employ antenna arrays. (See **ARRAY, phased.**) Radars with electronic scanning are widely used, primarily in military systems. Examples of radars with electronic scanning are AN/TPS-59, W-2000, AN/FPS-108, AN/FPS-115, and AN/FPS-85. *AIL*

Ref.: Skolnik (1980), Ch. 8; Leonov (1988), pp. 89–140; Barton (1988), Ch. 4.

fighter guidance radars (see **airborne intercept radar**).

A **fire control radar** is a radar whose purpose is to direct the firing of weapons. An early example of a fire control radar is the SCR-584 (Fig. R19), an X-band, conically scanning radar used to track aircraft targets and direct (point) antiaircraft artillery (AAA). Airborne intercept (AI) radars, surface-to-air (SAM) system tracking and illumination radars, and certain multifunction array radars (MFARS) can all be considered fire control radars.

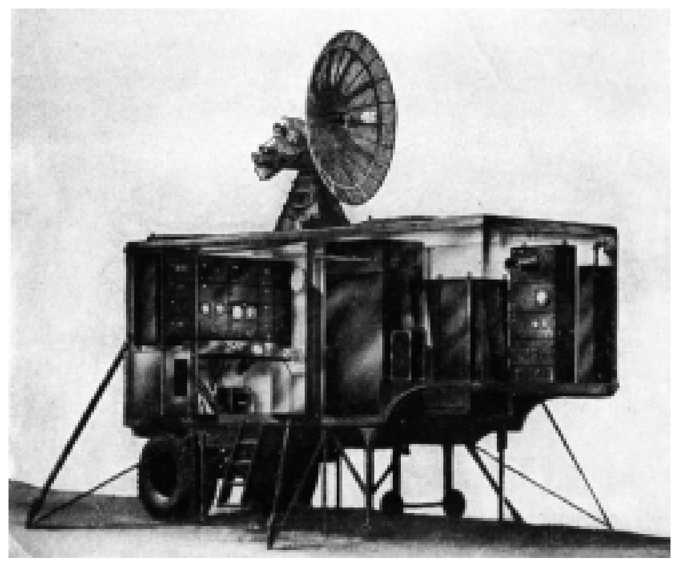


Figure R19 SCR-584 antiaircraft fire control radar, the first microwave fire control radar of World War II.

Antiaircraft fire control radars designed since the 1950s rely primarily on monopulse tracking techniques to reduce the target scintillation error encountered in conically scanned radars. A typical mobile system is the Roland, a command-guided, short-range missile system (Fig. R20). A comparable Russian system (Fig. R21) includes both guns and short-range missiles, controlled by a K_u-band monopulse tracker.

Figure R20 Roland short-range missile system, using monopulse tracking and guidance radar.

Figure R21 Russian antiaircraft gun and missile system, using S-band search and K_u-band fire control radar.

Fire control radars for long-range SAM systems of the 1950s through 1980 were of two types: the mechanically steered, single-target monopulse tracker such as the Nike Hercules (Fig. R22), and the electromechanically sector-scanned tracker such as Fan Song (Fig. R23). The Nike Hercules target tracker achieved accuracies approaching those of the best instrumentation radars (0.1 to 0.2 mrad in angle), and has since its retirement from tactical service been adapted for test range instrumentation. In the command guidance application, two separate radars were required, one for the target and a second for the interceptor missile. Accurate alignment of the two antenna pedestals was a critical problem. In subsequent Russian development of the Land Roll radar, the missile tracking antennas were mounted on the same pedestal as the larger target tracker, to avoid misalignment, and terminal guidance was performed using differential tracking within the beam of the larger antenna.

The sector-scanning process used in the Russian SA-2 and SA-3 SAMs has the advantage of differential tracking of target and missiles with the same scanning beams (one for azimuth and another for elevation). The tracking method uses pulse amplitude envelopes from the scanning beams in split-gate angle trackers, in a procedure similar to many Western ground-controlled approach radars. Target scintillation errors and response to AM jammers are the major source of error in these systems. The large warhead used in the SA-2 missile tends to compensate for these inaccuracies.

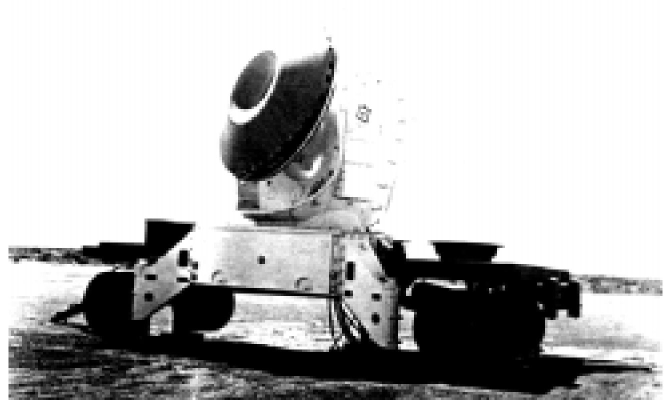


Figure R22 Nike Hercules monopulse fire control radar.

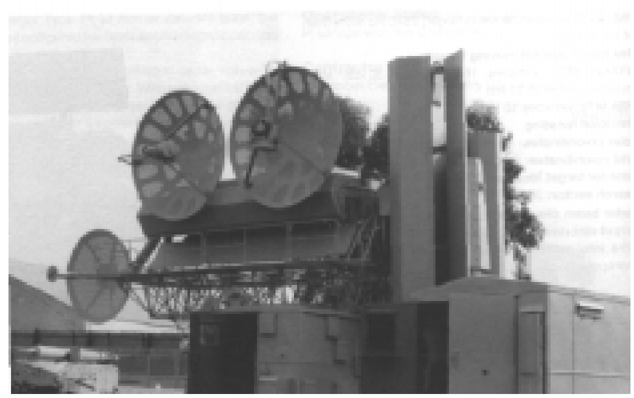


Figure R23 Fan Song (SA-2) sector-scanning fire control radar.

More recent Russian SAM fire control radars are multiple-target phased-array trackers, using space-fed lens arrays such as Flap Lid (Fig. R24), or reflectarrays such as Top Dome (Fig. R25). These radars operate in target-tracking modes similar to those of the Patriot and Aegis MFARs, but since the Russian approach dedicates the fire control radar to target (and sometimes missile) tracking, they provide higher data rates on engaged targets (typically six targets at 10 Hz update rate). The SA-10 Flap Lid tracks the interceptors for command midcourse guidance, and provides uplink data and target illumination for semiactive homing. In the SA-12 system, the similar Grill Pan fire control radar performs only target tracking, illumination and uplink data being provided by separate illuminators slaved in angle to the radar data.

The dedicated fire control radar approach permits the radar to operate at X-band for greater effective radiated power and accuracy, and also provides the long-dwell pulsed dop-

Figure R24 SA-10 Flap Lid fire control radar for the SA-10 SAM system.

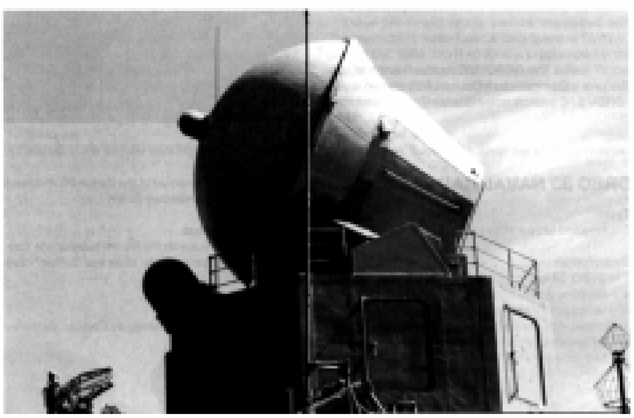


Figure R25 SA-N-6 Top Dome fire control radar for the SA-N-6 naval SAM system.

pler waveforms needed for clutter rejection. Scheduling and control of the radar are also simplified, avoiding the software problems usually encountered in MFAR design and operation. *DKB, PCH*

Ref.: Barton (1988), pp. 477–485; Jane's (1995), pp. 108–171, 231–254.

A **frequency-agile (FA) radar** is "a pulse radar in which the transmitter carrier frequency is changed between pulses or groups of pulses by an amount comparable to or greater than the pulse bandwidth." The frequency is usually jumped in a pseudorandom way over the available tuning band (usually 5 to 10% of the center frequency). When the jumping is on a pulse-to-pulse basis the use of doppler filtering for rejection of clutter is precluded, but the following advantages are gained:

(1) Avoidance of spot and repeater jamming, at least at ranges within that of the jammer platform.

(2) Imposition of wide-bandwidth processing requirements on intercept and antiradiation missile receivers.

(3) Improvement of detection probability by producing rapidly fluctuating target echoes (Swerling Case 2 in place of Swerling Case 1).

(4) Reduction of target glint error in tracking radars.

(5) Filling of propagation null in elevation coverage, at angles above the horizon null.

(6) Decorrelation of echoes from extended clutter sources, providing for signal integration gain with respect to such clutter.

(7) Decorrelation of multipath error components in tracking radar, permitting smoothing to reduce the error.

The use of burst-to-burst agility does not reduce vulnerability to ECM or ARM missiles by as great a factor, but offers the advantages outlined below under frequency-diversity radar.

The frequency-agile transmitter may use either a tunable magnetron (spin-tuned or plunger tuned), or a wideband power amplifier driven by a variable-frequency source. Tuning of the receiver can be a problem when a magnetron is used, requiring specialized techniques to sense the resonant frequency of the magnetron cavity just before the pulse is transmitted, or to tune the local oscillator rapidly after the transmission. When an amplifier transmitter is used, the variable-frequency drive can also provide the necessary shift of the receiving local oscillator. *DKB*

Ref.: IEEE (1993), p. 526; Barton (1977), (1988), p. 86; Skolnik (1990), p. 9.17.

A **frequency-diversity (FD) radar** uses two or more carrier frequencies transmitted simultaneously (or in rapid succession during each pulse), or in sequential bursts of pulses having constant frequency within the burst. The advantages of frequency diversity are

(1) To force a jammer to spread its power at least over those frequencies used simultaneously.

(2) To improve detection performance by decorrelating the target echo, creating for dual diversity a Swerling Case 3 signal and for high order of diversity a closer approach to the Case 2 signal.

(3) Reduction of glint and multipath errors for tracking and three-dimensional surveillance radar.

(4) Partial filling of propagation nulls in elevation coverage, above the horizon null.

(5) Decorrelation of echoes from extended clutter sources.

(6) Improved reliability through redundancy, when separate transmitters and receivers are used for dual diversity.

(7) Increase in available transmitter power, when duplicate transmitters are used with tubes having limited average power.

In cases where the different frequencies are transmitted in contiguous subpulses within each pulse, a common wideband receiver front end may be used, with subsequent filtering into the two or more diversity channels as shown in Fig. R26. A small delay is used in the frequency channel first transmitted, aligning the two video outputs for subsequent processing and display. *DKB*

Ref.: Barton (1977), (1988), pp. 85, 371, 529.

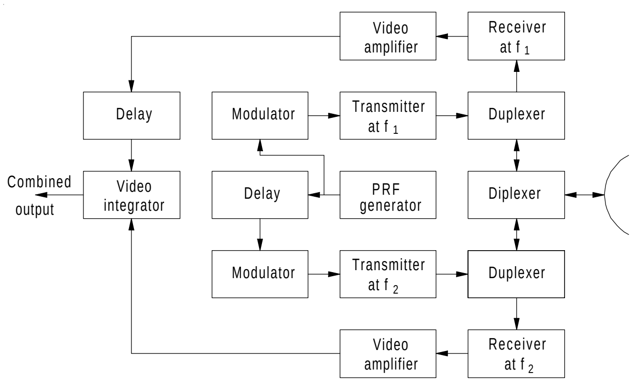


Figure R26 Block diagram of dual-diversity radar using parallel frequency channels.

A **frequency-scanning radar** is one in which electronic scan is implemented in a frequency sensitive feed illuminating a circular parabolic reflector or as a rectangular array of such radiating structures. (See **three-dimensional radar**; **ARRAY, frequency scanned**; **SCANNING, frequency**). *DKB*

A **gap-filler radar**, in a ground-based air defense network, is one that is used to augment coverage in a region or zone not properly covered by existing, deployed, fixed or mobile radar sites. For example, two surveillance radars separated by a distance of 100 km may provide coverage of the airspace out to some maximum range and altitude, and a degree of overlapping coverage in to some minimum range, but due to line-of-sight restrictions, terrain masking, or both, no coverage against low-altitude targets inside that limit. A low-altitude gap-filler radar could be deployed to cover the defined vulnerable region. It is more economical to deploy special-purpose radars to fill in small gaps in coverage than to field enough large radars to ensure that no gaps exist, in that the latter option typically would result in a high degree of unneeded overlapping coverage.

Figure R27 shows the Swedish *Giraffe*, a C-band, pulsed doppler low-altitude search-and-acquisition radar with a boom-mounted antenna. The *Giraffe* is made in several versions and used by many nations as a gap filler radar. *PCH*

Ref.: Barton (1964), p. 497; Jane's (1995), p. 75.

Radar in geographical research is radar used in studying natural resources in the environment. Radar in geographic studies allows one under any weather conditions to survey large areas of the earth, map-making, continuous observation of the snow cover, forests, and condition of agricultural crop plantings, to predict harvests. It is possible to obtain information in the search for new regions to develop the exploitation of minerals, and also for ice coverage and icebergs (their sizes

Figure R27 Swedish *Giraffe* gap-filler radar (from Jane's, 1995, p. 75).

and condition). Figure R28 shows a multifrequency airborne radar designed for remote sensing of earth resources.

Analysis of the radar information allows one to obtain data on the degree of sea swelling and direction of surface layer currents of seas and oceans and to study dispersion effects of shallow waters and much other information. The main radar equipment in geographic research are earth-surface-scanning radars. *AIL*

Ref.: Kondratenkov (1983), pp. 4–7; Glushkov (1981), p. 51; Skolnik (1988), Ch. 6; Ulaby (1981), (1982), (1986).

Figure R28 Multifrequency airborne earth resources remote sensing radar, at 1993 Moscow air show.

A **ground-attack radar** is an airborne radar used to locate fixed and moving surface targets for attack with bombs, missiles, or aircraft gunfire. Located in the nose of a fighter-bomber, the radar normally scans a limited sector from the nose of the aircraft, producing a high-resolution map of the area ahead of the aircraft. The ground-attack capability is often included as an operating mode of a multifunction airborne radar, sharing the equipment and time with airborne intercept, terrain following, and other radar functions.

Ground-attack modes of operation may include real-beam mapping, in which the antenna executes a sector scan to generate a sector map of the PPI type; squinted-beam SAR mapping, in which a strip map is produced by a beam at fixed angle (e.g., 30 or 60°) from the nose of the of the aircraft; and a spotlight SAR mapping mode, in which a high-resolution map is produced by holding the beam on a specified target region as the aircraft flies at an angle to that region. *DKB*

Ref.: Skolnik (1988), pp. 295–302; Barton (1991), pp. 9.12–9.17; Carrara (1995).

A **ground-controlled approach (GCA) radar** is "a ground radar system providing information by which aircraft approaches to landing may be directed via radio communications; the system consists of a precision approach radar (PAR) and an airport surveillance radar (ASR)." The ASR is similar to those used generally for air traffic control (ATC). The specialized PAR has no counterpart in civil ATC, but is used at military air bases to supplement or replace the instrument landing system (ILS). (See also **airport surveillance radar**; **precision approach radar**.) *DKB*

Ref.: IEEE (1993), p. 567.

A **ground-controlled intercept (GCI) radar** is "a radar system by means of which a controller on the ground may direct an aircraft to make an interception of another aircraft." The radar is usually a long-range surveillance radar, often of the three-dimensional type, whose data are used by ground controllers in an air defense system to vector manned interceptors against air targets. If a 2D surveillance radar is used, it will be supported by one or more height-finding radar to provide target altitudes at rates consistent with the requirements for interceptor guidance. Radar reports from each scan are formed into track files on the hostile penetrators and the interceptors to be controlled. The appropriate commands are generated by computer or judgment of the controllers, and transmitted by voice or data link to the interceptor. (See also **air urveillance radar**; **height finding radar**; **three-dimensional radar**; **two-dimensional search radar**.) *DKB*

Ref.: IEEE (1993), p. 567.

A **ground-mapping radar** (ground-target-detection radar) is an airborne or spaceborne imaging radar designed to form accurate, high-resolution images of the land surface. Synthetic aperture radar technology is usually applied, with resolutions from a few meters down to fractions of one meter. Mapping radar, originally applied to military reconnaissance and intelligence, is now used widely for civil remote sensing purposes. This type of radar is used for detecting ground targets (tanks, vehicles, self-propelled weapons, etc.). For determination of moving targets in order to distinguish them on a background of signals reflected from local objects (clutter), in such radar one usually uses MTI based on the doppler effect. (See **MOVING TARGET INDICATION.**) *DKB, AIL*

Ref.: Popov (1980), p. 342; Skolnik (1990); Harger, 1970; Burenin, 1972; Kovaly, 1976; Goryainov, 1988; Cantafio, 1989; Curlander, 1991; Carrara, 1995.

guidance radar (see **fire control radar**).

A **gunfire control radar** is a short-range tracking radar, or for naval targets a sector-scanning track-while-scan radar, used to control the aiming of antiaircraft or antiship artillery. Typical tracking ranges are from 5 to 15 km, angle accuracies are a fraction of one mrad, range accuracies are a few meters, and data rates 1 to 10 Hz. Most gunfire control radars operate at X- to K_a-band, although tank-mounted units may use W-band or laser radar. Advanced gunfire control radars may track the outgoing projectiles, forming a sampled-data control loop to bring subsequent rounds onto the target.

Figure R29 shows a tank-mounted X- and K_a-band antiaircraft gunfire control radar, located on the front of a turret between two rapid-fire guns. An X-band target-acquisition radar antenna is mounted behind the guns. *DKB*

Ref.: Jane's (1995), pp. 108–171, 231–254; Macfadzean (1992), p. 95.

Figure R29 Flakpanzer antiaircraft tank system. The K_a-band fire control tracker shares the forward antenna with a longer range X-band tracker, whose transmitter also serves the target acquisition radar at the rear of the turret.

A **height-finding radar** is one "whose function is to measure the range and elevation angle to a target, thus permitting computation of altitude or height; such a radar usually accompanies a surveillance radar which determines other target parameters." Thus it is a 2D radar that scans in the elevation plane rather than in azimuth. The accompanying search radar cues the height finder to the target azimuth of interest. Because of their cyclical up-and-down scanning motion, height-finding radar antennas are sometimes referred to as "nodding" antennas. Height-finders are gradually becoming obsolete as more three-dimensional, stacked-beam or phased-array search radars are being deployed.

Figure R30 shows the S-band Russian height-finding radar bearing the NATO designation *Side Net.* The radar was designed for coverage to about 180 km in range and altitudes up to 32 km. This radar has been widely deployed in conjunction with early warning search radars such as *Bar Lock* and *Tall King* within the former U.S.S.R., and has been exported

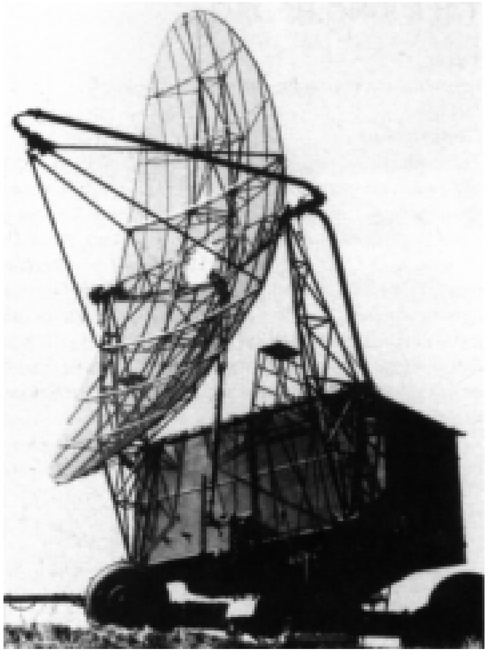


Figure R30 A Russian S-band nodding height-finder radar (from Jane's, 1995, p. 49).

to several other nations. The unit is mobile and the entire van can be rotated in azimuth. *PCH*

Ref.: IEEE (1990), p. 18; Skolnik (1990), Ch. 20; Jane's (1995), p. 49.

A **helicopter radar** is a radar installed on a helicopter. Radars used for different purposes can be installed on helicopters. earth-surface-scanning radars, passive radars for determination of direction to an object and direction-finding systems, and onboard radars for solving a broad array of problems are installed on them in particular.

One version of a helicopter passive radar for determination of direction to a "beacon" is a radar in which rotating antennas are located on the propeller blades (based on number of blades) and one stationary antenna on the airframe fuselage.

Determination of the bearing to a source of pulse radiation is accomplished by using the doppler frequency shift that occurs during reception of a signal of a turning antenna located on the helicopter rotor blade. The frequency increases when the antenna moves in the direction of the signal source and it decreases when the antenna moves away from the signal source. The doppler frequency shift passes across the zero value when the blade is oriented directly on the same azimuth as the signal source.

An omnidirectional antenna, installed on the fuselage, also receives the signal from the emitting source. This signal and a signal which has a doppler frequency shift are processed so that frequency modulation is isolated and transferred to the reference frequency. The frequency modulated reference signals are compared with the reference unmodulated signal and when the frequency difference passes across zero there is sensing of the bearing value.

Such helicopter radars are used in navigation to determine direction of one moving object with respect to another, to maintain one's position strictly in a direction to a source of emission. *AIL*

Ref.: Skolnik (1988), p. 247; Gribanov A. S., "Electronic Observation Equipment Installed on Helicopters," *Zarubezhnaya Radioelektronika*, no. 12, 1991; Jane's (1995), pp. 259, 263, 267, 273, 277, 286, 300, 323.

high-PRF radar (see **PULSE REPETITION FREQUENCY, high**).

A **high-range-resolution (HRR) radar** uses wideband waveforms to resolve individual scatterers within the target, presenting a range profile for use in target recognition or precise guidance. Pulse compression is normally required to provide adequate average power with wideband waveforms. The range resolution to qualify as an HRR is typically in the order of 1m (signal bandwidth ≈200 MHz, when weighting is used to control range sidelobes).

In one experimental application, HRR was combined with monopulse difference channel data to yield two-dimensional information on the shape of aircraft targets (Fig. R31). *DKB*

Ref.: Skolnik (1990), pp. 18.30–18.32; Wehner (1994).

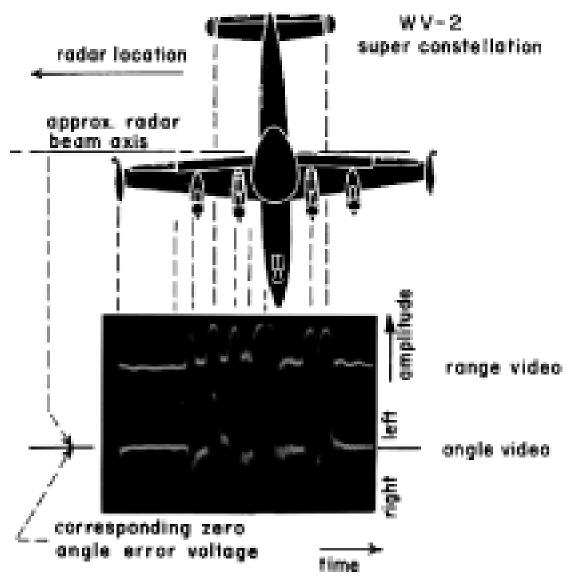


Figure R31 High-range-resolution monopulse range and angle video from a Super-Constellation aircraft in flight (from Skolnik, 1990, Fig. 18.28, p. 18.32, reprinted by permission of McGraw-Hill).

Holographic radar is an imaging radar based on application of the principles of holography to radar signals, and designed to obtain a three-dimensional radar presentation of objects. The following is the operating principle of holographic radar: an echo signal reflected from an object is analyzed to define wavefront amplitude and phase at points on a surface of observation. These data are recorded in computer memory, forming a radio hologram. Then, using a computer and display system, the wavefront is restored and an output image of the object is shaped. Radio-optical arrays and optical information processing methods may be used in holographic radars. Holographic radars must have high range and angular resolution. They are used to identify objects at great ranges. *AIL*

Ref.: Popov (1980), p. 95; Skolnik (1980), p. 523; Safronov (1973), p. 166.

A **home-on-jamming (HOJ) radar** is a homing seeker mode for guided missiles in which jamming emissions provide the source of angle data, range data generally being denied by the jamming. All radar homing seekers are designed to switch to the HOJ mode when jamming is strong enough to obscure the semiactive or active target signal. Absence of range (or time-to-intercept) data prevents the guidance loop from optimizing its gain, but the strong, point-source signal ensures accurate homing even with nonoptimum gain settings. *DKB*

Ref.: Skolnik (1990), pp. 19.18–19.20.

Hybrid radar is a term applied to a bistatic radar designed to operate directly with a cooperative monostatic radar. Such a system can sometimes improve on the performance available from two netted monostatic radars, primarily when operating with retrodirective jamming. *DKB*

Ref.: Willis (1991), pp. 224–234.

An **imaging radar** is a high-resolution radar whose output is a representation of the intensity (scatter coefficient) of the radar signals from the object or scene resolved in two or three spatial dimensions. The radar may be real or synthetic aperture and may be monostatic or bistatic. However, the impulse response must be smaller than the object or scene. (See **synthetic-aperture radar**; **TARGET RECOGNITION AND IDENTIFICATION.**) *DKB*

Ref.: IEEE (1990), p. 18.

An **impulse radar** develops its transmitted signal by shock exciting a low-Q RF element, often the antenna itself, with a short video pulse (Fig. R32). Shock excitation of the transmitting antenna allows one to generate a very short ultrawideband pulse, the spectrum of which can extend from tens to thousands of megahertz.

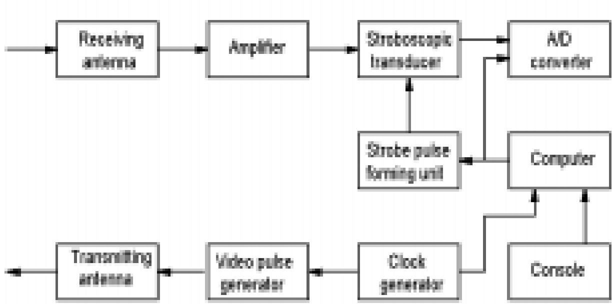


Figure R32 Impulse radar with ultrawideband signal.

An impulse radar that uses video pulses differs from ordinary radars in several ways. The pulse generator sends a short video pulse for shock excitation of the wideband transmitting antenna (typically a dipole). By changing the dimensions of the antenna one can switch to a different center frequency.

Pulses, with duration from tenths to units of nanoseconds, pulse repetition frequency up to several hundred hertz, and amplitudes of 1,000V, are emitted by the transmitting antenna in the direction of the target. The signal reflected by the target is received by the receiving antenna, amplified by the wideband amplifier, and quantized for time in the gating convertor (i.e., it is converted to a sequence of sampled values, which are constant for the time of the analog-digital conversion (ADC)). Triggering of the gating converter and ADC is accomplished by the gate-pulse shaping circuit. Digital codes from the ADC output go to the computer for digital processing.

These radars provide excellent range resolution and, are mainly used in subsurface radar. They are sometimes called *short-pulse radars*, and have a number of nonmilitary applications. (See also **ultrawideband (UWB) radar**.) *AIL*

Ref.: Bennett, C. L., and Ross, G. F., "Time Domain Electromagnetics and Its Applications," *Proc. IEEE* ***66***, no. 3, Mar. 1978, pp. 229–318; Astanin (1989); Ross, G. F., in Barton (1991), pp. 7.48–7.54; Noel (1991), pp. 125–140, 491–518.

An **instrumentation radar** is a tracking radar used at a test range to gather trajectory and other data on targets under test. For instrumentation of guided missile test ranges, high-precision monopulse trackers have been developed, starting with the AN/FPS-16 in the 1954 to 1956 period (Fig. R33).

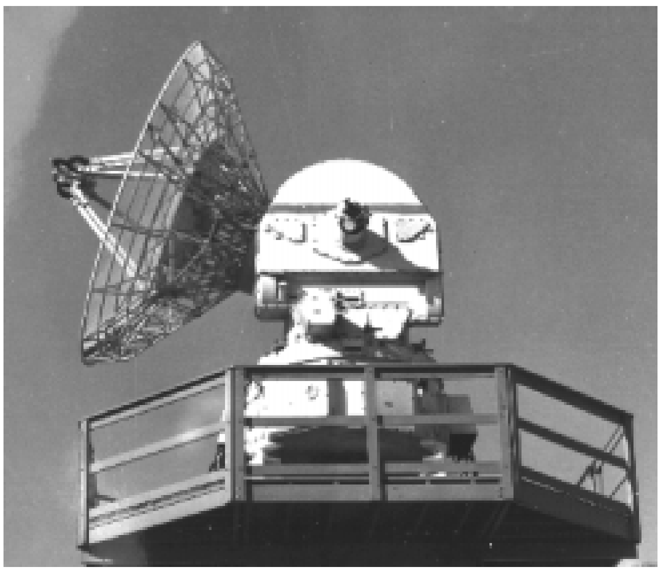


Figure R33 AN/FPS-16 instrumentation radar (RCA photo).

The new class of monopulse instrumentation radars replaced the modified SCR-584 fire control radars that had been adapted for instrumentation use following World War II. Accuracies approaching 0.1 mrad in angle and 2m in range were achieved, leading to the adoption of radar as the primary source of metric data on most test ranges, in preference to optical systems and various types of interferometers and baseline electronic instruments.

During the 1960s, larger radars with more power and somewhat greater accuracy were developed for both land-based and shipboard use (Figs. R34, R35).

Figure R34 AN/FPQ-6 instrumentation radar (RCA photo).

Figure R35 Range instrumentation ship (RCA photo).

By 1990, a multiple-target phased-array instrumentation radar had been developed to provide accurate trajectory data on several targets simultaneously.

Special instrumentation radars for observation of reentry objects were developed, starting with the Tradex (Fig. R36) in 1962. That radar was an adaptation of the AN/FPS-49 search-track radar of the BMEWS system. Subsequently, a series of specialized reentry measurement radars was designed and installed at such points as the Kwajalein atoll, as well as on ships that could be moved to positions near the impact points of missile sunder test. Both the United States and Russia used

Figure R36 Tradex reentry instrumentation radar.

such radars to gather data on the other's missile programs. The largest such radar is the Cobra Dane (Fig. R37), a phased-array system installed on the Aleutian island of Shemya. *DKB*

Ref.: Barton (1964), pp. 342–347, 505–517; Skolnik (1988), pp. 458–468.

Figure R37 Cobra Dane phased array reentry instrumentation radar (Raytheon Co. photo).

intercept radar (see **airborne intercept radar**).

A **laser radar** is "a radar whose carrier frequency is usually in the infrared or visible light region and whose transmitter is a laser." Laser radars operate in the wave band $\lambda = 10^{-4}$ to 10^{-7} m. A laser radar consists of the following main parts (Fig. R38a): optical transmitter, transmitting and receiving optical antennas, optical receiver, guidance system, and data processing and target coordination system. A laser radar transmitter can operate in pulse or continuous mode. As a transmitting optical antenna a lens telescoping antenna is used in a laser radar (Fig. R38b).

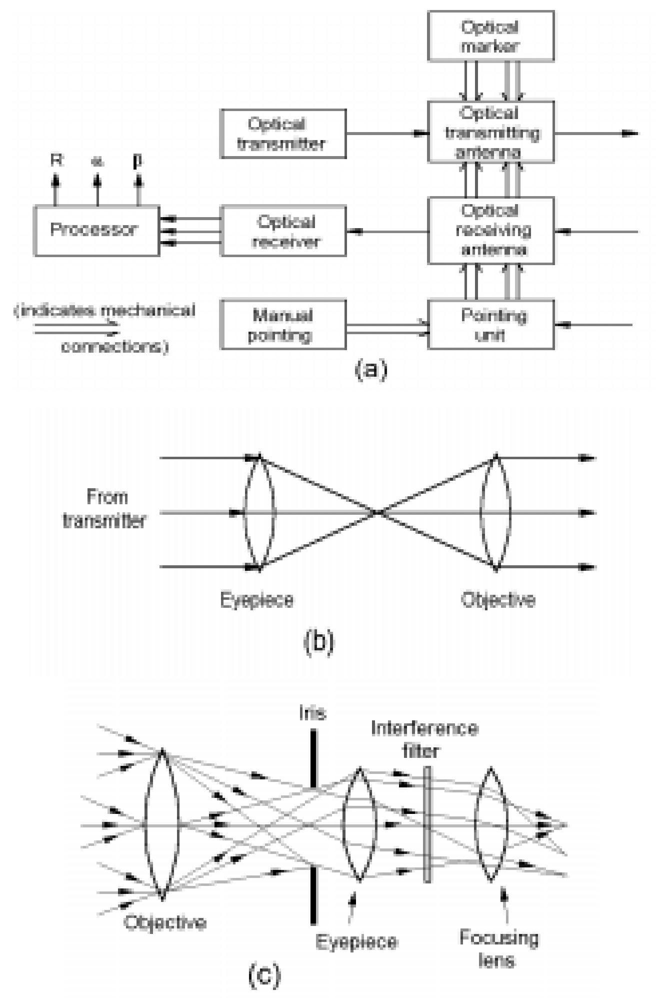


Figure R38 Optical radar (after Dulevich, 1978, Fig. 181, p. 498).

As receiving antennas one uses lens, telescope and mirror antennas. In the receiving antenna detection of optical signals by frequency is accomplished by using narrow band interference filters (Fig. R38c). Spatial detection of external background radiation is accomplished by selecting the field of the receiving antenna, which is characterized by the angle of the field of view, which is determined by the size of the diaphragm and the focal length of the objective lens.

The receiver in the general case can be multichannel and contain range and angular coordinate channels. As a rule one uses optical direct gain receivers. They consist of a photo detector and resolving unit. At the output of the photo detector a video pulse is generated that is sent for further processing to the processing and coordinate determination system. Ranges in optical radars are measured by all known methods. (See **MEASUREMENT, range**.) The pulse method of range measurement is most widely used.

For initial pointing of the optical radar beam, one usually uses target indication from infrared for direction-finding equipment. Advantages of optical radars are excellent accuracy in measuring range and angular coordinates and excellent resolution for these coordinates, possibility of measuring very low speeds; good accuracy for determination of velocity and excellent velocity resolving capability; and good noise immunity. Disadvantages of optical radars include strong attenuation of laser emissions during rain, snow, fog, that sharply reduces the effective range of the optical radar under ground conditions; complexity of guiding a narrow beam of the optical radar and long scanning time; and need for target designation. Optical radars are used in military equipment, measurement equipment, space technology, meteorology and geodesy.

One type of laser radar is the floodlight laser radar, a bistatic laser system in which the transmitting beam is considerably larger than the target area. *AIL*

Ref.: Dulevich (1978), pp. 497–502; Bachman (1979); Skolnik (1980), pp. 564–566; Hovanessian (1988), Ch. 6; Jelalian (1992).

A **lobe-on-receive-only radar** uses a fixed transmitter beam while switching the position of receiving lobes to sense target angle. It is the discrete lobe-switched version of the conical-scan-on-receive-only tracking radar. *DKB*

Ref.: Lothes (1990), p. 86; Chrzanowski (1990), p. 141.

Low-probability-of-intercept (LPI) radar refers to one using some combination of low peak power, wide transmitted spectrum, low antenna sidelobes, and control of emission times to achieve significant reduction in the probability of signal intercept by an ESM receiver or ARM seeker. The objective is to permit the radar to detect a target at ranges greater than that of signal intercept by the enemy receiver. Success can be achieved through exploitation of low transmit sidelobes if the intercept receiver is well separated in angle from intended radar targets, if the intercept receiver fails to invoke signal processing measures appropriate to the LPI waveform, or if the emission period can be held short enough to deny recognition and reaction by the user of the intercept receiver.

An important property of LPI waveforms is their bandwidth-time product $B\tau$. This can be seen to equal the pulse compression ratio, or matched-filter processing gain, where B is the instantaneous waveform bandwidth, and it can exceed that ratio when frequency agility is used to occupy, sequentially, a broader band than is subject to matched filtering. Since the emitted waveform can use pseudorandom coding or frequency stepping, it is often assumed that the intercept receiver must use the wide bandwidth B, without processing gain on the signal, thereby providing the radar with a relative gain approaching $B\tau$. However, as pointed out by Wiley and others, intercept receivers can be designed to achieve processing gain approaching $\sqrt{B\tau}$, leaving the LPI radar with a net advantage equal to $\sqrt{B\tau}$.

A major potential advantage of LPI waveforms using extended pulse widths, in a complex military environment, is that the intercept receiver may be unable to resolve multiple, overlapping emissions and thereby obtain angle measurements on individual emitters. *DKB*

Ref.: Schleher (1986), p. 291; Wehner (1987), pp. 424–438; Chrzanowski (1990), p. 47; Wiley (1985), pp. 4, 15–38.

low-PRF radar (see **PULSE REPETITION FREQUENCY, low**).

mapping radar (see **ground-mapping radar**).

marine radar (see **shipborne radar**; **navigation radar**).

Meteorological radar is used to gather data for weather forecasting, storm and flood warning, and atmospheric research. The target objects are usually precipitation particles (rain, snow, and hail) which give significant radar cross sections without excessive attenuation in S- through X-bands. Cloud and fog particles are more readily detectable in the millimeter-wave bands, while clear-air turbulence and wind shear may be detectable only through scattering by embedded particles of dust and debris.

The conventional meteorological radar uses a pencil beam to scan an azimuth sector (up to the full 360°), an elevation sector, or to perform a spiral scan in azimuth and elevation for three-dimensional data. Slowly changing weather patterns permit use of frame times measured in minutes. (See also **doppler weather radar**, **airborne weather avoidance radar**.) *DKB*

Ref.: Battan, (1959); Atlas, in Landsberg, (1964); Stepanenko, (1966), 2nd ed. (1973); Korol', (1971); Battan, (1973); Belov, (1976); Finkel'shteyn, (1977); Doviak and Zrnic (1984); Brookner, (1988); Bogush, (1989); Skolnik, (1990); Kachurin, (1992); Sauvageot, (1992).

Microwave radar uses frequency bands from 2 GHz to 30 GHz (wavelengths 15 cm to 1 cm), generally in the allocated portions of S-band (2.7 to 3.5 GHz), C-band (5.25 to 5.925 GHz), X-band (8.5 to 10.68 GHz), and K_u-band (13,4 to 14 and 15.7 to 17.7 GHz). Radars at L-band and below are not generally considered as microwave radars. Above 30 GHz the term millimeter-wave radar is applied, although K_a-band, 33.4 to 36 GHz, is sometimes considered a microwave band, lying at the transition between microwaves and millimeter-waves. *DKB*

Ref.: Skolnik (1990), p. 1.14.

Millimeter-wave radar operates at frequencies from 30 GHz up to about 300 Ghz, the term indicating the use of electromagnetic waves measuring between 10 and 1 mm in length. Figure R39 shows common terminology relating to millimeter and submillimeter waves and their relationship to other regions of the spectrum.

Millimeter waves are severely attenuated by absorption in the earth's atmosphere, and the frequencies near the attenuation minima, or RF "windows" at 35, 94, 140, and 200 GHz are where most practical radar applications can be found. Rain, heavy fog, and dense, extended cloud formations further attenuate millimeter-wave signals to such an extent that weather considerations are the primary factor in radar system design at these frequencies. Practical millimeter-wave radar developments are, therefore, relatively short-range systems that are built to take advantage of the high angular resolution properties inherent at millimeter-wave frequencies. Modern applications include terminal, active-radar, homing seekers; dual-band (usually with X-band), low-altitude, tracking radars; and air-to-ground terrain avoidance and terrain mapping radars. *PCH*

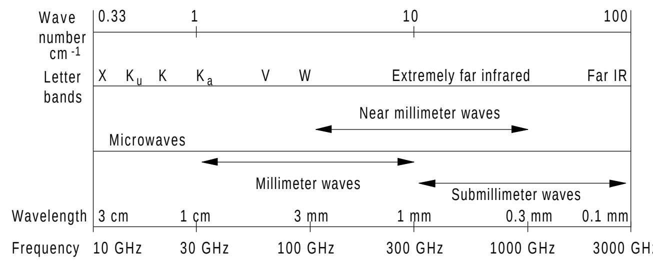


Figure R39 Millimeter waves and their relationship to other frequencies (from Currie, 1987, p. 8).

Ref.: Johnston (1980); Currie (1987).

A **missile guidance radar** is a tracking radar used to provide target or interceptor missile coordinate data, or both, to support guidance of antiair or surface attack missiles. In radar-guided surface-to-air missile (SAM) and air-to-air missile (AAM) systems, the target is acquired and tracked by a radar before missile launch, to determine when it will become engageable. After launch, midcourse guidance is provided, generally based on data from the same target-tracking radar (TTR). Midcourse command guidance uses the TTR data to formulate commands to the missile, in conjunction with data from a separate missile-tracking radar (MTR as in the Nike system), a mode of the TTR that gives missile coordinates (as in the SA-2 and Patriot systems), or with data from inertial instrumentation on the missile (as in the SA-12 system).

If semiactive homing is used, the TTR provides target illumination directly with its tracking waveform (as in Hawk), by injecting a separate waveform into the tracking antenna, or by slaving a separate target illuminator (as in Aegis). Terminal command guidance uses direct comparison of TTR and MTR data (Nike), or differential track of both target and missile within the same TTR beam (SA-2, SA-8) or with two beams from the same array antenna (SA-10, SA-15). The data requirements for midcourse command guidance are sometimes met by sequential tracking of two or more targets with a mechanically steered TTR (e.g., Phoenix), but generally require a dedicated TTR or a tracking channel of a multifunction tracking radar (Patriot, SA-10, SA-12).

Monopulse tracking radar was applied to the earliest generations of intermediate-range ballistics missiles (IRBMs), generally in combination with inertial instrument packages. Later generations of IRBMs and ICBMs used pure inertial guidance.

Air-to-surface missiles (ASMs) can also be radar guided. A typical ASM system uses a multimode airborne radar to develop a high-resolution spotlight map of the target area, within which the target object is recognized and located in a local coordinate system. The ASM is then guided to those coordinates by inertial means. (See also **fire control radar**.) *DKB*

Ref.: Macfadzean (1992).

Monopulse radar is "a radar technique in which information concerning the angular location of a source or target is obtained by comparison of signals received in two or more simultaneous antenna beams, as distinguished from techniques such as lobe switching or conical scanning in which the beams are generated sequentially. The simultaneity of the beams makes it possible to obtain a two-dimensional angle estimate from a single pulse (hence the name monopulse), although multiple pulses are usually employed to improve the accuracy of the estimate or to provide doppler resolution. The monopulse principle can be used with continuous wave as well as pulsed radar."

Monopulse radar, due to its high accuracy in measuring angular coordinates, is widely used in modern radars for different purposes. (See **MONOPULSE**.) *AIL*

Ref.: IEEE (1993), p. 821; Skolnik (1962), Ch. 5; Barton (1964), Ch. 9; Leonov (1984), pp. 5–11.

Monostatic radar is "A radar system that transmits and receives through either a common antenna or through adjacent antennas." A majority of modern radars are built using the principles of monostatic radar. *AIL*

Ref.: Dulevich (1978), p. 10; Skolnik (1980), p. 553.

moving-target-indication radar (see **MOVING-TARGET INDICATION**).

A **multiband radar** has transmitter, receiver, antenna, and microwave components capable of supporting operation in two or more radar bands. For tactical use, a typical dual-band tracking radar (Flycatcher, Flakpanzer) uses X-band for initial target acquisition and tracking, making the transition to K_a-band for control of gunfire during the actual engagement. The X-band radar channel may continue to provide range data, while the angle data, especially in elevation, from the K_a-band channel is needed for fire control accuracy. In this case, a common antenna is used with a dual-band feed system and separate transmitter and receiver channels.

For instrumentation purposes, radars operating simultaneously in three or more bands have been developed. The U.S. Naval Research Laboratory developed a tracking radar to measure target characteristics simultaneously at L-, S-, and X-bands. Data from this radar were used to compare RCS statistics on aircraft targets (Fig. R40), showing that typical aircraft gave similar values of mean RCS and fluctuation statistics (Swerling case 1) for the three bands as a function of azimuth.

Clutter was measured in several bands by airborne radar during the Overland Radar Test (ORT) program, which preceded AWACS development, and by ground-based instrumentation radars in the Lincoln Laboratory Air Vehicle Survivability program. *DKB*

Ref.: Skolnik (1990), p. 18.32; Galati (1993), pp. 550–558; Freeman (1995), pp. 176–178.

A **multibeam radar** is one in which the antenna generates several independent beams that perform simultaneous scanning of space and have independent signal processing channels. Multiple-beam radars are subdivided into those with

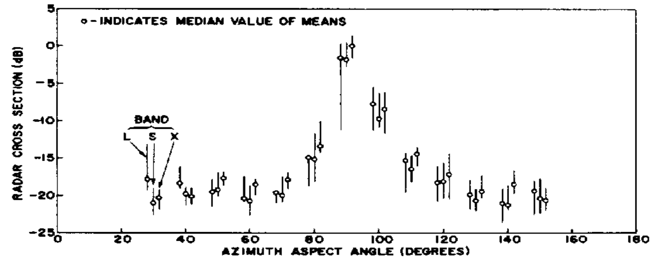


Figure R40 Target RCS results from NRL multiband radar. The data were averaged over 10° azimuth sectors, and the circles represent averages of several runs; the bars indicate the standard deviations of the data. Note the minor variation in RCS over the bands from L-band to X-band (from Barton, 1988, Fig. 3.3.2, pp. 112–113).

scanning and without scanning of the space. An example of multiple beam radars are three-dimensional radars in which the antenna pattern in the vertical plane is generated from several narrower patterns spaced to provide some overlapping. Another example is monopulse radar. Multiple-beam radars also include some space target detection radars. *AIL*

Ref.: Leonov (1988), pp. 94–100; Barton (1988), pp. 196–197, 348–358; Skolnik (1990), pp. 20.7–20.14.

A **multichannel radar** is one in which signal processing is carried out simultaneously in several channels, each of which is used to process signals coming from a target located in a single cell which resolves in range, velocity, and angular coordinates. Information for each target in these coordinates is isolated as a result of simultaneous processing of signals received in all channels. Among multichannel radars one can distinguish multichannels for range, velocity, space-multichannel, and frequency-multichannel. In multichannel radars for range, signals come from different range cells consecutively in time, and each range resolution cell has its own channel. In velocity multichannel radars, a set of filters is installed which completely cover the possible range of variations of the doppler frequencies. In frequency-multichannel radars, there are several channels for radiation and reception, which are tuned to different frequencies and which use identical or different modulation patterns. Such radars also include radars in which several beams are generated, within the limits of which oscillations of only the carrier frequency are emitted. Spatial multichannel radars have several points of reception and radiation of signals, which are separated in a certain manner in space. As a rule, multichannel radars use phased antenna arrays that carry out practically instantaneous scanning of space and generate multichannel signal processing algorithms. In addition to the listed types of multichannel radars, improved versions have been built which represent their combinations. Multichannel radars also include radar sets in which signal processing is performed simultaneously in several channels, each of which is intended for processing signals from a single target. Multichannel radar is widely used in modern radar for different purposes. Examples are monopulse radars, radars for detecting space targets, and multistatic radars. *AIL*

Ref.: Lukoshkin (1983).

multifrequency radar (see **frequency diversity radar**).

A **multifunction radar (MFR)** is one designed to perform both search and tracking functions with the same antenna in a single frequency band. As applied to surface-to-air missile (SAM) systems, such radars are implemented with phased array antennas (see **multifunction array radar**), which provide not only search and track of targets but also missile-guidance functions. In airborne applications, mechanically steered antennas are commonly used, and the functions typically include search and tracking of aerial targets, surface target search using both real-beam and synthetic-aperture mapping, terrain and weather avoidance, and weapon control.

Advantages of MFR include minimizing space and weight (essential in the airborne applications), adapting the operating modes in real time to changing requirements, and eliminating errors in transferring from detection to tracking modes. Disadvantages include the need to choose a single, compromise frequency, scheduling the multiple modes and target updates within a limited time period, and complexity in generation, transmission, and signal processing of the different waveforms required. *DKB*

Ref.: Skolnik (1988), pp. 284–294, 295–302.

A **multifunction array radar (MFAR)** is an electronically scanned radar designed to perform search, tracking, and (usually) missile guidance with the same array in a single frequency band. The primary application for such radars has been in Western SAM systems such as Patriot and Aegis, where the complexity and high initial cost have been justified by the critical requirements of firepower and life-cycle cost.

In the Patriot system, a single array is carried on the radar vehicle (Fig. R41), and this array is mechanically trainable to cover any 90° azimuth sector, over elevations from the horizon to zenith. The radar operates in C-band, that having been selected as the best compromise between search, tracking,

Figure R41 AN/MPQ-53 Patriot multifunction array radar (Raytheon photo).

and guidance functions within the aperture size restrictions of the vehicle. All radar functions of the SAM system are performed, in automatically scheduled sequences, by the MFAR: volume search, horizon search, multiple-target tracking, missile tracking and midcourse command guidance, target illumination for semiactive terminal homing, and kill assessment. A separately steered array is provided to receive missile downlink signals in support of the target-via-missile (TVM) type of semiactive guidance.

The Aegis MFAR (Fig. R42) performs volume search, horizon search, multiple-target tracking, and midcourse command guidance, all at S-band. Four arrays on the ship provide 360° azimuth coverage. Target illumination for terminal guidance is provided by separate X-band illuminators, slaved mechanically to MFAR target data. On some ships, a separate volume search radar supplements the MFAR search function. The two arrays on the forward superstructure share transmitters and receivers, as do the two on the aft superstructure.

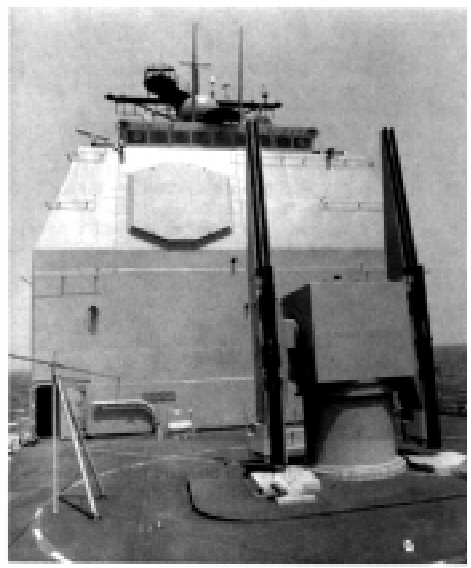


Figure R42 AN/SPY-1 Aegis multifunction array radar (from Brookner, 1988, Fig. 3.A.1-3, p. 257).

In both Patriot and Aegis, as in other MFARs, the many tasks to be scheduled dictate heavy reliance on single-pulse dwells, which lack the ability to reject clutter. Multiple-pulse dwells provide MTI capability over limited portions of the search volume and for high-priority targets in clutter regions. High-energy burnthrough dwells may also be scheduled in regions subject to strong jamming. The adaptive scheduling required to balance the MFAR resources in time and power with changing requirements and environments constitute a major software problem. Provision of alternative modes to defend against tactical ballistic missiles has been added to both systems, supplementing the air defense modes originally implemented.

An example of a very large MFAR is the *Don* radar (Fig. R43), developed during the period 1973–1989 and installed north of Moscow near the city of Pushkino as part of the Moscow area ABM system. The main functions of this radar are to detect, track, and discriminate (classify) ballistic targets, and to guide the interceptor missiles. On each face of the pyramidal building are space-fed transmitting and receiving phased arrays, the circular receiving arrays having a diameter of 16m and the rectangular transmitting arrays dimensions 7 × 8m. Each receiving antenna can form several independent groups of monopulse beams. The radar operates in the centimeter-wave band with ranges up to several thousand kilometers, and has high capabilities in target resolution, accuracy, jamming immunity, and target capacity.

Figure R43 The *Don* radar, designed to support the Moscow ABM system and installed at Pushkino, northeast of the city.

Airborne MFAR applications have been limited to the B-1B bomber and the MiG-31 fighter (Figs. R44, R45). In both cases, the nose radome position has been allocated to a mechanically fixed, electronically scanned array, which performs the several functions more commonly served by gimballed flat-plate array or reflector antennas. The use of electronic scan avoids the time wasted in reversing scan direction in mechanically scanned systems, and permits multiple targets to be tracked without lost time in transition from one to another. *DKB, AIL*

Ref.: Kahrilas (1976); Billetter, D. R., (1989); Sabatini (1994).

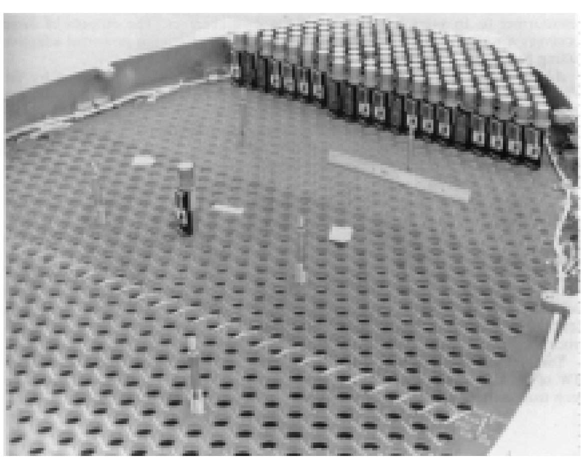


Figure R44 Multifunction AN/APQ-164 phased array for B-1B radar, during assembly (from Skolnik, 1990, Fig. 7.48, p. 7.71, reprinted by permission of McGraw-Hill).

Figure R45 Multifunction phased array for MiG-31 Flash Dance airborne radar.

A **multilateration radar** system uses range, range-sum, or range-difference measurements from two or more sites to establish target coordinates. In the range multilateration system, radars at two sites are capable of providing two-coordinate target location, representing approximate x and y coordinates in the ground plane for targets at low or medium altitude. Both stations must have line of sight to the target. An example of this approach was the *Oboe* system used for blind bombing during World War II. The aircraft flies at a constant range from the first station until it reaches a predetermined range from the second, at which point the bombs are released. The accuracy of the release point was determined by the standard deviation σ_r of the range measurements, degraded by a factor k_r, the geometrical dilution of precision (GDOP):

$$\sigma_{x,y} = k_r\sigma_r \approx \frac{\sigma_r}{\sin\alpha}$$

where α is the intersection angle of the two ranging paths. For an aircraft flying at known altitude, the effect of that altitude could be precomputed to yield accurate x, y data.

When three monostatic radars are used, all three target coordinates can be measured, although the accuracy of altitude data is likely to be poor because the paths will intersect at shallow angles in the vertical plane.

Multilateration systems using range-sum or range-difference require an additional site: three for two-coordinate determination and four for three-coordinate. An example of such systems is Loran, in which pairs of transmitter stations are used to establish lines of position for a platform equipped with a receiver measuring range-difference. Two pairs of transmitting stations (which may include a common station) establish two intersecting lines of position on the earth's surface. Although there will, in principle, be two points at which these intersect, one is so distant from the known receiver position that it can be rejected.

A modern range multilateration system is the Global Positioning System (GPS), which uses four transmitters to solve for three position coordinates and a time reference at the receiver. A constellation of 24 satellites is operational, ensuring path intersection angles such that the GDOP is minimized. Position data to 100m are available at most points on the earth, and simultaneous observation of five or more satellites can improve this accuracy. The potential accuracy is 10m or better when the system operates in a mode permitting receivers to fully exploit the ranging modulations. *DKB*

Roberts (1947), pp. 14–17; Barton (1964), pp. 507–512; Kayton (1990).

Multistatic radar is one in which information on a target is obtained by means of simultaneous processing of signals of several spatially separated transmitting, receiving or transceiving positions. When using a multistatic radar one can achieve energy gain and, consequently, increase of effective range, reduction of angular coordinate measurement errors, possibility of measuring three coordinates and velocity of emitting sources (including noise), improvement of noise immunity, increase of transmitting capability, and improvement of survivability. But all these advantages are achieved at the cost of substantial increase of complexity and cost of the radar multistatic system and expenditures for its operation. The simplest type of multistatic radar is the bistatic radar. *AIL*

Ref.: Skolnik (1970), Ch. 36; Chernyak (1992), pp. 7–32.

A **navigation radar** is one used to provide position data on a mobile platform by display of fixed surface features in map form. Shipborne and airborne systems are in use, with the maritime application most common. During World War II the airborne “bomb-nav” radar was used extensively, and modern, multimode fighter-bomber radars have ground-attack modes based on mapping and detection of fixed and moving targets. Important properties of a navigation radar are high resolution in both azimuth and range and the ability to form images of fixed surface features in the presence of sea and precipitation clutter.

The typical marine navigation radar operates at X-band with a beamwidth of 0.5° to 1.5°, rotation rate of 30 to 60 rpm, and display ranges from 0.25 to 32 nmi (metric units have not yet received wide acceptance in navigation applications). For use on ships, radars such as the Raytheon Model 3900 (Fig. R46) are widely used. The parameters of this radar are given in Table R5.

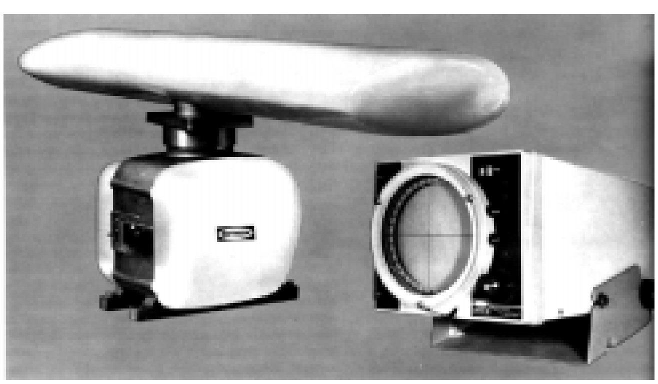


Figure R46 Raytheon navigation radar, Model 3900, designed for small commercial ships.

Table R5
Raytheon Model 3900 Radar

Parameter	Units	Value
Frequency	MHz	9,375±30
Peak power	kW	7
Pulse widths	μs	0.1, 0.67
PRF	Hz	3,000 or 1,500
Beamwidths	deg	$2° \times 22°$
Scan rate	rpm	30
Antenna width	ft	4
Receiver bandwidth	MHz	12, 4
System noise factor	dB	11
Range scales	nmi	0.5 to 32

The choice of operating band for shipboard navigation radar depends on the requirements for range of operation and weather conditions. For short-range use, X-band or K_u-band are suitable, the latter in less than heavy rain. For long-range use under difficult weather conditions, S-band is preferred. The current regulations of the International Standards Organization (ISO) require large ocean-going ships to have two radars, one at X-band and the other at S-band. *DKB*

Ref.: Skolnik (1970), Ch. 31; Barton (1988), pp. 364–366.

Noiselike-waveform radar transmits noiselike waveforms, rather than pulses or regular modulated CW signals. To detect signals reflected from the target on a background of radio interference and determine its distance and speed, correlation signal-processing is used.

In such radars, the wideband, CW operation and the random nature of noise signals ensures excellent resolution for simultaneous determination of range and velocity, secrecy of operation, and excellent noise immunity. Practical implementation is in the form of pseudonoise signal radar, a CW radar using a pseudonoise signal. In Fig. R47 we show a simplified block diagram of such a radar. The pseudorandom video signal S_1 (an M-sequence) from the pseudorandom waveform generator goes to the microwave modulator and to a mixer.

A signal from the sine-wave generator is also supplied to the modulator. At the output of the modulator, a phase-coded signal is generated. The signal phase changes abruptly by π, based on the pattern of the pseudorandom sequence S_1. The generated signal is applied through a circulator to the antenna and emitted.

The received signal passes through the circulator to the balanced mixer, to the second input of which part of the emitted signal is fed. The received signal is modulated by the pseudorandom sequence S_τ, which differs from the modulating sequence S_1 by the shift for the propagation time of radio

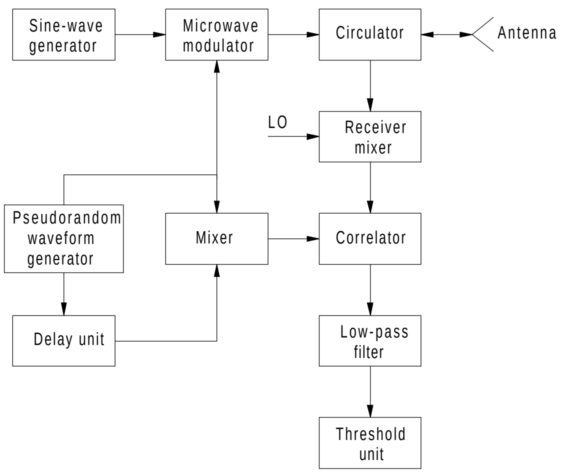


Figure R47 Radar with pseudonoise waveform.

waves to the target and back. At the output of the balanced mixer, a pseudorandom signal is obtained, which is the result of beating of the emitted and received random sequences. The resulting beats are compared with the "local beats" that are formed by the addition of the sequence S_1, with the sequence S_0 that corresponds to the anticipated radio wave propagation time. The newly generated and local beats are supplied to a correlator. If there is equality of the anticipated and actual delays, the voltage at the correlator output is maximum in value and changes in time sinusoidally with the frequency f_d, equal to the doppler change of the emitted signal frequency.

Radars with pseudonoise signals are usually used in close-in radar to improve accuracy and range resolution. *AIL*

Ref.: Popov (1980), p. 481; Skolnik (1990), pp. 14.28–14.30.

A **noncoherent radar** is one in which the phase relationships between transmitted and received signals are not maintained. In a noncoherent radar, the signal returns from a target must be detected separately and added together after detection and before the application of thresholding. This process is referred to as *postdetection integration*. Unlike a coherent radar, a noncoherent radar cannot make optimal use of system information redundancy. The most widely used is the noncoherent pulse radar that transmits a signal and receives a reflected echo in the form or a train of RF pulses, and in which there is no coherent oscillation for synchronization of the emitted and received signals. Frequently in the name of this radar the word "noncoherent" is dropped and the radar is simply called a pulse radar, as opposed to a coherent-pulsed radar. The block diagram of a pulse radar is given in Fig. R48.

The radar transmitter consists of an RF generator and pulse modulator, which determine the pulse duration and repetition frequency of the transmitted pulses. The antenna switch ensures operation of the receiver and transmitter with one antenna. Upon reflection the reflected signal goes to the low-noise RF amplifier. Next it is mixed with the local heterodyne signal. After the intermediate frequency amplifier, the signal goes to the amplitude protector and then to the output unit (display or computer).

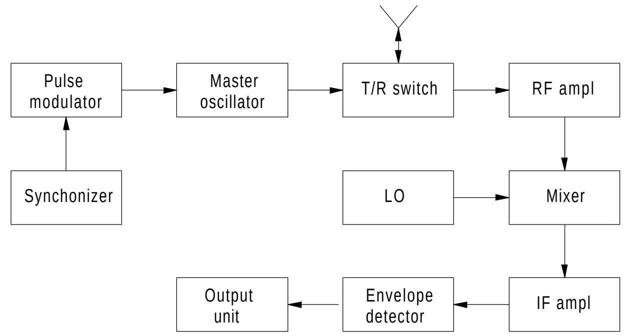


Figure R48 Pulse radar (after Leonov, 1988, Fig. 1.5, p. 12).

The advantage of pulse radars is that for comparatively simply equipment one can simultaneously measure the range of many objects. A disadvantage is limited possibilities of measuring target radio velocity. Pulse radars are used to solve a wide variety of detection and measurement problems. *AIL*

Ref.: Ridenour (1947), pp. 38–47; Leonov (1990), p. 12.

Oil slick detection by radar is possible because the clutter characteristics of the sea surface depend on the surface tension of the water, and this is modified in the presence of oil. High-resolution X-band radar can detect small quantities of oil leaking from a small boat at anchor. Airborne synthetic aperture radar can monitor oil slicks over large areas. *DKB*

Ref.: Skolnik (1980), p. 482.

Optical (band) radars (see **laser radars**).

An **over-the-horizon (OTH) radar** is "a radar using sufficiently low carrier frequencies, usually in the high-frequency (HF) band, so that ground-wave or ionospherically refracted sky-wave propagation can allow detection far beyond the ranges allowed by line-of-sight propagation." OTH radar can be carried out using single-bounce reflection or multiple reflections of waves from the ionosphere and the earth's surface. OTH radars that take advantage of sky-wave propagation, or refraction from the ionosphere, can realize detection ranges on the order of thousands of kilometers. The illustration to the left in Fig. R49 shows, for a simple "one-hop" case, that at high angles, the radar radiation escapes, causing a "skip zone" in which there is no radar coverage. The figure on

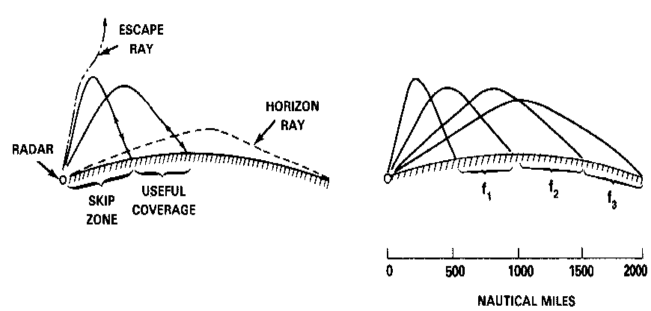


Figure R49 Over-the-horizon sky wave propagation (from Skolnik, 1990, Fig. 24.1, p. 24.2).

the right illustrates that by using several different frequencies, the OTH radar can fill in all but the nearest range coverage gaps attributable to any given frequency. In actuality, due to the various layers in the ionosphere, there will be a multitude of "hops," and the radar energy may, in fact, circle the entire earth. The performance of OTH radars is sensitive to the state of the ionospheric environment and to determine this, in real time, appropriately instrumented sounders (as well as the radar itself) are typically deployed.

A second type of OTH propagation at HF occurs due to diffraction of energy around the earth's curvature. OTH radars that use this ground-wave technique are limited to ranges on the order of a few hundred kilometers due to the propagation losses that increase exponentially with range.

OTH radar can be used to detect launches of intercontinental ballistic missiles and to detect these missiles and satellites in flight, nuclear explosions, and so forth. In addition to detecting traditional targets of interest, OTH radars can provide useful information about sea state, land features, and nonterrestrial events such as aurora and meteors. OTH radars designed to detect aircraft targets at ranges out to 4,000 km might typically employ average powers of hundreds of kilowatts. To realize useful angle resolution at HF, the azimuth dimension of the radar antenna (or antennas, if a bistatic radar design, using separate transmit and receive antennas were employed) required to achieve reasonably narrow beamwidths would be on the order of 300m or more. For OTH radars to achieve adequate target resolution, pulse-compression waveforms and long coherent processing times are generally required. *AIL*

Ref.: IEEE (1993), p. 908; Skolnik (1980), Sec.14.2; Kolosov (1987), pp. 6–10; Skolnik (1990), Ch. 24.

Passive radar is one that does not have a transmitter and that accomplishes detection and measurement of target coordinates based on their own emissions. Sources of radiation could be working radio transmitters of the targets, and the targets themselves, which have thermal or other radiation contrasting with the surrounding medium. Because of the absence of information on the radiation time, the range to the radiation source can be determined from data of no less than two receiving points.

Antennas and receivers of passive radars do not differ from antennas and receivers of active radars, which receive analog signals (pulse or continuous-wave), in the same waveband. An exception is reception of thermal signals. In reception of thermal signals, the frequency detection and temporal gating cannot be received, because the signals have a noise-like nature. Therefore, receivers for thermal signals come in the form of noise meters. Such receivers are conventionally called *radiometers*. (See **RADIOMETER**.)

At the present time single-channel and multichannel thermal-microwave imagers are being used. They cannot determine the range and velocity of a target, but they determine angular coordinates and intensity of target radiation. Thermal-microwave imagers (RTL) can be scanning or tracking. In scanning thermal-microwave imagers (Fig. R50), one fre-

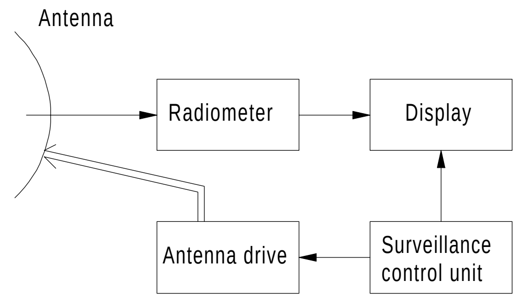


Figure R50 Scanning thermal-microwave imager (after Nikolaev, 1970, Fig. 28, p. 72).

quently uses pencil-beam antennas, which ensure a possibility of determining the two angular coordinates of the target.

In a thermal-microwave imager one can use different kinds of scanning: line, spiral, and cycloidal. Scanning thermal-microwave imagers can be ground-based, or onboard (installed on aircraft). Tracking thermal-microwave imagers are used for automatic tracking of solitary radio thermal targets based on angular coordinates. The operating principle and functional diagram of a tracking thermal-microwave imager are quite similar to the operating principle and functional diagrams of tracking radars. Advantages of passive radars consist in secrecy of their operation and, consequently, excellent noise immunity. A disadvantage is the impossibility of measuring range to target from a single point and poor resolution. Passive radars are used in navigational and measurement systems, for mapping terrain, in military complexes, for ice-cover reconnaissance, geologic and geophysics studies, and so forth. *AIL*

Ref.: Shirman (1970), p. 494; Nikolaev (1970), pp. 71–83, 94; Skolnik (1970), Ch. 39; Popov (1980), p. 278.

A **personnel-detection radar** is one used to detect humans moving on a battlefield or other area subject to hostile intruders. These radars are typically small units, carried by individual soldiers, on small vehicles, or mounted near ground level along the periphery of the protected area. Broad, fixed beams may be used for an alerting function, while narrower, scanning beams provide location data within the sector. Both pulsed and CW radars are used in personnel-detection radar, the latter often including FM to provide range data. In low-power systems, a single antenna may be used with FM waveforms, which are then economical in size, weight, cost, and complexity.

A typical battlefield personnel-detection radar is shown in Fig. R51. *DKB*

Ref.: Skolnik (1988), pp. 453–457, (1990), p. 14.39.

A **phased-array radar** uses an electronically steered (phased-array) antenna, in place of or mounted on a mechanically moving pedestal. The electronic scan permits the beam to be steered rapidly (within a millisecond and often within a few microseconds) to scan a volume or move between multiple targets in track. (See **artillery and mortar location radar**; **fire control radar**; **instrumentation radar**; **multi-**

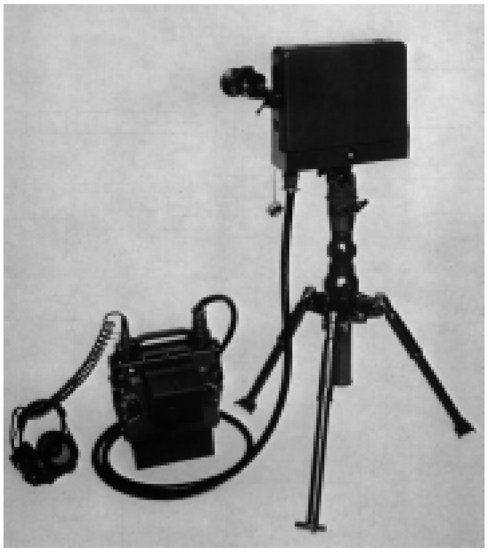


Figure R51 Typical man-portable personnel detection radar for battlefield use.

function array radar; **precision approach radar**; **space-target acquisition radar**; **three-dimensional radar.**)

Ref.: Brookner (1988).

Pincushion radar is a term sometimes used to denote a radar with a reflector antenna that generates a cluster of many beams. SAL

Ref.: Skolnik (1980), p. 316.

Polarimetric radar exploits the differences in target RCS at different polarizations to extract information on target characteristics, or to discriminate targets from background clutter. At least two receiving channels having different polarizations are required, and two transmitting channels are sometimes used as well, multiplexed in time or frequency. The ideal polarimetric radar would determine the complete polarization matrix of each potential target and use this information to select and recognize particular targets. Applications of radar polarimetry include noncooperative target recognition (NCTR), adaptive polarization for clutter rejection, and survey of surface features such as crop conditions. *DKB*

Ref.: Currie (1987), pp. 298–310.

A **polarization adaptive radar** is one in which the receiving polarization, or both transmitting and receiving polarizations, are continuously adapted to maximize the signal-to-interference ratio (*S/I*). To do this, it is necessary to develop a "polarization error" channel that measures the departure from the optimum polarization and initiates changes in the polarization of the main channel. The error can be measured by sequentially varying the relative phase and amplitude of two orthogonal polarization channels feeding the system output, and centering the interval over which they vary to maximize the output S/I. Alternatively, a channel orthogonal to that of the main channel may be formed, and signal components in this channel minimized (thereby maximizing the signal in the main channel). If transmitted polarization is varied as well as that in the receiving system, there will be four degrees of freedom, two for transmitting and two for receiving, leading to considerable complexity. For this reason, fully adaptive systems of this sort are not usually found. *DKB*

Ref.: Nathanson (1991), pp. 157–165.

A **polarization diversity radar** is one using two receiving channels to improve target detection. Most targets of interest return signals in two orthogonally polarized components. An optimum receiving system operating in a noise background would combine these with proper amplitude and phase to recover the total target RCS. In a clutter background, the optimum system would combine the received signals in such a way as to maximize the signal-to-interference (S/I) ratio. Approximations to this performance can be used to improve detection performance: selecting the channel with the larger S/I; adding the two channel outputs at video, to integrate the two components; or forming a multiplicity of output channels matched to different target polarizations, and choosing that with the greatest S/I ratio. Polarization diversity can also be implemented with two transmitting channels, multiplexed in time or frequency, but the cost of such an approach is appreciably higher than that using two receiving channels. *DKB*

A **police radar** is a small (often hand-held) radar used to measure the speed of roadway vehicles. Unmodulated CW waveforms are most commonly used, at X- or K-band. Modern units also use laser transmissions to achieve angular resolution and to reduce the effectiveness of intercept receivers used to warn drivers of the radar's presence. *DKB*

Ref.: Skolnik (1990), p. 14.21.

A **precision approach radar (PAR)** is "a radar system located on an airfield for observation of the position of an aircraft with respect to the approach path and specifically intended to provide guidance to the aircraft in the approach." It is an element of a ground-controlled approach (GCA) radar system. Ground-based PAR normally operates X-band with antenna beamwidths of 1° or less to preserve accuracy on low-elevation targets near touchdown. Three distinct design approaches have been used:

(1) Sector-scanning fan-beam PAR. In this system, two channels (beams) are used, one scanning in azimuth and the second in elevation. The beamwidth in the nonscanning coordinate is made broad enough to encompass the approach sector (Fig. R52). Scanning may be accomplished mechanically,

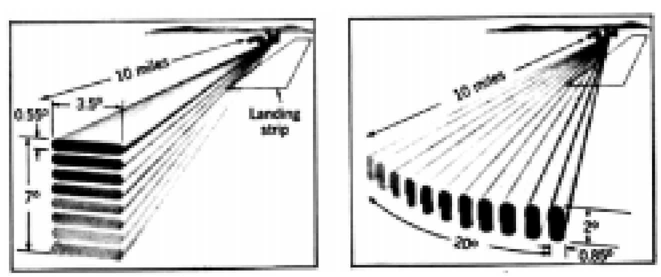


Figure R52 Sector-scanning GCA system (from Hall, 1947, Fig. 7.37, p. 242).

electromechanically, or electronically, the usual rates being two to four scans per second in each coordinate.

An early electromechanical scanning technique, still used in many systems, is the Eagle scanner (Fig. R53), in which a line of dipole radiators illuminates a parabolic cylinder reflector. The phase difference between the dipoles is varied linearly by varying the width of the waveguide through which they are fed, using a cam-controlled linkage. This scans the beam in the plane of the feed line. During the retrace time of one antenna beam, the transmitter and receiver are connected to the other antenna, providing one scan in each coordinate during the mechanical cycle.

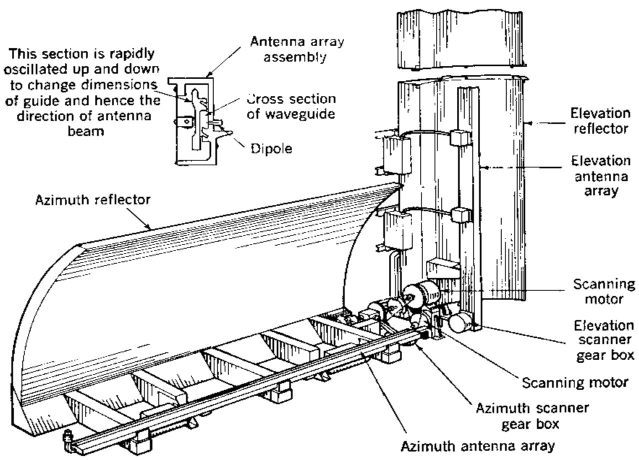


Figure R53 Eagle scanner used in sector-scanning PAR (from Hall, 1947, Fig. 7.38, p. 244)

Separate, mechanically scanning reflector antennas can also be used (Fig. R54).

Two display channels are used, often on a single cathode-ray tube, one giving the azimuth versus range and the other elevation versus range. The controller estimates visually the target range and angular departure from the approach path, which is indicated by a distinct reference trace on each display. The approach guidance loop is closed by voice advisories to the pilot (e.g., "One point five miles from touchdown, twenty feet below glide path, twenty feet right").

(2) Raster-scanning pencil beam PAR. In this system, a single beam, narrow in both coordinates, is scanned over the approach sector in a series of vertical or horizontal lines. To provide the necessary data rate, electronic scan is required. In the AN/TPN-25 (Fig. R55), phase scan in both coordinates is produced by some 800 phase shifters in a feed array that illuminates the main reflector, giving a beam 1.4° by 0.75° with a scan field 20 deg in azimuth by 15° in elevation. In the AN/TPN-22 frequency scan is used in elevation and phase scan in azimuth.

(3) Tracking PAR. In this system, a pencil beam tracker is locked onto the approaching target to provide high quality data in range, azimuth, and elevation. A mechanically scanned antenna can provide data on a single target, while multiple targets can be tracked in a time-multiplexed mode using electronic scan. The AN/SPN-42 uses a mechanically

Figure R54 AN/TPN-18 X-band PAR antenna system (ITT Gilfillan Co. photo).

pointed reflector at K_a-band for one aircraft at a time, with two complete radar systems providing for two closely spaced approaches. The AN/TPN-25 interlaces monopulse tracking beams on up to six targets within the raster-scanned approach sector, the latter scan providing redundancy and ensuring rapid operator reaction in case of tracker malfunction. The display for the AN/TPN-25 (part of the AN/TPN-19 GCA system) presents range, azimuth, and elevation data on target tracks, with track history to help the operator estimate the velocity vector of the aircraft, and with data from the raster scan overlaid for safety (Fig. R56).

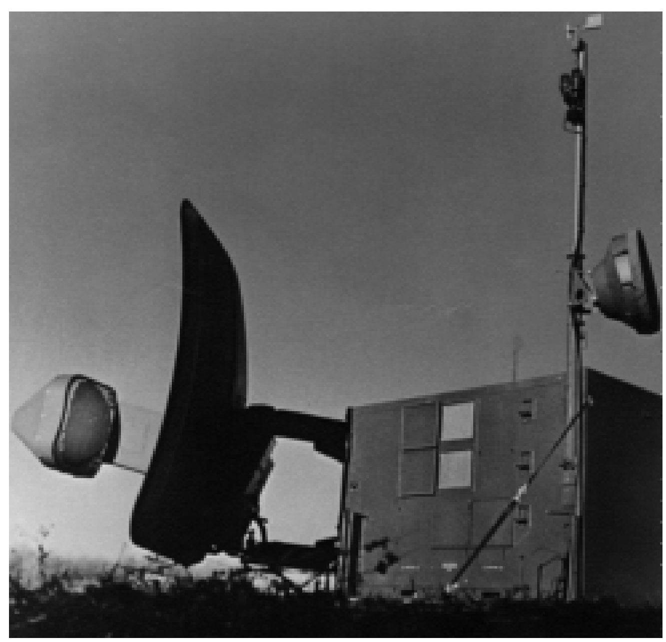


Figure R55 AN/TPN-25 reflector antenna with limited-scan phased-array feed for monopulse tracking PAR (Raytheon Co. photo).

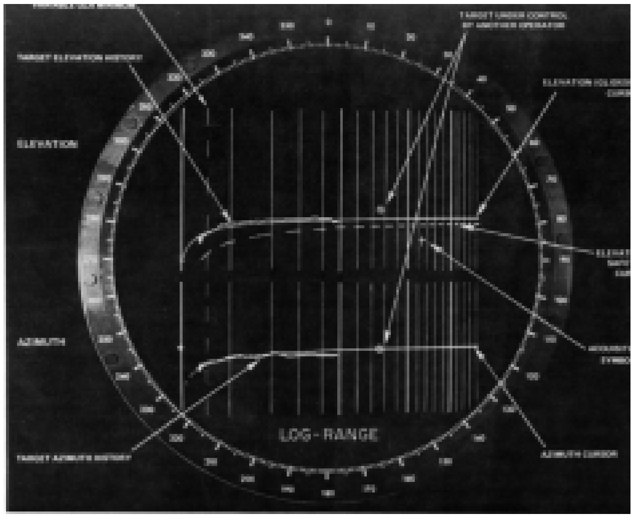


Figure R56 AN/TPN-25 PAR display (Raytheon Co. photo).

A common requirement of all types of PAR is high resolution, especially in elevation (to minimize multipath errors). A maximum beamwidth of 1° is necessary to maintain elevation angle accuracy of one mrad (corresponding to 10 ft in height at the end of a 10,000 ft runway). Most successful systems use elevation beamwidths between 0.5° and 1.0°. Attempts to apply superresolution techniques have had little success, due to the complex structure of diffuse multipath components. The other major source of error is target glint, which may exceed one mrad on large aircraft near touchdown. Unless controlled by some type of special processing, the resulting random error would exceed the 10-ft altitude requirement. Time averaging over several scans at 10 scans/sec can be used, along with frequency diversity or high range resolution to restrict the tracking point to the nose of the aircraft. In the carrier landing applications of PAR, a corner reflector is mounted on the nose gear of the aircraft to provide a strong point source at the K_a-band frequency. *DKB*

Ref.: IEEE (1993), p. 998; Barton (1988), p. 480; Ward, H. R., Fowler, C. A., and Lipson, H. I., "GCA Radars: Their History and State of Development," *Proc. IEEE*, June 1974, pp. 705–716.

A **primary radar** is "a radar system, subsystem, or mode of operation in which the return signals are the echoes obtained by reflection from the target. Since this is the normal method of radar operation, the word *primary* is omitted unless necessary to distinguish it from *secondary.*"

Ref.: IEEE (1993), p. 1005.

A **proximity fuze radar** is a radar installed in a projectile or guided missile for the purpose of determining the near-optimum time for warhead fuzing on the target. The proximity fuze used in antiaircraft artillery rounds was one of the first successful applications of radar was in World War II. Proximity fuze radar is basically a CW device that senses the change in doppler voltage when the antenna pattern intercepts the target. *PCH*

Ref.: Skolnik (1970), p.16-20.

A pulsed doppler radar is "a doppler radar that uses pulsed transmissions." It is understood that pulsed echo signals are processed to provide coherent integration and selection of targets having doppler frequencies within specific bands. In the search or target acquisition mode, such radars may use a bank of narrow doppler filters (implemented, for example, with an FFT processor) to cover a wide band of possible target dopplers. In pulsed doppler tracking radars a single filter (or a discriminator filter pair) may suffice, once target acquisition has been accomplished. The pulsed doppler radar is distinguished from an MTI radar in that the latter uses a bandstop filter to reject clutter and a bandpass filter to accept moving targets without coherent integration.

To process pulse train echoes efficiently, the pulsed doppler radar must precede the narrowband doppler filter with a range gate approximately matched to the pulse width (or to the compressed pulse width in a system using pulse compression). When the duty factor of the waveform is high (e.g., >25%), a single gate may cover the portion of the pulse repetition interval during which the transmitter is off, with acceptable mismatch loss. With such a waveform (sometimes known as interrupted CW) the sum of eclipsing and gate mismatch losses will be relatively high. As duty factor is decreased, multiple gates become necessary in search or target acquisition modes to provide efficient recovery of signal energy on targets of unknown range. Once a target has been placed in range track a single gate (or a split-gate pair) is sufficient.

Pulsed doppler radar may operate in high-, medium-, or low-PRF modes (as defined under **PULSE REPETITION FREQUENCY**). Modern airborne pulsed doppler radars operating in the high-PRF (HPRF) mode achieve clutter attenuations approaching 80 dB, as necessary to overcome land clutter entering the many range ambiguities within the main beam. This mode is preferred for approaching targets whose doppler shifts carry the signals into the clutter-free region above the mainlobe clutter and below the ambiguous response to receding clutter from the rearlobes of the antenna (Fig. R57). Detection of receding targets or those within the mainlobe clutter spectrum is not expected in the HPRF mode.

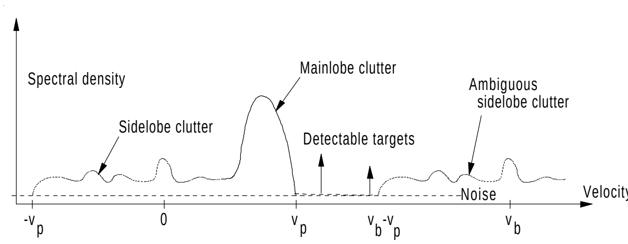


Figure R57 Spectrum of noise, clutter, and targets in HPRF airborne radar.

Against receding targets, whose doppler shifts place the signals within the sidelobe clutter region, MPRF operation is preferred because it reduces the total clutter input from range ambiguities, and hence the level of the "back porch" of the doppler spectrum (Fig. R57). The clutter attenuation available is controlled in this case by the antenna sidelobes and radome effects, and is generally less than that of the HPRF mode (e.g., 70 dB). Detection of targets in mainlobe clutter is not expected, and approaching targets are detected at shorter ranges than in the HPRF mode with the same average power.

LPRF operation is restricted generally to radars looking upward or operating from fixed or slowly moving platforms, such that the mainlobe clutter spectral spread is a small fraction of the velocity ambiguity of the waveform. As applied to this class of radar, a low-PRF pulsed doppler processor is often called a moving-target detector (MTD), a term originated by the MIT Lincoln Laboratory in the 1970s for the processor and waveform approach they used in improving performance of an airport surveillance radar. The clutter attenuation in MTD systems is typically near 60 dB, limited by oscillator and transmitter stability. The phase noise components within $f_r/2$ of the carrier, and aliased from many PRF lines at higher frequencies into that region, generally exceed the level experienced in MPRF and HPRF systems.

Russian engineers have successfully applied HPRF waveforms and processors to high-power, ground-based radars used in air defense, for which clutter attenuations approaching 100 dB are required. The achievement of such high clutter attenuations requires application of techniques and components originally developed for ground-based CW radars, which may require clutter attenuations approaching 130 dB in filters detecting targets with doppler shifts of 10 kHz or greater. *DKB*

Ref.: IEEE (1993), p. 1,029; Barton (1978); Morris (1988); Nathanson (1969), Ch. 11; Hovanessian (1984), pp. 21–24; Skolnik (1990), Ch. 17.

PRF-agility and -diversity radars are capable of rapid change in pulse repetition frequency, as needed to solve ambiguities in range or doppler, and to counter repeater jamming. If the radar can change the PRF from pulse to pulse on a random basis, while integrating over several pulses, it becomes impossible for a repeater to generate false targets at ranges within that of the jammer platform. This mode of operation also provides resolution of range ambiguities, but it does not support coherent integration or MTI rejection of clutter beyond the unambiguous range of the waveform. If pulsed doppler processing or rejection of echoes from beyond the unambiguous range is required, the PRF change must be applied from burst to burst, a process known as *PRF diversity.* Some range and velocity ambiguities on targets detected on more than one burst can be resolved. Clutter beyond the unambiguous range can be canceled if the number of pulses in the burst exceeds the number processed coherently by the number of ambiguous range intervals beyond the first (e.g., a nine-pulse burst permits eight-pulse coherent processing on clutter out to the end of the second range interval). Repeater jamming can generate leading false targets if the response is triggered before the end of the burst, but the false echoes do not receive the full processing gain in the radar. *DKB*

Ref.: Skolnik (1990), pp. 15.34–15.40, 17.19–17.25; Barton (1988), pp. 255–262, 330-336; Chrzanowski (1990), p. 113.

RCS instrumentation radar is used to measure the radar cross section (RCS) of targets on test ranges or in flight over instrumented regions. For greatest accuracy on flying targets, the instrumentation radar will be a monopulse tracker, maintaining a known antenna gain on the target while measuring its aspect angle (the direction of the velocity vector with respect to the radar line of sight). At test ranges where the target is rotated on a turntable, a fixed antenna can be used. The radar requirement in this case is to resolve the target scatterers from clutter originating at the turntable and in surrounding surfaces. Since the antenna sidelobes cannot generally be controlled sufficiently to eliminate such clutter, the current practice is to use wideband waveforms to resolve the target in range. For extreme sensitivity, inverse synthetic aperture (ISAR) techniques with ultrawideband (UWB) waveforms are used to provide a two-dimensional image of the target while resolving it from surrounding scatterers. (See also **RADAR CROSS SECTION measurement methods.**) *DKB*

Ref.: Currie (1989), Ch. 14; Mensa (1991).

A **real-aperture radar** uses an antenna to form its beam and provide angular resolution, as opposed to the use of a synthetic aperture.

Rendezvous radar is a term applied to radar on a space vehicle such as the lunar lander, which must join another spacecraft in orbit. The radar used on the U.S. space shuttle for this purpose is integrated with the communications system, operating in K_u-band. Its major parameters are listed in Table R6. *DKB*

Ref.: Cantafio (1989), Ch. 6.

Table R6
Space Shuttle Rendezvous Radar Parameters

Parameter	Units	Value
Frequency	GHz	13.75 to 14.02
Peak power	W	50
Pulsewidth	μs	0.122 to 66.4
Pulse repetition frequency	Hz	300 to 7,000
Antenna diameter	m	0.9
Gain	dB	38.4
Beamwidth	deg	1.68°
Receiver noise figure	dB	5
System noise temperature	K	1,585
System weight	kg	61
Prime power input	W	460

A **scan-on-receive-only radar** is a tracking radar using conical scan or other sequential lobing technique in the receiving pattern to obtain angle data on targets while holding the transmitting beam fixed during the scan cycle. In the case of conical scan, the technique is called *conical scan on receive only* (*COSRO*), or when two receiving channels are used it is called *conopulse*. In conopulse, the measurement plane of a monopulse pair is rotated about the tracking axis to sample the two coordinates sinusoidally with a 90° phase difference between the two channels. This is equivalent to conical scanning with two beams 180° out of phase with each other. The technique can also be applied to sector-scanning trackers such as the precision approach radar of a ground-controlled approach system, and similar fire control radars, by using a fixed, broadened pencil beam to illuminate the sector scanned by the two receiving antennas.

The advantage of scan-on-receive-only is that it deprives the target of signal data indicating the scan frequency, making it impossible to apply inverse gain jamming and difficult to apply spot scan-frequency jamming for angle deception. (See also **conical-scan-on-receive-only radar**; **CONOPULSE**; **LOBE-on-receive-only.**) *DKB*

Ref.: Lothes (1990), p. 86.

A **search radar** is "a radar used primarily for the detection of targets within a particular volume of interest." A surveillance radar "is used to maintain cognizance of selected traffic within a specified area, such as an airport terminal area or air route." The difference in terminology implies that a surveillance radar maintains track files on the selected traffic, while the search radar output may be simply a warning or one-time designation of a target for acquisition by a tracker or fire control radar. With the advent of modern, 3D phased-array and stacked-beam radars, however, the distinction between the two definitions has less meaning, and search radar is the more general term.

Search radars can be categorized in terms of their objective; that is, air search, surface search, horizon search, or by the number of spatial dimensions in which target data is provided; that is, 2D (range and azimuth) or 3D (angle, azimuth, and elevation). (See **airborne early warning radar**; **air surveillance radar**; **two-dimensional radar**; **three-dimensional radar**) *PCH*

Ref.: IEEE (1993), p. 1,176; Barton (1988), p. 315; Skolnik (1988), pp. 49–166.

A **secondary radar** is "(1) A radar technique or mode of operation in which the return signals are obtained from a beacon, transponder, or repeater carried by the target, as contrasted with primary radar, in which the return signals are obtained by reflection from the target; (2) A radar, or that portion of a radar, that operates on this principle."

Ref.: IEEE (1993), p. 1,178; Honold (1976).

A **secondary surveillance radar (SSR)** is a secondary radar used air traffic control, where it is the primary source of data for the controllers, providing reliable detection, identification of specific aircraft, and readings of the aircraft's barometric altimeter to replace height data unavailable from the associated 2D surveillance radar. Interrogation is at 1,030 MHz with response at 1,090 MHz, eliminating the clutter normally associated with radar operation. In military operations the SSR is often called an IFF (identification, friend or foe), providing an encoded interrogation and processing the properly encoded response from a transponder having the decoding algorithm. The intention is to prevent aircraft response to enemy radars and to avoid "friendly" responses from enemy aircraft.

The main problems of SSR are the limited azimuth resolution of the L-band antenna, sidelobe responses inherent in the one-way interrogation and response paths, and "garble" caused by overlap of coded response from different aircraft. When occupying the full width of an ASR antenna having 1.2° beamwidth at S-band, the SSR antenna will generate a 3° beam, with a width of perhaps 8° at its first nulls. To restrict the azimuth sector over which the beacon will respond, the SSR antenna may be equipped with an interrogator sidelobe suppression (ISLS) feature. The antenna array is divided in the center and the two halves are combined in-phase to produce the normal (Σ) pattern. A difference channel (Δ) is formed in a hybrid combiner, placing the two halves in antiphase and producing a pattern with a deep null at the center, two lobes surrounding the null, and a broad pattern of response extending from the –10-dB points of the lobes over the entire forward hemisphere (Fig. R58). The SSR transmission is divided into three pulses, the outer two transmitted through the sum pattern and the center through the difference pattern. In the transponder, the amplitudes of P1 and P2 are compared, and reply is inhibited if P2 > P1 (corresponding to Δ/Σ > 1. This narrows the response sector to the central portion of the main Σ lobe.

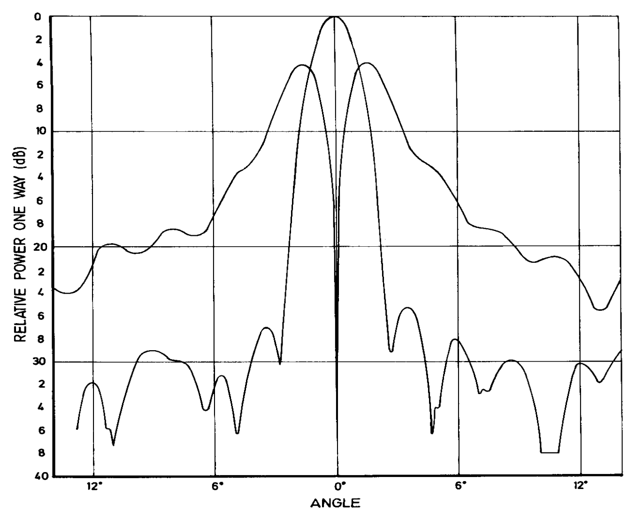


Figure R58 Antenna patterns for interrogator sidelobe suppression system (from Stevens, 1988, Fig. 4.5, p. 44).

Reduction in the total number of response trains reduces the probability of garble, but not enough for high-traffic situations. A new Mode S system (for selective interrogation) uses an interrogator code to request replies from a specific transponder at a time, avoiding overlap of response. (See also **RANGE EQUATION for secondary radar.**) *DKB*

Ref.: Stevens (1988); Skolnik (1970), Ch. 38.

A **sector-scanning radar** is one which scans a beam over a sector less than 360° in azimuth. Nose-mounted aircraft radars are necessarily sector scanners, as are planar arrays without mechanical rotation of multiple faces for 360° coverage. A sector-scanning tracking radar uses two fan beams, one narrow in elevation and scanning in that coordinate, the other narrow in azimuth and scanning in that coordinate. (See **precision approach radar**.) *DKB*

Ref.: Barton (1988), pp. 390–397.

A **semiactive radar** is a bistatic receiver depending on a separated transmitter to provide target illumination. The usual semiactive radar is a homing seeker on an antiair missile, launched from the site or platform on which the transmitter is based. The semiactive radar is distinguished from passive radar in that the latter depends on reflection or radiation of energy not controlled by the user of the radar system. (See also **hybrid radar**.) *DKB*

Ref.: Skolnik (1990), Ch. 19.

A **shipborne radar** is one mounted on a ship, and used for air surveillance, carrier-controlled landing, fire control, instrumentation, missile guidance, or navigation. For further data, refer to the entries on those subjects.

A **sidelooking radar** is "a ground mapping radar used aboard aircraft involving the use of a fixed antenna beam pointing out the side of an aircraft either abeam or squinted with respect to the aircraft axis. The beam is usually a vertically oriented fan beam having a narrow azimuth width. The narrow azimuth resolution can either be obtained with a long aperture mounted along the axis of the aircraft or by the use of synthetic-aperture radar processing." Sidelooking radars usually employ synthetic-aperture radar (SAR) techniques in order to achieve high cross-range resolution, but real beam techniques are applicable as well for mapping, as well as for other missions (e.g., air-to-ground navigation). In the case of real beam mapping, the actual, physical antenna beam is steered, either electronically or mechanically, to maintain observation on the same region of the ground for longer periods than permitted by the radar's physical beamwidth. Provided that system stability requirements are met, this generates higher resolution, or alternatively, more noncoherent integration, or both, in the resulting spotlight map. (See also **synthetic aperture radar**.) *PCH*

Ref.: Hovanessian (1988), p. 127.

Single-channel radar is radar in which the angular coordinate in each coordinate plane is determined by using a single measurement channel. This type of radar is widely used for tracking and nontracking measurement of angular coordinates. *AIL*

Ref.: Dulevich (1978), pp. 260, 360.

A **space-based radar (SBR)** is one mounted on a space vehicle, usually one in orbit about the earth. Three different types of SBR have been developed: (1) the rendezvous radar, a tracking sensor for a guidance system; (2) the remote sensing SBR, designed to survey the earth's surface for geodetic, agricultural, and similar applications; and (3) the surveillance SBR, designed to detect and track moving targets. *DKB*

Ref.: Cantafio (1989); Meneghini (1990).

A **space-target acquisition radar** is one designed as part of a defensive system for acquisition, coordinate determination, and identification of earth satellites and ballistic missiles. The following are the principal features of a space-target acquisition radar:

(1) Potential to acquire small targets at great ranges.

(2) Multifunctionality, such as the ability to combine the tasks of acquisition, measurement of three coordinates, tracking, and identification of targets in one radar.

(3) Multichannel capability, allowing simultaneous operation against a large number of targets.

(4) High sensitivity and use of complex waveforms to solve identification problems.

(5) Use of highly productive multiprocessor units to process radar information.

Space-target acquisition radars are categorized based on purpose, such as missile-warning radars (MWR), antiballistic missile defense (ABM) radars, and Spacetrack radars (STR). Missile-warning radars are designed to detect the fact of a nuclear missile attack and to determine flight time characteristics (number of missiles, their trajectories, impact points, and so forth). Composition, deployment sites, types, and number of missile-warning radars are presented in Table R7.

Table R7
Missile Warning Systems and Radars

Site	Radar complement
BMEWS	
Clear, Alaska	AN/FPS-50 (3) AN/FPS-92 (1)
Thule, Greenland	AN/FPS-50 (4) AN/FPS-49 (1)
Fylingdales Moor, England	AN/FPS-49 (3)
Pave Paws	
Beal AFB, California	AN/FPS-115 (1)
Goodfellow AFB, Texas	AN/FPS-115 (1)
Otis AFB, Massachusetts	AN/FPS-115 (1)
Robins AFB, Georgia	AN/FPS-115 (1)
Individual Radar Sites	
Grand Forks, North Dakota	PAR
McDill AFB, Florida	AN/FSS-1

The early BMEWS radars, the AN/FPS-49 and AN/FPS-50, used reflector antennas with mechanical or electromechanical scanning, and operated in that same band at 300-kW average power (Fig. R59).

Figure R59 AN/FPS-50 BMEWS missile-warning radars at Thule, Greenland (General Electric Co. photo).

During the 1980s, phased-array systems replaced these radars, and new systems were deployed within the continental U.S. to provide warning of intermediate range missiles launched from naval platforms. The Pave Paws radar (Fig. R60) is an example of this class of phased array.

Figure R60 Pave Paws missile-warning radar (from Brookner, 1988, Fig. 4.1, p. 280).

Missile-warning radars are designed to obtain trajectory information on ballistic missiles. When multifunction arrays are used, they may also provide data for warhead selection against a background of decoy targets, and antimissile missile guidance toward selected warheads. These radars include phased-array radars: PAR and MSR. Spacetrack radars are designed for acquisition, tracking, and cataloging of space objects in near-earth space, earth satellites primarily. They include the following radars: AN/FPS-17, AN/FPS-79, AN/FPS-80, and AN/FPS-85. (See also **multifunction array radar**.) *AIL*

Ref.: Leonov (1988), pp. 116–152; Brookner (1988), pp. 25–33.

A **space-target detection radar** is a high-power search radar designed to detect long-range ballistic missiles or objects orbiting the earth. Required detection ranges are measured in thousands of kilometers, and are required on targets whose RCS is 1 m^2 or smaller. An extreme example is an orbital debris detection radar, required to detect objects of centimeter size ($\sigma \approx 10^{-4}$ m^2) at 1,000 to 2,000 km range. Other examples are the space target acquisition radars discussed in the previous section, having ranges of 3,000 to 4,000 km on targets of 0.1 to 1.0 m^2. These radars operate in the 425-MHz band, using large active-module arrays to radiate average powers of hundreds of kilowatts. Several space target radars have also gathered data on missile tests. These include the Tradex radar and others on the Kwajalein Islands in the Pacific, the L-band Cobra Dane (AN/FPS-108) in the Aleutian Islands (Fig. R61), and the S-band Cobra Judy shipboard radar, AN/SPQ-11 (Fig. 6R2). *DKB*

Ref.: Leonov (1988), pp. 116–152; Brookner (1988), pp. 25–33.

Figure R61 Cobra Dane reentry observation radar (Raytheon photo).

Figure R62 Cobra Judy reentry observation radar (Raytheon photo).

Spacecraft-landing-support radar is an onboard radar used to measure altitude and speed during spacecraft landing. Normally in such radars, CW radiation is used to determine the speed of closing with the landing surface, and FMCW for measurement of altitude. An example of such radars are the Apollo spacecraft radar units. On the Apollo spacecraft there are two landing support radars; surveyor for direct control of the craft to a soft landing on the moon and the Apollo lunar cabin radar to provide information to the astronaut on speed and elevation. Specifications of these radars are given in Table R8. *AIL*

Ref.: Skolnik (1970), p. 34.5; Cantafio (1989), pp. 262–268.

A **stacked-beam radar** is "a radar that forms two or more simultaneous beams at the same azimuth but at different elevation angles. The beams are usually contiguous or partly overlapping. Each stacked beam feeds an independent receiver." See **three-dimensional (3D) radar**.

Ref.: IEEE (1993), p. 1272.

Table R8

Parameters of Apollo Spacecraft Landing Radar

Parameter	Surveyor	Lunar module
Frequency	K_u-band	X-band
Altitude capability, m:		
Velocity meter	0 to 15,000	0 to7,500
Altimeter	0 to 12,000	0 to 12,000
Limits of velocity change, m/s	±90	±150
Accuracy, %		
Velocity meter	3	1.5
Altimeter	5	1.4
Modulation:		
Velocity meter	CW	CW
Altimeter	FMCW	FMCW

A **subsurface radar** is one concerned with investigation of a medium by means of radar sounding. Radar methods allow one not only to detect deeply concealed objects and to measure the surface of subsurface layers but also to obtain certain structural and electrical characteristics of media. Subsurface radar is used to measure thickness of layered media (e.g., coal seams), to detect pipes in the ground, cables, and their damage; to evaluate quality of welding work; to detect defects on metal surfaces and objects made of dielectric materials; and to conduct deep sounding and engineering geology, and so forth. Radar subsurface sounding from on board an aircraft allows one to make aerial icy region surveys; to measure thickness of glaciers, ice cover and snow; to make a search of underground structures and supply lines; and to monitor the condition of airfield landing strips, construction platforms, highways, and so forth.

Subsurface radar has a number of specific differences from traditional radar applications. The transmitting and receiving antennas of the subsurface radar, as a rule, are in a single medium, and the object of study is in a different medium, with different parameters of propagation of electromagnetic oscillations. For example, moist clay at frequency of 100 MHz during propagation of waves in one direction yields attenuation of about 30 dB/m, which at a frequency of 1 GHz increases to 100 dB/m. The propagation velocity of electromagnetic waves depends on the medium, and for specific subsurface soundings is approximately known.

An increase of attenuation with increase of frequency makes it impossible to use decimeter and centimeter waves. In subsurface radar one uses metric waves. The transition to the low frequency wave band increases the depth of sounding, and at the same time degrades the resolution capability for range. Therefore, to improve it one must generate ultrawideband pulses. The use of the metric waveband also has led to considerable degradation of directional properties of the antenna; that is, to reduction of angular resolution. In subsurface radar one employs different methods of sounding: pulse, continuous, frequency modulation. Most popular in practice is pulse sounding. (See **impulse radar**.) Sounding in subsurface radar as a rule is carried out by moving antennas along a single coordinate, and the objects of study are stationary. *AIL*

Ref.: Daniels, D. T., Gunton, D. T., and Scott, N. E., *IEE Proc.*, 1988, ***F135***, no. 4.; ERA, *Technology News*, Sept. 1986; Peters, L., Daniels, J. J., and Young, J. D., "Ground Penetrating Radar as a Subsurface Environmental Sensing Tool," *Proc. IEEE* ***82***, No. 12, Dec. 1994, pp. 1,802–1,822.

Surface search [surveillance] radar is used to scan an assigned region of the land or sea surface for any of several objectives: air or sea navigation, target detection or tracking, survey of natural resources, or determination of wind fields over the sea. The objectives and methods depend to a large extent on the platform supporting the radar.

An airborne surface surveillance radar is used primarily for navigation and target detection and tracking. (See **navigation radar**; **ground attack radar**; **personnel-detection radar**.) Survey of earth resources is also an objective in some systems. The methods used include both real-aperture and synthetic-aperture radar, often combined as operating modes of a single military radar system. For location of moving targets, airborne moving-target indication, and related pulsed doppler techniques are used.

Ground-based surface surveillance radar is used primarily for target detection and tracking. (See **personnel-detection radar**.) However, it also sees service as a navigation radar for harbor and waterway ship traffic control. Only real-aperture techniques are available.

Shipborne surface surveillance radar is most frequently applied to navigation of ships and smaller craft on which the radar is mounted, but in naval applications it can also provide fire control data for guns and antiship missiles.

Space-based surface surveillance radar has been used primarily for remote sensing of earth resources, but military and intelligence applications can be assumed. Both real-aperture and synthetic-aperture techniques are applied. *DKB*

Ref.: Barton (1991), pp. 9.10–9.17.

A **surveillance radar** is one "used to maintain cognizance of selected traffic within a selected area, such as an airport terminal area or air route." (See **air surveillance radar**; **search radar**.)

Ref.: IEEE (1993), p. 1315.

A **synthetic-aperture radar (SAR)** is "a coherent radar system that generates a narrow cross range impulse response by signal processing (integrating) the amplitude and phase of the received signal over an angular rotation of the radar line of sight with respect to the object (target). Due to the change in line-of-sight direction, a synthetic aperture is produced by the signal processing that has the effect of a longer antenna." The synthetic array is formed by pointing a real antenna (whose maximum size is restricted by the physical dimensions of the carrier vehicle) broadside to the direction of forward motion of the platform, and coherently processing (summing) the returns from successive pulses. The points at which successive pulses are transmitted can be thought of as the antenna elements of the synthetic array. Synthetic aperture radars, installed aboard aircraft and space vehicles, are capable of achieving very high angular resolution, which, when combined with the high range resolution achievable through pulse compression techniques, make the SAR a nearly ideal all-weather terrain or sea surface mapping system.

The achievable azimuth resolution (crossrange) of a real array antenna is approximately equal to the ratio of the wavelength to array length, multiplied by the range, or

$$\Delta_x \cong \frac{\lambda}{L}R = R\theta_a$$

where θ_a is the beamwidth of the physical antenna of width L. For a SAR, the azimuth resolution is

$$\Delta_x = \frac{\lambda}{2L}R$$

where L is now the length of the synthetic array. The factor of 2 in the denominator comes about because the two-way pattern of the synthetic array has the same shape as the one-way pattern of a real antenna of twice the length. Coherent processing over an extended interval requires adjustment to the phase data as a function of range. When this focusing is accomplished, the maximum cross-range resolution for a SAR is found to be equal to one-half the length of the actual antenna.

The concept of the SAR has been explained, for simplicity, in terms of a broadside-pointing array. For air-to-ground applications, the real beam is also depressed below the horizontal by some angle α, and for real-time mapping, the beam may be squinted forward to map territory ahead of the aircraft. If the real antenna beamwidth is sufficiently wide, the identical target surface area may be mapped several times without a change in the antenna look-angle. With this *multilook* mode, the successive maps are superimposed (noncoherent integration of map data) and the characteristic random fluctuations of amplitude (scintillation) in any single map are averaged, thus improving image quality. Other operational modes common to SAR include moving target display, spotlight mode, and doppler beam-sharpening.

There are two basic types of SAR: focused and unfocused. A focused SAR is one in which the length of a synthetic array is an appreciable fraction of the range to the ground being mapped, the lines of sight from a point at that range to the individual elements of the array will not be equal in length. These range differences translate to phase errors that degrade the quality of the SAR map. In a focused SAR, the phase errors are corrected by applying a phase correction to the return signal at each element of the array. The phase error is proportional to the square of the distance d_n of the element from the center (boresight) of the array, and the phase correction $\Delta\phi$ is the negative of this error, or

$$\Delta\phi = -\frac{2\pi}{\lambda R}d_n^2$$

The primary advantage of a focused array is that the limitation on array length (and object range) due to phase errors is essentially removed, and by increasing the length of the synthetic array, the same resolution may be achieved at any range.

An unfocused SAR is one for which no compensation is made for phase errors in the received signals that arise due to the differences in range from the individual elements of the array to a point on the target surface (ground). To be usable, the synthetic array of an unfocused SAR is limited to an effective length L_{eff} that is a small fraction of the radar range. The maximum allowable length of an unfocused array $L_{eff} = 1.2\,(\lambda R)^{1/2}$. The best achievable cross-range resolution for an unfocused SAR is approximately equal to $0.4L_{eff}$.

Arrested SAR refers to the use of a DPCA canceler preceded by an FFT for doppler beam sharpening.

Bistatic SAR has been used to reduce the contribution of corner reflector targets within the strip map, while preserving essentially the same resolution as for monostatic SAR. Another application of bistatic SAR is to permit the receiving aircraft to approach the target area directly, achieving crossrange resolution by the motion of the other platform. A problem with this mode of operation is that the master oscillators on the two platforms must be held to very tight tolerances.

Other techniques include:

(1) Inverse-synthetic-aperture radar (ISAR), which is "a synthetic-aperture radar in which the angular resolution of the object within the radar beam is caused by the object's own angular rotation, rather than translational motion, relative to the radar."

(2) Spotlight SAR, a technique used by multifunction airborne radars to produce high resolution maps for ground target detection and location. The observation time can be extended, relative to sidelooking SAR, by tracking the target area with the gimballed antenna, and also by flying a circular arc around the target. (See also **ground attack radar.)** *PCH*

Ref.: IEEE (1993), p. 1,328; Skolnik (1970), Ch. 23; Kovaly (1976); Hovanessian (1980); Stimson (1983), pp. 527–48; Wehner (1987), Chaps. 7, 8; Barton (1988), pp. 361–362.; Curlander (1991); Cantafio (1989), p. 438; Currie (1989), pp. 385-396; Willis (1991), pp. 46–49, 234–238, 260–264; Mensa (1991); Barton (1991), pp. J.16–J.25; Schleher (1991), pp. 497–516; Carrara (1995).

Target-detection radar can refer either to a target acquisition radar or to a fire control or missile-guidance radar in its acquisition mode. *DKB*

Terminal-guidance radar can refer either to a homing seeker on a missile or to a missile-guidance radar during the terminal phase of guidance. *DKB*

A **terrain-avoidance radar** is an airborne radar "which provides assistance to a pilot for navigation around obstacles by displaying obstacles at or above the pilot's altitude."

Ref.: IEEE (1993), p. 1,354; Hovanessian (1984), p. 323.

A **terrain-following radar** "works with the aircraft flight control system to provide low level flight following the contour of the earth's surface at some given altitude." This type of radar scans a sector of the surface ahead of the aircraft to measure the height of surface features (ground, trees, and structures) and provides a profile of surface height versus range in the direction of the present or projected flight path of the aircraft. Climb commands or advisories are initiated if the projected path does not meet a preset criterion of adequate clearance. When an obstacle has been passed, the aircraft reduces its altitude until it again has the preset clearance above the surface. The terrain avoidance function may be provided in a special radar or as an operating mode of a multimode military radar. The usual band of operation is X- or K_u-band, and scanning is performed mechanically over a sector consistent with the turn rate of the aircraft (typically ±10° to ±30°). Measurement of terrain height is usually performed using a monopulse technique in elevation. The beam is depressed relative to the aircraft velocity vector to place a level surface in the beam center at a selected range R_0 from the aircraft (Fig. R63). Elevation scan about this angle provides monopulse receiver data identifying the range gate corresponding to the center of the beam, generating the needed surface profile. *DKB*

Ref.: IEEE (1993), p. 1,354; Skolnik (1988), pp. 277–283; Hovanessian (1984), p. 323.

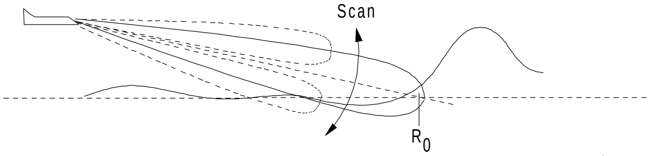


Figure R63 Geometry of terrain-following radar beams.

Terrain-observation radar is installed aboard airborne vehicles and designed to obtain a radar image of the earth's surface and objects located on it. Earth-surveillance radars are categorized as sidelooking radars with fuselage-mounted antenna and sidelooking synthetic-aperture radars. Panoramic radars survey the earth's surface through circular rotation or sectoral wobbling of the antenna beam in the azimuth plane. The antenna shapes a beam narrow in the horizontal (azimuth) plane and sufficiently broad in the vertical plane. As the antenna rotates, terrain sectors are examined sequentially in various directions such that a radar image in azimuth-range coordinates is formed on the indicator screen.

The main drawback of a panoramic radar is low azimuth resolution. A requirement to significantly improve resolution led to creation of new types of earth-surveillance radars, so-called sidelooking radars. A radar with a fuselage-mounted antenna is a radar with a long transceiving antenna placed along the side of the fuselage or in a pod fairing. Terrain is surveilled through displacement of the antenna relative to the earth's surface as the airborne vehicle flies along a straight trajectory. The antenna shapes a beam directed perpendicular to the aircraft flight path; that is, in a lateral direction. High azimuth resolution, as well as prolonged accumulation of the energy of the reflected signals during passage of the resolved sector of the earth's surface in the antenna beam, make it possible to obtain a high-quality image of the terrain and objects. Usually, radars with fuselage-mounted antennas are millimeter- or centimeter-band pulse sets with ranges up to R = 100 km or more. Range resolution is 5 to 30m, azimuth resolution 1.7 to 7.7 mrad. However, despite the significant increase in angular resolution compared with a panoramic radar, radar with a fuselage-mounted antenna do not ensure effective accomplishment of all surface surveillance tasks at great ranges from the airborne vehicle.

High resolution relative to flight-path range at great distance from the airborne vehicles is possible using radars with a synthetic-aperture antenna (SAR). The principle of operation of a SAR is based upon use of the straight-line movement of the radar antenna for sequential shaping of an antenna array on the flight trajectory. SARs usually are used by centimeter-band pulse radars with a range of up to 100 km and surveillance band of up to 20 km. Range and azimuth resolution is 3 to 15m. As a rule, they have moving target indicators. earth-surveillance radars are used to reconnoiter the battlefield, areas where military equipment is concentrated, airfields, moving objects, rail lines and highways; for cartography; in engineering and geological reconnaissance; for determination of the ice situation; compilation of vegetation and snow-cover maps; ecological reconnaissance; and so on. The most typical examples of radars with a fuselage-mounted antenna include the AN/APS-94D and AN/APQ-97. Examples of synthetic aperture radars include the AN/APQ-102A and AN/ARD-10. (See also **ground-mapping radar**.) *AIL*

Ref.: Kondratenkov (1983), pp. 6–27, 136-149; Cantafio (1989), p. 28; Ulaby (1981), (1982), (1986); Skolnik (1988), Ch. 6.

A **three-dimensional (3D) radar** is "a radar capable of producing three-dimensional position data on a multiplicity of targets." In addition, the 3D radar has a number of other advantages over the two-dimensional (2D) radar: (1) the antenna aperture height is not limited by the required elevation coverage; (2) an increased aperture height allows better control of the elevation coverage pattern, permitting a sharp cutoff at the horizon and more uniform coverage above the horizon; (3) if the entire receiving aperture is used for upper elevation beams, the increase in effective search sector for csc^2 coverage is less than that for a 2D radar; (4) surface clutter can be eliminated or greatly reduced in all but the lower beam position; (5) rain clutter and chaff are received at reduced ranges in the upper beams, and with reduced spectral spread in all beams; (6) data from adjacent 3D radars can be combined to yield range information on the jammers through triangulation, with less ambiguity (ghosting) than in the 2D radar case; (7) the correct ground range (and, hence, x and y coordinates) can be derived for targets at high altitude.

There are three fundamental approaches to implementation of 3D radars: (1) a single pencil beam may be scanned in elevation rapidly enough to cover the entire vertical sector during the time it takes for the azimuth scan to cover one azimuth beamwidth; (2) the required vertical coverage may be obtained simultaneously by stacked beams, wherein every elevation beam realizes the full azimuth-scan-determined time-on-target t_o = (search frame time × azimuth beamwidth) ÷ (azimuth search sector width); or (3) a set of n beams may be used in parallel to cover a portion of the vertical sector, the set scanning over the entire vertical coverage region during the azimuth-scan-determined time-on-target. The choice of method used to implement the 3D search coverage is largely dependent upon the dwell time required to support doppler processing and diversity gain on the target. The stacked-beam approach provides the maximum dwell time for a given search frame time and aperture width. Whether the other methods described can be used successfully depends largely on the expected clutter environment.

An example of the scanning pencil beam is the AN/TPS-59 solid-state, L-band, 3D radar (Fig. R64). The tall antenna provides a narrow beam for accurate measurement of target altitude, and the narrow width provides sufficient azimuth dwell time for the sequencing of the scanning elevation beam over the coverage sector.

The stacked-beam approach is illustrated by the Martello (Fig. R65), also an L-band radar with a tall antenna to provide accurate target altitude. In this case the narrow width provides time on target for multiple-pulse integration and doppler processing.

The hybrid approach in which multiple beam are scanned to cover different elevation sectors is illustrated by the AN/SPS-48 S-band, frequency-scanned, 3D radar (Fig. R66). Nine narrow beams are generated at a time, using a transmitter pulse with nine subpulses at different frequencies, covering an elevation sector width of approximately 10°. A wideband receiver accepts echoes from all beams and routes them to separate frequency channels. During each azimuth dwell there are three transmissions covering an elevation sector of some 30°.

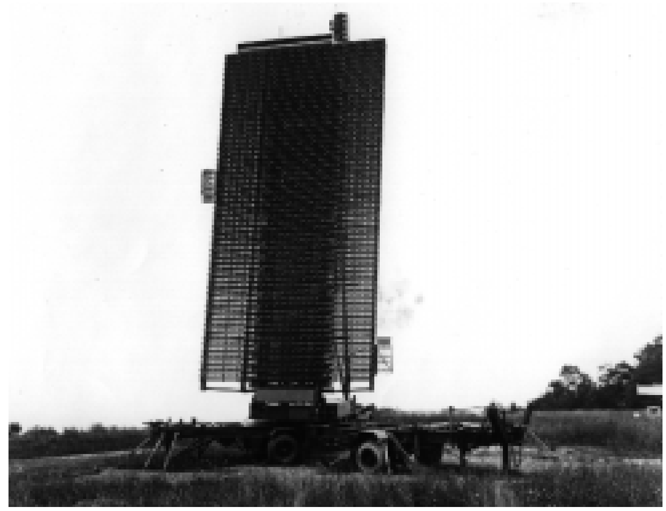


Figure R64 AN/TPS-59 scanning-beam 3D radar. In this design, each row of elements has its own RF module, containing a power amplifier, a low-noise receiving amplifier, and a phase shifter (General Electric Co. photograph).

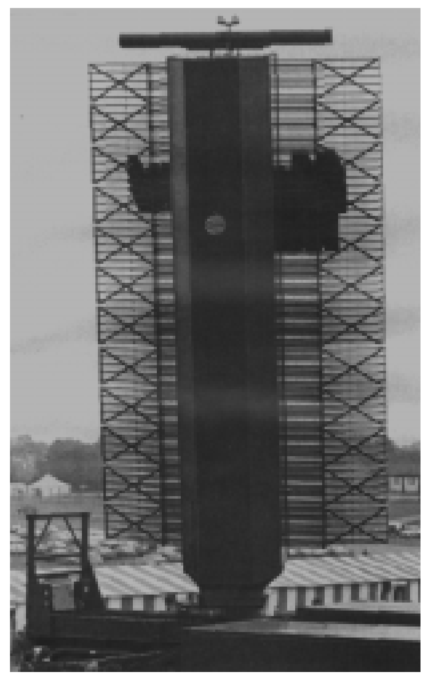


Figure R65 Martello stacked-beam 3D radar. In this design, each row of elements has its own receiver front end, and an IF beam-forming matrix is used to generate the multiple elevation beams (Marconi Co. photograph).

In theory, any of the 3D techniques previously described can be implemented in a fully phased array radar (i.e., no mechanical scanning in any coordinate), but the high cost associated with obtaining 360° coverage in azimuth in this manner has led to the deployment of rotating stacked-beam radars that use conventional line or horn-fed antenna arrays. In several “hybrid” search radar designs, the mechanical scan in azimuth is retained, and the elevation search is implemented using phase- or frequency-scanning techniques. Full

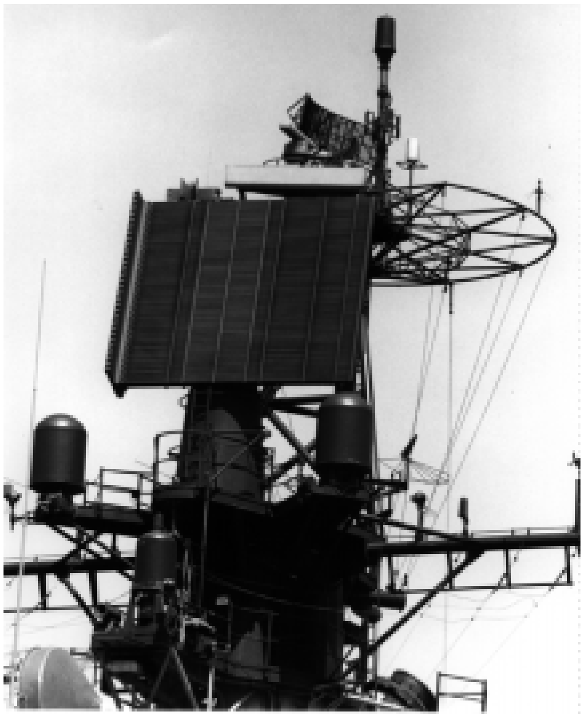


Figure R66 AN/SPS-48 multiple-beam scanning 3D radar (ITT Gilfillan photograph).

phased-array implementations have generally been restricted to date to those applications, such as air defense system target acquisition and tracking radars, that do not require a single radar to cover an entire 360° azimuth sector. Special-purpose multiface phased-array radars have been built for long-range surveillance (e.g., the AN/FPS-115 Pave Paws), but these are basically for strategic defense and are relatively few. *PCH*

Ref.: Barton (1988), pp. 345–349; Skolnik (1990), Ch. 20.

A **tracking radar** is one whose "primary function is the automatic tracking of targets." The conventional type uses a parabolic reflector or flat-plate array to generate a narrow pencil beam, maintaining the axis of this beam on the target by sequential lobing or monopulse sensing to provide error signals to an electromechanical servomechanism. Modern multitarget and multifunction tracking radars use electronically scanned array antennas to produce monopulse patterns, sequencing these among several targets at a rate that supports sampled-data tracking loops in an associated computer. This mode of operation may be distinguished from track- while-scan by the fact that the beam positions are controlled by the tracking loops.

The lobing technique used in a tracking radar may be conical scan, conical-scan-on-receive-only (COSRO), sector scan, or monopulse (simultaneous lobing), the last being most common in systems designed since 1960. Antenna techniques include parabolic reflectors, fixed lenses, and flat-plate arrays with fixed-phase radiating elements, for single-target trackers, and space-fed or corporate-fed phased arrays for multitarget trackers. Applications of tracking radar include fire control, missile guidance, and test range instrumentation.

The most advanced type of tracking radar is a four-coordinate tracking radar, one using pulsed doppler techniques, in which the target radial velocity is tracked as well as the three spatial coordinates. Advantages of this over conventional, three-coordinate trackers include the ability to reject clutter to a greater degree than is found in trackers using moving target indication, provision of accurate radial velocity data, reduced vulnerability to countermeasures, and ability to maintain track on signals of low single-pulse signal-to-noise ratio, as long as the signal energy ratio in the tracking time constant is adequate. *DKB*

Ref.: IEEE (1993), p. 1389; Hovanessian (1984), p. 98; Barton (1964), pp. 387–389, (1988), Chs. 8–11; Skolnik (1990), Ch. 18; Neri (1991), p. 135.

A **track-via-missile radar** (also known as *target-via-missile*) is a type of semiactive seeker in which the target signals are retransmitted to the illuminating radar for processing. The guidance commands are then generated in the radar's computer and uplinked to the missile. Originally intended to relieve the seeker of having to perform complex signal processing, the technique has become almost obsolete with the advent of modern microelectronics and digital signal processing. *DKB*

Ref: Skolnik (1990), p. 19.20; Chrzanowski (1990), p. 167.

A **track-while-scan (TWS) radar** is one using the TWS process, defined as "an automatic target tracking process in which the radar antenna and receiver provide periodic video data from a search scan, together with interpolation measurements, as inputs to computer channels that follow individual targets." Any surveillance or sector-scanning radar may qualify when equipped with suitable signal and data processing equipment. However, to maintain given tracking accuracy on maneuvering targets the radar must scan at an adequately high rate, with high single-scan probability of detection, to avoid excessive lag errors in the tracking loops.

The tracking process involves the extraction of single-scan reports on target position in each coordinate, by interpolation within the range and angle resolution cell in which the target is detected; correlation of these reports with existing track files; updating of track files using the α-β type of recursive filter; initiation of new track files on detections that fail to correlate with an existing track; and terminating tracks when no correlated detection occurs within a given number of scans (see **FILTER, α-β, Kalman**). *DKB*

Ref.: IEEE (1993), p. 1389; Skolnik (1990), pp. 8.23–8.40; Blackman (1986); Barton (1988), pp. 391–397, 465–466.

A **two-dimensional (2D) radar** mechanically scans a fixed beam, either in the azimuth plane (the conventional air search radar) or in the elevation plane (the nodding height-finder radar). A fundamental property of 2D air search radar is that the height of the antenna aperture must not be greater than that required to produce an elevation beamwidth that is matched to the required vertical coverage sector. Smaller values may, for mechanical reasons, be used, but at the expense of wastefully spreading energy above the required elevation coverage sector. The width of the antenna is determined by the azimuth resolution required and the requirement for obtaining a given minimum observation time (or time on target) while scanning. As in the case of the 3D search radar, this minimum observation time is largely determined by the doppler processing requirements, which in turn, are a function the clutter environment, and the diversity gain needed on the target.

For air search operations over broad elevation sectors, the minimum power-aperture product strongly favors use of the lower radar frequencies in 2D radar, although higher frequencies are often used for surface search, where the target is a land or sea-vehicle, a surface (for navigation), or a fixed structure. Operation at microwave frequencies is feasible here because, for surface-based radars, the required elevation sector is generally small, while for airborne radars the required elevation search sector is either small enough to match the beamwidth of the available antenna aperture height, or such as to require only a few overlapping "bar" scans in elevation. (See also **COVERAGE, radar**; **POWER-aperture product**; **search radar**; **three-dimensional (3D) radar.**) *PCH*

Ref.: Barton (1988), pp. 319–327; Skolnik (1988), pp. 49–115; Neri (1992), p. 118.

An **ultrawideband (UWB) radar** is one whose instantaneous signal bandwidth exceeds 50% of its center frequency. Waveforms used in UWB radar may be short impulses, producing fewer than two cycles of the carrier at the antenna output, or swept FM signals for which the sweep width Δf meets the UWB criterion. Radars of this type, unless their power is very low, violate the frequency allocations of the International Telecommunications Union (ITU), but may be used in specialized military and civil applications. The latter include ground penetrating radars for locating pipelines and similar buried features, intrusion alarms, and other search radar applications in which radiated power does not create excessive interference with other services. Military applications are primarily for measurement of radar cross section, although target-detection applications have been claimed. A report summarizing the similarities and differences between UWB and conventional radar has been published. *DKB*

Ref.: Fowler, C. A., Entzminger, J., and Corum, J., "Assessment of Ultra-Wideband (UWB) Technology," *IEEE AES Magazine*, Nov. 1990, pp. 45–49; Taylor (1995).

V-beam radar is "a ground-based, three-dimensional radar system for the determination of distance, bearing, and, uniquely, the height or elevation angle of the target. It uses two fan-shaped beams, one vertical and the other inclined, that rotate together in azimuth so as to give two responses from the target; the time difference between these responses, together with distance, are used in determining the height of the target." (See **HEIGHT-FINDER, V-beam**).

IEEE (1993), p. 1,453.

volumetric radar (see **search radar**; **three-dimensional radar**; **two-dimensional radar**).

weather-navigational radar (see **airborne weather-avoidance radar**)

weapon location radar (see **artillery and mortar location radar**).

A **wideband radar** is one whose signal bandwidth B is less than 50% of its center frequency f_0 but is such as to resolve expected targets into more than one range cell. Most such radars are limited to $B < f_0/10$ by the bandwidth of RF components in the transmitter, antenna, and receiver. *DKB*

Ref.: Wehner (1994).

RADAR-ABSORBENT MATERIALS (see **ABSORBER, radar**).

RADAR ALTIMETER (see **ALTIMETER, 'radar**).

RADAR ASTRONOMY (see **ASTRONOMY, radar**).

RADAR BANDS (see **FREQUENCY**).

RADAR BEACON (see **RADAR, secondary**).

RADAR BUOY (see **BUOY, radar**).

RADAR CONTRAST (see **CONTRAST, radar**).

RADAR COORDINATES (see **COORDINATES, radar**).

RADAR COVERAGE (see **COVERAGE, radar**).

RADAR CROSS SECTION is "a measure of the reflective strength of a radar target." The usual notation is σ. For engineering evaluations it may be defined as

$$\sigma = 4\pi\frac{P_s}{P_i}$$

where P_s is the power per unit solid angle scattered in a specific direction and P_i is the power per unit area in a plane wave incident on a scatterer from a specified direction. The units for RCS are square meters. As RCS can span a wide range of values (from about 10^{-5} m^2 for insects and low RCS missiles warheads to 10^6 for large ships), a logarithmic power scale is also used with a typical reference value σ_{ref} equal to 1 m^2:

$$\sigma_{dBsm} = 10\log\left(\frac{\sigma}{\sigma_{ref}}\right)$$

Typically two notations are used interchangeably: dBsm and dBm^2 (the latter is less commonly used). A comparison of the linear and logarithmic scales are shown in Fig. R67.

Typical RCS values for common artificial and natural objects are shown in Fig. R68.

There are several parameters used to describe RCS:

(1) In theoretical calculations three types of RCS are distinguished: absorption RCS, extinction RCS, and total RCS.

(2) From the standpoint of the direction of backscattered energy also three cases are distinguished: backscattering or monostatic RCS, forward-scattering RCS and bistatic (or in general case multistatic) RCS.

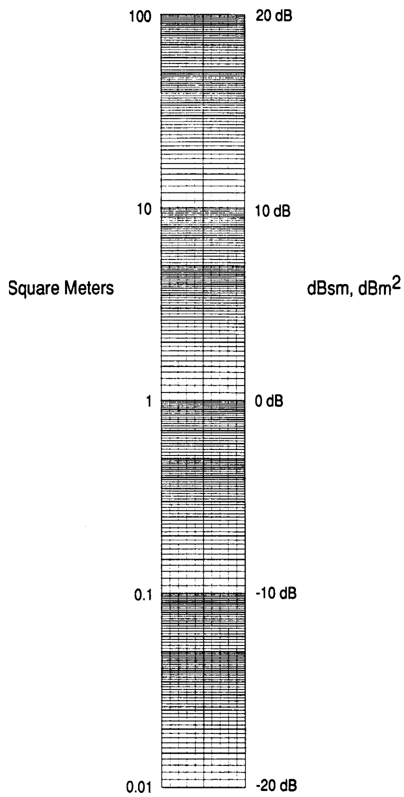


Figure R67 RCS linear square meter and logarithmic decibel scales compared (from Knott, 1993, Fig. 3.3, p. 69).

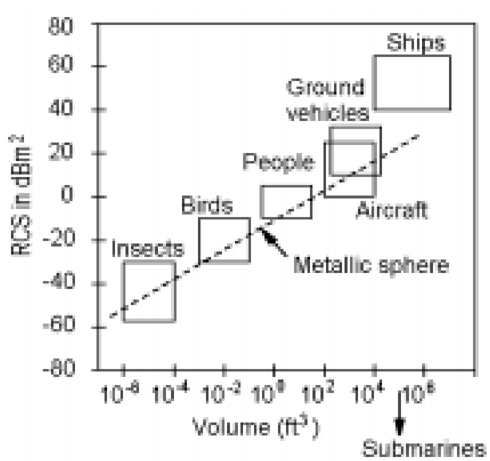


Figure R68 Typical RCS (after Brookner, 1988, Fig. 8.23, p. 436).

(3) From the point of view of the motion of the target relative to radar dynamic and static RCS may be defined.

(4) As to the type of polarization used on receive in respect to the transmitted one copolarized and cross-polarized RCS may be introduced.

(5) Another common feature used to classify RCS is whether it is related to simple or complex objects (RCS of simple shapes and RCS of complex shapes). First are the elementary profiles (cones, disks, plates, spheres, wires, etc.) whose RCS can be evaluated using the exact (analytical) methods of RCS prediction technique. Practically all natural (ground, sea, birds, hills, trees, etc.) and artificial (aircraft, ships, missiles, satellites, and others) targets are the complex objects in the most cases involving the experimental methods of RCS prediction. In general the performing of reliable RCS measurement for real targets is rather complicated problem as a variety of factors that influence target reflectivity in different ways must be considered thoroughly and in a proper manner.

The main factors that influence the value of RCS are: radar wavelength, radar polarization, propagation and multipath effects, target size, shape, orientation (aspect angle), material composition, surface coating and roughness, and moisture content. In many cases of radar applications RCS augmentation or conversely RCS reduction are desirable. The synonyms often used for RCS are simply cross section and effective echoing area. *SAL*

Ref.: IEEE (1993), p. 1,052; Ruck (1970); Barton (1988), pp. 99–122; Barton (1991), pp. 5.2–5.13; Skolnik (1990), pp. 11.1–11.51; Knott (1993); Bhattacharyya (1991); Currie (1989); Morchin (1993), pp. 89–130; Long (1992), pp. 131–204.

Absorption RCS is the amount of power absorbed by the target normalized to the incident power density. The usual notation is σ_a. It is the measure of the absorbed incident power that is not available for reradiation for imperfect conductor targets (e.g., those using radar absorbing materials) when the part of the energy is turned into heat. It is may be calculated as

$$\sigma_a = \sigma_e - \sigma_t$$

where σ_t and σ_e are the total RCS and extinction RCS respectively. *SAL*

Ref.: Bhattacharyya (1991), p. 15; Knott (1993), p. 68.

The **RCS of aircraft and helicopters** depends on the type, size, radar band, and the aspect angle. The data obtained in a series of experiments for a variety of aircraft, aspect angles, and radar bands shows that in many cases the Rayleigh distribution is a good approximation for aircraft RCS statistics, although there were exceptions, especially for smaller aircraft and for all aircraft at broadside. The main echo regions that characterize RCS signature are the nose, broadside, and tail. A simple approximation for the nose-on aircraft RCS is

$$\sigma = 0.01L^2$$

where L is the aircraft length.

The classical picture of RCS pattern of a B-26 bomber showing the lobe structure of this pattern for two different frequencies are shown in Fig. R69.

Median RCS for various classes and aspect regions for three aircraft with the typical size and nose-on RCS for a variety of different aircraft are given in the Tables R9 and R10. By using special measures of RCS reduction, cross section of aircraft can be reduced considerably (see **RCS reduction**) *SAL*

Ref.: Skolnik (1980), p. 38; Nathanson (1990), p. 155; Morchin (1993), p. 110.

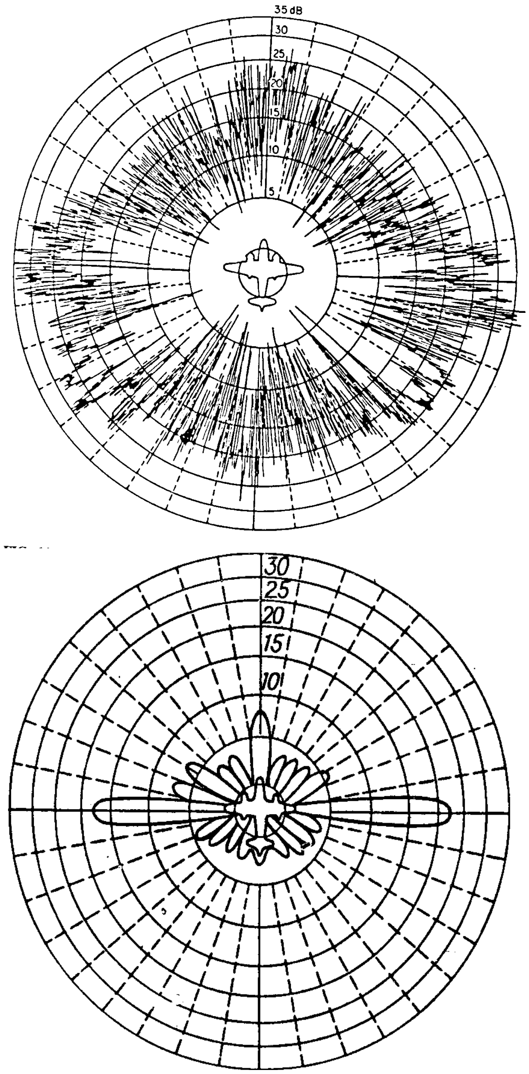


Figure R69 RCS pattern of B-26 bomber in the horizontal plane: (a) at wavelength 10 cm; (b) at 3 to 5m (from Leonov, 1988, Fig. 1.2, p. 10).

Table R9
Median Aircraft RCS
(for various classes and aspect regions, values in m^2)

Class	Aspect	VHF 0.03-0.3	UHF 0.3-1.0	L 1-2	S 2-4	C 4-8	X 8-12
Large, heavy bomber or 707, DC-8, size jet	Nose			32	40	10-40	28
	Tail			15/63	250		25/500
	Broad			550/500	500		300/300
	Av		11	24/40	23/40		60
Medium attack bomber, 727, DC-9, size jet	Nose	60/100	1.2/20/30	6/10/20			4
	Tail	100	6	20	20		
	Broad	300	200	300	280		800
	Av			12	12		11
Small fighter or four-passenger jet	Nose	10	6	0.3-7.5	0.2-9.5	0.7-2	1.2
	Tail	10	3	2	0.6-15		
	Broad			5-90	35-300		20/30/65
	Av			0.7/1.3	1.3/2		1.3/5

(from Nathanson, 1969, Table 5-3, p. 159)

Table R10
Nose-on RCS of Aircraft

Designation	Name	RCS (m^2)
Aero 500	Aero Commander	6.5
B-1A	prototype	10
B-1B		1
B-26		100
B-29	Super Fortress	100
B-47	Stratojet	16
B-47		60
B-52	Strato Fortress	125
B-57B	Canberra	10
707	Boeing 707	32
707	at UHF	10
707	at L-band	25
707	at S-band	40
707	at C-band	50
707	at X-band	60
727	Boeing 727	10
727	(broadside)	125
720, 727	Boeing 720, 727	25
737	Boeing 737	3
737	Boeing 737	600
Britannia	Bristol Britannia	25
C-54		60
C-121	Constellation	100
C-130	Hercules	80
	Caravelle	16
	Comet	40
	Cessna 180	1.5
	Cessna 310B	4
240, 340, 440	Convair Metropolitan	40
DC-3	Dakota	20
DC-8		32
DC-9		25
F-1/FJ	Sabre	5
F-4	Phantom	10
F-9F	Cougar	12.5
F-27	Fokker Friendship	25
F-86	Sabre	5
F-104	Starfighter	5
IL-28	Beagle	8
	Javelin	8
	Lamps	25
L-1011	(broadside)	49
Mig-21	Fishbed	4
P-3A	Orion	80
P-3B	Orion	95
T-38		1
TU-16	Badger	25
TU-20	Bison	40
TU-95	Bear	125
	Viscount	16

(from Morchin, 1993, Table 4.1, p. 111).

The **RCS of antennas** is contributed by two components: a structural mode defined by the antenna as a scatterer of a given shape, size, material (this mode is independent of the fact that antenna is a device specially designed to transmit and receive RF energy), and antenna mode relating to the antenna as a device radiating and receiving energy with the specific pattern. If no special measures of RCS reduction are taken, the RCS of an antenna can be as large as that of a flat plate with the area of antenna. In this case antenna RCS is equal to

$$\sigma = \frac{\lambda^2}{4\pi} G^2 \Gamma$$

where G is antenna gain and Γ is the power reflection coefficient of the antenna feed. If the antenna is perfectly matched, the antenna mode component is equal to zero. *SAL*

Ref.: Morchin (1993), p. 119; Knott (1993), pp. 407–447.

RCS augmentation is the enhancement of RCS. Primarily this technique is required when it is necessary to enhance the returns from small vehicles (boats, light aircraft, radar buoys, and etc.) in civil applications, to enhance deception echoes in passive jamming in military applications, and to calibrate radar systems in the period of test. There are three basic methods to achieve RCS enhancement: proper shaping of the target to achieve a large echo area; using of impedance loading at some points on the target to disturb the induced current; usage of a special augmentation device or RCS augmenter. The popular RCS augmenters are corner reflectors and Luneburg lenses. *SAL*

Ref.: Bhattacharyya (1991), pp. 115–140.

The **RCS of an automobile** typically increases with increasing frequency, and at X-band is generally greater than that of an aircraft or boat. At frontal aspect the RCS may vary from 10 to 200 m^2, and a typical value is 100 m^2. *SAL*

Ref.: Skolnik (1980), p. 44.

backscattering RCS (see **monostatic RCS**).

RCS of birds and insects. Birds and insects can be significant sources of angel echoes, especially when they travel in flocks. Typical mean RCS of single birds and insects is given in the Table R11. *SAL*

Ref.: Skolnik (1980), p. 509; Nathanson (1990), p. 169.

bistatic RCS (see **multistatic RCS**).

RCS of clutter. Most natural targets, if they are not the objects of interest for the radar, may be considered as clutter. The primary sources of clutter in most radar applications are the ground and sea surfaces, precipitation, birds and insects. (See **CLUTTER**; **RCS of birds and insects.**) *SAL*

RCS of complex shapes. Complex shapes are shapes that may be decomposed into the ensemble of simple components. (See **RCS of simple shapes.**) Most of real radar targets are complex shapes (see **RCS of antennas**, **RCS of aircraft and helicopters**, **RCS of birds and insects**, **RCS of marine targets**, **RCS of missiles and satellites**). The RCS of complex targets can be determined either experimentally (see **RCS prediction methods**) or by decomposition into a series of simple shapes and estimation the RCS of ingredients theoretically, followed by summing of the RCS of separate components. The example of the last approach is shown in Fig. R70, where letters on the RCS curve indicate the predominant contribution to the RCS of each angular region. *SAL*

Ref.: Barton (1991), p. 5.8; Skolnik (1990), pp. 11.13–11.18; Morchin (1993), pp. 110–122; Currie (1989), p. 50; Stone (1990); Bhattacharyya (1991) pp. 85–96.

RCS of cone (see **RCS of simple shapes**).

Table R11
Typical Mean RCS of Single Birds and Insects, dBsm

Type, length	Aspects observed	Frequency Band (vert. polarization) UHF 0.45 GHz	S 3.0 GHz	X 9.0 GHz
Sparrow	Head			−46
	Broad-side			−32
	Tail			−47
	Average	−56	−28	−38
Pigeon	Head			−40
	Broad-side			−20
	Tail			−40
	Average	−30	−21	−28
Duck	Head	−12		
Grackle		−43	−26	−28
Hawkmoth, 5.0 cm		−54	−30	−18
Worker bee, 1.5 cm		−52	−37	−28
Dragonfly		−52	−44	−30

(from Nathanson, 1969, Table 5-6, p. 169, reprinted by permission of McGraw-Hill)

The **RCS of convex objects** varies with angle (except for the sphere), but has an average value approximately equal to one-fourth the total surface area. This relationship applies when specular reflection dominates the RCS, when there are no corner-reflector type surfaces. As the aspect ratio (the ratio of length to width) of the object increases, the major contributions to average RCS are large lobes centered at the normal to the longitudinal axis, leaving a low value for broad sectors about the nose and tail and giving a low value of median RCS. *DKB*

Ref.: Kerr (1951), p. 467.

Copolarized RCS is the radar cross section when the polarization of the scattered field coincides with the polarization of the incident field.

Cross-polarized RCS is the radar cross section describing the cross-polarization phenomenon when the polarization of

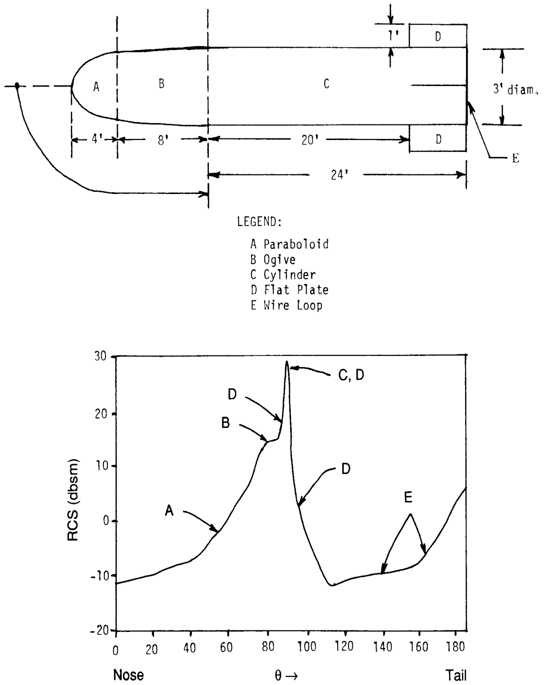


Figure R70 Complex target and resulting estimated RCS (from Morchin, 1993, Fig. 4.2, p. 92).

the scattered field orthogonal to the polarization of the incident field is of interest in radar applications. Typically, real radar targets are asymmetrical and the materials used for their construction are nonideal, so depolarization phenomena occur in the scattered field. Cross-polarized components of radar return contain detailed information about the target. They can be useful in different tasks of radar target discrimination and classification. *SAL*

Ref.: Bhattacharyya (1991), p. 97.

RCS of cylinder (see **RCS of simple shapes**).

RCS density is the RCS per unit area or volume, used to describe the backscattering properties of distributed targets (e.g., clutter), and also called reflectivity or specific RCS. *SAL*

Skolnik (1980), p. 471.

Differential RCS is the same as the general RCS definition (see **RADAR CROSS SECTION**), but the name stresses the fact that it gives the angular distribution of the scattered power. *SAL*

Ref.: Knott (1993), p. 68.

The **RCS of diffusely scattering objects** tends toward a local Rayleigh distribution about a local mean value which approximates the projected area of the object on a plane normal to the incident ray. *DKB*

Dynamic RCS is the term used for RCS of the target when there is the relative motion between the target and the radar causing doppler shift and rapid change of aspect angle. Typically, under these conditions the radar signal is more complicated that complicates the signal processing. The example of dynamic RCS for the satellite 1957β (Sputnik II) is given in the Fig. R71. *SAL*

Ref.: Bhattacharyya (1991), p. 30.

RCS of disk (see **RCS of simple shapes**).

RCS of edges (see **RCS of simple shapes**).

Effective RCS is that of a point target σ modified by the pattern-propagation factor F^4 which makes a correction for the effect of interference lobes in the case of multipath propagation. The usual notation is σ_{eff}, and the evaluation formula is

$$\sigma_{\text{eff}} = \sigma F^4$$

Under the assumption of a target located below the first reflection lobe over the flat, smooth earth

$$F^4 = 16\sin^4\left(\frac{2\pi h_r h_t}{\lambda R}\right)$$

where h_r = antenna height; h_t = target height; R = range; and λ = radar wavelength. *SAL*

Ref.: Barton (1988), pp. 337–339; Long (1992), p. 118.

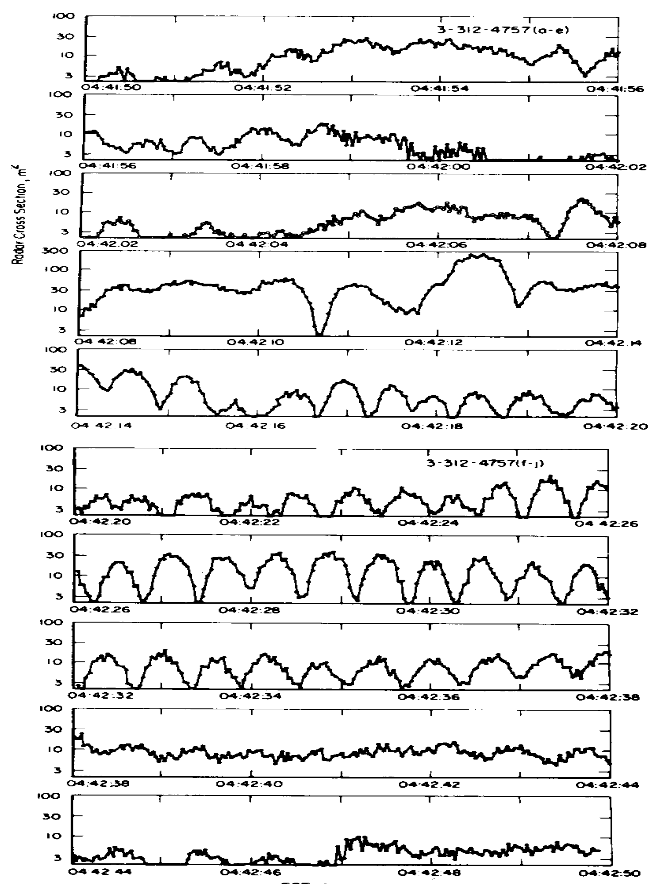


Figure R71 Plot of RCS in m^2 versus time for satellite, 1957 measured at 440 MHz (from Bhattacharyya, 1991, Fig. 1.18, p. 31).

RCS of ellipsoid (see **RCS of simple shapes**).

RCS evaluation (see **RCS prediction methods**).

Extinction RCS is the measure of the total power extracted from the incident wave and is equal to the sum of the total RCS and the absorption RCS. The usual notation is σ_e. *SAL*

Ref.: Knott (1993), p. 70; Bhattacharyya (1991), p. 15.

RCS fluctuations are the variations of target cross section with time. Primarily these variations arise from changing of the target aspect relative to radar and they result in variations in radar return additional to those caused by meteorological conditions, equipment instabilities and so on. Typically, the value of RCS changes with time in a stochastic manner as the type of the target, its trajectory, and aspect angle in any moment of time are not exactly known. So usually the threshold is set and statistical parameters are used to describe the RCS in the terms of being above this threshold with some probability during some fraction of time (0.95 and 0.99 are the typical values of the thresholds). For the detailed and more correct description of RCS behavior probability density function and correlation properties (autocorrelation function and spectral density) of RCS should be taken into account. There are two general ways to obtain information about the statistical properties of RCS fluctuations: to get the experimental data about the behavior of the targets of interest in different dynamic situations, and to introduce some theoretical analytic models that can describe this behavior in a satisfactory manner.

For radar performance analysis the Swerling models are commonly used (Table R12). Each Swerling case corresponds to a set of conditions that approximate some real target. The Swerling models correspond to chi-square distributions with $2K$ degrees of freedom, where the value of I is shown in the table. Radar detection performance for $P_d > 0.35$ is best for Case 0, with cases 4, 2, 3, and 1 following in that order (see **LOSS, fluctuation**). Broader distributions, applicable to certain types of target giving even poorer detection performance than Case 1, can be represented by chi-square distributions with $K < 1$, also known as Weinstock distributions.

The statistics of RCS fluctuations can also represented by the gamma or Rice distributions.

The relationship between the RCS fluctuation spectrum and target size can be used as the recognition feature to determine the size of the observed target. *SAL*

Ref.: Skolnik (1980), pp. 46–52; Dulevich (1978), pp. 36–48.

Forward-scattering RCS is the RCS when the incident and scattering directions and senses are the same. When added to the incident field, the electric field scattered in the forward direction forms a shadow behind the target. *SAL*

Ref.: Barton (1991), p. 5.2; Knott (1993), p. 71.

RCS of ground surface (see **CLUTTER, land**).

RCS for high-resolution radar differs from RCS for low resolution radar by the fact that in this case many individual scatterers of the target can be resolved. In this case an average narrowband RCS, σ, under the assumption that the target contains a large number of equal backscattering elements, can be given by formula

$$\bar{\sigma} = n\sigma_e$$

where n is the number of the elements and σ_e is the average RCS of each element. *SAL*

Ref.: Wehner (1987), p. 27.

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Table R12
Summary of Swerling Target Models

Case	Application	Type of fluctuation	PDF	Correlation time	Chi-square duo-degrees of freedom, K
0	Steady target	None		∞	∞
1	Many comparable scatterers (basic model for all complex targets)	Slow, correlated from pulse to pulse but uncorrelated from scan to scan	$W(\sigma) = \frac{1}{\bar{\sigma}}\exp\left(-\frac{\sigma}{\bar{\sigma}}\right)$	$t_c >> t_o$	1
2	Targets rotating very rapidly, or observed by radars with pulse-to-pulse frequency agility	Fast, uncorrelated from pulse to pulse	$W(\sigma) = \frac{1}{\bar{\sigma}}\exp\left(-\frac{\sigma}{\bar{\sigma}}\right)$	$t_c < t_r$	t_o/t_r
3	(a) Target with one predominant scatterer; (b) Case 1 target observed by dual- diversity radar	Slow fluctuations like Case 1	$W(\sigma) = \frac{4}{(\bar{\sigma})^2}\exp\left(\frac{2\sigma}{\bar{\sigma}}\right)$	$t_c >> t_o$	2
4	Case 2 target observed with dual-polarization radar	Fast, like Case 2	$W(\sigma) = \frac{4}{(\bar{\sigma})^2}\exp\left(\frac{2\sigma}{\bar{\sigma}}\right)$	$t_c < t_r$	$2t_o/t_r$

Notes: t_o is observation (integration) time; t_c is correlation time; t_r is pulse repetition interval.

Instantaneous RCS is that of a target taken at any particular moment of time. Dynamically, instantaneous RCS is the function of aspect angle. *SAL*

Ref.: Skolnik (1990), p. 2.13.

The **RCS of a man** varies from fractions of a meter to several meters, depending on aspect angle, frequency, and polarization. Some typical figures are shown in Table R13. *SAL*

Ref.: Skolnik (1980), p. 44.

Table R13
RCS of Man

Frequency (MHz)	RCS (m^2)
410	0.03–2.3
1,120	0.01–1
2,890	0.14–1.0
4,800	0.4–2.0
9,375	0.5–1.2
(after Skolnik, 1980, p.44)	

RCS of marine targets. The reflecting properties of a maritime vessel depend on its displacement and at low elevation angles can be found from an approximate expression where σ

$$\sigma = 1644 F^{\frac{1}{2}} D^{\frac{3}{2}}$$

is RCS in square meters, F is the radar frequency in gigahertz, and D is the ship displacement in kilotons. Generalized values of RCS for a variety of ships is given in Table R14. *SAL*

Ref.: Skolnik (1980), p. 42; Nathanson (1990), pp. 164–168; Morchin (1993), p. 119.

RCS measurement methods are based on the measurement of RCS of the target or its accurate model. The measurement facilities are typically classified into two major categories: the radar range and field radar. Radar ranges have the advantage of simplicity and low cost of measurement. In this case, the radar wavelength is scaled by the same factor as the dimensions of the model and the power received from the target with a known illuminating power density is measured by special equipment, which is much simpler then a standard radar because one does not need to measure the coordinates, and the measurement time has no stringent limitations. The ranges can be indoor or outdoor, the latter usually having to meet the requirements for full-scale RCS measurements on large targets. Accurate RCS measurements require that the background signal be maintained at a low level to reduce measurement errors, about 20 dB less then measured RCS value. Field radars provide measurements of the full-scale target under real condition (e.g., for aircraft the RCS can be measured during normal flight). In this case, the target range and actual gain of whole radar channel from the transmitter to the receiver output must be known and RCS is calculated from the radar range equation. These measurements are termed dynamic and have limitations in that the requirements for detailed knowledge of the target trajectory and attitude relative to radar will affect the resultant accuracy of RCS measurement. *AIL*

Ref.: Skolnik (1970), pp. 27.5–27.18; Currie (1989); Mayzel's (1972), pp. 144–177

RCS of missiles and satellites. Missiles and satellites generally fall into the class lying between simple geometric shapes and complex targets (e.g., large aircraft). The resulting distribution of RCS typically is neither that of a point scatterer nor a random array of scatterers. The RCS pattern of a generic missile, displaying relative contribution of the nose, fuselage, wing, and tail, is given in Fig. R72. *SAL*

Ref.: Nathanson (1990), p. 159; Morchin (1993), p. 113.

Monostatic RCS is the RCS when the incident and scattering directions coincide, but are opposite in direction. The term *backscattering RCS* is used interchangeably. *SAL*

Ref.: Barton (1991), p. 5.2; Knott (1993), p. 16.

Table R14
Generalized Maritime Target RCS Values

Target Type	Size (ft)	RCS (m^2)
Small open boat		0.02
Periscope		1
Small pleasure boat		2
Fishing vessel	20	6
Submarine snorkel		10
Cabin cruiser	40	10
Wooden mine sweeper	144×25	80
Patrol boat	84	100
Small cargo ship		140
Trawler	525	150
Surfaced submarine	306×27	10–200
Freighter	1,200	500
Freighter	2,000	1000
Tanker		2,230
Navy picket ship	441	200–3,160
Loaded tanker		630–2,130
Loaded freighter		7,400
Navy cruiser	664	14,000
Cargo ship	278×40	$100–10^6$
Coast Guard, 3rd class		65
Coast Guard, 8×26		370
Coast Guard, 1st class		670
(from Morchin, 1993, Table 4.2, p. 120)		

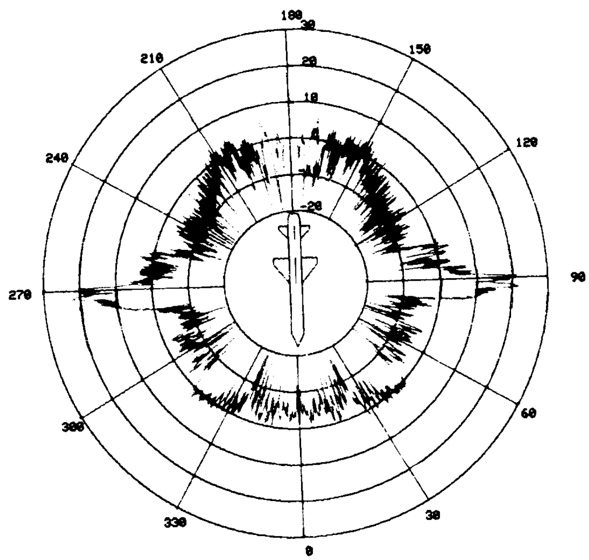


Figure R72 RCS pattern of a generic missile (from Morchin, 1993, Fig. 4.22, p. 114).

Multistatic RCS depends on the direction of the target to the transmitting and receiving positions of the multistatic radar. The angle between directions from the target to the transmitting and receiving positions is termed the *bistatic angle* β_b. The approximate formula to estimate multistatic RCS σ_m:

$$\sigma_m = \sigma_{mon}[1 + \exp(n_b|\beta_b| - 2.4n_b + 1)]$$

where $\beta_b = \operatorname{acos}\{(R_1^2 + R_2^2 - b^2)/2R_1R_2\}$, R_1 and R_2 are the range to transmitting and receiving positions, respectively, b is the baseline, $n_b = 7$ to 10 is the empirical coefficient defined by the target configurations and complexity, and σ_{mon} is conventional monostatic RCS. *AIL*

Ref.: Willis (1991), p. 145; Chernyak (1993), pp. 32–56; Aver'yanov (1972), pp. 17–27.

Normalized RCS is the RCS referred to any constant value. For example, in describing the properties of scattering regions, the RCS of a conducting sphere is typically normalized to the area of its surface. (See **SCATTERING regions.**) *SAL*

Ref.: Currie (1987), p. 197.

RCS of ogive (see **RCS of simple shapes**).

RCS of paraboloid (see **RCS of simple shapes**).

Passive RCS is the RCS of a target in passive radar when the detector of the energy radiated by the target operates in radiometric mode. (See **RADIOMETER.**) *SAL*

Ref.: Currie (1987), p. 848.

The **RCS pattern** is a graphical representation of RCS as a function of aspect angle (Figs. R69, R70, and R72). *SAL*

Ref.: Skolnik (1990), p. 11.16.

RCS of plate (see **RCS of simple shapes**).

RCS of precipitation (see **CLUTTER, rain, snow**).

RCS prediction methods can be based on the accurate or approximate theories. In the first case, the RCS can be computed through the formal solution of Maxwell's equations for the boundary conditions appropriate to the body of specified shape that is typically done by separation of variables and integral-equation formulation. This can be done only in special cases for some simple bodies such as a sphere, a spheroid, a torus, a cone, and some others. The approximate theories typically are classified into three major classes: the geometric optics approach, which is the simplest and treats ray bundles by the laws or reflection and refraction; the physical optics approach, for which the local current density at each point on the illuminated portion of the body is assumed to be identical to the current density that would flow at that point on an infinite tangent plane; and the geometrical diffraction theory that is the extension of geometric optics that accounts for diffraction. Geometric optics is the simplest approach, but it cannot distinguish the effects of polarization and the wave nature of the problem. The physical approach is more sophisticated; it makes it possible to some extent to distinguish the effects of wave interference, but it is not valid for applications requiring the specifications of polarization and depolarization effects. The geometrical diffraction theory combines the simplicity of the ray approach with the necessary considerations of wavelengths and phases, and it enables the evaluation of the RCS with the consideration of RCS diagram fluctuations, the preservation of the polarization dependence in evaluated result, and the prediction of the dependence of radar scattering upon bistatic angle. *AIL*

Ref.: Skolnik (1970), p. 27.19-27.38.

The **ramp function of RCS** is a time characteristic of scattering of a radar target, representing a response to a linearly increasing signal. The values of the ramp function $F_r(t)$ are equal to the area of the cross section of the illuminated target intersected by the plane of the incident plane wave, which moves along the line of sight at speed of $c/2$ (c is the speed of light), multiplied by a constant coefficient:

$$F_r(t) = \left(\frac{-1}{\pi c^2}\right)A(x)\Big|_{x = ct/2}$$

where $A(x)$ is the profile function, which shows the change in the area of a target cross section along the coordinate x. The ramp function is equal to the integral of the transient characteristic of the target $a(t)$ (see also **STEP FUNCTION**):

$$F_r(t) = \int_0^\infty a(t')dt'$$

The ramp function is an effective means of describing the shape of a radar target through synthesis of a three-dimensional model of the target in measured ramp functions. (See **IMAGING, three-dimensional.**) *IAM*

Ref.: Kennaugh, E. M., and Moffatt, D. L., "Transient and impulse response approximations," *Proc. IEEE* ***53***, no. 8, Aug. 1965, pp. 893–901.

RCS reduction. In many military and civilian applications it is desirable to reduce the RCS of a target. In general, there are four basic ways to reduce the RCS: target shaping, distributed loading (using of absorbers), discrete loading (passive cancellation), and active loading (active cancellation). Target shaping refers to the special selection of target surface shapes to minimize the amount of energy scattered back to the radar, typically at the cost of redirection of the scattered energy from the direction of interest to another region of little or no interest. In distributed loading the target is coated with lossy material soaking up incident energy and therefore reducing the energy scattered back to the radar. (See **ABSORBER.**) Active and passive cancellation methods require the loading of the target with discrete antenna-like elements (e.g., slots or dipoles) to reduce the overall cross section. These techniques are rather difficult to implement in practice, are severely limited in bandwidth, require the control of possible self-oscillations, and in some cases may even revert to reinforcement when some parameters differ from the nominal values for which the loads were designed. So target shaping and radar absorbing materials are the most attractive measures of RCS reduction. The most advanced state-of-the-art attainments of RCS reduction are concentrated in stealth technology. (See **STEALTH**. *SAL*

Ref.: Skolnik (1990), p. 11.43; Bhattacharyya (1991), pp. 141–239; Knott (1993), pp. 269–296; Morchin (1993), p. 122.

RCS of reflectors (see **RCS of simple shapes**).

RCS of sea surface (see **CLUTTER, sea**).

RCS of simple shapes. Simple shapes are those that permit the use of the analytic (exact) methods of RCS evaluation (prediction). The shapes typically considered in the tasks of RCS evaluation are the following: cones, cylinders, disks, edges, ellipsoids, ogives, paraboloids, plates, corner reflectors, spheres, and wires. The basic approximation formulas for these objects are given in the Table R15.

Table R15
RCS of Simple Shapes

Scattering feature	Incidence	Approximate RCS
Cone (infinite)	Nose-on	$\sigma = \dfrac{\lambda^2}{16\pi}$
Cylinder (circular)	Normal to the axis	$\sigma = \dfrac{2\pi a L^2}{\lambda^2}$
Disk (circular)	Normal	$\sigma = \dfrac{4\pi A^2}{\lambda^2}$
Edge (curved)	Perpendicular to the edge	$\sigma = a\lambda/2$
(straight)		$\sigma = L^2/\pi$
Ellipsoid of revolution ($a, b \gg \lambda$)	Along the major axis	$\sigma = \dfrac{\pi b^4}{a^2}$
Ogive	Nose-on	$\sigma = \dfrac{\lambda^2}{16\pi}(\tan\delta)^4$
Paraboloid (circular)	Nose-on	$\sigma = \pi a^2$
Plate (flat)	Normal	$\sigma = \dfrac{4\pi A^2}{\lambda^2}$
Reflector (corner):		
(dihedral)	$0 < \theta < 45°$	$\sigma = \dfrac{16\pi w^4}{\lambda^2}(\sin\theta)^4$
	$45° < \theta < 90°$	$= \dfrac{16\pi w^4}{\lambda^2} \times \{\sin[(\pi/2) - \theta]\}^4$
	$\theta = 45°$	$\sigma_{max} = \dfrac{8\pi w^4}{\lambda^2}$
(trihedral)	From the direction of maximum triple-bounce	$\sigma = \dfrac{4\pi w^4}{\lambda^2}$
Sphere	Any	
(Rayleigh region, $a < \lambda/\pi$)		$\sigma = 9\pi a^2\left(\dfrac{2\pi a}{\lambda}\right)^4$
(resonant region, $\lambda/\pi < a < 5\lambda/\pi$)		$\sigma = \sqrt{\sigma_0^2 + \sigma_c^2}$; $\sigma_0 = \pi a^2$; $\sigma_c = \pi a^2\left(\dfrac{2\pi a}{\lambda}\right)$
(optics region, $a > 5\lambda/\pi$)		$\sigma = \pi a^2$
Wire	Normal	
(parallel polarization, $ka \ll 1$)		$\sigma = \dfrac{\pi L^2}{(\pi/2)^2 + c}$; $c = [\ln(\lambda/1.78\pi a)]^2$
(perpendicular polarization)		$\sigma = \dfrac{9}{4}\pi L^2 (ka)^4$

δ = half-angle formed by the ogive axis and the tangent line at the ogive axis; A = area of plate or disk, a = radius or semimajor axis, w = width of each plate, L = length, θ = angle relative to the bisector angle of the dihedral, $k = 2\pi/\lambda$.

Targets of simple shapes may be used as the elementary components from which targets of complex shapes may be decomposed to estimate their reflectivity characteristics (see **RCS of complex shapes**) and for calibration of radar systems. *SAL*

Ref.: Barton (1991), p. 5.4; Skolnik (1990), pp.11.4–11.13; Morchin (1993), pp. 93–110; Currie (1989), pp. 35–50; Bhattacharyya (1991), pp. 60–75.

RCS of sphere (see **RCS of simple shapes**).

Static RCS is the RCS of a stationary target without consideration of its motion relative to radar (a concept opposite to dynamic RCS). *SAL*

Ref.: Bogush (1989), p. 177; Long (1992), p. 139; Bhattacharyya (1991), p. 25.

Surface RCS is that of the surface-distributed target, e.g., land and sea clutter. It is characterized by radar reflectivity σ^0, or RCS per unit area A illuminated by radar (see **CLUTTER**):

$$\sigma^0 = \sigma / A$$

where σ is the RCS of the illuminated area. *SAL*

Ref.: Currie (1992), p. 9.

Total RCS is the measure of the total power scattered by the target in all spatial directions (4π steradians) and is equal to the sum of absorption RCS and extinction RCS. The usual notation is σ_T. *SAL*

Ref.: Knott (1993), p. 70.

Volume RCS is that of a volume-distributed target (e.g., precipitation or chaff). It is characterized by reflectivity η_v or RCS per unit volume V illuminated by radar:

$$\eta_v = \sigma / V$$

where σ is the RCS of the illuminated volume (See **CLUTTER.**) *SAL*

Ref.: Currie (1992), p. 8.

RCS of wires (see **RCS of simple shapes**).

RADAR DATA (see **DATA**).

RADAR DECEPTION (see **DECEPTION; ELECTRONIC COUNTERMEASURES**).

RADAR DECOY (see **DECOY**).

RADAR DETECTION (see **DETECTION**).

RADAR DISPLAY (see **DISPLAY**).

RADAR ECHO (see **ECHO**).

RADAR ECHOING AREA (see **RADAR CROSS SECTION**).

RADAR ERROR (see **ERROR**).

RADAR (RANGE) EQUATION (see **RANGE EQUATION**).

RADAR ESTIMATOR (see **MEASUREMENT, radar**; **TRACKER**).

RADAR FREQUENCY BANDS (see **FREQUENCY**).

RADAR FUZE (see **FUZE, radar**).

RADARGRAMMETRY refers to the use of radar imaging for terrain mapping. The main method of obtaining images in radargrammetry is synthetic-aperture radar. *IAM*

Ref.: Levine (1960), pp. 55, 96; Anovetskiy (1979), p. 5.

RADAR GUIDANCE (see **GUIDANCE).**

RADAR HEIGHT FINDER (see **HEIGHT FINDER**).

RADAR HOLOGRAM (see **HOLOGRAM, HOLOGRAPHY**).

RADAR HOMING (see **HOMING**).

RADAR HORIZON (see **HORIZON**).

RADAR IDENTIFICATION (see **TARGET RECOGNITION AND IDENTIFICATION**).

RADAR IMAGE (see **IMAGE, IMAGING, radar**).

RADAR INTELLIGENCE (see **INTELLIGENCE**).

RADAR LOCK-ON (see **TARGET lock-on**).

RADAR MAP (see **MAP, MAPPING, radar**).

RADAR MEASUREMENT (see **MEASUREMENT, radar**).

RADAR METEOROLOGY (see **METEOROLOGY, radar**).

RADAR MODELING (see **MODEL, radar**).

RADAR NAVIGATION (see **NAVIGATION**).

RADAR NOMENCLATUR E (see **NOMENCLATURE**).

RADAR OPERATOR (see **OPERATOR, radar**).

RADAR PERFORMANCE see **PERFORMANCE**).

RADAR RANGE EQUATION (see **RANGE EQUATION**).

RADAR RANGE FINDER (see **RANGE FINDER**).

RADAR RECEIVER (see **RECEIVER, RECEPTION**).

RADAR RECOGNITION (see **TARGET RECOGNITION AND IDENTIFICATION**).

RADAR RELIABILITY (see **RELIABILITY**).

RADAR RESPONDER (see **TRANSPONDER**).

RADAR SEEKER (see **SEEKER).**

RADAR SERVICE (see **SERVICE**).

RADAR SHADOW (see **SHADOW**).

RADAR SIGHT (see **SIGHT**).

RADAR SIGNAL (see **SIGNAL**)

RADAR SIGNATURE (see **TARGET RECOGNITION AND IDENTIFICATION**).

RADAR SILENCE (see **SILENCE**).

RADAR SPEEDOMETER (see **RADAR, police**).

RADAR TARGET (see **TARGET**).

RADAR TEST (see **TEST**).

RADAR THROUGHPUT CAPABILITY (see **THROUGHPUT CAPABILITY**).

RADAR TRACKING, TRACKER (see **TRACKING**).

RADAR TRAINER (see **TRAINER**).

RADAR TRANSMITTER (see **TRANSMITTER**).

RADAR WAVEFORM (see **WAVEFORM**).

RADIATION

Atmospheric radiation is thermal radiation caused by the presence in the dense layers of the atmosphere of water vapors and oxygen of relatively high concentrations. It is a component of sky radiation Sky radiation includes not only atmospheric radiation, but also cosmic RF radiation (cosmic noise, radiations of the Sun, Moon, planets, stars) and is described by sky temperature. *AIL*

Ref.: Nikolayev (1970), pp. 25–27; Skolnik (1990), pp. 2.26–2.31.

Blackbody radiation is electromagnetic radiation from all objects which are not at a temperature of absolute zero. A body upon which electromagnetic radiation is incident may transmit, reflect, or absorb portions of this radiation. If the body absorbs all radiation incident upon it, and radiates all the radiation it absorbs, thus maintaining a constant ambient temperature, it is known in physics as a *perfect blackbody*. Blackbody radiation follows the Stefan-Boltzmann Law:

$$E = \sigma T^4$$

where E, the total emissive power, or radiant emittance, is expressed in terms of power per unit area, T is the absolute temperature in kelvins, and σ is the Stefan-Boltzmann constant, equal to 5.668×10^{-8} W/m^2/K^4. Blackbody radiation is an important concept in the determination of radar system noise temperature. *PCH*

Ref.: Merritt, T. P., and Hall, F. F., Jr., "Blackbody Radiation," Paper 3.1.2, *Proc. IRE* ***47***, no. 9, Sept. 1959.

Radiation efficiency is the ratio of P_r, the power radiated by an antenna, to P_o, the power accepted by the antenna from a connected transmitter:

$$\eta = \frac{P_r}{P_o}$$

and is the reciprocal of antenna loss. (See **LOSS.**) *SAL*

Ref.: IEEE (1993), p. 1,057; Johnson (1993), p. 1.5.

Radiation intensity is "the power radiated from an antenna per unit solid angle in that direction."

Ref.: IEEE (1993), p. 1057; Johnson (1993), p. 1.5.

Isotropic radiation is electromagnetic radiation that has the same power density (W/m^2) in any direction at the same distance R from a point source. *PCH*

Radiothermal radiation is an inherent radiation of objects caused by thermal motion of electrons. Radiothermal radiation is a field of thermal noise fluxes penetrating into the thickness of a radiating body. The intensity of the electromagnetic radiation, spectral composition, and degree of polarization depend upon the temperature and the physical properties of the radiating body. Characteristic special features of radiothermal radiation are wide frequency band, fluctuating power, and spectral density. The power of radiothermal radiation is directly proportional to the temperature of the radiator and inversely proportional to the square of the wavelength. Therefore, it is more advantageous for thermomicrowave imaging radar to use millimeter-band waves than it is to use centimeter- and decimeter-band waves. Sensors operating on the basis of thermal radiation are called *radiometers*.

Radio-brightness temperature, $T_a = T_p\varepsilon$, is the quantitative characteristic of radiothermal radiation, where ε is the emissivity of the body, characterizing a body as a source of radiothermal radiation, and T_p is its absolute temperature (usually around 300K). Table R16 shows the values of the emissivity for various types of surfaces in the centimeter waveband. *AIL*

Table R16
Microwave Emissivities

Type of Object	ε, for angle of observation, θ	
	$\theta = 30°$	$\theta = 45°$
Water	0.41	0.34
concrete	0.88	0.80
Asphalt	0.89	0.82
Short cut grass	0.94	0.94

Ref.: Nikolayev (1970), pp. 6–15; Skolnik (1970), p. 39.8.

Radiation resistance is "the ratio of the power radiated by an antenna to the square of the rms antenna current referred to a specified point." *SAL*

Ref.: IEEE (1993), p. 1,058; Johnson (1993), p. 2.12.

sky radiation (see **atmospheric radiation**).

Spurious radiation is auxiliary radiation, an onset of which is not linked to formation of carrier frequency oscillations. Spurious radiation is determined by the power of harmonics (relative to power on the main frequency) and the power of spurious noises in the operating band. Special filters combat radiation of harmonics. Spurious noise in the operating frequency band results from transmitter operation, mostly affects the radar receiver, and is usually impossible to filter out. *AIL*

Ref.: Skolnik (1970), p. 7.42.

RADIATOR, RADIATING ELEMENT, of an antenna. *Radiator* is a general term for a basic antenna that, in transmitting, ensures conversion of the energy of RF electric current into the energy of electromagnetic waves. In a receiving antenna, the opposite process occurs. The most common elements are waveguide, slotted, horn, and dipole radiators. Waveguide, dipole, and horn radiators are used in reflector antennas and lens antennas. Along with these, slot radiators and printed radiators are used in antenna arrays. Helical and log periodic antennas are used as broadband radiators.

AIL

Ref.: Skolnik (1970), p. 10.7; Sazonov (1988), p. 384; Voskresenskiy (1981), pp. 177–190.

A **dipole radiator** is an driven dipole or a reflector in the form of a disk or passive dipole. When used in the capacity of passive dipole reflector, it is placed at distance $\lambda/8$ from the active dipole and is somewhat longer than the latter. In the second case, the reflector comprises a round flat disk at distance $\lambda/4$ from the active dipole. A coaxial line or square waveguide may be used to feed an active dipole. Individual dipoles rarely are used as radiators since they have unsatisfactory radiation patterns and polarization characteristics. Dipole radiators with disk or passive dipole reflectors are used as feeds in reflector antennas and as antenna array elements. *AIL*

Ref.: Sazonov (1988), p. 380; Skolnik (1970), p. 10.7; Fradin (1977), pp. 17–20.

A **horn radiating element** is a waveguide connected to a horn. Horn radiators are simple in design, are broadband, and make it possible within broad limits to select the requisite radiation pattern. (See **ANTENNA, horn.**) A shortcoming of horn radiators is the necessity to impart great length to them to obtain a highly directional radiation pattern. In practice, they are small and comparable to aperture width. Pyramidal horns are used as feeds in lens antennas and reflector antennas. Sectoral horns are used as elements in phased-array antennas. *AIL*

Ref.: Sazonov (1988), p. 382; Mailloux (1994), p. 275.

An **isotropic radiating element** is a hypothetical radiator radiating uniformly in all directions. *AIL*

Ref.: Popov (1980), p. 143.

Linear radiating elements are identical elements distributed continuously or discretely along a given axis. Multislotted waveguide arrays, sectoral horns, microstrip arrays, and the like may be used a linear radiators. They often are manufactured in the form of a pillbox antenna (of the "cheese" type), in turn excited by the open end of a square waveguide. Linear radiators may be used as feeds in reflector antennas in the form of parabolic cylinders, as well as antenna array elements. *AIL*

Ref.: Sazonov (1988), p. 383.

A **printed radiating element** is a radiator made of printed elements. These radiators are positioned at a low height above a flat conducting screen (approximately $\lambda/20$). Their special feature is the capability to use printed technology when manufacturing multielement SHF subarrays with radiators located on both sides of a support. In its simplest variant (Fig. R73), a printed radiator is a disk 1 on metal screen 2 on slightly thick dielectric support 3. The disk is excited with the aid of two pins to which energy is supplied by means of a plane line located on the opposite side of the screen. The pin is excited

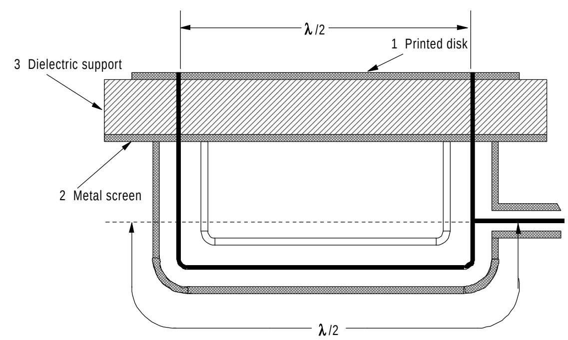


Figure R73 Printed radiator (after Voskresenskiy, 1981, Fig. 2.18a, p. 37).

in antiphase, which ensures maximum radiation in the direction of the normal to the plane of the screen. A printed radiator is used as elements of printed antenna arrays. *AIL*

Ref.: Voskresenskiy (1981), p. 37.

A **slotted radiator** is a radiator comprising a cylindrical resonator and square waveguide (Fig. R74). The constrained shape and mutual positioning of slots cut in the end of a cylindrical resonator close to the reflector, as well as the diameter of the round screen, are selected to obtain identical radiation pattern width in planes E and H. A shortcoming of such a radiator is the narrow operating frequency band (about 3%). Slotted radiators may be used as feeds in reflector antennas and as array elements. *AIL*

Ref.: Sazonov (1988), p. 381.

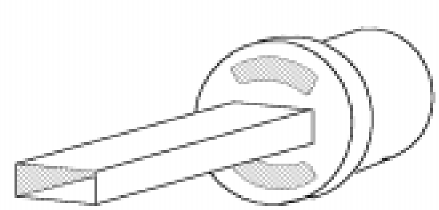


Figure R74 Double-slotted Cutler radiating element (after Sazonov, 1988, Fig. 14.20, p. 382).

A **waveguide radiating element** is a waveguide with an open end. A square waveguide with wave type TE_{01} and a round waveguide with wave type TE_{11} are used. The round waveguide has advantages over the square one because it makes it possible to create more uniform radiation of the mirror and, therefore, ensures formation of identical radiation patterns in planes E and H. At half power, a square waveguide has radiation pattern width $\theta_3 = 80° - 120°$, while a round waveguide has a width of 75° – 80°. Waveguide radiators are used as feeds in reflector antennas or lens antennas, as well as in the capacity of antenna array elements. A shortcoming of waveguide radiators is the relatively low level of waveguide matching with free space. *AIL*

Ref.: Skolnik (1970), p. 10.7.

RADIOMETER, RADIOMETRY, microwave. A radiometer is the passive radar that detects radiothermal signals in a noise background, and radiometry is the measurement of radiothermal signals using passive radar. The simplest radiometer consists of high-frequency amplifier, detector, and smoothing low-pass filter, (i.e. a receiver with the direct amplification). To increase the power of the radiometric signal, the bandwidth, Δf, of the RF amplifier should be as large as possible, and to suppress the noise component after the detector, the bandwidth ΔF of the subsequent low-pass filter should be as small as possible. In modern radiometers, $\Delta f / \Delta F \approx 10^7$ to 10^9, giving a high output signal-to-noise ratio when the input signal-to-noise ratio is close to unity. The simplest radiometer cannot determine angular coordinates of the target and detect whether it moves or not, so it is not used in practice.

The main component of a radiometer is the radiometer receiver. The main types of radiometers are noise-adding radiometers (Fig. R75), in which additional noise is introduced to increase the sensitivity, correlation radiometers with two independent receiving channels that eliminate the influence of radiometer internal noise (Fig. R76), total-power radiometers, measuring the power after compensation of internal noise (Fig. R77), null balancing (Fig. R78), Dicke (Fig. R79), and Graham's (parallel-channel) radiometers (Fig. R80).

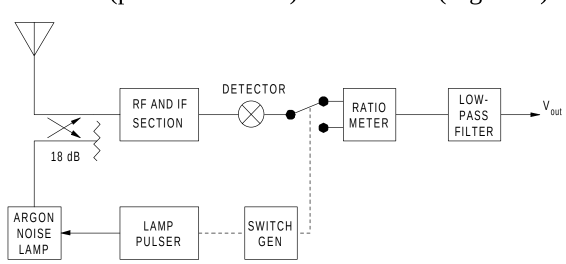


Figure R75 Noise-adding radiometer (from Skolnik, 1970, Fig 19, p. 39.19, reprinted by permission of McGraw-Hill).

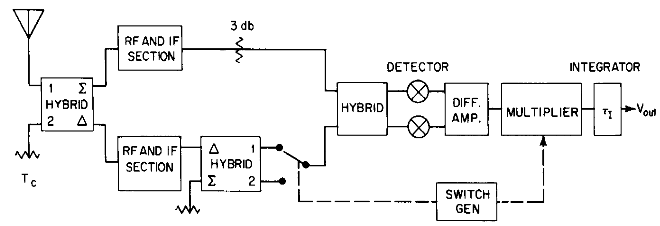


Figure R76 Correlation receiver with phase switching (from Skolnik, 1970, Fig. 18, p. 39.18, reprinted by permission of

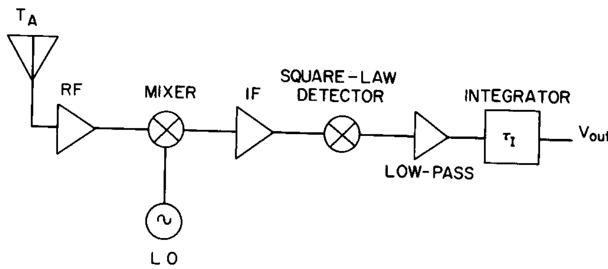


Figure R77 Total power radiometer (from Skolnik, 1970, Fig. 14, p. 39.15, reprinted by permission of McGraw-Hill).

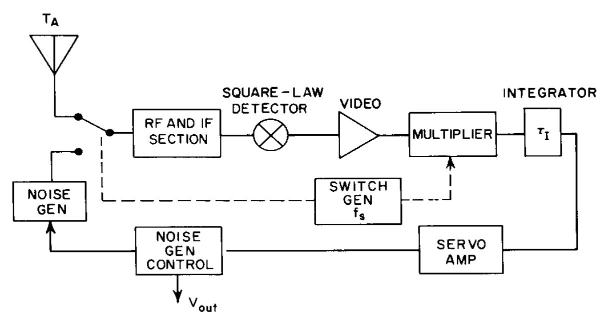


Figure R78 Null-balancing radiometer (from Skolnik, 1970, Fig. 16, p. 39.17, reprinted by permission of McGraw-Hill).

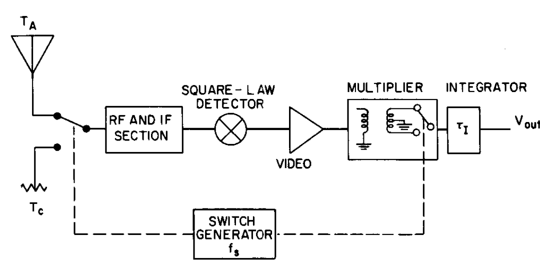


Figure R79 Dicke receiver (from Skolnik, 1970, Fig. 15, p. 39.16, reprinted by permission of McGraw-Hill).

One of the main characteristics defining the capability of a radiometer to receive weak radiothermal signals is radiometer sensitivity δT. Numerically, it is equal to the temperature of the signal for which the output signal-to-noise ratio is equal to unity, and is defined as

$$\delta T = \frac{T_N}{\sqrt{\Delta f \cdot \tau}}$$

where T_N in the noise temperature of a high-frequency amplifier, Δf is the RF amplifier passband, and τ is the signal duration. The radiometers primarily are used for mapping and target identification. *AIL*

Ref.: Evans (1977); Skolnik (1970), p. 39.1-39.35; (1989).

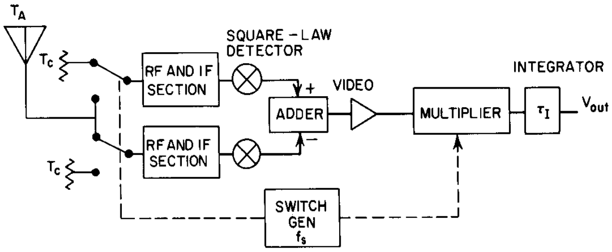


Figure R80 Parallel-channel (Graham's) receiver (from Skolnik, 1970, Fig. 17, p. 39.17, reprinted by permission of McGraw-Hill).

RADOME. A radome is "a cover usually intended to protect an antenna from the effects of its physical environment without degrading its electrical performance." Radomes are divided into several categories:

(1) From the viewpoint of location, into surface (ground or shipborne) or airborne.

(2) From the viewpoint of the structural approach, into rigid and inflatable.

(3) From the viewpoint of wall cross section, into single-layer and multiple-layer (the latter sometimes called sandwich-type).

The main purpose of the radome is to protect the antenna from environmental exposure: rain, snow, ice, and high winds. Ideally, the radome should not degrade the electrical performance of the antenna, but in practice it introduces

(1) Transmission loss, due to reflection and absorption (see **LOSS, radome**).

(2) Antenna pattern distortion, both by changing the mainlobe width and increasing sidelobe levels.

(3) Beam deflection, resulting in boresight error. *SAL*

Ref.: IEEE (1993), p. 1,063; Walton (1970); Skolnik (1970), Ch. 14, (1980), pp. 264–270.

A-, B- C-sandwich radome (see **multiple-layer radome**).

An **airborne radome** is used on an aircraft, missile, or space vehicle and can take different forms, ranging from a small, flat-panel configuration to highly streamlined shapes; from flush-mounted, normal-incidence structures to streamlined, high-incidence-angle nose radomes. The requirements for airborne radomes are mainly set by aerodynamic loads, of which the determining factors are vehicle speed, altitude, radome shape, and location on the vehicle. The main factors in electrical performance are the rain-erosion-resistant coating, antistatic coating, and use of lightning-protection strips, along with formation of water films during operation in precipitation. Typical airborne radar performance requirements are given in Table R17. *SAL*

Ref.: Skolnik (1970), p. 14.21.

A **dielectric-metal radome** uses dielectric layers with metal inclusions to reduce the thickness or to provide broadband performance. This structure is typically used in reinforced radomes, and the metal inclusions take the form of a wire grid. *SAL*

Skolnik (1970), p. 14.4.

Table R17
Typical Airborne-radome Performance Requirements

Parameter	Unit	General purpose	Fire control or guidance
Transmission:			
Average	%	90	85–90
Minimum	%	85	75–80
Reflection	%	2	2
Beam deflection	mrad	-	2–4
Beam deflection rate	mrad/deg	-	0.5–1.0
Beamwidth change	%	10	10
Sidelobe increase, at −20-dB level	dB	3	3
(from Skolnik, 1970, Table 2, p. 14.21)			

A **foam(-shell) radome** is a multilayer dome in which a foamy dielectric is used as the filler between layers. The thickness of the foamy filler is selected to ensure compensation for reflections in a broad band and angles of incidence. Individual panels forming the dome may either be glued or welded at their joints. These domes have good electrical characteristics in a broad frequency band. *AIL*

Ref.: Skolnik (1970), p. 14.23; Lavrov (1974), p. 281.

ground-based radome (see **surface-based radome**).

An **inflatable [air-supported] radome** is made of nylon fabric with a special covering transparent for the radiated and received energy of radio waves and having a spherical shape due to inside air pressure. Special equipment is used to maintain constant internal air pressure. Failure of this inflation system can cause the dome and antenna to be put out of service. Separate inflation partially solves this problem in domes with a dual wall. Such a dome falls in the category of domes with a dual-layer structure having thin skins (A-sandwich). *AIL*

Ref.: Skolnik (1970), p. 14.23.

A **multi(ple-)layer radome** is a multilayer structure comprising thin envelopes and an internal filler. They may comprise three, five, seven, nine, or more thin layers of transparent material and filler of a material with low density ensuring high gain over a broad frequency band. Paper fiber or honeycomb fiberglass usually is used as filler. If three layers are used (two dense skins with a thicker low-density filler) it is called an *A-sandwich radome*. The inverse of this is a *B-sandwich*, where the skins are of lower density than the core. A five-layer structure with two outer and a center skin, separated by two cores, is a *C-sandwich*. Multilayer domes are widely used in modern radars. *AIL*

Ref.: Skolnik (1970), p. 14.13.

Radome parameter measurement methods are designed for measurement of the transmission coefficient and beam deflection errors. Direct measurement methods and compensation may be used to measure the transmission coefficient of domes. In the direct measurement method, the protected antenna's mainlobe is oriented in the direction of a probe and the amplitude of signal V_d or power P_d is measured.

The dome is then put in place and amplitude $V(\theta,\phi)$ or power $P(\theta,\phi)$ are measured for different directions relative to the dome axis in the antenna beam scan sector, The transmission coefficient is found using this ratio:

$$K(\theta, \phi) = \left(\frac{V(\theta, \phi)}{V_d}\right)^2 = \frac{P(\theta, \phi)}{P_d}$$

In the compensation method, amplitude (power) measurement is replaced by maintenance of a fixed level V_d or P_d of the signal with the aid of an attenuator. The transmission coefficient is determined from the difference in attenuations L (on a logarithmic scale) or from the relationship of absolute values X (on a linear scale):

$$K(\theta, \phi) = \frac{X(\theta, \phi)}{X_d} = L(\theta, \phi) - L_d$$

where X_d and L_d are attenuator settings without the dome, and $X(\theta,\phi)$ and $L(\theta,\phi)$ are attenuator settings in direction θ,ϕ with the dome.

Beam deflection errors are determined using automated measurement complexes. Here, a probe initially is installed in the direction of the beam axis in the absence of the dome. After the dome is installed, the direction of the beam changes its position and deviates from the initial direction by an angle $(\delta\theta,\delta\phi)$. Displacing the probe, it is possible to determine the beam deflection error in each direction. *AIL*

Ref.: Hirsch (1987); Strakhov (1985), p. 20.

A **reinforced radome** is made from a dielectric sheet and a reinforcing metal grid of parallel wires. Grid reflectivity depends upon the ratios of the diameter of the wires and the distance between them to wavelength. It is possible to achieve mutual compensation for reflections from the mesh and the dielectric sheet through proper selection of these ratios. *AIL*

Ref.: Lavrov (1974), p. 281.

A **rigid radome** has structural members to carry the primary loads imposed on it. It is usually a space-frame radome having a network of beams in the shape of a truncated sphere. When these beams are metal, rather than plastic, the dome is called a metal-space-frame radome. This type of dome is usually superior in electrical performance; cheaper and easier to fabricate, transport, and assemble; and available in larger diameters (up to 150m has been considered). An example of a rigid radome for ground-based radar is shown in Fig. R81. *SAL*

Ref.: Skolnik (1970), p. 14.4, (1980), p. 266.

A **rotodome** is the structure incorporating a radome and antenna rotating together as an entire unit. It is used in ground-based and airborne radars. An example is the 7.3m diameter rotodome of the U.S. Navy's E-2C aircraft (Fig. R82). *SAL*

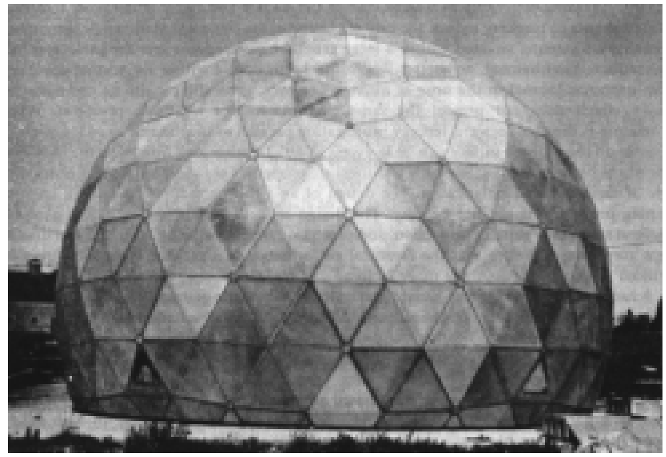


Figure R81 Rigid radome for ground-based antenna (from Skolnik, 1980, Fig. 7.29, p. 266, reprinted by permission of McGraw-Hill).

Figure R82 E-2C AEW aircraft with rotodome antenna.

Ref.: Skolnik (1980), p. 268.

A **single-layer radome** is manufactured from such materials as fiberglass, ceramic, elastomer, or monolithic foam materials. The thickness s of a single-layer dome depends upon wavelength. Single-layer domes have the following shortcomings: thin-walled domes ($s \ll \lambda$) may not be rigid enough, while those with a half-wave wall are exceptionally heavy. *AIL*

Ref.: Skolnik (1970), p. 14.5; Lavrov (1974), p. 279.

space-frame radome (see **rigid radome**).

A **surface- (ground- or ship-)based radome** is usually an inflated or rigid radome of multilayer construction (sandwich or space-frame). The most important factors are radome size, shape, and mounting structure; layout of nearby structures; and surrounding terrain. An example of a metal-space-frame radome is shown in Fig. R83, and the performance of rigid-surface radomes is given in Table R18. *SAL*

Ref.: Skolnik (1970), p. 14.22.

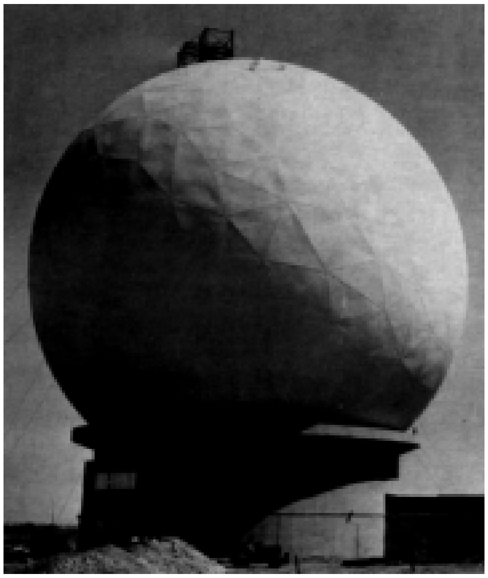


Figure R83 A 110-ft-diameter metal-space-frame radome (from Skolnik, 1970, Fig. 31, p. 14.26, reprinted by permission of McGraw-Hill).

Table R18
Performance of Rigid Radomes

Radome type	Diameter (m)	Radar band	Trans-mission (%)	Boresight error (mrad)	Sidelobe increase	
					dB	dB level
Tw[a]	8.1	S	88	0.3	2	-30
F[b]	5.2	K		0.0 - 0.6	2.5	-18
F	7.9	C	93-98	0.1		
F	7.9	X	94-97	0.2-0.4	0.7	-24
DSF[c]	16.8	S	79-89	0.23	2.0	-24
Sd	16.8	S	87-94	0.25-2.0	1.5	-
S	18.3	C	96	0.6	2.7	-3-
S	42.7	UHF	98	0.1-0.3	0.9	-20
MSF[e]	28.3	L	83	0.44	0.8	-21
MSF	33.5	UHF	87	0.1-0.3	0.9	-20
MSF	45.7	UHF	85	0.2	6.3	-30

[a] thin wall, [b] foam, [c] dielectric space frame, [d] sandwich, [e] metal space frame
(from Skolnik, 1970, Table 4, p. 14.27, reprinted by permission of McGraw-Hill)

A **tangent ogive radome** is shaped as a body of revolution about the axis of the vehicle, formed by a circular arc tangent to the cylindrical body diameter of the vehicle (Fig. R84). The equation for the ogive is

$$r = \sqrt{R^2 - (x - L)^2} - b$$

where the symbols are defined in the figure. The tangent ogive will have $R = b + D/2$. The ratio of length L to diameter D is the *fineness ratio* of the radome. *DKB*

Ref.: Currie (1987), pp. 566–570.

A **thin-wall radome** is made from blanks with screwed-on flanges to hold them in place. Thin rigid dome dimensions are limited to a diameter < 9m due to design considerations. *AIL*

Ref.: Skolnik (1970), p. 14.23.

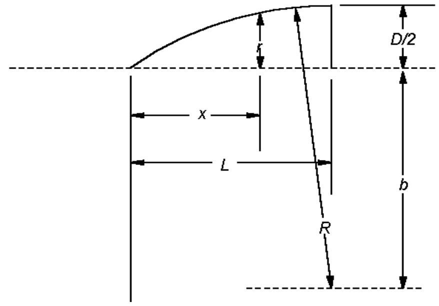


Figure R84 Ogive radome shape.

radome transmission coefficient (see **LOSS, radome**).

A **Von Karman radome** is one having a tapered shape, similar to the tangent ogive (Fig. R84) but having the equation

$$r = \frac{D}{2\sqrt{\pi}}\sqrt{\phi - \frac{1}{2}\sin 2\phi}$$

where $\phi = \arccos[1/(1-2x/L)]$. Its advantage is that it produces the maximum ratio of volume to drag for a given fineness ratio. Figure R85 shows the comparative profiles of five types of radome, each with a fineness ratio of 2.5. *DKB*

Ref.: Currie (1987), pp. 566–570.

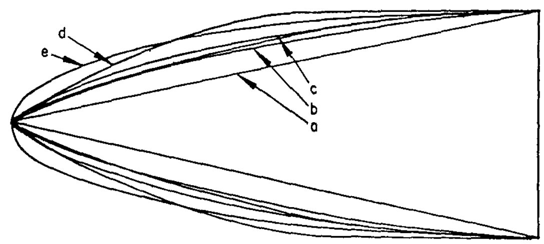


Figure R85 Five radome shapes, each having a fineness ratio of 2.5. (a) Cone, (b) Von Karman, (c) tangent ogive, (d) Haack, (e) 60° log spiral (from Currie, 1987, Fig. 12.1, p. 569).

RAILING is pulsed jamming using a pulse repetition frequency significantly higher than that of the victim radar. Its name derives from the appearance on an A-scope display, where it forms a closely spaced series of vertical lines resembling the bannisters of a railing. *DKB*

Ref.: Johnston (1979), p. 65.

RANGE

A **blind range** is "a range corresponding to the time delay of an integral multiple of the interpulse period plus a time less than or equal to the transmitted pulse length. A radar usually cannot detect targets at a blind range because of interference by subsequent transmitted pulses." The problem of blind ranges can be solved or largely mitigated by employing multiple PRFs. *PCH*

Ref.: IEEE (1990), p. 7.

burn-through range (see **JAMMING, barrage**).

(Maximum) detection range is the maximum range for which radar detection of a target (of specified radar cross section and Swerling fluctuation characteristics) occurs with a specified probability, and with a given probability of false alarm. For a scanning search radar, the maximum detection range is conventionally specified on a per-scan or on a cumulative-probability-of-detection basis (see also **DETECTION probability**; **RANGE EQUATION**). *PCH*

Ref.: Blake (1980), pp. 12–13.

Range-doppler coupling is a property of linear frequency modulated signals, for which target doppler shift produces an advance or delay in the matched filter output, indistinguishable from a decrease or increase in target range delay. For a frequency sweep rate $\dot{f}$ the delay effect of a doppler shift Δf is

$$\Delta t_d = \frac{-\Delta f}{\dot{f}}$$

which translates into an apparent range shift

$$\Delta R = \frac{\Delta t_d}{2} = \frac{-c\Delta f}{2\dot{f}} = \frac{v_t c}{\dot{f}\lambda} = \frac{v_t f_0}{\dot{f}}$$

where c is the velocity of light, v_t is the target velocity (measured outbound from the radar), λ is the radar wavelength, and f_0 is the frequency.

For example, using a pulse width τ = 50 μs, a linear FM bandwidth B = 5 MHz (sweeping upwards during the pulse), at f_0 = 3 GHz, the range shift is

$$\Delta R = 0.3 \text{ m per m/s}$$

A target moving outward at 1,000 m/s would show a range increase of 30m.

Rihaczek has shown that the effect of range-doppler coupling can be compensated exactly by assigning the measured range to a point later in time by $f/\dot{f}$ from that of the actual measurement (or 0.03 sec in the example). *DKB*

Ref.: Cook (1967), pp. 310–313; Rihaczek (1969), p. 247

range-gate pull-off (see **ECM, range-measurement**).

Ground range is the "distance along the ground between the points directly beneath the radar and the target." For short ranges (i.e., to 100 km or so), the ground range can be approximated as $G = R \cos E$, where R is the slant range, and E is the target elevation angle. For accurate determination of the ground range of long-range targets (Fig. R86), the effect of atmospheric refraction on the measurement of slant range and effect of earth's curvature on ground range must be taken into account. This is sometimes called *ground distance*. *PCH*

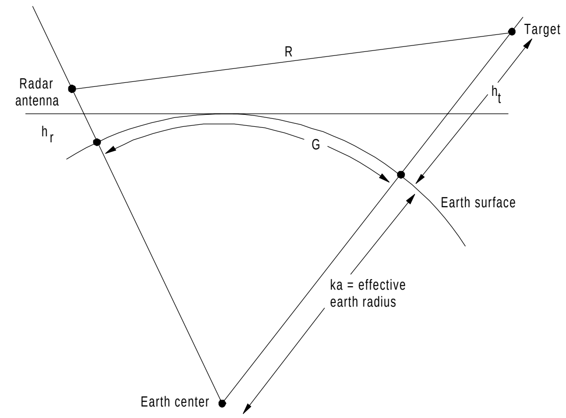


Figure R86 Ground range *G* in curved-earth reference frame.

Instrumented range is the range over which a radar display or other detection device is permitted to detect echoes. *PCH*

Ref.: Barton (1991), p. A-10.

Line-of-sight range is the range along the optical line of sight from a radar to the target, also referred to as the *slant range*. When "range" is used in radar discussions without a qualifier, it is understood to refer to slant range. The maximum physical line-of-sight range is determined by the radar horizon, the range from the radar to the point at which the radar wave, refracted by the atmosphere, is tangent to the earth's surface. (See also Fig. R83.) *PCH*

Ref.: IEEE (1990), p. 18.

A **range mark** is "a calibration marker used on a display to aid in measuring target range (distance from the radar)." Range marks may be generated at regular intervals, or a manually variable mark may be used to track the echo, permitting range readings to be obtained from a dial associated with the mark control. *DKB*

Ref.: IEEE (1993), p. 1066.

Maximum unambiguous range. If the round-trip time delay of the target signal exceeds the pulse repetition interval (PRI) of the radar, determination of the range to the target will be ambiguous (i.e., the target will appear at apparent ranges $R' = R - iR_u$, where i is the largest integer for positive R' and R_u is the unambiguous range of the waveform). Since the pulse repetition frequency f_r is the reciprocal of the PRI, the maximum unambiguous range of a radar $R_u = c/2f_r = c(\text{PRI})/2$. Range ambiguities are resolved through the use of multiple PRFs and the use of computational algorithms such as the Chinese Remainder Theorem. *PCH*

Ref.: Skolnik (1980), p. 2; Barton (1988), p. 234.

Range noise is the random error in range estimation or tracking caused by random interference in the radar receiver or by modulations on the target signal. The latter is also termed *range glint*. (See **ERROR, range.**) *DKB*

Ref.: Skolnik (1970), p. 28.3; Barton (1969), pp. 59–89.

Range offset processing is "synthetic-aperture processing in which the spectrum is translated from IF to a carrier offset from zero frequency by approximately half the IF bandwidth."

Ref.: IEEE (1990), p. 25.

Self-screening range (see **RANGE EQUATION with jamming**).

Range sidelobes (see **SIDELOBE**).

Slant range is the radial distance to the target (e.g., as measured along the radar antenna boresight (line of sight) of a tracking radar). This is sometimes called slant distance. (See also **line-of-sight range.**) *PCH*

unambiguous range (see **maximum unambiguous range**).

Range walk is "the migration of a point scatterer from range cell to range cell during the signal integration period. Range walk, for example, can occur in synthetic-aperture radar typically caused by range curvature and/or target rotational or radar line-of-sight translational motion. Range walk can also occur in a surveillance radar if the relative range rate between the target and the radar is high, relative to the ratio of the range cell and the integration period."

Ref.: IEEE (1990), p. 25.

Variable range rate refers to an image aberration in inverse synthetic aperture radar, wherein target motion through the range cells causes a range-dependent focus error in the cross-range coordinate. *DKB*

Ref.: Schleher (1991), p. 510.

RANGE EQUATION. The radar (range) equation permits calculation of the maximum range of radar operation in a specified mode (search, tracking, beacon, etc.). It is "a mathematical expression for primary radar that, in its basic form, relates radar parameters such as transmitter power, antenna gain, wavelength, effective echo area of the target, distance to the target, and receiver input power. The basic equation may be modified to take into account other factors, such as receiver noise, signal processing, attenuation caused by a radome, attenuation due to atmospheric losses or precipitation, and various other losses and propagation effects."

The equation is derived from the Friis transmission formula and the relationship between power density incident on the target and that returned to the radar, which give the ratio of received to transmitted power as

$$\frac{P_r}{P_t} = \frac{G_t G_r \lambda^2 \sigma F^4}{(4\pi)^3 R^4}$$

This leads to the basic equation for maximum detection range R_m in an environment of thermal noise can be written, for all types of radar, in terms of the energy transmitted during the coherent processing interval t_f:

$$R_m^4 = \frac{P_{av} t_f G_t A_r \sigma F^4}{(4\pi)^2 k T_s D_x(n') L_t L_\alpha} = \frac{P_{av} t_f G_t G_r \lambda^2 \sigma F^4}{(4\pi)^3 k T_s D_x(n') L_t L_\alpha} \quad (1)$$

where P_{av} is average transmitter power, G_t is transmit antenna gain, A_r is effective receiving aperture area, G_r is receive antenna gain, σ is target RCS, F is the pattern-propagation factor (which should, in general, include polarization effects), k is Boltzmann's constant, T_s is system input temperature, $D_x(n')$ is the effective detectability factor for integration of n' pulses or samples, L_t is transmission line loss, and L_α is two-way atmospheric loss (attenuation).

For a noncoherent pulsed radar, t_f is a single pulse repetition interval, $t_r = 1/f_r$, and the transmitted energy becomes $P_{av} t_r = P_t \tau$, while the number of integrated pulses $n' = n = f_r t_o$, where t_o is the observation time (time-on-target). For a fully coherent radar, integrating coherently over the entire observation time, the transmitted energy is $P_{av} t_o$ and $n' = 1$.

For definitions of the pattern-propagation factor, system noise temperature, detectability factor, and losses see **LOSS**; **DETECTABILITY FACTOR**; **PROPAGATION**.

Table R19 compares different radar range equations. (For notations used and more detailed information on each type of radar, see the corresponding entries in this section.) *DKB*

Ref.: IEEE (1993), p. 1,052; Blake (1980); Skolnik (1970), Ch. 2; Barton (1988), pp. 9–24.

Table R19
Radar Range Equations

Basic equation	$R_m^4 = \frac{P_{av} t_f G_t A_r \sigma F^4}{(4\pi)^2 k T_s D_x(n') L_t L_\alpha} = \frac{P_{av} t_f G_t G_r \lambda^2 \sigma F^4}{(4\pi)^3 k T_s D_x(n') L_t L_\alpha}$
for noncoherent pulsed radar	$R_m^4 = \frac{P_t \tau G_t G_r \lambda^2 \sigma F^4}{(4\pi)^3 k T_s D_x(n) L_t L_\alpha}$
in decibel notation	$40\log R_m = 10\log P_t + 10\log\tau + G_{t(dB)} + G_{r(dB)} + 20\log\lambda + 10\log\sigma + 40\log F - 10\log[(4\pi)^3 k] - 10\log T_s - D_{x(dB)} - L_{t(dB)} - L_{\alpha(dB)}$
in practical units	$R_m^4 = K\frac{P_{t(kW)} \tau_{(\mu s)} G_t G_r \sigma F^4}{f^2_{(MHz)} T_s D_x(n) L_t L_\alpha}$ where: K = 129.2 for R_m in nm; = 148.7 for R_m in st m; = 239.3 for R_m in km; = 261.7 for R_m in kyds; =185.0 for R_m in kft.
for altimeter (beamwidth limited case)	$R_m^2 = \frac{P_t t_f G_t A_r \sigma^0 \theta_3^2}{(4\pi)^2 k T_s D_x(n') L_t L_\alpha L_p^2}$
for altimeter (pulsewidth limited case)	$R_m^3 = \frac{P_t t_f G_t A_r \sigma^0 \tau_n (c/2)}{8\pi k T_s D_x(n') L_t L_\alpha L_p^2}$
for bistatic radar	$R_1^2 R_2^2 = \frac{P_{av} t_f G_t G_r \lambda^2 \sigma_b F^4}{(4\pi)^3 k T_s D_x(n') L_t L_{\alpha 1} L_{\alpha 2}}$
for laser radar (coherent)	$R_m^4 = \left(\frac{P_t \rho A_t A_r \eta_t \eta_r}{\Omega_t \Omega_r}\right)\left(\frac{\eta_d}{hfBD}\right)$
for laser radar (noncoherent)	$R_m^8 = \left(\frac{P_t \rho A_t A_r \eta_t \eta_r}{\Omega_t \Omega_r}\right)^2\left(\frac{1}{2hfBD\rho_{bk}}\right)$
for meteorological radar	$R_m^2 = \frac{\pi^2 P_t \tau G_t G_r \theta_a \theta_e \tau_n c F^4 \lvert K^2 \rvert a \times 10^{-18} r^b}{128 k T_s D_x(n) L_t L_\alpha L_p^2 \lambda^2}$
for secondary radar (beacon)	$R_m^2 = \frac{P_b \tau G_b G_r \lambda^2 F^2}{(4\pi)^2 k T_s D_x(n) L_b L_{\alpha 1}}$

for search radar	$R_m^4 = \dfrac{P_{av}A_r t_s \sigma}{4\pi\psi_s kT_s D_0(1)L_s}$
for synthetic aperture radar	$R_m^4 = \dfrac{P_{av}\lambda\Delta_r\Delta_x^2\gamma h_r}{4\pi\theta_e^2\gamma_p K_\theta' kT_s D_0(1)L_t L_\alpha L_n \sin\alpha}$
for tracking radar	$R_m^4 = \dfrac{P_t\tau G_t G_r\lambda^2\sigma F^4}{(4\pi)^3 kT_s D_t(n)L_t L_\alpha}$
with surface clutter	$R_m = \dfrac{I_m\sigma F^4 L_p}{\theta_a(\tau_n c/2)\sigma^0 F_c^4 D_{xc}(n_c)}$
with volume clutter	$R_m^2 = \dfrac{I_m\sigma F^4 L_p^2}{\theta_a\theta_e(\tau_n c/2)\eta_v F_c^4 D_{xc}(n_c)}$
with noise jamming	$R_m^4 = \dfrac{P_{av}t_f G_t A_r\sigma F^4}{(4\pi)^2(kT_s + qJ_0)D_x(n')L_t L_\alpha}$
stand-off noise jamming	$R_m^4 = \dfrac{P_{av}t_f G_t A_r\sigma F^4}{(4\pi)^2 k(T_s + T_j)D_x(n')L_t L_\alpha}$
self-screening noise jammer	$R_m^4 = \dfrac{P_{av}t_f G_r B_r\sigma F^2}{4\pi q P_j G_j D_x(n')L_t L_\alpha}$
escort-screening noise jammer	$R_m^4 = \dfrac{P_{av}t_f G_r B_r\sigma F^4}{4\pi q P_j G_j D_x(n')L_t L_\alpha F_j^2}$

The **altimeter range equation** is found by substituting in (1) the appropriate equation for surface RCS at vertical incidence. There are two specific types of operation for a pulse altimeter:

(1) The beamwidth limited case, where a narrow beam is used and the antenna pattern has a major influence on the build-up of the echo pulse.

(2) The pulsewidth limited case, where the beam is relatively wide and the pulse is short. Case (a) is valid for most practical applications, except for high resolution waveforms at high altitude.

For the beamwidth-limited case:

$$\sigma = A_c\sigma^0 = \frac{R^2\theta_3^2\sigma^0}{L_p^2} \approx \frac{\pi R^2\theta_3^2\sigma^0}{4}$$

where the approximation for effective beam footprint, $\theta^2/L_p^2 = \pi\theta^2/4$ is commonly used in the literature. For the pulsewidth limited case:

$$\sigma = 2\pi\sigma^0 R\tau_n c/2$$

where A_c is the area of the beam footprint on the surface, θ_3 is the half-power beamwidth, σ^0 is the surface reflectivity, $\tau_n c/2$ is the range cell depth, and $L_p^2 = 1.77$ is the two-coordinate beamshape loss. Substituting for σ in (1) leads to

$$R_m^2 = \frac{P_{av}t_f G_t A_r\theta^2\sigma^0}{(4\pi)^2 kT_s D_x(n')L_t L_\alpha L_p^2} = \frac{P_{av}t_f A_r\sigma^0}{4\pi kTD_x(n')L_t L_\alpha L_p^2 L_n}$$

where $G_t = 4\pi/\theta_3^2 L_n$, where $L_n \approx 1.3$ is a constant depending on aperture illumination and efficiency. For many pulsed altimeters, coherent integration is absent and $P_{av}t_f = P_{av}t_r = P_t\tau$, $n' = n$. For the pulsewidth-limited case,

$$R_m^3 = \frac{P_{av}t_f G_t A_r(\tau_n c/2)\sigma^0}{8\pi kT_s D_x(n')L_t L_\alpha L_p^2}$$

where τ_n is the processed pulse width that determines the width of the range resolution cell. *DKB*

Ref.: Cantafio (1989), p. 240.

The **range equation with atmospheric attenuation** [in the radar-target path] is a transcendental equation, having the form:

$$R_m = R_0 \times 10^{-L_\alpha(R_m)/40}$$

$$L_\alpha(R_x) = \int_0^{R_x} k_\alpha(R)dR = \bar{k}_\alpha R_x \text{ dB}$$

where L_α is the two-way path attenuation, k_α is the path attenuation coefficient in decibels/unit length R_0 is the range in the absence of path attenuation R_m is the range with attenuation, and the approximation models the attenuation as uniform over the path.

In Blake's procedure (see **CHART, Blake**) the initial estimate of range R_0 is made without path attenuation, and a one- or two-step iteration is used to obtain the final estimate, R_m:

$$R_1 = R_0[1 - 10^{-L_\alpha(R_0)/40}]$$

$$R_m \approx R_1\{1 + 10^{[L_\alpha(R_1) - L_\alpha(R_0)]/40}\}$$

When the attenuation coefficient at the target is high, additional iterations may be required to achieve adequate accuracy. Alternatively, a computer program may be used to tabulate the actual received signal power or energy at ranges encompassing the final value of R_m, interpolation being used to find the value of R_m at which the signal requirements for detection are met. *DKB*

Blake (1982), pp. 197–221.

The **range equation for bistatic radar** is derived from (1) by replacing R^4 by $R_1^2 R_2^2$, the product of the squares of the two path lengths, target RCS σ by σ_b, its bistatic equivalent, F^4 by the product $F_t^2 F_r^2$ for the two paths, and L_α by the product $L_{\alpha 1}L_{\alpha 2}$ for the two paths. Thus

$$R_1^2 R_2^2 = \frac{P_{av} t_f G_t G_r \lambda^2 \sigma_b F^4}{(4\pi)^3 k T_s D_x(n') L_t L_{\alpha 1} L_{\alpha 2}}$$

DKB

Ref.: Willis (1991), p. 68.

The **range equation in classical form** (for the thermal noise background) was usually expressed in terms of the available and required signal power:

$$R_m^4 = \frac{P_t G_t G_r \lambda^2 \sigma F^4}{(4\pi)^3 S_{min} L_t L_\alpha} \tag{2}$$

where the symbols are as defined for (1). Blake's procedure improved on this by providing a clearer relationship between range and transmitter pulse energy $P_t\tau$ and receiving system noise temperature T_s:

$$R_m^4 = \frac{P_t \tau G_t G_r \lambda^2 \sigma F^4}{(4\pi)^3 k T_s D_x(n) L_t L_\alpha}$$

where $D_x(n)$ is the detectability factor (required single-pulse energy ratio at the receiver input) for video integration of n pulses. Note that this equation does not contain the waveform bandwidth or processed pulse width, but only the transmitted pulse width. Pulse compression only affects the result through introduction of a possible weighting loss component within the detectability factor D_x.

The generalization by Hall, used in Barton's adaptation of the Blake method, applies equally to noncoherent pulsed radar, pulsed doppler radar, and CW radar:

$$R_m^4 = \frac{P_{av} t_f G_t G_r \lambda^2 \sigma F^4}{(4\pi)^3 k T_s D_x(n') L_t L_\alpha}$$

where coherent integration is carried out over the coherent processing interval t_f with subsequent noncoherent integration of n' output samples from the coherent processor during the total observation time t_o: $n' = t_o/t_f$. Note that for noncoherent radar the transmitted pulse energy $P_t\tau = P_{av}t_f$ may be substituted in any of the equations of this section, with $n' = n$.

Thus, the classic form of the equation can be refined to the forms recommended in current radar practice. Note that the uncritical application of (2), setting $S_{min} = kT_sB(\mathrm{SNR})_{min}$ can lead to erroneous results when attempts are made to establish an "equivalent bandwidth" for pulse compression and pulsed doppler radars, and when effects of mismatched filters on $(\mathrm{SNR})_{min}$ are calculated.

A term sometimes used for the classical range equation when on-axis values of antenna gain are used is *searchlight range equation*: the target is said to be "searchlighted" in the beam, and no beamshape loss is included. *DKB*

Blake (1980); Skolnik (1970), Ch. 2; Barton (1974), pp. 71–81, (1988), pp. 17–21.

The **range equation with clutter** can take many forms, depending on the nature of the clutter and whether noise and other interference are included as well. In general, the clutter range equation using signal energy and "effective spectral densities" of clutter C_0 and jamming J_0 must be used:

$$\begin{aligned} R_m^4 &= \frac{E}{D_{xc}(n')(N_0 + J_0 + C_0)} \\ &= \frac{P_{av} t_f G_t G_r \lambda^2 \sigma F^4}{(4\pi)^3 D_x(n') L_t L_\alpha \left[k T_s \left(1 + \frac{C_0}{N_0} + \frac{J_0}{N_0} \right) \right]} \end{aligned} \tag{3}$$

Even without jamming, the evaluation of C_0 remains a problem, since it is usually the powers of clutter C, signal S, and noise N that are calculated. However, if the detectability factor can be found for operation in the noise background $D_x(n')$ and for the clutter background $D_{xc}(n')$, the effective spectral densities can be written as

$$\frac{C_0}{N_0} = \frac{C}{N} \times \frac{D_{xc}(n')}{D_x(n') I_m} \approx \left[\frac{C}{N}\right]\left[\frac{n' L_i(n_c)}{n_c L_i(n')}\right]\left[\frac{L_{cd}}{I_m}\right]$$

where $N_0 = kT_s$, I_m is the signal-to-clutter improvement factor resulting from use of MTI, doppler processing, or polarization, n_c is the number of independent clutter samples integrated, $L_i(n_c)$ is the corresponding integration loss, and L_{cd} is the clutter distribution loss caused by non-Rayleigh clutter amplitude distribution. Typically, 1 mW of clutter power at the processor output has the effect of about 5 mW of noise. When clutter is the dominant background component, simple forms for the clutter range equation can be derived, based on the requirement that the input value $C/S = I_m/D_{xc}(n')$, the sub-clutter visibility of the system, at the maximum detection range.

For volume clutter filling the beam in a system with no range-ambiguous clutter, the input C/S ratio may be written,

$$\frac{C}{S} = \frac{R^2 \theta_a \theta_e (\tau_n c/2) \eta_v F_c^4}{\sigma F^4 L_p^2} \tag{4}$$

which can be solved for maximum detection range:

$$R_m^2 = \frac{I_m \sigma F^4 L_p^2}{D_{xc}(n') \theta_a \theta_e (\tau_n c/2) \eta_v F_c^4} \tag{5}$$

For surface clutter, using the constant-γ model, there is no range dependence for C/S except that caused by the pattern-propagation factors of the signal and the clutter:

$$\frac{C}{S} = \frac{h_r \theta_a (\tau_n c/2) \gamma F_c^4}{\sigma F^4 L_p} \tag{6}$$

Hence, no straightforward solution for detection range is available. At short ranges, where the F factors are near unity, the clutter may or may not overcome the signal, but detection probability remains constant at whatever value is available. If the constant-σ^0 model is used for clutter reflectivity, the product γh_r in (19) may be replaced by $R\sigma^0$, giving a direct range dependence and permitting a solution for F/F_c = constant:

$$R = \frac{\sigma F^4 L_p}{\theta_a(\tau_n c/2)\sigma^0 F_c^4 D_{xc}(n')}$$

Equations applicable when clutter appears in range ambiguities can be derived from (3) using (4), (5), and (6) modified as necessary for each ambiguity containing clutter. *DKB*

Ref.: Barton (1988), Ch. 1.

The **range equation in decibel notation** is a representation of the range equation as the sum of decibel values of factors in the basic equation Practically it is embodied in the Blake chart, and can be summarized as in Table R20. *DKB*

Ref.: Barton (1988), pp. 20–23.

Table R20
Range Equation in Decibel Notation

Parameter	Symbol	Units	+dB	–dB
Peak power	P_t	dBW	$10\log P_t$	
Pulse width	τ	dBs		$-10\log\tau$
Transmit gain	G_t	dB	$10\log G_t$	
Receive gain	G_r	dB	$10\log G_r$	
Wavelength	λ^2	dBm^{-2}		$-20\log\lambda$
Target RCS	σ	dBm^2	$10\log\sigma$	
Patt-prop factor	F^4	dB	$40\log F$	
Basic det factor	D	dB		$10\log D$
Transmit loss	L_t	dB		$10\log L_t$
Match factor	M	dB		$10\log M$
Beamshape	L_p	dB		$10\log L_p$
Misc loss	L_x	dB		$10\log L_x$
Path atten	L_α	dB		$10\log L_\alpha$
Constant	*	*	75.6	
+dB sum		**	X	
–dB sum		***		Y
Combined	$R_m{}^4$	$\text{dB}(\text{km}^4)$	$X–Y$	
Range	$R_m = 10^{(X-Y)/40}$ km			

*The constants represents $[(4\pi)^3 k(\text{m/km})^4]^{-1}$ in units $\text{dB}(\text{W-s})^{-1}$;
**The +dB sum is in units $\text{dB}(\text{m}^2\text{s})$
***The -dB sum is in units $\text{dB}(\text{m}^{-2}\text{s}^{-1})$

The **range equation with interference** determines the maximum operational range of the radar in a background of interference. Interference consists of three main types:

(1) Thermal noise.
(2) Clutter (from a surface, a volume, or both).
(3) Jamming.

For each of these types of interference appearing individually, range equations are given under the corresponding entry. (See **range equation in classical form**, **range equation with clutter**, **range equation with jamming**.) When more than one source contributes a significant level of interference, the maximum range can be determined by evaluating the output signal-to-interference ratio:

$$\frac{E}{I_0} = \frac{E}{N_0 + C_0 + J_0}$$

where E is signal energy, and I_0, N_0, C_0, and J_0 are effective spectral densities of interference, noise, clutter, and jamming, respectively. The maximum range is that for which $(E/I_0) \geq D_x$, where D_x is the detectability factor for the specified target type and integration process in a noise background. Before summing the spectral densities, it is necessary to correct the clutter and jamming components to account for correlation and amplitude distribution, when these differ from white, Gaussian noise. *SAL*

Ref.: Barton (1988), p. 166.

The **range equation with jamming** can be written for the special case of noise jamming over arbitrary bandwidth B_j, for the general case of a jammer separated from the target:

$$R_m^4 = \frac{P_{av} t_f G_t G_r \lambda^2 \sigma F^4}{(4\pi)^3 (kT_s + qJ_0) D(n') L_t L_\alpha}$$

where q is the jamming noise quality factor and J_0 is the jamming power spectral density, given by

$$J_0 = \frac{P_j G_j G_r \lambda^2 F_j^2}{(4\pi)^2 R_j^2 B_j L_{\alpha j}}$$

where P_j is the jammer power, G_j is the jammer antenna gain, F_j^2 is the pattern-propagation factor for the jammer-to-radar path (including radar sidelobe level as appropriate), R_j is the jammer range, B_j is the jammer noise bandwidth, and $L_{\alpha j}$ is the one-way path attenuation from jammer to radar.

Special cases of interest can lead to simple range equations. For the standoff jammer (SOJ), R_j = constant, it is convenient to express the jammer noise as an equivalent temperature T_j, which may be added directly to T_s, forming a modified receiving system temperature, $T_s' = T_s + T_j$:

$$T_j = \frac{J_0}{k} = \frac{P_j G_j G_r \lambda^2 F_j^2}{(4\pi)^2 R_j^2 k B_j L_{\alpha j}}$$

The value T_s' may then be substituted in any of the classical radar equations to solve for range in the presence of the SOJ.

For the self-screening jammer (SSJ), $R_j = R_m$, $F_j = F$, $T_j >> T_s$, and

$$R_{m_{ssj}}^2 = \frac{P_{av} t_f B_j G_r \sigma F^2}{4\pi q P_j G_j D_x(n') L_t L_{\alpha j}} \quad (7)$$

For an escort screening jammer (ESJ), the term F^2 in (7) is replaced by F^4/F_j^2, where F applies to the radar-to-target path and F^j to the jammer-to-radar path.

The repeater jamming equation gives the effectiveness of a repeater jammer in terms of the jammer-to-signal (J/S) ratio that it is capable of enforcing against the victim radar. Repeater jammers are mostly used in deceptive ECM (DECM) systems on board aircraft and other type of penetrators. Sometimes referred to as self-protection jammers, they protect the penetrating host platform from ground and airborne air defense systems. There are two classes of repeater jammers:

(1) The "true" repeater, which receives the victim radar signal, modulates the received signal to achieve the appropriate deceptive effect, and retransmits the modified (and amplified) version of the signal back to the victim radar.

(2) The "transponder" repeater or "pseudo" repeater, which uses the input signal simply to identify the signal characteristics and trigger a programmed response, or stores the signal to later amplify a delayed, synthesized version of the signal.

For the true repeater case, neglecting jammer losses, the *J*/*S* ratio is expressed as:

$$\frac{J}{S} = \frac{G_e G_j^2 \lambda^2}{4\pi\sigma}$$

where G_e is the jammer electronic gain, G_j is the jammer antenna gain in the direction of the victim radar, λ is the radar signal wavelength, and σ is the radar cross section of the jammer platform (aircraft). Jammer electronic gains on the order of 70 to 80 dB are typical. The resultant jammer output power P_j is equal to $S_j G_e$, where S_j is the input signal level.

For the transponder repeater case, the *J*/*S* ratio can be expressed as

$$\frac{J}{S} = \frac{4\pi R^2 P_j G_j}{P_T G_T \sigma}$$

For the transponder repeater, *J*/*S* is limited by the jammer transmitter, which typically operates in the saturated mode.

The self-screening range equation for a search radar can be written as

$$R^2_{m_{ssj}} = \frac{P_{av} t_o \sigma}{\left(\frac{P_j G_j}{B_j}\right)\psi_s D_0(1) L_s}$$

where the symbols are as defined above. The radar receiving aperture does not contribute to increasing range in this case because it is equally effective on the signal and the jamming. (See also **self-screening jammer**; **search radar equation**.) *DKB*, *PCH*

Ref.: Schleher (1986), pp. 138, 418; Barton (1988), pp. 33–37.

The **range equation for laser radar** can be derived from the expression for received signal power P_r:

$$P_r = \left(\frac{P_t}{R_t^2 \Omega_t}\right)\left(\frac{\rho A_r}{\Omega_r}\right)\left(\frac{A_c}{R_r^2}\right)\eta_t \eta_r$$

where P_t is the transmitted laser power, R_t is the range from transmitter to target, Ω_t is the solid angle of the transmitted beam, ρ is the average reflectivity of the target, $A_r = (\pi/4)R_t^2\Omega_t^2$ is the projected area of the illuminated part of the target, Ω_r is the solid angle of the reflected beam, A_c is the receiving aperture area, R_r is the range from receiver to target, and η_t and η_r are transmission coefficients for the transmitted and received beams, including optics and atmospherics.

The signal-to-noise ratio at the output of the coherent (superheterodyne) thermal-noise-limited receiver can be written as

$$\frac{S}{N} = \frac{\eta_d P_r}{hfB}$$

and for the noncoherent receiver as

$$\frac{S}{N} = \frac{\eta P_r^2}{2hfBP_{bk}}$$

where η_d is the detector quantum efficiency, $h = 6.626 \times 10^{-34}$ J-s is Plank's constant, f is the transmission frequency, B is the electronic bandwidth of the receiver, and P_{bk} is the background power.

Combining the equations and setting the result equal to a required $S/N = D$, we obtain equations for maximum detection range, for a monostatic coherent system as

$$R^4 = \left(\frac{P_t \rho A_t A_r \eta_t \eta_r}{\Omega_t \Omega_r}\right)\left(\frac{\eta_d}{hfBD}\right)$$

and for the noncoherent system as

$$R^8 = \left(\frac{P_t \rho A_t A_r \eta_t \eta_r}{\Omega_t \Omega_r}\right)^2\left(\frac{1}{2hfBDP_{bk}}\right)$$

The required SNR may be found for given detection and false-alarm probabilities using the same procedures as for conventional radar. (See **DETECTABILITY FACTOR**.) *DKB*

Ref.: Jelalian (1992), pp. 8–21.

The **range equation for meteorological radar** can be derived by replacing the target RCS with that of the desired precipitation:

$$\sigma = V_c \eta_v$$

where

$$V_c = \left(\frac{R^2 \theta_a \theta_e (\tau_n c/2)}{L_p^2}\right)$$

is the volume of the radar resolution cell filled with precipitation and

$$\eta_v = \left(\frac{\pi^5}{\lambda^4}|K|^2 \sum_{i=1}^{N} D_i^6\right) = \frac{\pi^5}{\lambda^4}|K|^2 Z$$

is the reflectivity of precipitation. For definitions of the factors in these equations, see **CLUTTER**.

Using this RCS in the radar equation, we obtain the meteorological radar equation for received echo power

$$S = \frac{P_t G^2 \lambda^2 \theta_a \theta_e c \tau_n}{512(2\ln 2)\pi^2 R^2} \sum_{i=1}^{N} D_i^6 = \frac{P_t G^2 \lambda^2 \theta_a \theta_e c \tau_n \pi^3 |K|^2 Z}{512(2\ln 2)\lambda^2 R^2} \tag{8}$$

where $L_p^2 = (8/\pi)\ln 2$ has been calculated accurately for a Gaussian beam and can be applied also to low-sidelobe beams commonly used. In (8) the units of length must be consistent (e.g., meters). Values of Z stated in mm^6/m^3 must be multiplied by 10^{-18} for use in the equation. Empirical equations relating Z in mm^6/m^3 to precipitation rate r in mm/h are:

Stratiform rain: $Z = 200r^{1.6}$
Orographic rain: $Z = 31r^{1.71}$
Thunderstorm rain: $Z = 486r^{1.37}$
Snow: $Z = 2{,}000r^2$

Note that the constants used in these relationships are dimensional. The meteorological radar equation (8) may be rewritten in many forms, including substitution of relationships for Z as a function of precipitation rate. Care must be taken in converting units of length.

A form of the meteorological range equation in which these various relationships are combined is

$$R_m^2 = \frac{\pi^2 P_t \tau G_t G_r \theta_a \theta_e \tau_n c F^4 |K^2| a \times 10^{-18} r^b}{128 k T_s D_x(n) L_t L_\alpha L_p^2 \lambda^2}$$

where r is the precipitation rate in mm/h, and $|K|^2$, a, and b are parameters of the precipitation. (See **CLUTTER, rain.**) *DKB*

Ref.: Atlas (1964); Nathanson (1969), pp. 199–201; Skolnik (1990), pp. 23.2–23.5; Meneghini (1990), Ch. 2; Sauvageot (1992), pp. 145–150.

The **range equation for MTI radar** is written separately for targets in regions containing clutter and for clear regions, where the target competes only with noise at the output of the MTI processor. The effect of MTI in the latter case is to introduce extra processing loss components in the miscellaneous signal-processing loss factor, L_x.These loss components are functions of the required detection probability:

(1) MTI noise correlation loss:

$$L_{mti(a)} = \frac{D_x(an)}{D_x(n)}$$

where $a < 1$ is a factor reducing the number of independent noise samples at the output of the canceler.

(2) MTI target fluctuation loss:

$$L_{mti(b)} = \frac{L_f(bKn_e)}{L_f(Kn_e)}$$

where $L_f(x)$ is the fluctuation loss for a target with $2x$ degrees of freedom, and $0.5 < b < 1$ is a factor expressing the possible loss resulting from processing of only the in-phase component of the signal channel.

(3) MTI velocity response loss: When targets lie within the velocity region with low MTI response, there is a reduction in detection probability for some fraction of the targets, which may not be compensated by increase in probability for targets near the maximum MTI response. The velocity response loss is the required increase in signal-to-noise ratio to regain the required average detection probability.

The three components of MTI loss increase the value of detectability factor D_x, used in several forms of the radar equation, even when only noise is competing with a target in a clear region. When clutter is also present, the appropriate equations must also include the increased value of D_{xc}, along with the MTI improvement factor I_m. (See **range equation in classical form**; **range equation with clutter.**) *DKB*

Ref.: Nathanson (1969), pp. 354–358; Barton (1988), pp. 250–252.

range equation for multistatic radar (see **range equation for bistatic radar**).

The **range equation for over-the-horizon radar** takes the classical form (1), but its evaluation is made difficult by the need to evaluate the pattern-propagation factor F. in all propagation regions. (See **PROPAGATION.**) The uncertainty in ionospheric transmission conditions introduces greater errors for this class of radar than for conventional, line-of-sight systems. *DKB*

Ref.: Skolnik (1990), p. 24.3; Kolosov (1987), Chaps. 2, 3.

The **range equation for polarimetric radar** is the basic radar equation (1) but must include factors expressing the polarization properties of the target RCS in relation to transmitted and received antennas. This requires that the pattern-propagation factor F include as a component a two-way polarization factor F_p^4, or alternatively that this polarization factor be included to multiply the conventional σ in any of the range equations. Approximate values of F_p^4 for different transmitter and receiver polarizations are in Table R21 for targets (typical aircraft and missiles), rain, chaff, and jammers.

Table R21
Polarization Factors for Targets and Clutter

Polarization		Factor F_p^4			Factor F_p^2
Trans.	Rec.	Targets	Rain	Chaff	Jammer
H	H	1.0	1.0	1.0	1.0
H	V	0.1	0.01	0.25	0.1
H or V	R or L	0.5	0.25	0.5	0.5
V	V	1.0	1.0	0.5	1.0
V	H	0.1	0.01	0.25	0.1
R	R	0.5	0.01	0.5	1.0
R	L	1.0	1.0	0.5	0.1
R or L	V or H	0.5	0.25	0.5	0.5
L	L	0.5	0.01	0.5	1.0
L	R	1.0	1.0	0.5	0.1

The polarization effect for bistatic meteorological radar is discussed by Meneghini. *DKB*

Ref.: Barton (1993); Meneghini (1990), p. 58.

The **range equation in practical units** is the same as the classical form, but includes conversion factors for the units in which radar parameters are often expressed:

$$R_m^4 = K \frac{P_{t(\mathrm{kW})} \tau_{(\mu\mathrm{s})} G_t G_r \sigma F^4}{f^2_{(\mathrm{MHz})} T_s D_x(n) L_t L_\alpha}$$

where K = 129.2 for R_m in nmi;
= 148.7 for R_m in st mi;
= 239.3 for R_m in km;
= 261.7 for R_m in kyds;
= 185.0 for R_m in kft.

The wavelength, λ, has been replaced by c/f and appropriate values for c and other constants have been grouped into K. *SAL*

Ref.: Skolnik (1990), p. 2.7.

The **search radar equation** relates the detection range of a radar that scans uniformly over a solid angle ψ_s in a frame time t_s to the radar and target parameters:

$$R_m^4 = \frac{P_{av} A_r t_s \sigma}{4\pi \psi_s k T_s D_0(1) L_s}$$

where $D_0(1)$ is the basic detectability factor for a single sample of a steady signal, and losses from integration, fluctuation, and signal processing are included as components of L_s. The key radar parameter in this equation is the product of a radar's average transmitter power and the effective aperture area (i.e., the power-aperture product = $P_{av}A_r$). For a continuous-wave (CW) radar, the peak and average power are identical. In a pulsed radar, the average power is equal to the product of the peak pulse power P_t and the radar duty factor. Since the duty factor is equal to the product of the pulse width τ and the pulse repetition frequency f_r, $P_{av} = P_t \tau f_r = P_t(\tau/t_r)$, where t_r is the pulse repetition period.

The antenna effective aperture area is equal to

$$A_r = \frac{G_r \lambda^2}{4\pi}$$

where G_r is the antenna gain, and λ is the radar wavelength. Other relationships used in deriving the search radar equation are

(1) Relationship of transmitting antenna gain to beamwidths:

$$G_t = \frac{4\pi}{\theta_a \theta_e L_n}$$

(2) Relationship of search solid angle ψ_s, beamwidths θ_a and θ_e, and frame time t_s, to available integration time t_o:

$$t_o = t_s \frac{\theta_a \theta_e}{\psi_s}$$

In the search radar equation, σ is the target's RCS, k is Boltzmann's constant, $D_0(1)$ is the detectability factor (required SNR) for a single sample, and L_s is the total search loss.

The search radar equation is significant because it indicates that radar transmitting gain, wavelength and waveform are important only in that they affect the receiving system temperature, the search loss factor, and the value of receiving aperture A_r that may be used. The same equation can be used to find the acquisition range of a tracking radar while performing its acquisition scan over the solid angle ψ_s. *PCH, DKB*

Ref: Barton (1988), pp. 12–26.

The **range equation for** secondary radar describes the range at which the radar can successfully interrogate a beacon (transponder, or can detect the response. Two equations are necessary to describe the range performance of a beacon system: one characterizes the range of the interrogator and the other the range of the reply (transponder) link. The one-way radar beacon equation for detection of the response is

$$R_{mb}^2 = \frac{P_b \tau G_b G_r \lambda^2 F^2}{(4\pi)^2 k T_b D_x(n) L_b L_{\alpha 1}}$$

The subscript b represents the beacon transponder, the subscript r represents the interrogating radar, and $D_x(n)$ includes the effect of integration for the beacon-to-radar link. For the interrogation link, R_{mi} is found when $P_b G_r / T_b L_b$ is replaced by the corresponding radar values, $P_t G_t / T_s L_t$, and $D_x(n)$ is a single pulse value (n = 1) appropriate to the beacon triggering process. The pattern-propagation factor F and atmospheric loss $L_{\alpha 1}$ are those of the one-way path from beacon to radar and vice versa. The shorter of the two ranges determines the system operating range. When the same antennas are used for interrogation and response, at essentially the same frequency, $G_r = G_t$, and the ratio of squared ranges becomes

$$\frac{R_{mb}^2}{R_{mi}^2} = \frac{P_b T_b D_x(1)}{P_i T_s D_x(n)}$$

where T_i is the radar receiving noise temperature, $D_i(n)$ is the radar detectability factor, and similar terms with subscript b apply to the beacon receiver. The radar interrogation power normally exceeds the beacon power by a factor of 1,000 or more, and the minimum required beacon signal for triggering similarly exceeds the detectable signal in the radar, producing a rough balance in range capability of the two links. *PCH*

Ref.: Barton (1988), pp. 12, 13; Skolnik (1962), pp. 594–596.

The **range equation for** synthetic-aperture radar is given as

$$R_m^3 = \frac{P_{av} G^2 \lambda^2 \sigma^0 \Delta_r}{(4\pi)^3 k T_s D_0(1) L (2v_p/\lambda) \sin\alpha}$$

where σ^0 is the surface reflectivity, Δ_r is the range resolution cell width, v_p is the radar platform velocity, α is the beam angle from broadside, L is a combined loss factor, and other terms are as defined above. When expressed in terms of the cross-range resolution Δ_x, the surface reflectivity parameter γ, and the elevation beamwidth θ_e, this becomes

$$R_m^4 = \frac{P_{av} \lambda \Delta_r \Delta_x^2 \gamma h_r}{(4\pi) \theta_e^2 v_p K_\theta' k T_s D_0(1) L_s L_n \sin\alpha}$$

where h_r is the radar altitude above the surface, $K_\theta' \approx 0.5$ is the synthetic-aperture beamwidth constant, L_s is the total search loss, and L_n is a factor relating gain to beamwidth. This last form can be derived directly from the search radar equation. *DKB*

Ref.: Hovanessian (1984), p. 225; Barton (1988), p. 362.

The **equation for** tracking radar is calculated from the required accuracy of track or the minimum S/N power ratio required to maintain the AGC and tracking loops. When nonlinear detector effects are considered, the tracking error in the x coordinate can be expressed, normalized to the half-power width of the resolution cell, as

$$\frac{\sigma_x}{x_3} = \frac{\sqrt{(1+S/N)(\beta_n t_f)}}{k_x(S/N)} = \frac{\sqrt{\beta_{no} t_f}}{k_x\sqrt{1+S/N}}$$

where β_n is the tracking loop bandwidth, β_{no} is the design bandwidth for strong signals, t_f is the coherent processing interval (equal to t_r for a noncoherent radar), and k_x is the tracking error slope. The bandwidth achieved for weak signals is

$$\frac{\beta_n}{\beta_{no}} = \left(\frac{S/N}{1+S/N}\right)^2 \qquad (9)$$

A requirement for S/N for accurate tracking on a nonaccelerating target may be found as

$$\frac{S}{N} \geq D_t = \frac{x_3^2 \beta_{no} t_f}{\sigma_x^2 k_x^2} - 1$$

For example, for a loop that integrates $(1/2\beta_{no}t_f) = 50$ samples under high signal conditions, to track to an accuracy of 5% of the resolution cell width ($\sigma_x/x_3 = 0.05$), with a normalized slope $k_x = 1.4$:

$$D_t = \frac{400 \times 0.01}{2\sigma_x^2 k_x^2} - 1 = 1.0$$

unity single-sample SNR is required. This value may be used in the classical radar equation to find the maximum tracking range for this accuracy on nonaccelerating targets, but (9) indicates that the loop bandwidth is reduced by a factor of four for unity SNR.

For accelerating targets, the tracking lag error will be

$$\varepsilon_a = \frac{a_t}{K_a} = \frac{a_t}{2.5\beta_n^2} = \frac{a_t}{2.5\beta_{no}^2}\left(\frac{1+S/N}{S/N}\right)^2$$

In the example given, this will be 16 times the lag for high SNR. Considering a target acceleration a_t = 1g = 10 m/s^2, with a design loop bandwidth β_{no} = 2 Hz, the lag error for high SNR is ε_a = 1m, increasing to 16m for S/N = 1. For a range tracker using τ_3 = 1 ms, Δ_r = 150m, this lag is more than double the allowable $\sigma_t/\tau_3 = 0.05$ and the required $D_t \approx 2$. Loss of track will occur when $\varepsilon_a > x_3/2$, normally occurring first in the high-resolution range coordinate, $x_3 = \Delta_r$. In our example with a_t = 1g, ε_a > 75m for S/N < 0.55. *DKB*

Ref.: Barton (1988), pp. 467–472.

RANGEFINDER, RANGING, radar

Continuous-wave ranging is measurement of range to a target using continuous-wave (CW) signals. Depending on the measurement method, CW ranging is based on phase or frequency. Phase rangefinding is a technique applied to sinusoidally modulated signals in a CW radar system. The phase delay between transmission and reception is

$$\Delta\phi = t_d\dot{\phi} = t_d \cdot 2\pi f_m = \frac{2R}{c} \cdot 2\pi f_m$$

where t_d is the range delay and f_m is the modulating frequency. The resulting range estimate is

$$R = \frac{c}{4\pi} \cdot \frac{\Delta\phi}{f_m}$$

and the accuracy of the estimate is

$$\sigma_r = \frac{c}{4\pi} \cdot \frac{\sigma\phi}{f_m}$$

Measurement of the $\Delta\phi$ requires that the delayed echo be resolved from the transmission, either through the use of a transponder to shift the echo carrier frequency out of the transmission band, or by exploitation of target doppler shift. If the latter process is used, the ranging modulation frequency must be kept small enough that the transmitter sidebands from direct feedthrough and clutter reflections do not overlap the echoes from short-range targets: $f_m \ll f_{d\min} = 2v_{t\min}/\lambda$. In addition, the modulation frequency sidebands must remain within the bandwidth B_f of the target doppler filter. As a result, phase rangefinding in CW doppler radars is relatively inaccurate, although the technique can be applied with greater accuracy to altimeters. Assume, for example, that $f_{d\min}$ = 3 kHz to resolve targets from short-range clutter, and B_f = 500 Hz. The ranging frequency is limited to f_m < 200 Hz. If the phase measurement is accurate to σ_f = 0.1rad, we find

$$\sigma_r = \frac{3 \times 10^8}{4\pi} \cdot \frac{0.1}{200} = 12 \text{ km}$$

Range to the target is determined in a phase rangefinder through measurement of the difference in phases between the modulating oscillations of a scale frequency, which are extracted from emitted and received continuous signals (Fig. R87a).

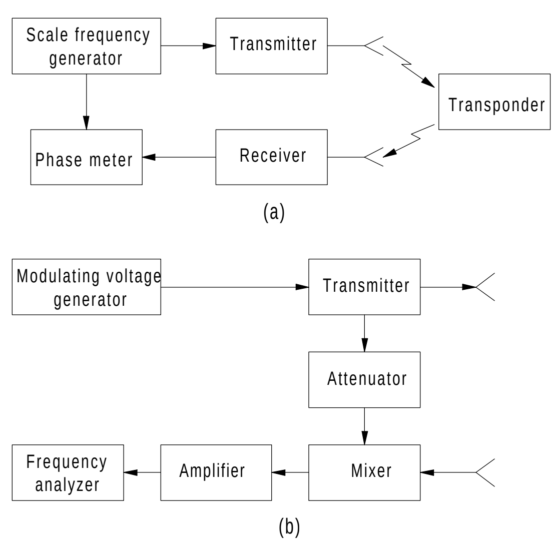


Figure R87 CW range finder: (a) phase; (b) frequency (after Dulevich, 1978, Figs. 7.2, p. 218, and 7.4, p. 221).

A scale frequency oscillator modulates the continuous signal generated in a transmitter relative to amplitude or frequency. The signal is extracted at the point of receipt and is used for modulation of the oscillations of the transmitter of a repeater. Scale frequency oscillations are extracted from a received signal in the rangefinder receiver. Phase shift between transmitter and receiver scale frequency oscillations is extracted in a phase meter and makes it possible to determine the range to an object. (See **MEASUREMENT, range.**) The main advantage of a phase rangefinder is high range-measurement accuracy. Along with this, there are a number of shortcomings, including inability to simultaneously measure the range of several objects and range-measurement ambiguity, requiring use of multiscale measurements. Due to these shortcomings, a phase rangefinder is used for measurement of the range of objects equipped with transponders.

Range to the target is determined in a frequency rangefinder through measurement of the difference in frequencies (beat frequency) of received and emitted continuous signals (Fig. R87b). The rangefinder device is a frequency analyzer (spectrum analyzer). A parallel spectrum analyzer, which is a set of a certain number N narrowband filters covering the band of possible beat frequency values, is usually used. Range is determined based on the number of the spectrum analyzer in which the signal appeared.

Ranging on FMCW waveforms is based on the frequency shift between a delayed signal and that being transmitted. This range-delay shift is

$$\Delta f_r = -t_d \dot{f}$$

where t_d is the range delay and $\dot{f}$ is the frequency sweep rate. The range delay-shift may be converted to target range using

$$R = \frac{ct_d}{2} = -\frac{c\Delta f_r}{2\dot{f}}$$

On moving targets, the process is complicated by the target doppler shift, f_d, which adds to the range-delay shift, introducing ambiguity in the reading (see **range-doppler coupling**). Resolution of this ambiguity requires use of two different FM sweep rates, giving two equations from which unknown R and v_t can be determined:

$$R = \frac{ct_d}{2} = -\frac{c\Delta f_r}{2\dot{f}} = \frac{c}{2}\left(\frac{\Delta f_1 - \Delta f_2}{\dot{f}_1 - \dot{f}_2}\right)$$

$$v_t = \frac{\lambda \Delta f_v}{2} = \Delta f_1 + \left(\frac{\Delta f_1 - \Delta f_2}{\dot{f}_1 - \dot{f}_2}\right)$$

The accuracy of range data using two sweep rates is related to the frequency measurement accuracy σ_f:

$$\sigma_r = \frac{c}{2} \cdot \frac{\sqrt{2}\sigma_f}{\left|\dot{f}_2 - \dot{f}_1\right|}$$

The main advantage of a frequency rangefinder is low emitted signal power compared with that of a pulse rangefinder, as well as high range-measurement accuracy. Shortcomings of a frequency rangefinder include equipment complexity, where distance measurement of many objects is concerned, and high demands for linearity of the frequency sweep. *AIL, DKB*

Ref.: Dulevich (1978), pp. 217–223; Skolnik (1980), pp. 95–98; Hovanessian (1988), pp. 62–64.

Laser rangefinding is measurement of range to a target in the optical waveband. Depending on range measurement method, laser rangefinders are categorized as pulse and continuous-wave (CW) radars. A laser rangefinder comprises a laser, Q-switch, synchronizer, photoreceiver, frequency counter, crystal oscillator, gated amplifier, and terminal (Fig. R88a). In this rangefinder, range is measured through measurement of the time interval between the moment the laser generates a pulse and the moment the reflected signal is acquired. The time delay is measured using a time interval counter (frequency counter) and terminal. Indicators or computerized automatic meters are used as the terminal. The diagram depicted in Fig. R88b explains the principle behind a continuous laser rangefinder.

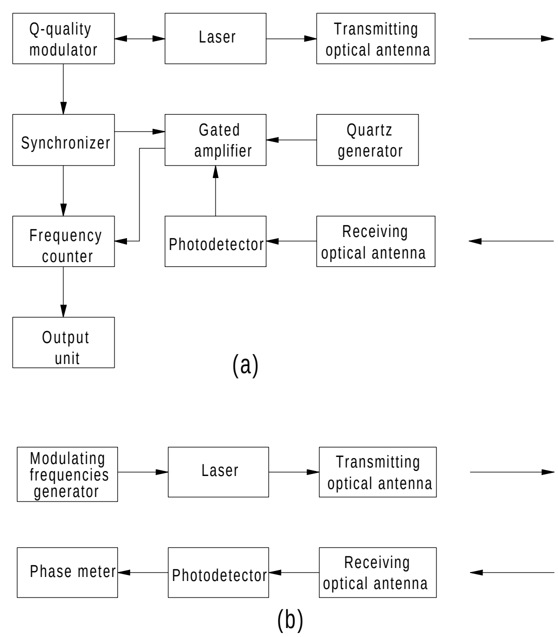


Figure R88 Laser range finders: (a) pulsed; (b) phase (after Dulevich, Fig. 3.7, p. 360; Fig. 7.3, p. 218).

A phase method is used to measure range. Laser transmitter radiation is modulated by harmonic oscillations on several frequencies. Range is determined from the phase difference between the transmitted and reflected signals. A phase laser rangefinder has higher accuracy compared with that of a pulse rangefinder. But it has a shortcoming linked with range determination ambiguity. Therefore, pulse-phase laser rangefinders are often used. Laser transmitter pulses are modulated by a harmonic radio signal such that there are at least 5-10 oscillations per pulse. A pulse rangefinder determines absolute range. A phase rangefinder evaluates changes in range. The result is high range-determination accuracy.

Laser rangefinders are used in laser radars, in artillery, when firing tank guns, in various meters, and so forth. *AIL*

Ref.: Skolnik (1970), p. 37.60; Vorobzhev (1983), p. 153; Dulevich (1978), p. 218.

Multiple-PRF ranging is the process of converting two or more measurements of apparent (ambiguous) range, $R_a = R - jR_u$, in a range ambiguous (high- or medium-PRF) system, to true target range. The total instrumented range R_s over which the ambiguity can be successfully resolved in a dual-PRF system depends on the product of the two unambiguous ranges and their difference:

$$R_s = \frac{R_1 R_2}{\Delta} = \frac{R_1 R_2}{R_1 - R_2}$$

where R_1 is the longer unambiguous range, R_2 is the shorter, and Δ is the difference. This equation is based on the fact that a target beyond the maximum instrumented range lies at range $jR_1 = (j+1)R_2$, and hence gives the same apparent ranges as one in the first range interval.

Close spacing of R_1 and R_2 is desirable to extend R_s, but to resolve the ambiguity (by determining j) the two readings of apparent range must differ by more than the maximum expected error in the difference measurement:

$$\Delta \geq 3\sigma_{\Delta r} = 3\sqrt{2\sigma_r} = 4.2\sigma_r$$

The result is

$$R_s \leq \frac{R_1 R_2}{4.2\sigma_r} = \frac{R_1(R_1 - 4.2\sigma_r)}{4.2\sigma_r}$$

For example, assume a high-PRF system operating with R_1 = 1,500m, capable of measuring range to σ_r = 50m. The maximum instrumented range for ambiguity resolution is 9,200m. To extend this range, a third and possibly a fourth PRF is needed, and targets must be detected on all these PRFs. *DKB*

Ref.: Skolnik (1990), Ch. 17; Hovanessian (1988), pp. 64–68.

Pulsed ranging is measurement of range to a target through fixing the moments of transmission and reception of pulse signals. A pulsed rangefinder is typically a simple pulsed radar that comprises a synchronizer, transmitter, antenna switch, antenna, receiver, and terminal (Fig. R89). The synchronizer generates of train of video pulses for synchronization of the transmitter and output unit. The transmitter forms RF modulated pulses that reach the antenna via the antenna switch and are radiated into space. Received signals from receiver output reach the terminal where time lag of these pulses relative to the transmitted pulses is measured. (See **MEASUREMENT, range.**) A pulsed rangefinder uses as its output unit visual indicators or automatic computerized meters which simultaneously convert magnitude into digital code.

The main feature of a pulsed rangefinder is the capability to measure the range of many objects using relatively simple equipment. Shortcomings include limited capabilities to measure radial velocity. *AIL*

Ref.: Dulevich (1978), p. 215.

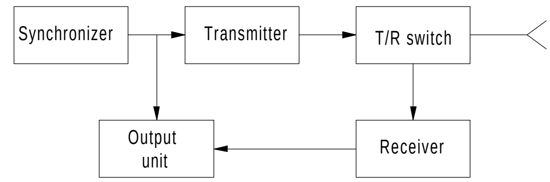


Figure R89 Pulsed range finder (after Dulevich, 1978, Fig. 7.1, p. 216).

RECEIVER, RECEPTION, radar. The receiver is a radar subsystem whose function is to receive radar returns, amplify them, convert in frequency, and filter in a manner to provide maximum discrimination between the desired echoes and undesired interference. The main RF characteristics of a receiver are sensitivity, selectivity, bandwidth, dynamic range, and interference immunity. The main type of a receiver in virtually all modern radars is the superheterodyne receivers.

Other types receivers, such as crystal video, tuned radio frequency, and superregenerative receivers are seldom used except in the simplest radars or radar beacons. A general configuration of radar receiver is shown in Fig. R90. The first

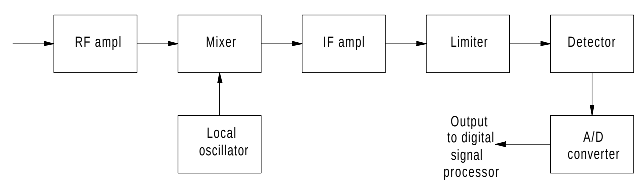


Figure R90 Basic elements of radar receiver.

component of the configuration is an RF amplifier (in Russian radar literature it is more frequently termed *high-frequency amplifier*) performing amplification of the signal at its carrier frequency. Typically it has three functions:

(1) To decrease noise figure and therefore to increase receiver sensitivity (it is first in an amplification chain and so to the greatest extent defines noise figure of the whole receiver, see **NOISE**). That is why low-noise amplifiers are preferable to serve as an RF amplifier.

(2) To amplify the received echo.

(3) To provide RF selectivity.

Frequency selectivity is ensured by using resonant loads, in the radar band typically by bandpass filters. The RF amplifier is followed by the downconverter, including a mixer and a local oscillator (LO) whose function is to convert the carrier to intermediate frequency, giving sufficient gain and the desired shape of the frequency response. As the requirements for the value of intermediate frequency are often contradictory (see **CONVERTER, frequency**), double or threefold frequency conversion may be employed in the radar receiver. In this case more than one LO and mixer are used. In modern radar receivers, LOs not only supply a signal to convert radar frequency, but also act as timing standard to extract range information from echo delay. The first LO, generally termed a *stable local oscillator (STALO)*, is very important for processing performance and defines the stability and coherence performance of the whole radar system. The final LO, generally referred to as a *coherent oscillator (COHO)*, is often used to introduce phase correction for the purposes of improving output signal performance.

After amplification at IF, the signal is filtered, passed through a limiter that prevents saturation of subsequent stages, and applied to an amplitude, phase, or synchronous detector. In most modern receivers the analog processing is complete at this stage, and the signal is then A/D converted. Subsequent processing, including matched filtering and CFAR processing, are performed in a digital signal processor, considered to be a separate radar subsystem. *PCH, SAL*

Ref.: Skolnik (1980), Ch. 9, (1990), Ch. 3; Fink (1982), pp. 25.66–25.72.

The **acousto-optic (Bragg-cell) receiver** is a receiver using a Bragg cell as the key element. The main elements of acousto-optic receivers are the laser, illuminating the Bragg cell with coherent light, the Bragg cell, deflecting the light in proportion to the frequency of the applied signal, and a transform lens focusing the light onto a photodiode detector array (Fig. R91). In principle this device performs as a Fourier-transform spectrum analyzer and the spatial distribution across the photodiode detector array represents the instantaneous Fourier transform of the input signal across the Bragg cell aperture. The frequency resolution is approximately equal to the acoustic transit time through the crystal. The acousto-optic receivers primarily are used in electronic support measures systems to ensure good performance under the conditions of dense signal environment. The performance specifications of a typical acousto-optic receiver are given in the Table R22. This receiver is also termed a *Bragg-cell receiver*, *acousto-optic spectrum analyzer (AOSA)*, or *instantaneous Fourier transform (IFT) receiver. SAL*

Ref.: Schleher (1986), p. 89; Neri (1991), p. 296; Wiegand (1991), p. 176.

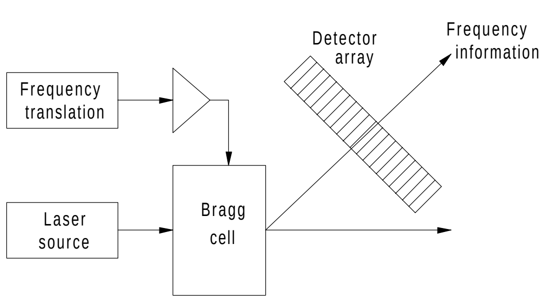


Figure R91 Bragg-cell instantaneous Fourier transform IFT (after Schleher, 1986, Fig. 2-9, p. 76).

Table R22
Performance Specification of Advanced Bragg-Cell Receiver

Parameter	Specification	Units
Center frequency	3.0	GHz
Instantaneous bandwidth	2.0	GHz
Diffraction efficiency	3	% per RF Watt
Dynamic range:		
Noise level	50	dB
Intermodulation	45	dB
Sensitivity	–60	dBm
Resolution:		
Frequency	10	MHz
Simultaneous signals	10	MHz
Laser wavelength	6328/8300	Å
Laser power	10–20	mW (CW)
(from Neri, 1991)		

A **back-bias receiver** is an ECCM receiver used against narrowband CW or spot noise jamming. It has a special circuit detecting CW signals and after these signals are detected they are suppressed, so only pulse signals appear at the receiver output. The process of suppression of CW jamming is shown in Fig. R92. Although the insertion of the back-bias receiver reduces the radar sensitivity by the order of 3 dB, jamming suppression about 50–60 dB can be obtained. *SAL*

Ref.: Neri (1991), p. 432.

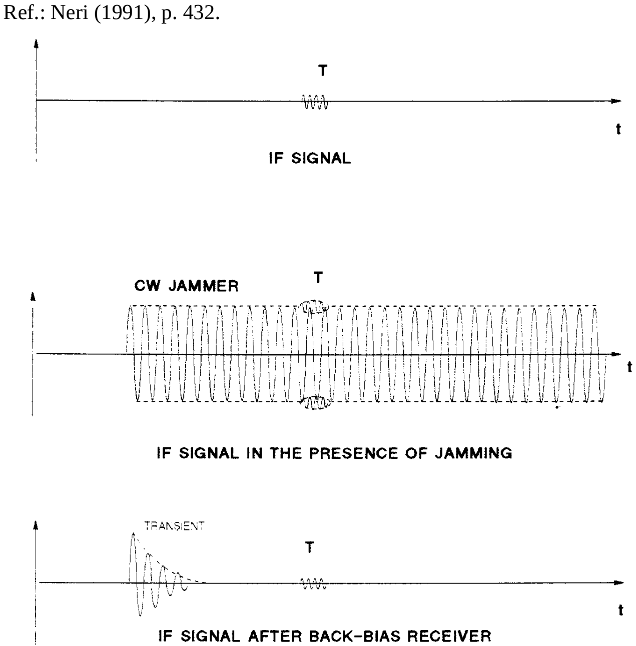


Figure R92 The suppression of CW signal with back-bias receiver (from Neri, 1991, Fig. 6.9, p. 433).

Receiver bandwidth is the band over which the receiver operates with the specified quality. It characterizes the selectivity capabilities of the receiver and is usually defined as the difference between the frequencies at which the gain drops by

3 dB. For pulsed radars, the optimum bandwidth (from the viewpoint of signal-to-noise ratio) depends on pulse width and shape but lies between 0.8/τ for a Gaussian pulse and 1.4/τ for a rectangular pulse, where τ is the pulse width. Many receivers use a somewhat wider bandwidth to preserve the near-rectangular shape of a pulse and provide better resolution between closely spaced targets. *SAL*

Ref.: Druzhinin (1967), p. 376; Barton (1969), pp. 70–72: Davydov (1988), p. 119; Leonov (1988), p. 54; Skolnik (1990), p. 35.

A **channelized receiver** is an electronic support measures (ESM) receiver in which the total coverage bond of the receiver is divided into many channels by means of contiguous filters. Each channel employs the circuits for detection, frequency measurement, analog-to-digital conversion, so the number of the components it uses is rather high. To reduce the cost of such types of the receiver advanced integrated circuit technology must be used for producing filtering, frequency converting, and amplifying circuits. *SAL*

Ref.: Neri (1991), p. 295; Wiley (1982), p. 234, Wiegand (1991), p. 143.

Coherent reception (see **DETECTION, coherent**).

The **compressive receiver** is electronic support measures receiver based on the use of dispersive delay lines producing the delay proportional to the frequency of the signal. In this receiver, the frequency information is transferred into delay information; it can discriminate and analyze signals of different frequencies overlapping in time with high probability of intercept. *SAL*

Ref.: Neri (1991), p. 299; Schleher (1986), p. 81.

A **constant-false-alarm-rate (CFAR) receiver** is a receiver employing constant-false-alarm-rate technique to maintain the constant level of interference background at its output; that is, automatically adjust its sensitivity as the intensity of interference varies. (See **CFAR.**) *SAL*

Ref.: Skolnik (1990), p. 3.46.

A **continuous-wave (CW) receiver** is used in CW radar. A typical block diagram is shown in Fig. R93.

Thanks to the double heterodyne, the frequency of the output signal in this circuit does not depend on f_t or f_0. The frequency and stability of the transmitter and heterodyne do not affect accuracy of measurements f_Σ.

The converted frequency f_Σ is amplified in the difference-frequency amplifier and goes to the processor, which ensures separate measurement of the range f_r and doppler f_d frequency increments. As converters one uses frequency counters or frequency discriminators. Following distinguishing of the frequencies, f_r goes to the spectrum analyzers for range measurement, and f_d goes to narrowband filters for determination of target radio velocity. (See **RADAR, CW.**) *AIL*

Ref.: Vinitskiy (1961), pp. 258–266; Skolnik (1980), p. 85; Skolnik (1990), pp. 14.15–14.19.

correlation receiver (see **DETECTION, coherent**).

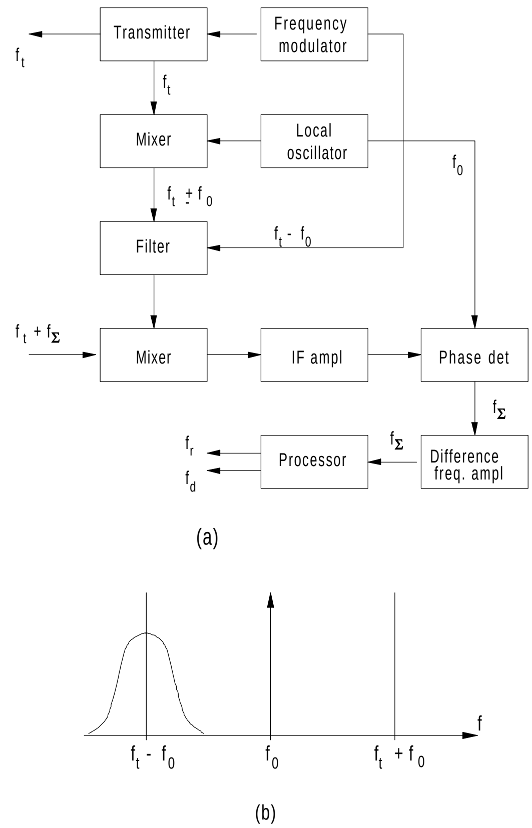


Figure R93 Block diagram (a) and frequency response, (b) of typical CW radar receiver.

A **crystal[-video] receiver** is the simplest receiver using a crystal detector as the basic component for reception. It is used mainly in EW applications because of a broad RF bandwidth, fast video output response, simplicity, and low cost. The main types of crystal detector receivers are the crystal video receiver and the log video amplifier receiver. This type of receiver is also used as an ESM receiver employing a crystal detector and high-gain video amplifier. Since there are no tuned circuits used, the selectivity of such a receiver is determined only by antenna. It has been widely used in ESM and ECM techniques because of its relative simplicity and low cost. *SAL*

Ref.: Wiegand (1991), p. 139; Johnston (1979), p. 57.

A **Dicke-fix receiver** is the receiver used as an ECCM technique against high-intensity wideband jamming. It consists of the broadband filter followed by the hard limiter that passes all wideband signals and limits them within a specified value, and the narrowband matched filter matched to the pulse. The gain in signal-to-jamming ratio is equal to B_w/B_n, where B_w is the bandwidth of broadband filter and B_n is the bandwidth of narrowband matched filter. *SAL*

Ref.: Neri (1991), p. 431.

receiver dynamic range (see **DYNAMIC RANGE**).

An **electronic support measures (ESM) receiver** is one used "to search for, intercept, locate, record, and analyze electromagnetic energy for the purpose of exploiting such radiations in support of military operations."

Ref.: IEEE (1993), p. 427.

A **frequency-measurement receiver** is an ESM receiver performing the measurement of the frequency of the input signal. These receivers can either have wide-open structures (called *wide-open receivers*) that instantaneously cover the whole ESM spectrum, or narrowband structure. In the latter case the entire spectrum coverage is ensured via sweeping. An instantaneous frequency measurement receiver is capable of measuring the frequency of each individual pulse in real time. The basic approach is to split the signal into two paths, one of which contains a time delay chosen to be a fraction of the period of the carrier frequency to be measured (Fig. R94). The relative phase θ of the two signals is then a measure of the frequency: $\theta = 2\pi f_0 t$.
Voltage outputs are generated proportional to the sine and cosine of the phase, and these can be digitized to produce a report:

$$E_1 = 2\sin(2\pi f_0 t) = 2\sin\theta$$

$$E_2 = 2\cos(2\pi f_0 t) = 2\cos\theta$$

A disadvantage of the IFM receiver is that it can only measure frequency over a limited band, determined by the delay used in one path. Parallel channels may be provided with different values of delay to cover the expected bands of input signals. Another disadvantage is that it cannot correctly measure frequency when two signals overlap at the input. In dense signal environments, it may prove necessary to use several parallel channels, each preceded by a bandpass filter, to reduce the probability of such overlap. *DKB, SAL*

Ref.: Wiley (1985), Ch. 9; Vaccaro (1993), Ch. 6; Tsui (1995), pp. 20, 371–376.

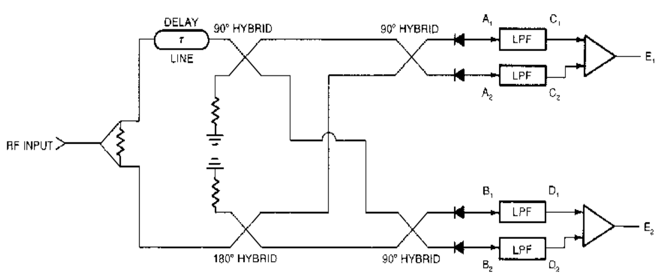


Figure R94 A typical basic IFM receiver (from Vaccaro, 1993, Fig. 6.1, p. 128).

The **receiver frequency response** describes the gain and phase of the receiver output, relative to the input (Fig. R95). The dependence of the absolute value of gain on frequency is called the amplitude-frequency response $K(f)$, and that of phase shift (between U_{in} and U_{out}) is the phase-frequency response $\phi(f)$. *SAL, AIL*

Ref.: Neri (1991), p. 287; Wiegand (91), p. 149.

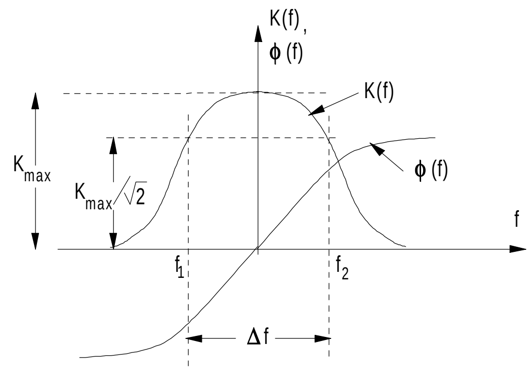


Figure R95 Receiver frequency response (after Druzhinin, 1988, Fig. 8.2, p. 346).

The **receiver front end** is the first stage of the receiver consisting of an RF amplifier or a bandpass filter. The noise figure of the receiver front end is defining considerably the noise figure of the entire receiver since the overall noise figure of the receiver consisting of N stages is given by the formula

$$F_r = \left(F_1 + \frac{F_2 - 1}{K_{p1}} + \frac{F_3 - 1}{K_{p1}K_{p2}} \dots + \frac{F_N - 1}{K_{p1}\dots K_{pN}}\right)$$

where F_i, K_{pi} are noise figures and power gain of the ith stage correspondingly.

Because early microwave RF amplifiers had a greater noise figure than when the mixer alone was employed at the input stage, they were not used in receiver front ends of early radars. When low-noise amplifiers providing a suitable noise figure were designed it became possible to use amplification at the receiver front end. The noise figure as a function of frequency for several receiver front ends used in radar applications is given under **low-noise receiver**. *SAL*

Ref.: Skolnik (1980), p. 351; Leonov (1988), p. 56.

The **receiver gain** characterizes the amplification properties of the receiver. One can distinguish power gain and voltage gain. The power gain K_P is the ratio of power at the output of the receiver P_{out} to the power at its input P_{in}:

$$K_p = \frac{P_{out}}{P_{in}}$$

Similarly one determines the voltage gain K_u:

$$K_u = \frac{U_{out}}{U_{in}}$$

The gain is a complex quantity because between the input and output voltage of the receiver there is a phase shift due to reactive components. (See **receiver frequency response**.)

The gain of modern radar receivers usually is $K_u = 10^6$ to 10^{10}; $K_P = 10^{12}$ to 10^{14}. It is frequently stated in decibels. *AIL*

Ref.: Davydov (1988), p. 119; Druzhinin (1967), pp. 345–346; Skolnik (1980), pp. 344–346.

Receiver gating is the "application of enabling or inhibiting pulses to one or more stages of a receiver only during part of a cycle of operation, when reception is either desired or undesired, respectively." (See also **GATING.**) *SAL*

Ref.: IEEE (1993), p. 549.

Heterodyne reception is "the process of reception in which a received high-frequency [RF] wave is combined in a nonlinear device with a locally generated wave, with the result that in the output there are frequencies equal to the sum and difference of the combining frequencies." If the resultant signal is superaudible and additional operations are required to reproduce the original wave, the process is called *superheterodyne reception*, and it is this process that is used in most radar receivers. It requires a local oscillator (LO) and frequency converter (mixer) to bring the input RF to intermediate frequency (IF) for further processing. Heterodyne reception offers excellent selectivity and sensitivity, but has disadvantages: (1) spurious reception channels (image response), (2) signal distortion (nonlinearity), and (3) spurious radiation from the LO. *SAL*

Ref.: IEEE (1993), p. 594.

Homodyne reception is "a system of reception by the aid of a locally generated voltage of carrier frequency." The received signal is thereby reduced directly to baseband, rather than to an intermediate frequency as in the case of heterodyne reception.

The homodyne principle is the technique used in a continuous-wave radar when the oscillator of the transmitter simultaneously serves as the local oscillator. The transmitter signal goes to the first LO of the receiver through the line of direct connection or through the controlled leakage link. This principle is used in some small CW radars such as police radars, traffic-light control radars, proximity fuse, and so forth. *IAM*

Ref.: IEEE (1993), p. 605; Skolnik (1970), p. 16.20; Schleher (1991), pp. 419–420.

Incoherent reception is the reception based on the use of incoherent signals. (See **INTEGRATION, noncoherent**.) *SAL*

An **inverse receiver** is one in which the conventional sequence of the stages features (wide IF, narrower doppler amplifier, final narrowband speedgate) is inverted and the narrow banding is placed at the early stages of the receiver to exclude interference from the subsequent stages (Fig. R96). It is used in radar seekers. *SAL*

Ref.: Skolnik (1990), p. 19.12.

An **inverse probability receiver** uses as a detection criterion the formation of the best estimate of the cause of the signal actually observed. The method of inverse probability requires the knowledge of *a priori* probabilities associated with hypotheses explaining the signal (e.g., noise alone or a combination of target signal and noise). Typically, in practice, it is impossible to specify *a priori* probabilities with good confidence, and in this case a likelihood-ratio receiver is usually employed. *SAL*

Ref.: Skolnik (1980), p. 377.

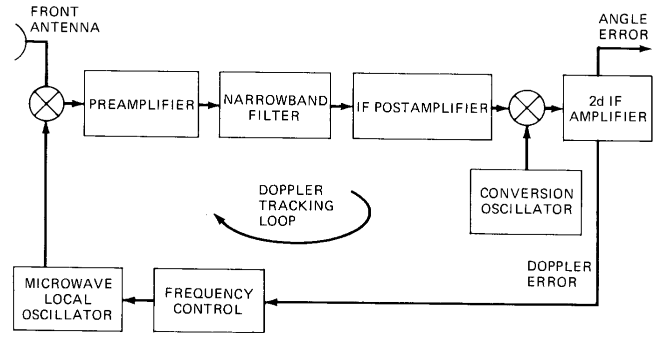


Figure R96 Block diagram of inverse receiver (from Skolnik, 1990, Fig. 19.8, p. 19.13, reprinted by permission of McGraw-Hill).

A **likelihood-ratio receiver** uses the likelihood ratio criterion for target detection. (See **DETECTION criteria**.) *SAL*

Ref.: Skolnik (1980), p. 377.

A **lin(ear)-log(arithmic) receiver** is one "having a linear amplitude response for small-amplitude signals and logarithmic response for large-amplitude signals." The law linking output V_{out} and input V_{in} voltages is given by a formula

$$V_{out} = a \log(1 + bV_{in})$$

where a, b are some constants. *SAL*

Ref.: IEEE (1993), p. 727; Skolnik (1980), p. 507.

A **logarithmic receiver** has logarithmic dependence of output signal V_{out} from input signal V_{in}

$$V_{out} = a \log bV_{in}$$

where a, b are some constants. The true logarithmic characteristic cannot be maintained down to zero input because in this case V_{out} goes to negative infinity. In practice, a logarithmic receiver has a law:

$$V_{out} = a \log(1 + bV_{in})$$

and turns into a lin-log receiver. *SAL*

Ref.: Skolnik (1980), p. 507.

A **log-FTC receiver** has a logarithmic input-output characteristic followed by a fast-time constant (FTC) circuit. It provides CFAR performance when input interference is described by a Rayleigh distribution. *SAL*

Ref.: Skolnik (1980), p. 506.

A **log(arithmic)-log(arithmic) receiver** is a logarithmic receiver in which the slope of the logarithmic characteristic progressively declines from noise level to +80 dB by a factor of 2 to 1 as opposed to the standard logarithmic receiver having logarithmic characteristic from –20 dB below noise level to +80 dB (lin-log receiver). In this receiver, the higher interference values are subject to greater suppression and it can be useful for reducing false alarm rate in non-Rayleigh clutter. *SAL*

Ref.: Skolnik (1980), p. 488.

A **logarithmic video amplifier (LVA) receiver** is a crystal-detector ESM receiver used to measure input RF amplitude over a wide dynamic range. It includes a crystal detector, a video logarithmic amplifier, and an analog-to-digital converter (Fig. R97). *SAL*

Ref.: Wiegand (1991), p. 139.

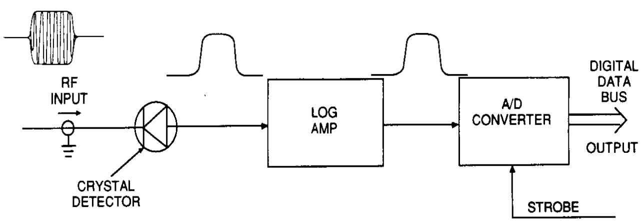


Figure R97 LVA receiver (from Wiegand, 1991, Fig. 4.1, p. 139).

A **low-noise receiver** is one employing an RF low-noise amplifier in its front end. The noise figure (noise factor) of a radar receiver establishes the receiver sensitivity, or the minimum detectable signal. Since the first stage of a multistage receiver generally establishes the noise figure (and the system noise temperature), provided the noise contribution from RF losses ahead of the receiver is small in comparison, a highly sensitive receiver can reduce the power-aperture requirements of a radar if this first stage consists of a low-noise amplifier or "front end." Figure R98 shows a number of available options. The parametric amplifier has the lowest noise figure of all the devices shown, and since thermal noise generation is proportional to temperature, even lower noise figures than those shown can be realized through cooling.

A cooled parametric amplifier front-end might make sense in a radar astronomy application, where the extra cost and complexity may be acceptable for one or a few radars so equipped, and where only the natural environment need be considered. For most radars, however, many additional factors must be considered, including receiver dynamic range, instantaneous bandwidth, tuning range, and phase and amplitude stability. For tactical air defense radars that must operate in a noise-jamming environment, the benefits of a low-noise receiver may seldom be exploitable, and the extra sensitivity that a low-noise receiver provides actually makes the radar more vulnerable to ECM. *SAL, PCH*

Ref: Skolnik (1980), pp. 351–353; Schleher (1986), p. 520.

Matched-filter reception uses a matched filter to maximize signal-to-noise ratio. (See **FILTER, matched.**) *SAL*

A **microscan receiver** is an ESM receiver capable of measuring the frequency of each individual pulse in real time, over a wide bandwidth. The basic approach is to downconvert the input signals with a frequency-swept local oscillator, and to use a dispersive delay line at IF to convert frequency into time delay relative to the start of the sweep (Fig. R99). The LO sweep rate can be made very high, permitting the receiver to scan the entire band within the pulse width of the expected signals. *DKB*

Ref.: Wiley (1985), pp. 238–246; Neri (1991), pp. 299–302.

monopulse receiver (see **MONOPULSE**).

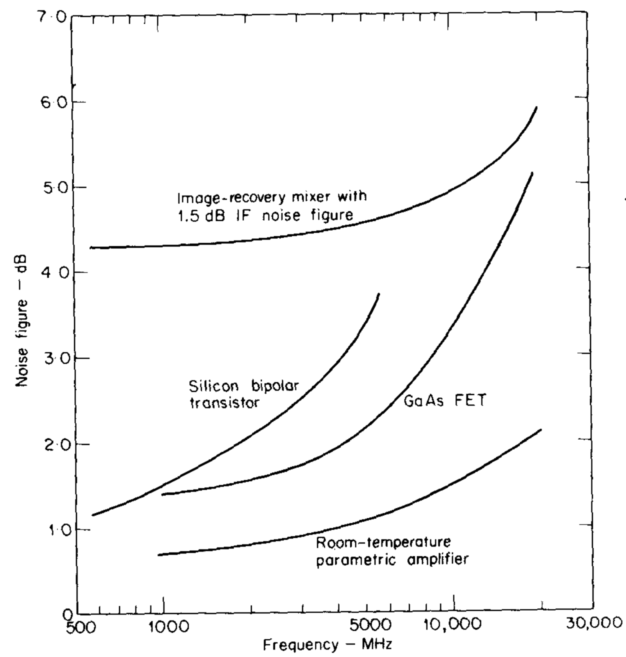


Figure R98 Noise figures for typical receiver front ends (from Skolnik, 1980, Fig. 9.4, p. 351, reprinted by permission of McGraw-Hill).

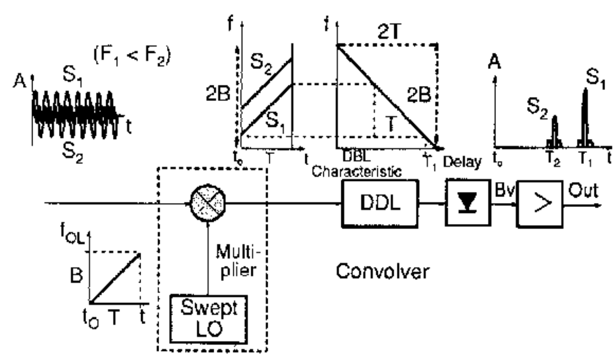


Figure R99 Microscan receiver (from Neri, 1991, Fig. 4.19, p. 301).

A **multichannel receiver** receives and processes several signals simultaneously (as opposed to the single-channel receiver, which processes only one signal at a time). Typically, the configuration of each channel is close to that of a single-channel receiver. Depending on the type of radar, the output video signals from each channel can be processed separately, summed, or combined by special circuits. An example of a multichannel receiver is a monopulse receiver. *AIL*

Ref.: Pereverzentsev (1981), p. 231.

Receiver operating characteristic (ROC) curves are "plots of probability of detection versus probability of false alarm for various input signal-to-noise ratios and detection threshold settings." They provide the same information plotted in figures under **DETECTION curves** but in different format. *DKB*

Ref.: IEEE (1993), p. 1,084.

The **optimum receiver** is one that ensures the maximum signal-to-noise ratio for the specified interference background (see **FILTER, optimum**). *AIL*

Ref.: DiFranco (1968), pp. 144–147, 185–199, 325–327; Dulevich (1978), pp. 68–86.

A **receiver protector** is a device that supplements the duplexer in diverting or absorbing transmitter power or pulses from external sources before they reach the sensitive circuits of a radar receiver. The basic function is provided by the duplexer, which passes most of the transmitter power to the antenna. Residual leakage of power toward the receiver must sometimes be controlled by inclusion of a receiver protector. The main protector devices are solid-state limiters (PIN or PN diodes), or ferrite limiters followed by diodes, placed in the receiver arm of the duplexer. Such limiters can also protect against high-power pulses from external sources, which are passed directly to the receiver from a circulator type of duplexer, and which may not be blocked by a gas-tube duplexer, especially if the radar to be protected is not operating. Another device used to protect the nonoperating radar is a waveguide shutter, which opens only when the radar power is on. (See also **DUPLEXER**.) *DKB*

Ref.: Skolnik (1980), pp. 359–366.

A **pulse receiver** is a superheterodyne receiver for the reception of pulse signals (Fig. R100). A signal arrives at the pulse receiver input from the antenna via the T/R switch, which allows use of one antenna for both reception and transmission. From the antenna switch the signal goes to the radio-frequency (RF) amplifier.

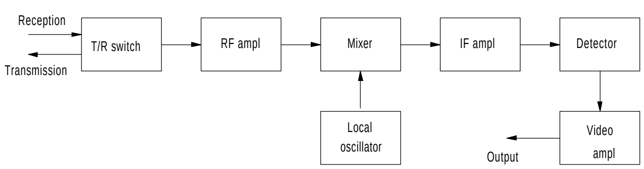


Figure R100 Typical receiver for pulsed radar.

In receivers of some radars the RF amplifier might not be present. Next the signal goes to the mixer in which RF is converted to an intermediate frequency (IF) by mixing with the local oscillator (LO) signal. After amplification at IF it is detected and sent to the video amplifier, from which the video signal goes to the output device, that may be a display screen or a computer. In many cases, the radar operates under conditions when the intrinsic noise of the receiver is not the dominant source of interference. Additional interference can noticeably reduce effectiveness of target detection. To eliminate the influence of interference, automatic gain control or a CFAR circuit is used, sometimes in the form of a log IF amplifier. The logarithmic amplifier provides maintenance of the false-alarm level in the receiver at an acceptable level with a change of interference intensity. Receivers that have these features are called *CFAR receivers*. *AIL*

Ref.: Skolnik (1970), p. 5.29; Leonov (1991), p. 12.

A **radar warning receiver (RWR)** is one designed to detect threatening emission and warn the crew of the aircraft of weapon systems guided by radars known to produce such signals. The technical parameters of the radars (e.g., frequency, pulsewidth) are stored in the memory of RWR and called *libraries*. When the receiver detects an emission corresponding to the type of data in the library, it identifies the threat and gives the warning signal. *SAL*

Ref.: Neri (1991), p. 274; Schleher (1986), p. 47.

radiometer receiver (see **RADIOMETER**).

An **unambiguous receiver** is a homing seeker receiver providing unambiguous (unfolded) doppler spectrum by offsetting the frequency of the rear reference before it is mixed with the front signal. It is used in radar seekers to eliminate clutter harmonic distortion, noise foldover, and approach-recede ambiguity (illuminator case). It is also termed the *offset video receiver*. *SAL*

Ref.: Skolnik (1990), p. 19.10.

RECIPROCITY. The principle of reciprocity is a concept applicable to most networks and transmission paths, stating that "if an electromagnetic force E at one point in a network produces a current I at a second point in the network, then the same voltage E acting at the second point will produce the same current I at the first point." In wave propagation, it is the property of "invariance of the complex amplitudes of the received signal to the interchange in location of transmitter and receiver."

With respect to antennas, it implies that the transmitting and receiving patterns for a given antenna, fed from the same port, are identical. Reciprocity is violated in actual antennas when a ferromagnetic device such as a circulator or ferrite phase shifter is included in the circuit, or when a nonlinear device (such as a gas-tube duplexer) is included. *PCH*

Ref.: IEEE (1993), p. 1085; Barton (1988), pp. 147,148.

RECIRCULATOR (see **INTEGRATOR, recirculator**).

REFLECTION is "a general term for the process by which the incident flux leaves a (stationary) surface of medium from the incident side, without change in frequency." Reflection of electromagnetic waves by the target is the basis for radar operation. Reflection can take either of two forms: (1) regular (specular, coherent) or (2) diffuse (noncoherent) reflection. Along with reflection from targets and clutter, radar is affected by reflection from the underlying surface, and effect described by the surface reflection coefficient in the pattern-propagation factor. (See **PROPAGATION**; **RADAR CROSS SECTION**; **SCATTERING.**) *SAL*

Ref.: IEEE (1993), p. 1,098; Barton (1988), pp. 288–296.

Creeping-wave reflection refers to energy returned to the radar by propagation along a surface on the target. *DKB*

Ref.: Knott (1993), pp. 8, 89, 246–250; Wehner (1987), p. 20.

Diffuse [noncoherent] reflection occurs when a radio wave is incident on an irregular surface. Using the Gaussian surface model, where the standard deviation of surface height is σ_h, the fraction of coherently reflected power (specular reflection) is given by

$$\rho_s^2 = \exp\left[-\left(\frac{4\pi\sigma_h}{\lambda}\sin\psi\right)^2\right]$$

The remaining fraction is the diffuse reflection power:

$$\rho_d^2 = 1 - \rho_s^2$$

The diffuse reflections form a Gaussian shaped lobe centered on the specular ray, having a width at the $1/e$ power contour equal to $4\beta_0$, where $\beta_0 = \sqrt{2}$ times the rms slope of the surface facets. As a result the diffuse reflections reaching a radar target will come from a region on the surface surrounding the so-called specular reflection point (which is actually the first Fresnel zone on the surface). The total field created at the target is equal to the sum of the direct ray, the specular component, and the diffuse component, the latter two being called multipath components. (See also **Rayleigh reflection criterion.**) *DKB*

Ref.: Beckmann (1963), p. 316; Barton (1975), pp. 351–352

The **reflection factor** "between two impedances, Z_1 and Z_2 is

$$\frac{(4Z_1Z_2)^{1/2}}{Z_1 + Z_2}$$

Physically, the reflection factor is the ratio of the current delivered to the load, whose impedance is not matched to the source, to the current that would be delivered to a load of matched impedance." *SAL*

Ref.: IEEE (1993), p. 1,099.

The **Fresnel reflection coefficient** describes the amplitude of a radio wave reflected from the earth's surface, relative to that of the incident wave. For vertical polarization, it is

$$\Gamma = \rho_0 e^{j\phi} = \frac{(\varepsilon_r - j60\sigma\lambda)\sin\psi - \sqrt{\varepsilon_r - j60\sigma\lambda - (\cos\psi)^2}}{(\varepsilon_r - j60\sigma\lambda)\sin\psi + \sqrt{\varepsilon_r - j60\sigma\lambda - (\cos\psi)^2}}$$

while for horizontal polarization it is

$$\Gamma = \rho_0 e^{j\phi} = \frac{\sin\psi - \sqrt{\varepsilon_r - j60\sigma\lambda - (\cos\psi)^2}}{\sin\psi + \sqrt{\varepsilon_r - j60\sigma\lambda - (\cos\psi)^2}}$$

where ρ_0 is the magnitude of the coefficient, ϕ is the phase, ε_r is the relative dielectric constant of the surface material, σ is its conductivity, λ is the wavelength, and ψ is the grazing angle of the incident wave.

Figures R101 and R102 show the coefficients for horizontal and vertical polarizations as a function of grazing angle for different surfaces. The only significant wavelength effect is for sea water at frequencies below 2 GHz ($\lambda > 0.15$m), where the factor $60\sigma\lambda > 90$ is comparable to $\varepsilon_r \approx 75$. The values of ε_r and σ used for these curves are listed in Table R23. *DKB*

Ref.: Kirby, R. C., in Fink (1982), pp. 18.53–18.54; Meeks (1982), pp. 13–20; Barton (1988), pp. 291–293.

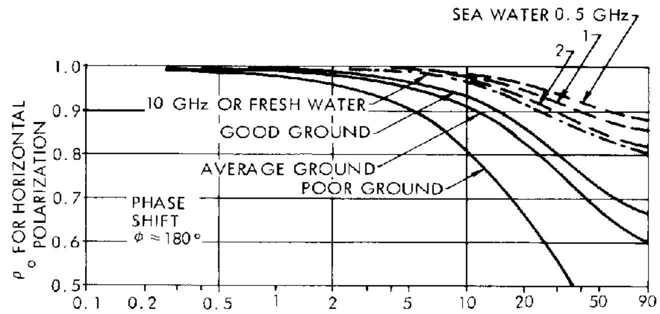


Figure R101 Fresnel reflection coefficients ρ_0 for horizontal polarization incident on different surfaces (from Barton, 1988, Fig. 6.2.2, p. 293).

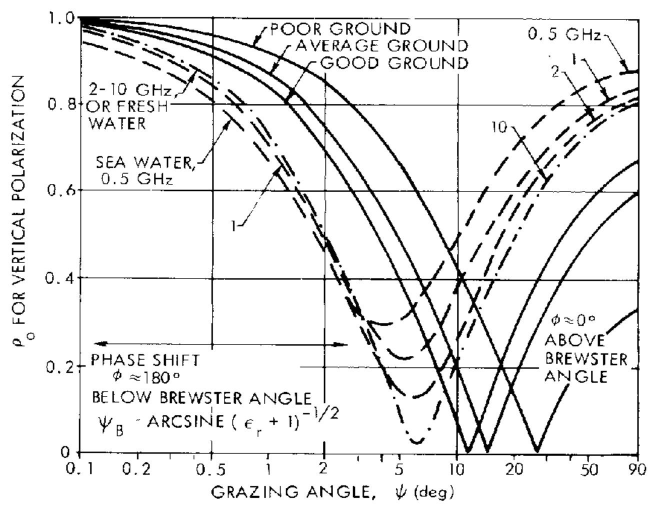


Figure R102 Fresnel reflection coefficients ρ_0 for vertical polarization incident on different surfaces (from Barton, 1988, Fig. 6.2.2, p. 293).

Table R23
Electrical Properties of Typical Surfaces

Material	ε_r	σ_e (mho/m)
Snow, ice	3	0.001
Poor soil (dry)	3	0.001
Average soil	15	0.005
Good soil (wet)	25	0.02
Fresh water (λ = 3m)	83	0.51
Fresh water (λ = 0.3m)	82.5	0.84
Fresh water (λ = 0.03m)	60	17
Salt water (λ = 3m)	76	4.1
Salt water (λ = 0.3m)	75	5
Salt water (λ = 0.03m)	55	20

Reflection lobes are the result of interference between the direct ray and specular reflections from the surface underlying the path. See **COVERAGE, radar**.

Multipath reflection refers to the combination of the direct ray with specular and diffuse reflections from the surface. (See **specular reflection**, **diffuse reflection**, **PROPAGATION**.)

The **Rayleigh reflection criterion** is the criterion determining whether the reflecting surface is smooth or rough. Surface irregularities (height deviation σ_h), wavelength λ, grazing angle ψ, of a wave, its polarization, and the dielectric constant of the ground affect the reflecting properties of the surface. As the criterion of roughness of the surface we usually use the inequality $\sigma_h > \lambda/(4 \sin \psi)$. A smooth surface is conditionally considered the one in which $\sigma_h < \lambda/(8 \sin \psi)$. Thus, as the limit of transition from a smooth surface to a rough surface we can assume

$$\psi = \arcsin(\lambda/6\sigma_h)$$

AIL

Ref.: Beckmann (1993), p. 9; Finkel'shteyn (1983), pp. 169–170.

reflection from a rough surface (see **diffuse reflection**).

Specular [regular] reflection is the coherent reflection of a radio wave from a smooth surface, either of the earth or of a target object. The laws of optics apply, stating that the angle of reflection is equal to the angle of incidence. When the surface is flat, the amplitude of the reflected wave, relative to the incident wave, is equal to the Fresnel reflection coefficient. The region of a flat surface contributing to the reflected wave is known as the first Fresnel zone, which surrounds the so-called specular reflection point. Curvature of the surface spreads the reflected wave, and only that portion of the surface that lies within $\lambda/4\pi$ from the wavefront as it first reaches the surface contributes to the backscattered wave.

For a partially rough surface, the specular reflection component can be described by the specular scattering coefficient

$$\rho_s^2 = \exp\left[-\left(\frac{4\pi\sigma_h}{\lambda} \cdot \sin\psi\right)^2\right]$$

where σ_h is the rms deviation from a smooth surface, λ is the wavelength, and ψ is the grazing angle. *DKB*

Ref.: Barton (1988), pp. 101–102.

Reflection from the spherical earth is calculated on the basis of a flat surface with modification in some cases resulting from the divergence of the reflected rays from the specular direction. The divergence factor D describing the reduction in field strength, relative to flat-earth reflection, is given by

$$D \approx \frac{1}{\sqrt{1 + \dfrac{2R_1R_2}{a(R_1 + R_2)\sin\psi}}}$$

where R_1 and R_2 are the ranges from radar to surface and from surface to target, $a = 6.5 \times 10^6$m is the radius of the earth, and ψ is the grazing angle.

The divergence factor appears in the equation for the pattern-propagation factor when propagation is governed by reflection-interference phenomena:

$$F = \left|f(\theta_t) + f(-\psi)\rho D\exp(-j\alpha)\right|$$

where $f(\theta_t)$ is the voltage gain at the target angle, $f(-\psi)$ is the gain at the depression angle to the reflection point, ρ is the magnitude, and α is the phase angle of the reflection coefficient. In many cases, especially with a low-sited radar and low-altitude target, propagation becomes dominated by diffraction phenomena before the reduction in D becomes effective, and application of the divergence factor in the equation for F leads to its being overestimated. *DKB*

Ref.: Kerr (1951), pp. 402–406; (1990), pp. 2.31–2.43; Barton (1988), pp. 290–294.

The **surface reflection coefficient** describes the strength of waves reflected from the underlying surface, usually that of the earth. The magnitude ρ of this coefficient, representing the ratio of voltage of the reflected wave to that of the incident wave, may be expressed as the product of three factors:

$$\rho = \rho_0\rho_s\rho_v$$

where ρ_0 is the magnitude of the Fresnel reflection coefficient for the surface material, ρ_s is the specular scattering factor, and ρ_v is a coefficient for vegetation absorption (the vegetation factor). The surface reflection coefficient is used in the equation for the pattern-propagation factor. *SAL*

Ref.: Barton (1988), p. 291.

REFLECTIVITY. The radar reflectivity of an object characterizes its ability to return part of the illuminating energy toward the receiver. Radar reflectivity numerically is characterized by the radar cross section (RCS) of a target, or the specific RCS of clutter. In radar applications, reflectivity of the targets is measured for several basic purposes. The main are basic RCS measurements (of ships, aircraft, missiles, military land vehicles, etc.); phenomenological measurements (determination of backscattering from land, sea, ice, precipitation, etc.); diagnostic measurements (high-resolution measurements to support the efforts in modeling and RCS reduction); and polarimetric measurements (measurements to support efforts in clutter suppression and target identification). (See also **CLUTTER**, **RADAR CROSS SECTION**.) *SAL*

Ref.: Currie (1989); Currie (1987), pp. 755–816.

Reflectivity calibration (see **CALIBRATION**).

The **(radar) reflectivity factor** Z for precipitation is defined as the sum of the sixth power of the diameters D of the scattering particles within a cubic meter of space, for volume clutter (e.g., rain) under the assumption of Rayleigh scattering:

$$Z = \sum_i D^6$$

For scattering from rain there is an empirical formula that is widely accepted:

$$Z = ar^b$$

where a and b are empirically determined constants, and r is the rainfall rate (see **CLUTTER**).

When the scattering is not Rayleigh, an equivalent radar reflectivity factor is defined as

$$Z_e = \frac{\lambda^4 \eta_v}{|K|\pi^5}$$

where $K = (\varepsilon - 1)/(\varepsilon + 2)$, ε is the dielectric constant of the scattering particles (usually $|K|^2$ is taken 0.93), and η_v is the backscatter cross section per unit volume. *SAL*

Ref.: Skolnik (1980), p. 500.

reflectivity measurement (see **RADAR CROSS-SECTION measurement**).

REFLECTOMETER. The reflectometer is an instrument for measuring the reflection factor of a microwave network.

Ref.: Gardiol (1984), pp. 332–337.

REFLECTOR. A reflector is "a device to redirect a flux from a source by a process of reflection." In radar applications the term is applied to various artificial reflectors and to reflectors (mirrors) of reflector antennas. *SAL*

Ref.: IEEE (1993), p. 1,099.

An **artificial reflector** is a target designed to achieve the specified backscattering characteristics (RCS, polarization characteristics, etc.). Typically, artificial reflectors have the simple geometric shapes. The widely used types of artificial reflectors are corner reflectors, Luneburg reflectors, and Van Atta reflectors. Since they do not contain an energy source, these reflectors are termed *passive reflectors*. *IAM*

Ref.: Boyd (1961), Ch. 18; Mishchenko (1966), p. 84; Johnson (1984), p. 17.24.

The **Bruderhedral reflector**, named after its developer, Joseph Bruder of the Georgia Institute of Technology, is a target for calibration of orthogonally polarized reflections. It is a dihedral, one surface of which is bent from a planar surface into a sector of a cylinder (Fig. R103). The reflection pattern is broader in the plane of the seam than for the planar dihedral. *DKB*

Ref.: Currie (1987), pp. 769–774.

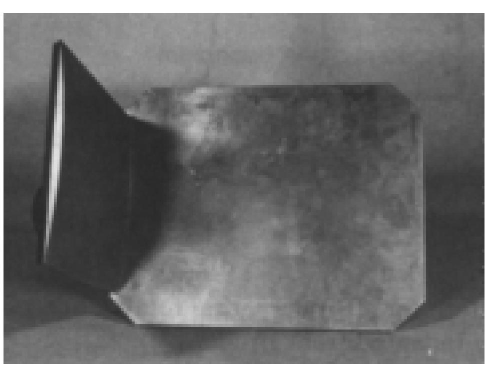


Figure 103 Bruderhedral reflector (from Currie, 1987, Fig. 17.9, p. 774).

A **conical reflector** is used as the main element of an antenna system (Fig. R104) consisting of three elements: a horn feed, a parabolic-arc subreflector, and the conical reflector. The feed wave with spherical wavefront passes from the subreflector to the conical reflector, from which it is reflected as a plane wave. This antenna is used when the complexity of a large doubly curved aperture is to be avoided. *SAL*

Ref.: Johnson (1993), p. 17.13.

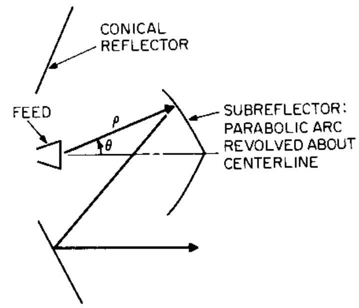


Figure 104 Conical reflector antenna (from Johnson, 1993, Fig. 17.13, p. 17.13).

A **corner reflector** consists of two or more metallic planes of arbitrary shape. The most common types of corner reflectors are dihedral reflectors with two planes and an arbitrary angle between them (a right angle for monostatic radar), biconical reflectors that are actually two-surface reflectors with a rib curved through the circumference, and trihedral reflectors containing three orthogonal planes, typically of the same shape (Fig. 105). Performance of corner reflectors is given in Table R24. The simplicity of manufacturing and the broad beamwidth of the scattering pattern determines the wide applications of corner reflectors as reference scatterers, passive beacons, target simulators, and so forth. *IAM*

Ref.: Hall (1947), pp. 324–331; Kobak (1975), p. 138.

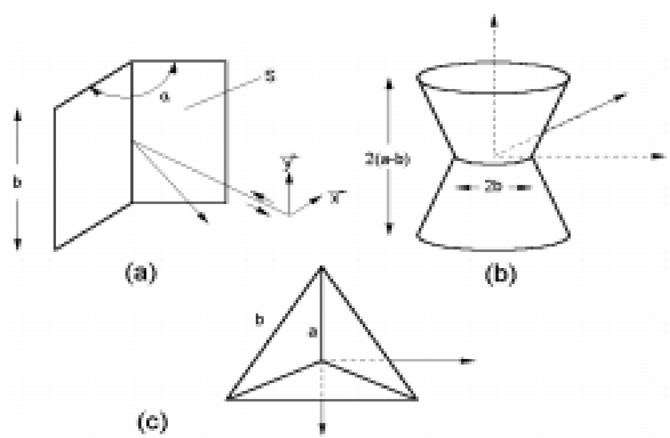


Figure R105 Corner reflectors: (a) dihedral, (b) biconical, (c) trihedral.

Table R24
Corner Reflector Performance

Corner reflector type	Maximum RCS	Beamwidth at a level of 3 dB
Biconical (b = 0)	$(4\pi a^3)/9\lambda$, $a \geq 10\lambda$	30° in vertical plane, circular in horizontal plane
Trihedral $b = \sqrt{2}\,a$	$(4\pi a^4)/3\lambda^2$	42° in vertical and horizontal planes

A **cylindrical reflector** is a lens reflector of cylindrical shape. It is made from a lossless dielectric for which the dielectric constant varies gradually along the radius according to either of two relationships:

(1) For the cylindrical Iton reflector (described in Russian literature), the dielectric constant varies from $\varepsilon = 1$ on the surface of the cylinder to $\varepsilon = \infty$ at its axis, according to the law

$$E(\rho) = \left(\frac{2a}{\rho} - 1\right)$$

where a is the cylinder radius and ρ is its radial coordinate. All rays, regardless of the incidence angle, propagate inside the lens along sections of the ellipses with a focus on the cylinder axis and return as a beam in the direction opposite to the direction of incidence.

(2) For the cylindrical Luneburg reflector, ε varies along the radius from $\varepsilon = 1$ on the surface to $\varepsilon = 2$ at the cylinder axis following the law

$$\varepsilon(\rho) = 2 - \left(\frac{\rho}{a}\right)^2$$

where a is the cylinder radius and ρ is the radical coordinate. This reflector is a cylindrical Luneburg lens and uses a metal coating a portion of the surface (Fig. R106).

The shape of the beam and the beamwidth at the 3-dB level in the plane going along the lens axis are analogous to those of the flat plate with dimension $l \times 2a$, l is cylinder length: $\theta_{0.5} = 25\lambda/l$, where λ is wavelength. Luneburg lenses are usually manufactured using multilayer configurations (10 to 25 layers). *IAM*

Ref.: Kobak (1975), p. 186.

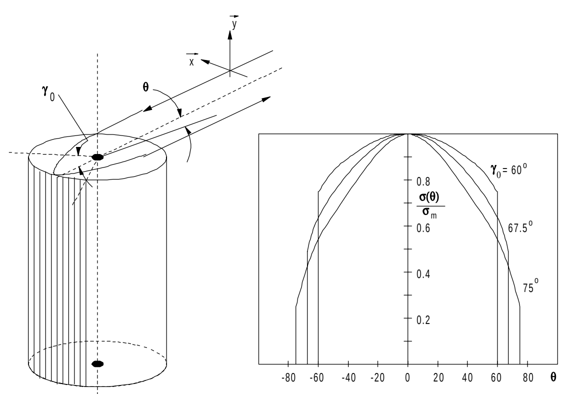


Figure R106 Cylindrical Luneburg reflector and its scattering pattern.

A **dielectric reflector** is based on a short-circuited dielectric antenna. The monostatic backscattering pattern within the beamwidth is proportional to the square of power antenna pattern for the dielectric antenna. The maximum monostatic RCS for this type of reflector is $G_m \approx 5.1L^2$, where L is the optimum length ($L \approx 4$ to 5λ). The dielectric reflector has a small transversal cross section (0.01 to 0.1 times the aperture area) and good capabilities for modulating and coding the scattered signal. This reflector is sometimes called a *polyrod reflector.* (See **ANTENNA, dielectric**.) *IAM*

Ref.: Kobak (1975), p. 204.

A **dipole reflector** is a passive device producing secondary reradiation from conductors. An example of such a dipole is the solid cylindrical conductor (Fig. R107). The relative scattering matrix for the cylindrical reflector is

$$M = \begin{bmatrix} \sigma_0 & 0 \\ 0 & 1 \end{bmatrix}$$

where σ_0 can be defined as

$$\sigma_0 = \frac{4\pi L^2 \cos\theta}{\pi^2 + 4\ln^2\left(\dfrac{0.178}{a\cos\theta}\right)}$$

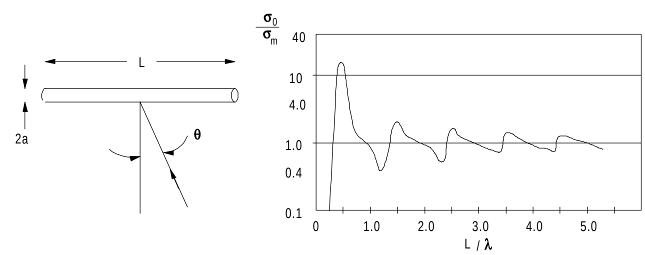


Figure R107 Dipole reflector and its RCS variation with length.

The maximum RCS ($\sigma_0 = 16.4\sigma_m$) corresponds to that of the half-wave dipole. Dipole reflectors are widely used as a source of passive jamming. (See **CHAFF**.) *IAM*

Ref.: King (1959), pp. 102–109; Popov (1980), p. 112; Kobak (1975), p. 200.

A **doubly curved reflector** is one with curvature in both coordinates throughout the aperture. Most common is the doubly curved parabolic reflector, which is often distorted to produce a cosecant-squared pattern by directing part of the energy upward, above the mainlobe (Fig. R108). *SAL*

Ref.: Skolnik (1980), p. 258; Barton (1988), p. 163; Johnson (1993), p. 17.14.

An **ellipsoidal reflector** is used as a subreflector in Gregorian antennas. *SAL*

Ref.: Johnson (1993), p. 17.33.

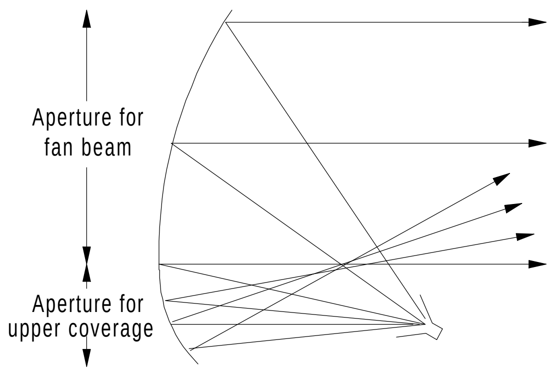


Figure R108 Shape of doubly curved reflector to produce cosecant-squared coverage pattern.

A **folding reflector** is one that has small dimensions in the folded condition, and is often used in the rescue devices. They can be made from the flat panels tied by the special fixing elements from the rubber or plastic materials. In the operational condition, it can have different desired shape: corner reflector with three edges, octahedron, and so forth. Inflated reflectors, used as false targets to penetrate through the antiballistic missile systems, can also be classified as folding reflectors. *IAM*

Ref.: U.S. patents 4119965, 1467481.

A **frequency-selective reflector** passes signals in a portion of the spectrum and reflects those in other portions. It consists of an array of closely spaced resonant elements (e.g., slots or dipoles). Slots transmit within a selected band, while dipoles reflect in that band. Waves with length large relative to the slots are almost totally reflected, while with dipoles they are transmitted. Such reflectors are also termed dichroic, or frequency-selective surfaces (FSS). (See **ABSORBER.**) *SAL*

Ref.: Johnson (1993), p. 17.30.

Group reflectors are groups of two, three, or more reflectors of one type (Fig. R109). They are used to imitate multilobe RCS diagrams as for real targets, or to obtain omnidirectional characteristics (e.g., for passive radar transponders). In the first case, usually a system of two reflectors called a dumbbell or a system of many reflectors are used. To obtain the omnidirectional pattern, the circular or spherical arrays of the reflectors with four or eight reflectors in a cluster can be used. *IAM*

Ref.: Kobak (1975), p. 214.

A **helispherical reflector** is a radio-transparent dielectric sphere with a polarization grid of wires or narrow strips put on its surface and reflecting a metal ring inside (Fig. R110). The wave going to the reflector along the equatorial plane is resolved into two components: one component is dissipated and another one (perpendicular to the grid wires) penetrates inside the sphere and after the reflection inside the sphere goes outside in the direction reverse to the direction of incidence. That is why the RCS diagram is circular in the equatorial plane. The maximum monostatic RCS for horizontal and vertical polarizations is $\sigma_m \approx r^4/\lambda^2$. *IAM*

Ref.: Kobak (1975), p. 211.

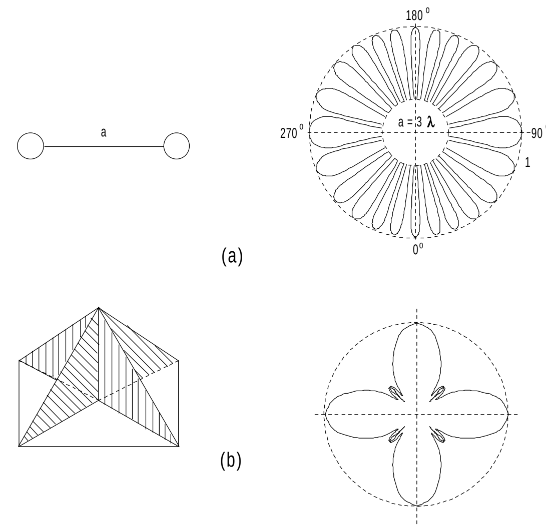


Figure R109 Group reflectors and their RCS patterns: (a) dumbbell; (b) cluster of four corner reflectors.

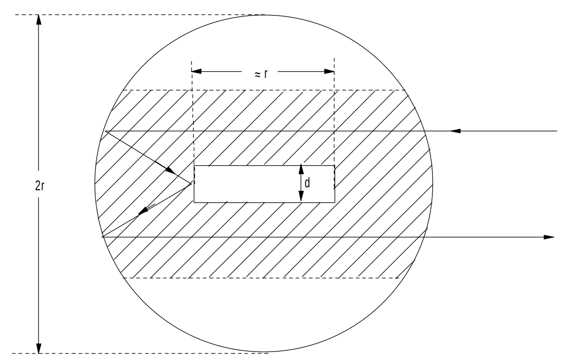


Figure R110 Helispherical reflector.

A **horn reflector** is a directional reflector having the shape of a horn (cylindrical, conical, etc.). In principle, it is a short-circuited horn antenna used as a reflector. For optimum antenna size and in the case when the incidence wave and the antenna have the same polarization, the maximum RCS is $\sigma_m \approx 5.1(S/\lambda)^2$, where S is the horn aperture area and λ is the wavelength. *IAM*

Ref.: Kobak (1975), p. 202.

An **hourglass reflector** is a parabolic reflector obtained by rotating a paraboloid about a vertical axis that is on the convex side of the arc. The feed system is typically a circular array. This type of reflector can give increased gain in the vertical plane. *SAL*

Ref.: Johnson (1993), p. 17.8.

An **isotropic reflector** is one having uniform RCS over all angles (4π sr). The sphere is the only shape capable of providing isotropic reflection. *DKB*

A **lens reflector** uses dielectric lenses of different sizes as its basic components. Typically, it may be classified as a cylindrical or a spherical reflector. A metal coating on the surface can be employed depending on the radial deviation of the dielectric constant. The scattering capabilities of lens reflectors are analogous to those of planar plates or corner reflectors (for cylindrical reflectors) or three-edged corner reflectors (for spherical reflectors). The main parameters of lens reflectors are given in Table R25. *IAM*

Ref.: Kobak (1975), pp. 186, 195.

Table R25
Properties of Lens Reflectors

Reflector	Maximum 3-dB beamwidth of the main beam
Cylindrical Luneburg lens	135° in the plane perpendicular to the axis, $(25\lambda/l)°$ in the axis plane
Cylindrical Iton reflector	Circular diagram in the plane perpendicular to the axis $(25\lambda/l)°$ in the axis plane
Spherical Luneburg lens	172°

A **mirror reflector** is a spherical or parabolic metallic mirror reflecting the incident energy in a specified direction. *IAM*

Ref.: Popov (1980), p. 135.

A **nonfolding reflector** is one with a rigid structure. It has sufficient mechanical rigidness and resistance to variation of environmental parameters. Corner reflectors, lens reflectors, horn reflectors, and some others are nonfolding reflectors. *IAM*

Ref.: Kobak (1975), pp. 138, 186, 200.

A **parabolic reflector** is one whose surface is generated by a parabolic arc (Fig. R111). The parabolic cylinder and parabolic torus are the types mainly used in radar. The paraboloidal antenna is formed by rotating the arc of a parabola about a line joining the vertex and the focal point and is the type most used for high-gain radar reflector antennas. (See **ANTENNA, reflector**.) *SAL*

Ref.: Johnson (1993), pp. 17.7, 17.17.

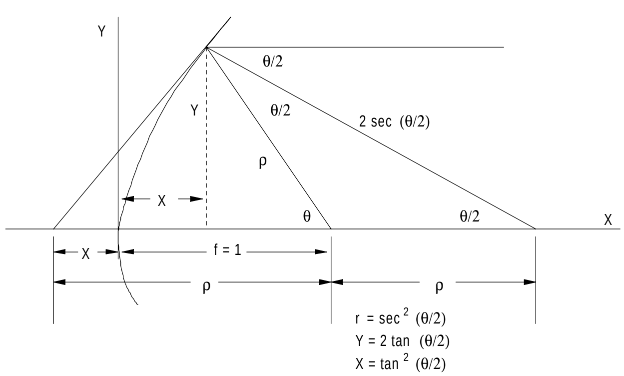


Figure R111 Parabolic reflector (from Johnson, 1984, Fig. 17.16, p. 17.17).

passive reflector (see **artificial reflector**).

polarization-sensitive reflector (see **transreflector**).

Simple-shape reflectors are reflectors such as spheres, plates, cylinders, and cones. The scattering characteristics of simple-shape reflectors are known and they are often used as the reference scatterers (See **RCS of simple shapes**.) *IAM*

Ref.: Kobak (1975), p. 101.

A **spherical reflector** is (1) a radar antenna reflector with spherical shape (see **ANTENNA, reflector**), or (2) a dielectric lens of a spherical shape. Of the latter, the most common is the spherical Luneburg lens, which is an inhomogeneous dielectric sphere large in comparison with the wavelength. It has metallic coating on a portion of the surface in the form of a "cap." The dielectric constant varies along the radius analogously to that of cylindrical Luneburg lens. The maximum RCS is $\sigma_m = 4\pi^3 a^4/\lambda^2$, where a is the lens radius. The maximum beamwidth of the reflected lobe at the –3-dB level $(0.5\sigma_m)$ is 172° for a cap angular dimension $2\gamma_0 = 172°$. *IAM*

Ref.: Kobak (1975), p. 195.

A **subreflector** is a "reflector other than the main reflector of a multireflector antenna." *SAL*

Ref.: IEEE (1993), p. 1,304.

A **top hat reflector** consists of a cylinder mounted on a circular plane, providing strong reflection over most of the hemisphere above the plane (Fig. R112). As a result of its broad pattern, its RCS is relatively small for a given physical size:

$$\sigma = \frac{2\pi a b^2}{\lambda(\cos\phi)^3}$$

DKB

Ref.: Currie (1987), p. 769.

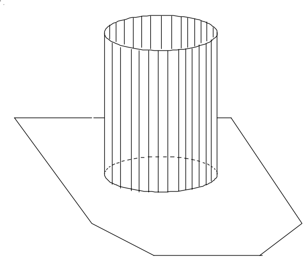


Figure R112 Top hat reflector (after Currie, 1989, Fig. 2.11, p. 45).

A **transreflector** is a wire grid designed to pass one linear polarization while reflecting the orthogonal polarization. It is used in polarization-twist reflector antennas. This reflector is also called a polarization-sensitive reflector (See **ANTENNA, Cassegorian.**)

Ref.: Skolnik (1980), p. 242; Johnson (1993), p. 17.28.

A **twistreflector** is a structure of wires or ridges mounted on a planar or parabolic reflector surface and designed to reflect a linear polarization rotated by 90° from the incident polarization. It is used in polarization-twist reflector antennas (See **ANTENNA, Cassegrainian.**) *DKB*

Ref.: Skolnik (1980), p. 242.

A **Van Atta reflector** consists of several pairs of the reflecting antennas connected by waveguide or feeder transmission lines. As a contrast to reflecting antennas (dipole, horn, or dielectric reflectors) that employ the short-circuit principle, the Van Atta reflector uses the principle of reradiation by an antenna of the signal that is received by another adjacent antenna. Typically, a planar array consisting of the several horizontal and vertical rows of dipoles spaced by $\lambda/4$ from a metallic plane is used. The plate is the reflecting screen. The RCS for the array consisting of $2N$ half-wave dipoles with the electrical length multiple to the wavelength is $\sigma_m = 4N^2\sigma_0$ in the normal direction where σ_0 is defined in the dipole reflector entry. The polarization characteristics of the Van Atta reflector are defined by the polarization characteristics of the employed antennas. The Van Atta reflector is also termed the Van Atta array. (See **ARRAY, Van Atta.**) *IAM*

Ref.: Kobak (1975), p. 206; Nebabin (1995), p. 17.

REFRACTION (see **ATMOSPHERE**).

ionospheric refraction (see **ERROR, propagation**).

refraction errors (see **ERROR, propagation**).

REFRACTIVITY. Refractivity N is the scaled-up value of the refractive index, n:

$$N = (n-1)\cdot 10^6$$

SAL

Ref.: Barton (1988), p. 304.

RELAY, radar. A radar relay is "an equipment for relaying the radar video and appropriate synchronizing signals to a remote location."

Ref.: IEEE (1993), p. 1052; Ridenour (1947), Chap 17.

RELIABILITY, radar. The general definition of reliability is "the ability of an item to perform a required function under stated conditions for a stated period of time." For radar it can be defined as the ability to perform assigned functions while retaining values of specified performance figures within assigned limits for each mode and condition of service, maintenance, repair, storage, and transportation. Typically, reliability is described by three components:

(1) Failure-free performance.

(2) Maintainability.

(3) Storability.

Failure-free performance is the ability to operate without failures for a specified operational time (see **FAILURE**). Maintainability is "the ability of an item, under stated conditions of use, to be retained or restored to a state in which it can perform its required functions, when maintenance is performed under stated conditions and using prescribed procedures and resources." Storability is the property of a radar to retain its operational conditions during storage and transportation. Reliability is an important radar performance characteristic, since it has significant effect on the effectiveness of radar operation and cost.

The main approaches to achieve reliability include

(1) The proper choice of technology and design efforts to avoid failures.

(2) Using a proper troubleshooting system (e.g., built-in test equipment) to determine and localize a failure as soon as possible after it happens.

The former approach includes the proper choice of reliable radar components and subsystems, and also the incorporation of the necessary redundancy in radar subsystems. The reliability of solid-state devices is usually much higher than that of vacuum-tube devices. As a result, a tube transmitter is usually one of the least reliable radar subsystems, and use of solid-state transmitters gives considerable improvement in radar reliability, permitting the manufacture of maintenance-free radars (at least in the sense that the permanent presence of maintenance personnel at the radar site is not required.). Multichannel solid-state transmitters and phased arrays offer fail-soft operation in case of failure of one or more modules (e.g., several amplifiers the transmitter or array can fail without degrading radar performance significantly).

Another common approach to reliability is to use redundancy. This approach is often used in radars where even a short-duration failure is critical (e.g., military or civil ATC radars). A common technique is to use two adequate receiver-signal-processor channels with automatic reconfiguration to place the standby channel in operation in case of failure. An example of high-reliability, solid-state radar with redundant receiver channels and built-in test equipment is the family of ATC radars including the Raytheon ASR-10SS and ASR-23SS. (See **RADAR, airport surveillance.**) *AIL, SAL*

Ref.: IEEE (1993), p. 1,117; Leonov (1991), p. 9; Fink (1982), Ch. 28.

REPAIR. Repair is "restoration or replacement of parts or components of material as necessitated by wear and tear, damage, failure of parts or the like in order to maintain the specific item of material in efficient operating condition." In radar applications it typically is a series of operations to restore serviceability or operability and to restore the operating life of a radar or of its component parts. The following types of repair are performed depending upon degree of aging, nature of damage, and the complexity and volume of operations: routine maintenance; medium repairs; and major overhaul. Routine maintenance is repairs made to restore radar operability and replacement or restoration of its individual units. It occurs following a failure and may be performed on site. These are unscheduled repairs. Repairs made to restore serviceability and for partial restoration of radar operating life with replacement or restoration of a limited number of component parts and inspection of the technical state of component parts are referred to as medium repairs. Repairs done to restore serviceability and for complete or nearly complete restoration of radar operating life with replacement or restoration of any of its parts, including fundamental ones, are referred to as major overhaul. Medium repairs and major overhauls are scheduled items. *AIL*

Ref.: IEEE (1993), p. 1,120; Aleksandrov (1976), p. 41.

REPEATER (see **JAMMING, repeater**).

RESISTANCE. Resistance is "that physical property of an element, device, branch, network, or system that is the factor by which the mean-square conduction current must be multiplied to give the corresponding power lost by dissipation as heat or as other permanent radiation or loss of electromagnetic energy from the circuit." The real part of impedance is sometimes termed resistance, but this is supplementary and not equivalent to the formal definition. *SAL*

Ref.: IEEE (1993), p.1,129.

Negative resistance is an element of an electrical circuit capable of generating alternating current due to conversion of feed source energy. Devices with negative resistance have a falling volt-ampere response sector, within the limits of which current and voltage increments are opposite in sign. Such devices include the tunnel diode, thyristor, and others. *AIL*

Ref.: Popov (1980), p. 273.

RESOLUTION. In radar, resolution is the ability to separate the signals from adjacent sources. The ability to distinguish one target from another is defined by a four-coordinate radar response, so angular, range, and doppler resolutions typically are distinguished. The common measure to consider two targets to be resolved in a particular dimension is when they are separated by a distance equal or more than half-power width of the radar response in this dimension: angle, range (time), or velocity (doppler frequency). A second target signal from the angle at the same frequency but with a different time delay can be resolved if the distance between the two targets along the line-of-sight is greater than the delay resolution (range resolution) of the radar.

Resolution, then, is determined by the relative response of the radar to targets separated from the target to which the radar is matched. The antenna and receiver are configured to match a target signal at a particular angle, delay, and frequency. The radar will respond with reduced gain to targets at other angles, delays, and frequencies. This response function can be expressed as a surface in a four-dimensional coordinate system. Because four-dimensional surfaces are impossible to plot, and because angle response is almost always independent of the delay-frequency response, these pairs of coordinates are usually treated separately.

In angle, the response function, $\chi(\theta,\phi)$, is simply the antenna pattern. It is found by measuring the system response as a function of the angle from the beam center. It has a mainlobe in the direction to which it is matched and side lobes extending over all visible space. Angular resolution (i.e., the mainlobe width in the θ and ϕ coordinates) is generally taken to be the distance between the 3-dB points off the pattern. The width, amplitude, and location of the lobes are determined by the aperture illumination (weighting) functions in the two coordinates across the aperture.

Because the matched antenna is uniformly illuminated, its response has relatively high sidelobes, which are objectionable in most radar applications. To avoid these, the antenna illumination may be mismatched slightly, with a resulting loss in gain and a broadening of the mainlobe.

Time delay and frequency can also be viewed as if they were two angular coordinates (i.e., as a two-dimensional surface which describes the filter response to a given signal as a function of the time delay t_d and the frequency shift f_d of the signal relative to some reference point). Points on the surface are found by recording the receiver output voltage while varying these two target coordinates. The response function, $\chi(t_d,, f_d)$, is given, for any filter and signal by

$$\chi(t_d, f_d) = \int_{-\infty}^{\infty} H(f)A(f-f_D)\exp[(j2\pi f)(dt)]df$$

or

$$\chi(t_d, f_d) = \int_{-\infty}^{\infty} h(t_D-t)a(t)\exp[(j2\pi f)(dt)]dt$$

where the functions $A(f)$ and $a(t)$, $H(f)$ and $h(t)$, are Fourier transform pairs describing the signal and filter, respectively.

These transform relationships, governing the time-frequency response function, are similar to those that relate far-field antenna patterns to aperture illumination. Hence data derived for waveforms can be applied to antennas, and vice versa, by interchanging analogous quantities between the two cases. There is a significant difference between waveform and antenna response functions and constraints, however, because the two waveform functions (in time delay and frequency) are dependent upon each other through the Fourier transform. The two antenna patterns (in θ and ϕ coordinates) are essentially independent of each other, depending on aperture illuminations in the two spatial coordinates x and y. Further differences arise from the two-way pattern and gain functions applicable to the antenna case.

The ability of the radar to form distinguishable response peaks on closely separated targets can be illustrated by plotting the composite waveform of two equal targets with varying if phase difference (Fig. R113). With the separation Δ equal to one-half the pulse-width τ_3 separate peaks appear only when the phase difference is near 180°. Separations between 1.0 and 1.25τ_3 give distinguishable peaks over most phase angles, and at 1.5τ_3 the two targets are always distinguishable (separately detectable and measurable with small interaction). For the smoothly shaped pulse with no sidelobes, shown in the figure, the targets become resolvable at separations between 1.25 and 1.5τ_3. In angle, with a scanning two-way antenna pattern, the corresponding resolution is between 0.9 and 1.1 times the half-power (one-way) beamwidth.

When a large target is present, greater separation is needed to resolve a smaller target. However, because of the steep skirts attainable in most radar response functions, the separation need seldom exceed twice the pulsewidth, or 1.5 times the beamwidth, unless the sidelobe response from the larger target obscures the smaller one. Thus the half-power width of the response function can provide a measure of resolution, subject to modification by a factor near unity to

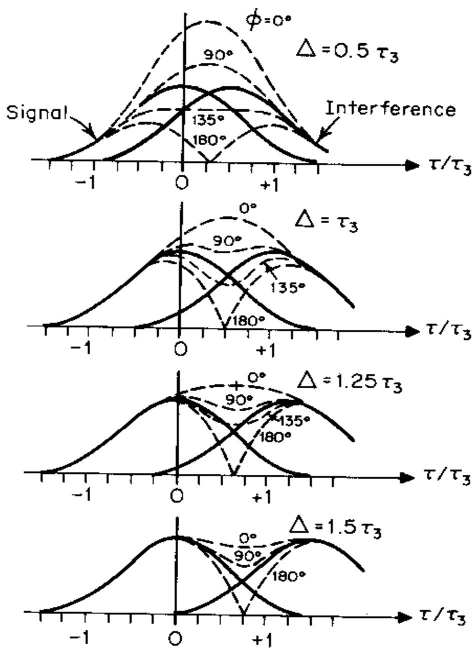


Figure R113 Resolution between adjacent targets with separations Δ and RF phase difference ϕ (from Barton, in Fink, 1982, Fig. 25-28, p. 25-23).

account for targets of different amplitudes. Other measures of resolution, such as Woodward's time and frequency resolution constants (extended, by analogy, to angle) can also be shown to be closely related to the half-power widths. (See also **AMBIGUITY FUNCTION.**)

Resolution, along with detection, measurement, and recognition is a fundamental radar procedure defining the performance of a radar. The requirements for resolution capability of the radar depend on the radar tasks (target detection, recognition, identification, etc.). A summary of resolution requirements for some tasks is given in Table R26. *DKB, SAL*

Ref.: Fink (1989), pp. 25.22–25.34; Morchin (1993), pp. 357–361.

Table R26
Resolution Required for Target Interpretation Tasks.

(Resolution in m)					
Object	**Detection**	**Recognition**	**Precise identification**	**Description**	**Technical intelligence**
Missile sites	3	0.5	0.6	0.3	0.08
Radar	3	0.9	0.3	0.15	0.04
Aircraft	4.5	1.5	0.9	0.15	0.03
Nuclear weapons	2.4	1.5	0.3	0.03	0.01
Components					
Surfaced submarines	30	6	1.5	0.9	0.03
Command					
Headquarters	3	1.5	0.9	00.15	0.03
Vehicles	1.5	0.6	0.3	0.05	0.03
(from Morchin, 1993, Table 17,2, p. 360)					

Angular resolution is "the minimum angular separation at which two equal targets at the same range can be distinguished from each other." Practically, two point targets at the same range are considered to be resolved if the angular distance between them exceeds the half-power beamwidth of the antenna. *SAL*

Ref.: IEEE (1993), p. 40; Long (1992), p. 101.

Cross-range [transverse] resolution in a surface search or mapping radar is the resolution in the direction of azimuth angle, measured in linear units (e.g., meters). For real-aperture radar it is defined as

$$W_{az} = R\theta_3$$

where R is the slant range, and θ_3 is azimuth 3-dB (half-power) beamwidth. *SAL*

Ref.: Brookner (1977), p. 232; Goj (1993), p. 2.

Doppler resolution is the minimum separation in targets radial velocities Δv_r that can be distinguished by the radar receiver. This resolution is fundamentally related to coherent integration time of radar return and can be evaluated as

$$\Delta V_r = \frac{\lambda}{2T}$$

where λ is the radar wavelength and T is the coherent integration time. *SAL*

Ref.: Wehner (1987), p. 39; Morchin (1993), p. 277.

The **resolution element** is a spatial and velocity region, contributing echo energy that can be separated from that of adjacent regions by action of the antenna or receiving system. In conventional radar, its dimensions are given by the beamwidths of the antenna, the transmitter pulsewidth, and the receiver bandwidth; its dimensions may be increased by the presence of spurious response regions (sidelobes), or decreased by use of specially coded transmissions and appropriate processing techniques. It is also called the *resolution cell. SAL*

Ref.: IEEE (1993), p. 1,133.

Frequency resolution is "the ability of the receiver or signal processing system to detect or measure separately two or more signals that differ only on frequency." The resolution typically is specified as the width of the frequency-response curve measured at specific value below the main response lobe (usually –3 dB). Frequency resolution is often used in the sense of doppler (frequency) resolution. *SAL*

Ref.: IEEE (1993), p. 530.

Ground range resolution is the resolution between targets located on the earth by airborne or spaceborne radars. The usual notation is W_r and is defined as

$$W_r = \frac{c\tau_n}{2\cos\psi}$$

where c is the speed of light; τ_n is the processed pulsewidth; and ψ is the grazing angle. *SAL*

Ref.: Goj (1993), p. 1.

The **pulse resolution volume** is the portion of space having the same cross section as the beam at distance R and a depth equal to the half-length of the pulse.The usual notation is V. For the 3-dB level, the approximate evaluation formula is

$$V = \frac{\pi}{4}(R\theta_a)(R\theta_e)(c\tau/2)$$

where θ_a and θ_e are the half-power beamwidths in azimuth and elevation, and τ is the pulsewidth. The factor $\pi/4$, while expressing the area within an elliptical beam, is more accurately represented by $1/L_p^2 = 0.565$ for beams approximating the Gaussian gain contour. *SAL*

Ref.: Sauvageot (1992), p. 35.

Range [radial, slant-range] resolution is "the ability to distinguish between two targets solely by the measurement of their ranges (distances from radar); usually expressed in terms of the minimum distance by which two targets of equal strength at the same azimuth and elevation angles must be spaced to be separately distinguishable." Resolution in the range domain Δ corresponds to the resolution τ_n in the time domain, and is set by the pulsewidth according to $\Delta R = c\tau_n/2$ (for pulse-compression waveforms τ_n is the pulsewidth after pulse compression).

The resolution associated with pulses having rectangular spectrum is shown in Fig. R114. Sometimes range resolution is also called slant-range resolution or radial resolution. *SAL*

Ref.: IEEE (1993), p. 1,067; Wehner (1987), p. 37; Morchin (1993), p. 277.

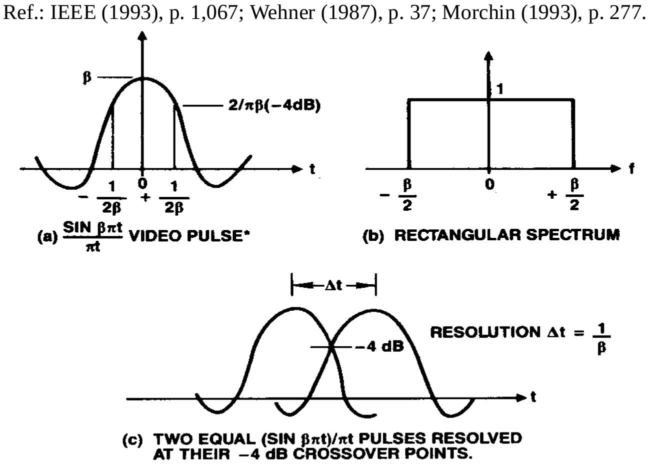


Figure R114 Pulse shape, spectrum, and resolution plot (from Wehner, 1987, Fig. 2.15, p. 38).

Range-ambiguity resolution is the process of achieving unambiguous range measurement in high- and medium-PRF ranging modes of pulse doppler radars. (See **AMBIGUITY, range.**) *SAL*

Ref.: Skolnik (1990), p. 17.21.

Combined **range-doppler resolution** is often discussed in terms of the combined time-frequency resolution of the radar waveform when processed with a matched filter. In time delay, the resolution cell width at the –3 dB (half-power) level is proportional to the reciprocal of the waveform: $\tau_3 = K_t/B$; while in frequency it is proportional to the reciprocal of the coherent processing time: $f_3 = K_f/t_f$. The values K_t and K_f are near unity, and we may write, for the area of the time-frequency resolution cell,

$$\tau_3 f_3 \approx \frac{1}{Bt_f}$$

For a single pulse, $t_f = \tau$, and the area is inversely proportional to the bandwidth-time product of the pulse. For simple (unmodulated) pulses, the product is near unity. That value may be increased without theoretical limits in pulse compression waveforms, the practical limit being set by the quality of the radar equipment and the task to be performed. However, for linear-FM pulse compression, the ambiguity ridge is a long, diagonal ridge in delay-doppler coordinates, and the area within the ellipse remains near unity, as is true for the unmodulated pulse.

In phase-coded pulse compression, elongation of the main response lobe can be avoided, and the two half-power widths may be reduced arbitrarily. When response beyond the half-power level is considered, however, the total volume under the ambiguity function is not changed by application of any modulation. Thus, the combined resolution volume under the ambiguity function remains unity for all waveforms, and the choice of modulation can only redistribute this volume in the time-frequency coordinates. (See also **AMBIGUITY FUNCTION.**) *DKB*

Ref.: Woodward (1953), Ch. 7; Rihaczek (1969), Ch. 5.

Resolution in SAR can be considered for two cases:

(1) Resolution in unfocused SAR is limited by the quadratic change in range as the radar platform moves past the points in the strip map. The maximum effective synthetic-aperture length to hold quadratic phase error to $\lambda/8$ is

$$L_{eff} = \sqrt{R\lambda}$$

where R is the range to the mapped point, and the resulting cross-range resolution is

$$\Delta_x = K'_\theta\sqrt{R\lambda}$$

(2) Resolution in focused SAR depends on the weighting applied to reduce cross-range sidelobes in the processed output, but when measured as the half-power width of the two-way processed response Δ_x it can approach half the width w of the real antenna used to illuminate the mapped surface:

$$\Delta_x = K'_\theta w$$

where $0.5 < K'_\theta < 1$ is the synthetic-aperture beamwidth constant. *DKB*

Ref.: Skolnik (1970), pp. 23.3–23.5.

***x*- and *y*-direction resolution** is the notation sometimes used for ground range resolution and azimuth resolution correspondingly achieved by airborne or spaceborne radars (Fig. R115). *SAL*

Ref.: Hovanessian (1988), p. 130.

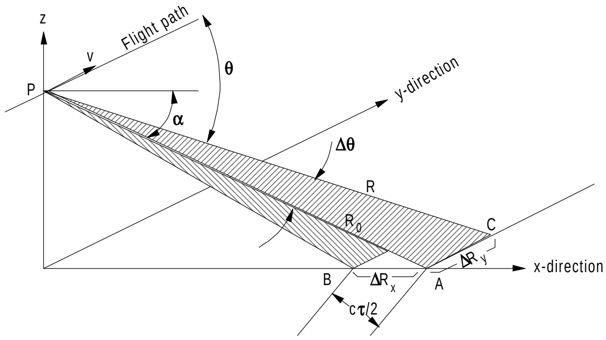


Figure R115 Flight path direction and x- and y-direction resolution (after Hovanessian, 1988, Fig. 3.1, p. 127).

RESONATOR, microwave. Microwave resonators are devices in which the energy of oscillations can be accumulated, and in which the phenomenon of resonance occurs.

At frequencies up to 1 GHz, circuits based on components with lumped parameters can be used as resonators, the main type at higher frequencies being a hollow resonator (cavity) with distributed parameters. Strip line resonators, coaxial resonator, hollow, three-dimensional resonators, and resonators based on surface acoustic waves are used. In terms of shape and design, cylindrical, toroidal and annular resonators, and resonators based on regular and irregular transmission lines are distinguished.

The accumulation of energy in resonators occurs in a time that is much longer than the period of oscillations. Electromagnetic oscillations in resonators exist in the form of standing or traveling waves. In the ideal standing wave resonator (without losses), many types of nonattenuating oscillations with natural frequencies ω_m exist. Each type is characterized by a specific field structure and periodic transition between electrical and magnetic energy. In a real resonator, the oscillations are supported through external energy, the energy of an oscillator, through connection elements, or energy that changes a parameter of the periodic resonator.

Along with large resonators (hollow resonators) capable of storing much energy, miniature resonators which are realized in integrated form can be designed for low-power devices.

Precise analytical determination of the natural frequencies and Q-factor is possible only for a few rudimentary shapes of resonators, but in the general case requires solution of a complex electrodynamic problem.

Resonators are components of oscillating systems of various types and are used as band-pass filters, and electrodynamic systems of high-power electronic microwave devices, for radar testing (see **echo box resonator**), and in other measurement applications. *IAM*

Ref.: Gassanov (1988), p. 46; Popov (1980), p. 372; Fink (1982), p. 9.12; Jordan (1985), pp.30.20–30.30.

An **annular resonator** is a toroidal resonator with a circular cross section in which a tube-like electron beam interacts with the resonator magnetic field. This configuration has greater current density, increasing efficiency when used in high-power microwave devices. Annular projecting pieces may also be present in a coaxial resonator, which differs from the annular resonator in the presence of a metallic rod along the principal axis of symmetry. Annular resonators are used as cells in multislot resonators, where they are located along the beam axis. *IAM*

Ref.: Grigor'ev (1984), p. 23.

A **coaxial resonator** is based on a section of coaxial transmission line. It is used in oscillating systems in the UHF to S-bands. Hollow coaxial resonators are used as annular resonators of high-power microwave electronic devices with excitation of the higher type of oscillations. The electron flow through such resonators is tubular in form. Such resonators have greater wave resistance than cylindrical ones due to the exclusion of energy stored in the near-axis region. *IAM*

Ref.: Grigor'ev (1984), p. 23; Popov (1980), p. 177.

A **crystal resonator** is a crystal plate that is placed between two metal electrodes. Operation of the resonator is based on the piezoelectric effect. Resonance sets in if a whole number of half-waves of mechanical oscillations fits along some axis of the crystal plate (in practice an odd number is realized, since only in this case are there charges of different signal at the electrodes). The resonant frequencies of the crystal resonator depend on the dimensions of the plate d and the speed of propagation of the flexible oscillations in the crystal (3 to 6×10^3 m/s) and are given by

$$f = (1.5 \text{ to } 3)\frac{(2n-1)}{d}$$

with d in mm.

Frequency range of crystal plates excited at a basic resonance (n = 1) is limited by their mechanical strength and as usual 20 MHz.

The resonant frequencies of modern crystal resonators operating at mechanical harmonics can be as high as 400 MHz.

Crystal resonators are used in crystal and frequency multiplier local oscillators. *IAM*

Ref.: Fink (1982), pp. 12.40–12.43; Petrov (1989), p. 157.

A **cylindrical resonator** has the shape of a cylinder. Hollow cylindrical resonators can be used in high-power microwave electronic devices based on the interaction of the electron flow with the electromagnetic field. Here the longitudinal component of the intensity of the electromagnetic field of the main type of oscillation E_{010} has its maximum on the axis of the resonator, where the electron flow passes. Usually more sophisticated toroidal resonators are used. Cylindrical resonators are used as echo-resonators and in miniature microwave devices, and cylindrical dielectric resonators are widely used. *IAM*

Ref.: Grigor'ev (1984), p. 22.

A **dielectric resonator** is a dielectric body of a specific shape, with a relative dielectric constant ε_r that is greater than that of the surrounding medium. We distinguish between open and screened resonator. The former are usually bodies of regular shape (a sphere, cylinder, torus), arranged in an unbounded isotropic medium. Screened dielectric resonator have a conducting surface close to the dielectric-air boundary, which partially or fully screens the dielectric resonator.

When materials with ε_r = 40 to 100 and more are used, the basic factor determining the resonance properties is the three-dimensional resonator in the dielectric model, and the conducting surfaces are viewed as a small disturbance. In a resonator with impermeable material with $\varepsilon_r < 20$, resonance phenomena are formed through partial reflection from the dielectric-air boundary and complete reflection from the conducting surfaces. Such resonators are called waveguide-dielectric resonators and constitute a waveguide with dielectric irregularity usually of simple shape. They have better temperature characteristics and smaller dielectric losses.

Dielectric resonators are widely used as miniature resonators. The advantages of dielectric resonators are their high quality (Q up to 5,000) and the temperature stability of the resonance frequency. A drawback is the impossibility of direct connection of semiconductor devices to the resonator, and the need to use for these purposes a wire turn or sections of microstrip and slot lines, and also the presence of parasitic passbands.

Dielectric resonators are used in microwave narrowband filters and in highly stable semiconductor oscillators. *IAM*

Ref.: Garault, Y., and Guillon, P., *IEEE Trans* ***MTT-25***, No 11, 1977, pp. 916–922.

An **echo box (resonator)** is a tunable three-dimensional resonator with high Q-factor, often used for general function testing of a radar. Usually cylindrical hollow resonators are used. Microwave oscillations from the transmitter are sent to the echo box through a directional coupling over an internal communication channel, or by using an external probe antenna placed in the emission field of the radar antenna.

The operating principle of the echo box consists of excitation of microwave oscillations in it upon emission of the probing radar pulse and emission of part of the oscillating energy of the echo box after the conclusion of the radar pulse. Here on the screen of the radar indicator a long blip will appear with a length along the range scale from zero to several kilometers.

The echo box is used to estimate the sensitivity of the receiver, the frequency spectrum of the emitted signal, and the average frequency of the transmitter, the time of restoration of sensitivity of the receiver, and a number of others. The Q-factor, with loading, of a three-dimensional resonator as an echo box is 90% of that without a load, and is between 40,000 and 200,000 in a frequency band of 1 to 25 GHz. *IAM*

Ref.: Skolnik (1970). p. 8.28.

The **eigenfrequency of a resonator** is the angular frequency at which the homogeneous Maxwell equations have nonzero solutions E_m and H_m. The eigenfrequency is associated with the eigenwavenumber, k_0, the dielectric constant (or permittivity), ε, and the permeability, μ, by the formula

$$\omega_0 = \frac{k_0}{\sqrt{\varepsilon\mu}}$$

The eigenfrequency ω_0 for a given resonator may assume an infinite number of values, which correspond to the different frequencies of the natural oscillations of the resonator.

The eigenfrequency is valid if there are no losses in the walls and medium filling the resonator. For example, a rectangular resonator with dimensions $a \times b \times$ c:

$$\omega_0 = \frac{\pi}{\sqrt{\varepsilon\mu}}\sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2 + \left(\frac{p}{c}\right)^2}$$

where m, n, and p are whole numbers. (See **eigenwavelength of resonator**.)

IAM

Ref.: Nikol'skiy (1964), p. 326.

The **eigenwavelength of resonator** is the wavelength of natural oscillations of a resonator, associated with the eigenwavenumber k_0 of the resonator by the formula

$$\lambda_0 = 2\pi/k_0.$$

For resonators of simple shape, the values of λ_0 are represented in Table R27. *IAM*

Ref.: Nikol'skiy (1964), p. 328.

Table R27
Eigenwavelengths of Resonators with Simple Shapes

Type of resonator	λ_0	Comment
Rectangular with dimensions $a \times b \times L$	$\dfrac{2}{\sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2 + \left(\frac{p}{L}\right)^2}}$	for an E-field and H-field
Cylindrical, length L, radius R	$\dfrac{2\pi}{\sqrt{\left(\frac{B_{nm}}{R}\right)^2 + \pi^2}}$	for H-field (H_{mnp})
	$\dfrac{2\pi}{\sqrt{\left(\frac{A_{nm}}{R}\right)^2 + \pi^2}}$	for E-field (E_{mnp})
Coaxial resonator of length L	$\dfrac{2L}{k}$, $k = 1, 2, \ldots$	for T-field (T_k)

B_{nm} are the roots of the equation $J_n(x) = 0$, where $J_n(x)$ is the Bessel function of the first order; A_{nm} are roots of equation; $J_n'(0) = 0$; m, n, and p can take the following values: for the E-field: $p = 0, 1, 2\ldots$; $m, n = 1$; $nm \neq 0$; for the H-field: $p = 1, 2, \ldots$; $m, n = 0, 1.$; $nm \neq 0$.

The **eigenwavenumber of a resonator** is a characteristic of a resonator associated with the intensity **E** of an electromagnetic wave field by the wave equation (following from Maxwell's equation)

$$\nabla^2\mathbf{E} + k_0^2\mathbf{E} = 0$$

and the boundary condition for the vector **E** on the surface S_0: $[\mathbf{E}, \mathbf{n}_0] = 0$, where $\mathbf{n}_0$ is the normal to S_0 and k_0 is the eigenwavenumber of the resonator. The values of k_0 form an infinite numerical sequence of real numbers in the absence of losses, each of which determines the type of field in the resonator.

Several different types of fields (degenerate oscillations) may correspond to one value of k_0, which is determined by the shape, dimensions and type of field of the resonator. *IAM*

Ref.: Nikol'skiy (1964), p. 324.

A **ferrite resonator** is a small volume (on the order of 0.2 to 2 mm^3) of monocrystal ferrite (usually a crystal of yttrium-iron-garnet, YIG) magnetized with a constant magnetic field to saturation and interacting with a variable microwave magnetic field whose frequency coincides with the frequency of ferromagnetic resonance. Ferrite resonators are marked by their high frequency selectivity, the possibility of tuning them to the resonance frequency over wide limits by changing the magnetizing field, by their different interaction with circular polarization fields of the opposite direction of rotation. The quality of a ferrite resonator reaches 10,000. A ferrite resonator is the only type of microwave resonator whose resonance frequency does not depend on dimensions.

These resonators are widely used in microwave filters. *IAM*

Ref.: Gassanov (1988).

A **hollow [empty] resonator** is a cavity bounded on all sides by a conducting surface and constituting an oscillating system with distributed parameters. Oscillations excited in a hollow resonator can be viewed as standing waves formed as a result of interference with multiple reflections from walls of the resonator. Usually a lower type of oscillation is used in hollow resonators. (See **eigenwavelength of resonator.**) Depending on the shape, we distinguish between toroidal, cylindrical, and coaxial annular resonators.

High Q-factor and the absence of radiation are the advantages of hollow resonators. The connection of the hollow resonator with an external circuit is made with a loop, slot, or hollow projection for interaction of the field with the flow of electrons in the cause of use of the resonator in oscillating systems of high-power microwave electronic devices. This resonator is often termed a cavity (resonator). *IAM*

Ref.: Popov (1980), p. 305.

An **irregular-line resonator** is one with a change in transverse section or electromagnetic properties of filling medium along the length of the line. The first are more widely used due to their higher precision. Depending on the nature of this change, we distinguish between resonators based on stepped and smoothly-irregular lines.

Stepped-irregular resonators usually are made on the basis of strip and coaxial lines, smoothly-irregular ones basically with striplines. Resonators of irregular transmission lines are used in tunable-band oscillating systems. *IAM*

Ref.: Voinov (1973), pp. 12, 79, 198.

Miniature resonators are those with volumes V an order of magnitude smaller than the volume V_0 of a half-wave metal resonator based on a standard rectangular waveguide (i.e., $n = V_0/V \geq 10$). The basic types of miniature resonators (Table R28) are microstrip, waveguide-slot, waveguide-strip, dielectric, and ferrite resonators.

Table R28
Types of Miniature Resonators

Type of resonator	*n*	*Q*
Microstrip	170	500
Waveguide-slot	10	3,000
Waveguide strip	10	1,000
Dielectric	40	5,000
Ferrite (permanent magnet)	9	3,000

Miniature resonators are used in hybrid integrated microwave circuits in the centimeter and millimeter wavebands. These resonators can be made directly at their location on a common substrate with other elements in the process of a single technological manufacturing cycle of the entire device. Modular microwave devices, for example, superheterodyne modules that perform the functions of preliminary selection and transposition of the spectrum of the input signal, are the most promising. *IAM*

Ref.: Shelamov, G. N., *Radiotekhnika*, no. 12, 1987, p. 7, in Russian.

A **periodic resonator** is one with a periodically changing parameter. The simplest example is an oscillating circuit with a periodically changing capacitance. In periodic resonators, at specific values of parameters, parametric excitation occurs. Here the oscillations in the resonator for which attenuation compensates for the external force changing the parameter at any moment are called the internal oscillations.

A specific feature of the parametric resonator is the internal oscillations are nonsinusoidal (so-called Hill functions), and the properties of the resonator are described not by individual eigenfrequencies ω_0 but by the eigenfrequency functions $\omega_0(t)$. In contrast with other linear oscillating systems, the periodic resonator has higher phase sensitivity. This is manifested in the fact that with a periodic change in detuning, it exerts a different action for different initial phases of the external force.

The use of a parametric resonator makes it possible to raise the selectivity of the receiver by establishing a specific connection between the modulating law of the oscillator and the modulating law of the corresponding parameter of the filter. *IAM*

Ref.: Vinitskiy (1961), pp. 99, 104.

The **resonator Q-factor** is a characteristic determined by the ratio of the energy reserve W_0 in the resonator to the loss L_n:

$$Q = \omega_0 W_0 / L_n,$$

where ω_0 is the eigenfrequency of the resonator. Losses in the resonator are due to two basic causes: losses in the metal components and dielectric medium of the resonator L_0 and losses as a result of radiation through the connecting elements L_c. In accordance with this, we distinguish between the natural Q_0 and the external Q_c:

$$\frac{1}{Q} = \frac{1}{Q_0} + \frac{1}{Q_c}; \quad Q_0 = \frac{\omega_0 W_0}{L_0}, \quad Q_c = \frac{\omega_0 W_0}{L_c}$$

The quality Q is also called the loaded quality. A component of quality due to losses in the dielectric is determined by the properties of the dielectric and does not depend on the frequency of oscillations or the dimensions of the resonator. A component that is done to losses in the metal, for a given type of oscillations, increases with an increase in dimensions proportionally to the square root of the eigenwavelength, and with excitation of higher types of oscillations in a resonator of fixed dimensions, increases as the square root of the eigenfrequency of the resonator.

The quality with a low value of losses is approximately equal to the ratio of the eigenfrequency of the resonator to half the width of the resonance curve the half-power level. (See also **Q(-FACTOR)**.) *IAM*

Ref.: Nikol'skiy (1964); Grigor'ev (1984).

A **reflective-obstacle resonator** is a transmission resonator formed by a pair of separated reflecting obstacles (irregularities) in a transmission line (or waveguide). The irregularities can be waveguide irises or other components with an analogous placement scheme. The exchange of energy between the resonator and conducting transmission lines, and consequently the passband of such a resonator depends on the reactive conductivity of the irregularities.

Resonators based on reflective obstacles are often used in passband microwave filters. *IAM*

Ref.: Sazonov (1988), p. 132.

A **stripline resonator** is one made of strip transmission line of any type: microstrip, slot, asymmetrical, symmetrical, etc. Determination of the eigenwavelength λ_r of the resonator is done using the Oliner model, which has a uniform dielectric filling and geometrical dimensions whose values are found from the equation of full energy of the field of the resonator and its model (Fig. R116).

For a resonator based on a asymmetrical line, λ_r for a wave E_{m0n} is determined from the formula:

$$\lambda_r = \frac{2\sqrt{\varepsilon_{ef}}}{\sqrt{\left(\frac{m}{a_{ef}}\right)^2 + \left(\frac{n}{L_{eff}}\right)^2}}$$

where m,n are the number of half-waves on the width (along the x-axis) and the length (along the z-axis) respectively. The lower types of oscillations in such a resonator are E_{001} and E_{101}.

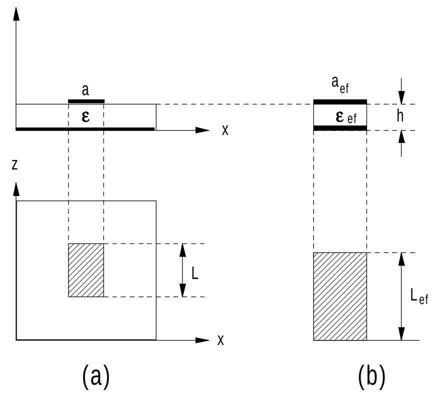


Figure R116 Resonators: (a) based on asymmetrical microstrip line; and (b) Oliner model (after Veselov, 1988, Fig. 2.17, p. 48).

Structurally the resonator can be made short-circuited or open at the end. The former has lower losses to radiation, but their realization is not always convenient from the technological standpoint. Besides rectangular resonators, other variants of strip resonators are used: round, elliptical, annular round and rectangular, resonators based on a slot line, and three-dimensional strip resonators. (See **waveguide-strip resonator**.)

Rectangular and round printed resonators are widely used as radiating elements of antenna arrays. *IAM*

Ref.: Veselov (1988), p. 47; Gassanov (1988), p. 48.

A **surface-acoustic-wave (SAW) resonator** is one which uses serial or parallel resonance between electrical leads of an opposing-pin converter of a SAW device. Resonance is possible when there is a converter with a high number of electrodes and strong electromagnetic connection.

SAW resonators can be made with high precision and are marked by high quality (Q-factor above 10^4), which cannot be obtained in a band lower than 150 MHz in conventional resonators.

SAW resonators can be used in narrowband filters and high-stable oscillators in the UHF band, analogous to crystal oscillators at frequencies of 30 MHz. *IAM*

Ref.: Kouamada, Y., et. al., *Proc. IEE* ***615***, 1976.

A **toroidal waveguide resonator** is one having a cylindrical sheath and hollow projecting piece inside it along the axis of the cylinder. Such hollow resonators are used in high-power microwave electronic devices. The projection (transit tube) increases the interaction factor of the electron flow with the electromagnetic field in comparison with a cylindrical resonator. The face of the projection can be covered with an electron-transparent grid that is opaque for the field, or left open. The shape of the projection can have a more complex structure.

Certain types of dielectric resonators have a toroidal shape. *IAM*

Ref.: Grigor'ev (1984), p. 21.

A **waveguide slot resonator** has one or several longitudinally oriented waveguides in the E-plane made of metal-dielectric plates with openings of a specific shape. Regardless of the type, a waveguide-slot resonator contains two sections of beyond-cutoff waveguides at the input and output, and between there is a resonance region in the form of a short-circuited or opened waveguide-slot line, in which propagation of one or both waves is possible.

Waveguide-slot resonators are used as miniature resonators in various types of microwave frequency-selective devices thanks to the extreme simplicity and planar nature of the design, the efficiency of manufacture, and the convenience of assembly of semiconductor device. *IAM*

Ref.: Lerev, A. M., and Shelamov, G. N., *Radiotekhnika*, no. 4, 1986, p. 7, in Russian.

A **waveguide-strip resonator** uses one or two vertically oriented metal strips in a waveguide. A strip resonator occupies a position in the longitudinal E-plane of the waveguide or in a transverse plane of the waveguide. Depending on the correlation of dimensions of the metal strip and the waveguide, the resonance frequencies of the waveguide-strip resonator are at regular or beyond-cutoff frequency ranges of the waveguide, which corresponds to realization of a bandstop or bandpass filter.

Compared with waveguide-slot resonators, waveguide-strip resonators make it possible to implement filters with greater frequency-selective capabilities. *IAM*

Ref.: Shelamov, G. N., *Radiotekhnika*, no. 10, 1986, p. 75, in Russian.

RESPONDER, radar (see **TRANSPONDER**).

RESPONSE

Amplitude-frequency response is the dependence of the gain modulus on the frequency. *AIL*

Ref.: Popov (1980), p. 28.

Angle-sensing response is the dependence of an output voltage of a direction finder on the angle of arrival. *AIL*

Ref.: Vasin (1977), p. 32.

Discriminator response is the dependence of the tracker output voltage on the mismatch error. As an example, a monopulse discriminator response in an x-plane is shown in Fig. R117, where u_x is the output voltage, and θ_x is the null axis offset angle. *AIL*

Ref.: Fel'dman (1978), pp. 197–199.

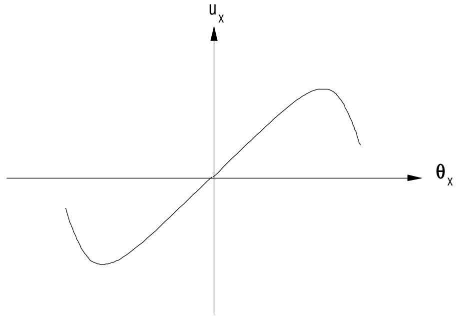


Figure R117 Discriminator response of monopulse system.

impulse response (see **FILTER characteristics**).

Phase-frequency response is the dependence of a phase shift between sinusoidal oscillations at the input and output of the linear circuit on the frequency of the input signal. *AIL*

Ref.: Popov (1980), p. 454.

RETURN LOSS is the term used to describe transmission line performance, where it is "the measure of the dissimilarity between two impedances, being equal to the number of decibels that corresponds to the scalar value of the reflection coefficient, Γ, and hence being expressed by the following formula:

$$20\log\left|\frac{Z_1 + Z_2}{Z_1 - Z_2}\right| \text{ decibel}$$

where Z_1 and Z_2 = the two impedances." *SAL*

Ref.: IEEE (1993), p. 1,143.

RING

Ring around is undesired self-oscillation caused by the recirculation of amplified noise energy. It is defined for transponders as "(1) the undesired triggering of a transponder by its own transmitter; (2) the triggering of a transponder at all bearings, causing a ring-type presentation ... on a plan-position indicator." Ring around can also be a problem in some repeater jammers if sufficient transmitter-to-receiver antenna isolation is not available. Some form of physical shielding, in addition to spatial separation between antennas, is often necessary to realize the required degree of isolation. *SAL*

Ref.: IEEE (1993), p. 1,148; Schleher (1986), p. 139.

Ring time is "the time during which the indicated output of an echo box remains above a specified signal-to-noise ratio." This time is a measure of radar performance. *SAL*

Ref.: IEEE (1993), p. 1,149.

S

SAMPLE, SAMPLING. Sampling is the process of converting an analog waveform into a discrete waveform by taking the analog values (samples) at a series of points in time (Fig. S1). This process is called sampling in time, or time-domain sampling, and it represents discretization of some input function of time. Any discretization of a continuous

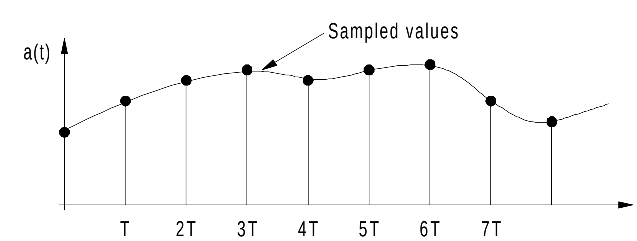


Figure S1 Sampling in time (after Barton, 1969, Fig. 7.1, p. 184).

function (e.g., sampling of a target by a single pulse or a pulse train) results in a periodic spectrum of the discrete waveform, at intervals $\omega = 2\pi/T$, where T is the time between successive samples, known as the sampling interval (Fig. 2).

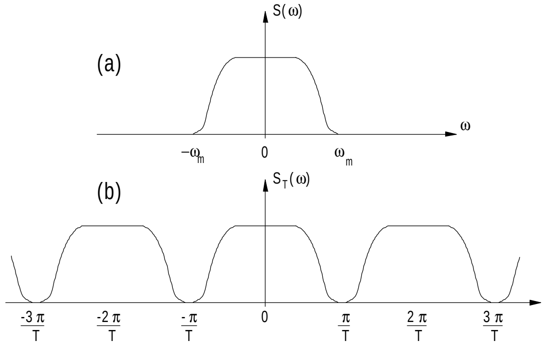


Figure 2 Spectra before and after sampling: (a) analog signal spectrum, (b) discrete waveform spectrum for $T = \pi/\omega_m$.

The optimum sampling interval is defined by the Nyquist sampling theorem:

$$T \leq \frac{\pi}{\omega_m} \quad (1)$$

where ω_m is the maximum frequency of the analog waveform spectrum. In this case the periodic spectrum components do not overlap and can be separated by filtering. The sampling rate (or sampling frequency), $f_s = 1/T$, corresponding to (1) is called the Nyquist rate, and sampling at this rate is sometimes called Nyquist sampling. A sampling interval $T > \pi/\omega_m$ results in overlapping of spectral components and distortions in the sampled data. (See **ALIASING**.) If the sampled signal is a random function of time, dependent samples are obtained for $T < t_c$, and independent samples for $T > t_c$, where t_c is the correlation time of the signal.

Frequency-domain sampling of the waveform spectrum is analogous to time-domain sampling. This approach is used in ultrawideband waveforms for high-resolution radars to decrease the number of samples required for unambiguous sampling. The term *space sampling* is sometimes used to describe the sampling of a signal received by an array antenna.

In radar circuits, sampling is usually performed by analog-to-digital converters. *SAL*

Ref.: Woodward (1953), p. 31; Barton (1969), p. 183; Wehner (1987), pp. 77–81.

A **complex sample** is an in-phase and quadrature pair of real samples of the radar waveform taken simultaneously. The radar signal can be unambiguously determined (with the accuracy up to unknown phase) by Δf complex samples per second, where Δf is the bandwidth of the signal outside which its spectral components have values that can be neglected (corresponding to ω_m/π in Fig. S2). *SAL*

Ref.: Wehner (1987), p. 78.

Subaperture sampling is a technique of jamming cancellation employed in adaptive phased arrays making it possible to avoid cancellation of the desired signal and to restore the jamming cancellation when correlation exists between the desired signal and interference. The approach is based on combining the measurements from overlapping subarrays and it works if the number of array elements at least twice the number of sources incident on the array (for uncorrelated sources N–1 jammers can be suppressed for N element array). This technique is also called *spatial smoothing*. *SAL*

Ref.: Farina (1992), p. 270.

SCAN, SCANNING. Scanning is "a programmed motion given to the major lobe of an antenna for the purpose of searching a larger angular region than can be covered with a single direction of the beam, or for measuring the angular location of a target; also the analogous process using range gates or frequency-domain filters." There are two basic ways of classifying scanning methods:

(1) From the viewpoint of principles of beam steering (another term for scanning), the methods are described as mechanical, electromechanical, or electronic.

(2) From the viewpoint of the type of beam motion introduced to scan a volume, the methods are described as helical scan, raster scan, Palmer scan, *n*-bar scan, and so forth.

For fan-beam radars a simple rotation in azimuth is sufficient. For pencil-beam radars, especially those used in fire control, a more complex type of scan is required to cover the volume around the designated target position. Figure S3 illustrates some of the more common scan types.

Figure S3(a) represents a helical scan pattern, so called because the center of the beam traces out a helix. It is shown here scanning in the elevation plane, but the orientation could just as easily be in azimuth plane. Figure S3(b) is a Palmer scan pattern, which, with the n-bar scan or raster scan shown in Fig. S3(c), is often associated with airborne fire control (or airborne intercept) radars. The nodding scan pattern shown in Fig. S3(d) is another, less popular option used to cover a search sector that is narrow in the elevation dimension. Other scanning techniques, particularly applicable to target tracking radars, include the conical scan and the spiral scan. In the former method, a nutating beam forms a cone in space that is designed to include a restricted volume of coverage based

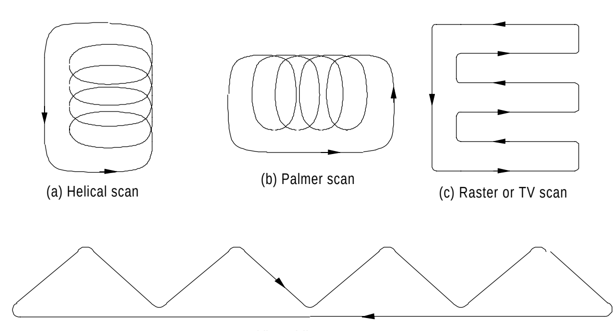


Figure S3 Common types of antenna search patterns (after Hovanessian, 1984, Fig. 2-10, p. 25).

upon a target designation from another (search) radar. In the case of the spiral scan, also implemented in response to a target designation by a separate search radar, the uncertainty volume is searched by spiraling the beam out from the designated coordinates to the limit of required coverage. In all of the single-beam scanning techniques described, antenna pattern overlap, generally at the antenna 3-dB points, is a requirement to avoid coverage gaps. *PCH*

Hovanessian (1984), pp. 25–27.

Conical scan is a form of sequential lobing in which the direction of maximum radiation generates a cone whose vertex angle is of the order of the antenna half-power beamwidth (see **ANGLE, squint**; **TRACKING, conical-scan**).

Electromechanical scanning moves the beam within some limited angle interval by motion of antenna structures relative to each other. It is typically implemented in reflector or lens antennas through one of the following means:

(1) Motion of the antenna feed or its phase center about the focal point of the reflector or lens.

(2) Motion of a subreflector in a multireflector antenna.

(3) Motion of the main reflector relative to a fixed feed position.

The first method is the most common (See **SCANNER.**) The last method is more difficult, due to the inertia of the reflector, and is seldom used.

The sector covered by electromechanical scanning is restricted by degradation of the antenna pattern. When feed motion is used, the gain of a paraboloid antenna with f/D = 0.25 is reduced by about 1 dB when the beam is scanned three beamwidths from the axis, and sidelobes increase significantly. While other types of scanner can achieve broader scan sectors, modern radars favor the use of electronically scanned arrays to preserve performance as the beam is scanned. *SAL*

Ref.: Skolnik (1980), p. 245; Johnson (1984), Ch. 18.

Electronic scanning moves the beam by electronically varying the phase of the aperture illumination. Three approaches are available: frequency scanning, phase scanning, or electronic feed switching, any of which can change the linear phase gradient across the aperture. (See **ARRAY**; **PATTERN, array.**) *SAL*

Ref.: Leonov (1988), p. 35; Skolnik (1990), p. 7.7.

Frequency scanning moves the beam by changing the carrier frequency of the transmitter and receiver. A frequency-sensitive feed structure is used (see **FEED, serpentine**), in which the phase shift between adjacent element ports varies as a linear function of frequency. The basic equation describing the angle θ_m of the beam axis is the frequency-scan equation:

$$\sin\theta_m = \frac{s\lambda}{d}\left(\frac{1}{\lambda_g} - \frac{1}{\lambda_{gm}}\right)$$

where s is the length of the feed line between element ports, λ is the free-space wavelength, λ_g is the wavelength in the feed line, d is the distance between radiating elements in the aperture plane, and m is an integer.

The frequency-sensitive feed line may excite horn feeds to illuminate a parabolic cylinder reflector, which forms the beam in the nonscanning coordinate, or may provide inputs to a number of waveguides containing slot radiators (see Fig. A82 in **ARRAY**).

Within-pulse scanning is one method of exploiting frequency scan. In this case the frequency-sensitive antenna is excited by a wideband chirp-type waveform, spreading the pulse energy over the elevation sector of a 3D radar. Upon reception, the echo signals are fed through a wideband receiver front end and then to channel filters, each corresponding to one elevation beam.

Changing the frequency of the carrier is a simple and convenient method of obtaining the phase shift required for scanning, but it introduces some limitations in radar design and operation:

(1) A major portion of the available tuning range is devoted to beam steering, reducing the possibility of exploiting frequency agility or wideband (range resolution) waveforms.

(2) Wideband signals (short pulses or pulse compression waveforms) cause distortion of the antenna pattern, since each spectral component is radiated in a different direction.

(3) The transmitter, receiver, and feed lines must have sufficient bandwidth.

(4) Only one coordinate can be scanned electronically.

(5) The slow-wave structures typically used in the feed are rather bulky and complicated. *DKB, SAL*

Ref.: Skolnik (1980), pp. 298, 314; Johnson (1984), Ch. 19.

Intermediate-frequency (IF) scanning is used in receiving arrays, where the phase shifting is implemented at IF instead of at RF. Separate receiver front ends are required for each controlled row or column of the array, and phase shifters are inserted before the channels are added. A linear change in phase shift with time may be obtained by using offset local oscillators in the several channels. *SAL*

Ref.: Skolnik (1990), p. 7.8.

Mechanical scanning moves the beam by rotation of the entire antenna structure. In conventional search radars this involves mounting the antenna system on a pedestal that rotates continuously in azimuth, the transmitter power being passed through a rotary joint. Sector scanning may also be used for height finding and precision approach radars, and for other special purposes. Tracking radars use two-axis antenna pedestals (or sometimes three-axis, in shipborne or airborne radars) to cover their assigned volumes. Mechanical scanning is the simplest and cheapest method of covering a volume of space, especially when near-hemispherical coverage is required, as in military or ATC surveillance radars and airborne early warning radars. Azimuth rotation can be combined with electronic scanning in elevation to provide 3D search radar coverage. *SAL*

Ref.: Leonov (1988), p. 80; Long (1990), p. 278.

Phase scanning moves the beam by controlling the phase of the antenna illumination, using phase shifters or delay lines.

This is the most flexible type of scanning, provides the highest accuracy, and is most often the intended meaning of the term *phased array.* (See **ARRAY**; **PATTERN, array.**) The phase shifters may be associated with each row or column in an array providing electronic scan in one coordinate (and relying on mechanical scan in the second coordinate, see **RADAR, three-dimensional**). It is also possible to combine phase and frequency scanning, using a set of phase shifters for one coordinate, feeding frequency-sensitive lines for scanning in the second coordinate. In a two-coordinate phase scanning array there must be a phase shifter associated with each element. *SAL*

Ref.: Johnson (1984), p. 20.2; Skolnik (1990), p. 7.7.

Scan rate is a measure of the rate of angular coverage of a scanning antenna beam, in deg/s or r/s. The antenna scan rate is designed to allow the radar to complete a search of the designated volume within the appropriate frame time while ensuring that the implemented scan rate permits enough time-on-target for clutter rejection waveforms to be effective. *PCH*

Sector scan is a search mode in which the radar antenna repeatedly scans a limited volumetric sector or solid angle. Sector scan is typical of airborne intercept radars, most surface-based fire control radars and all height-finders. The orientation and dimensions of the sectors are either established *a priori* or are defined, as in a fire control or target-tracking radar, by the uncertainty in the designated target location provided by a separate (search) radar. *PCH*

scan loss (see **LOSS, scanning**).

Time-delay scanning is a method of phase scanning in which variable time delay is used to shift phase continuously through many cycles, rather than applying the shift modulo 360°. The resulting beam position is independent of carrier frequency, but time-delay elements are usually too expensive and bulky to be applied to control of individual radiating elements. Time-delay control of subarrays or other grouped elements, supplemented by individual element phase shifters has proven practical in some cases, although much more expensive than conventional phase control. The advent of photonic devices, in which time delay is provided by a length of fiber-optic cable, may make time delay scanning more practical in future radars. *SAL*

Ref.: Skolnik (1990), p. 7.8.

Volume scanning is scanning performed simultaneously in azimuth and elevation angles to observe the required search volume in space. Typically, such scanning is performed by phased arrays having narrow pencil beams. Different algorithms of looking through the required search volume exist: helical scan, raster scan, TV scan, and so forth. *SAL*

Ref.: Long (1992), p. 278.

scan with compensation (see **CONOPULSE**).

within-pulse scanning (see **frequency scanning**).

SCANNER, antenna. An antenna scanner is a structure implementing the scanning of the beam. The term is typically applied to electromechanical scanning mechanisms. The simplest type of scanner consists of a focusing objective (aperture), usually a reflector or lens, and a point-source feed. Scanning occurs when the feed moves about the focal point, changing the phase gradient across the aperture (Fig. S4).

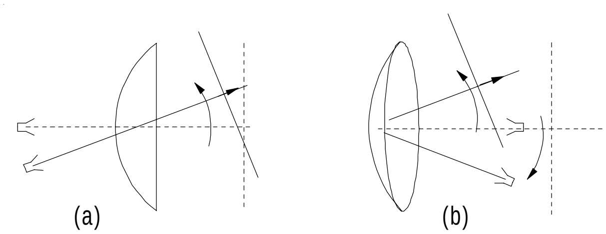


Figure S4 Scanning with a focusing objective: (a) lens; (b) reflector.

Another approach keeps both the feed and the objective fixed while changing the path between them (a method first implemented in the Foster scanner). The main types of feed-motion scanners (Lewis, Robinson, Foster, and Schwarzchild scanners) were developed during World War II. They are seldom used in modern radars because of their limitations in scan rate and pattern distortion (see **SCANNING, electromechanical**), and have been replaced since the 1960s by phased array antennas. ***SAL***

Ref.: Skolnik (1980), p. 248; Johnson (1984), Ch. 18.

A **concentric-lens scanner** is a parallel-plate scanner made up of a cylinder (or a half-cylinder) whose height is small compared with its diameter and contains a lens with surfaces concentric with the cylinder surface. *SAL*

Ref.: Johnson (1984), p. 18.13.

A feed-motion scanner generates spatial motion of the beam by mechanical motion of the feed. This motion is usually circular and is transferred to linear motion of the beam by special design of the electromagnetic path within the antenna structure. The main types of feed-motion scanners are virtual-source scanners, organ-pipe scanners, and moving-waveguide-wall scanners. The examples of feed-motion scanners are Foster, Robinson, Lewis, Schwarzchild scanners, and so forth, which were developed during World War II. Typically, feed-motion scanners are used to provide electromechanical scanning in reflector antennas. *SAL*

Ref.: Johnson (1984), pp. 18.2–18.26; Skolnik (1980), p. 248.

A **ferrite scanner** is a scanner using the phase-shifting abilities of ferromagnetic materials to steer the antenna beam. One of the applications of the ferrite scanner is the series ferrite scan technique. In this case electronic scanning of phased arrays is achieved by employing a waveguide, to which the radiating elements of the guide are connected, and which has one long series-connected phase shifter inserted lengthwise, (Fig. S5). This technique requires one driver circuit per column or row of phased array if each of them is made up of the

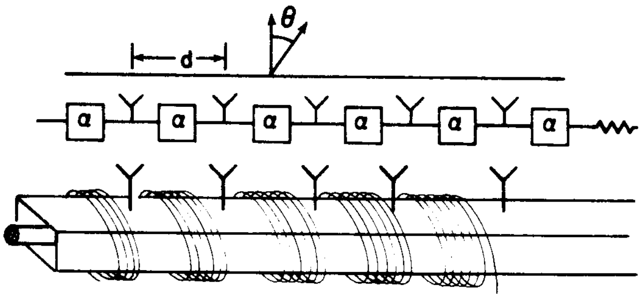


Figure S5 Ferrite scanner (from Brookner, 1977, Fig. 13, p. 332).

ferrite loaded waveguide. As many discrete phase shifters in this case are replaced with a single phase shifter (practically one scanner for each angular plane as in continuous aperture antennas), the type of scanning realized in such arrays sometimes is termed *continuous-aperture scanning*. *SAL*

Ref.: Brookner (1977), p. 330.

A **Foster scanner** is a feed-motion scanner of the prism type, in which the phase of the illuminating signal is varied by rotating the inner cone at the cross section of the device (Fig. S6). The phase variation is minimum when the input and aperture are aligned and is maximum when the energy travels counterclockwise or almost 180°. *SAL*

Ref.: Johnson (1984), p. 18.22.

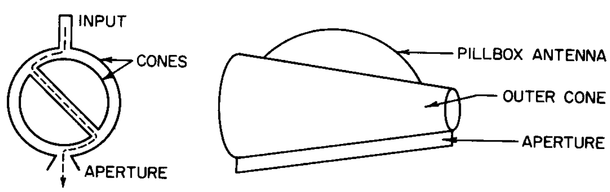


Figure S6 Foster scanner (from Johnson, 1984, Fig. 18.25, p. 18.22, reprinted by permission of McGraw-Hill).

A **helical-slot scanner** is a moving-waveguide-wall scanner in which one wall of the waveguide system can be removed, so the slot radiator is formed between the waveguide opening and a helical opening on a circular cylinder (Fig. S7). *SAL*

Ref.: Johnson (1984), p. 18.25.

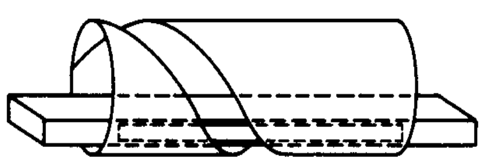


Figure S7 Helical-slot scanner (from Johnson, 1984, Fig. 18.29, p. 18.25, reprinted by permission of McGraw-Hill).

A **lens scanner** is the scanning system using lenses to scan the beam (see **LENS**). *SAL*

Ref.: Johnson (1984), p. 18.7.

A **Lewis scanner** is a feed-motion scanner in which rays from the real source strike a reflector strip at 45° to the antenna aperture producing a virtual source feeding the lens (Fig. S8). The feed rotates about the base of the cylinder (the straight-line feed path is formed into a circle by rolling the parallel-plate region into a cylinder). This produces a virtual source which appears to be moving along the straight line. *SAL*

Ref.: Johnson (1984), p. 18.20.

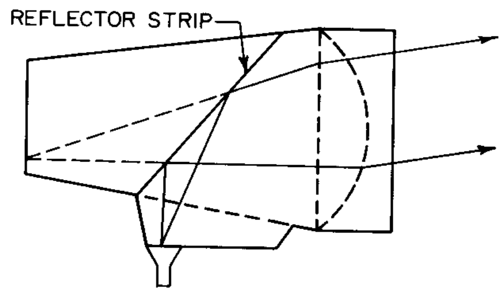


Figure S8 Lewis scanner (from Johnson, 1984, Fig. 18.20, p. 18.20, reprinted by permission of McGraw-Hill).

An **organ-pipe scanner** is one using a small rotating horn feed, which excites a row of waveguide sections of the identical length, whose output ends are arranged in a straight line and are used to irradiate a parabolic mirror (Fig. S9). In those cases when the required angular dimensions of the scanning zone exceed the capabilities of the parabolic mirrors (approximately five times the width of the directional pattern to both sides), the linear arrangement of the leads of the waveguides are converted into an arc of a circle to irradiate the operating section of a toroidal-parabolic mirror.

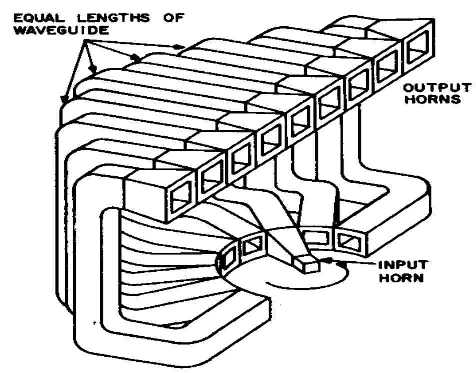


Figure S9 Organ-pipe scanner (from Brown, in Skolnik, 1970, Fig. 7, p. 22.7, reprinted by permission of McGraw-Hill).

The scanner consists of a fixed feed, fixed cassette of bent waveguides (stator) and fixed cassette of waveguide (rotor) (Fig. S10). The feed "illuminates" several adjacent waveguides of the rotor. Depending on the position of the rotor, the energy is transmitted to a specific group of waveguides of the stator. This is equivalent to mixing a phase center of a reflector or lens.

An advantage of the scanner is the absence of a rotating joint (the feed is not important). The speed of rotation of the rotor is half that of the Lewis scanner. However the rotary scanner has a narrower bandwidth and lower efficiency due to reflection of the wave at the faces of the rotor and stator. *IAM*

Ref.: Rakov (1970), p. 339; Skolnik (1970), pp. 22.6–22.7.

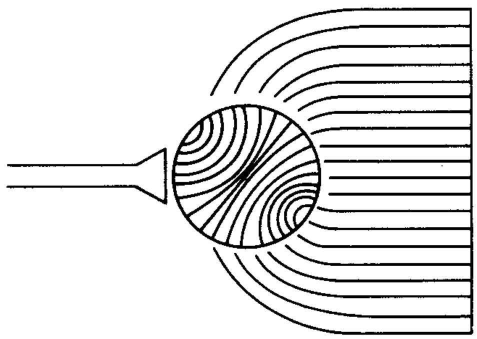


Figure S10 Organ-pipe (rotary) scanner (from Johnson, 1984, Fig. 18.27, p. 18.24, reprinted by permission of McGraw-Hill).

A **parallel-plate scanner** is the scanner whose mean surface is the surface of revolution. The main scanners of this type are the pillbox scanner, the concentric-lens scanner, the Schmidt scanner, the Schwarzchild scanner, and the zero-cardioid scanner. *SAL*

Ref.: Johnson (1984), p. 18.11.

A **pillbox scanner** is a parallel-plate scanner based on a pillbox antenna configuration. A cylindrical surface-of-revolution scanner is shown in Fig. S11. *SAL*

Ref.: Johnson (1984), p. 18-12.

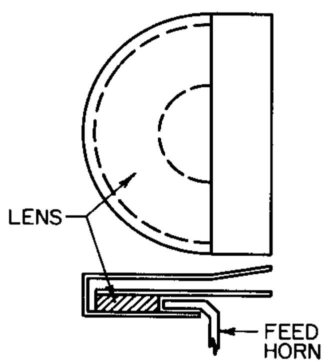


Figure S11 Pillbox scanner (from Johnson, 1984, Fig. 18.12, p. 18.13, reprinted by permission of McGraw-Hill).

A **prism scanner** is a feed-motion scanner employing an element placed between the feed and the objective. Figure S12 shows where the circular prism element (scanning obtained by the rotation of the element) and triangular prism element (scanning obtained by oscillation of the element) are located. *SAL*

Ref.: Johnson (1984), p. 18.22.

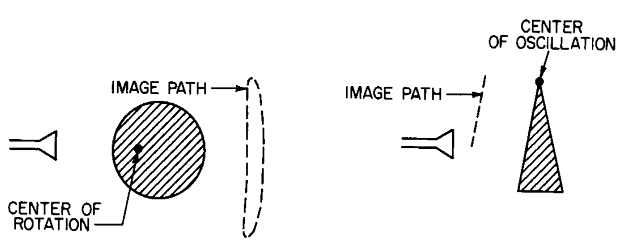


Figure S12 Prism scanners (from Johnson, 1984, Fig. 18.24, p. 18.22, reprinted by permission of McGraw-Hill).

A ring-switch scanner is a moving-waveguide-wall scanner made up of the circular section of waveguide with a fixed input and a number of moving outputs (Fig. S13). *SAL*

Ref.: Johnson (1984), p. 18.30.

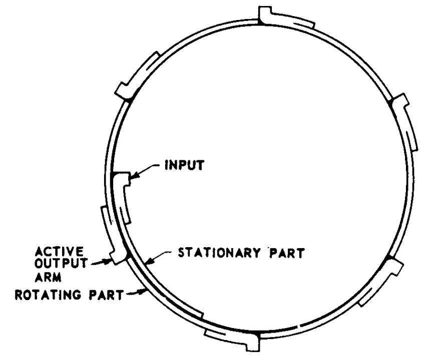


Figure S13 Ring-switch scanner (from Johnson, 1984, Fig. 18.30, p. 18.25, reprinted by permission of McGraw-Hill).

A **Robinson scanne**r is a feed-motion scanner made up in parallel plates similar to that of the Lewis scanner, except for the absence of the focusing lens and the use of the more complex image reflector. The surface used in this scanner is developed from a plane (Fig. S14), so rays paths can be found from the consideration of the ray in a plane. This scanner makes it possible to perform a sawtooth scan in the vertical plane within 10 or 20 beamwidths. To convert the linear movement of the feed into circular movement, more convenient for rapid scanning, a more complex design may be used through bending and convolution of the trapezoidal parallel plates. *IAM, SAL*

Ref.: Skolnik (1970), p. 22.6; Johnson (1984), p. 18.21.

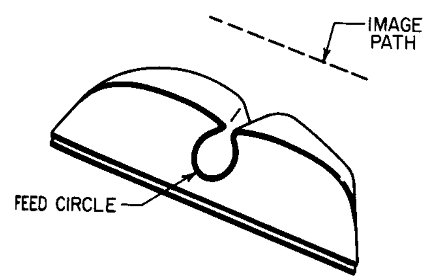


Figure S14 Robinson scanne**r** (from Johnson, 1984, Fig. 18.21, p. 18.21, reprinted by permission of McGraw-Hill).

The **Schmidt scanner** is a parallel-plate scanner consisting of a spherical reflector with a nearby planar correcting lens passing through the center of the sphere. *SAL*

Ref.: Johnson (1984), p. 18.13.

The **Schwarzchild scanner** is a parallel-plate scanner using multiple mirrors rather than a lens and a mirror. The only experimental system was built in the form of a triple-layer parallel-plate region.

Ref.: Johnson (1984), p. 18.13.

A **tape scanner** is one of the simplest type of moving waveguide-wall scanner involving the physical motion of the slots in one face of waveguide (Fig. S15). Energy is radiated through two slots that move along the waveguide length, and an effective waveguide short circuit has to be moved in synchronism with the slots to couple out all energy within the guide. *SAL*

Ref.: Johnson (1984), p. 18.24.

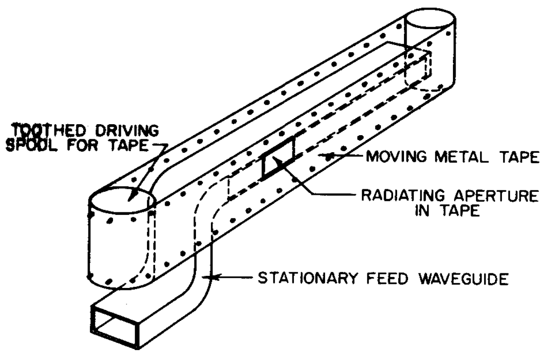


Figure S15 Tape scanner (from Johnson, 1984, Fig. 18.28, p. 18.24, reprinted by permission of McGraw-Hill).

A **tilting-plane scanner** is a feed-motion scanner in which the feed horn is held fixed when it directs the energy into the plane reflector and the orientation of the plate can be changed (Fig. S16). *SAL*

Ref.: Johnson (1984), p. 18.21.

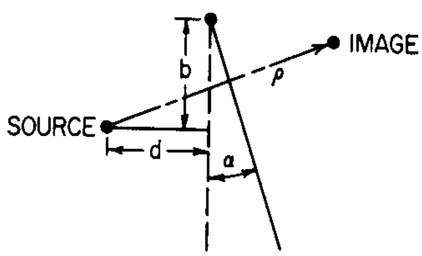


Figure S16 Tilting-plane scanner (from Johnson, 1984, Fig. 18.23, p. 18.22, reprinted by permission of McGraw-Hill).

A **virtual-source scanner** is a feed-motion scanner using a source produced by a reflection or a refraction at some optical component between a real source and the focusing objective. In most cases a real source (feed horn) is rotated continuously in front of the imaging reflector; for example, it rotates in the lower level of a a two-layer parallel-plate region (Fig. S17). In this case, energy from the feed reflects from a parabolic cylinder placed at the junction between the upper and the lower layers, and the reflected rays appear to come from a virtual source which moves along approximately linear path while the feed horn move through the arc of 120°. The Lewis scanner, Robinson scanner, and tilting-plane scanner are different virtual-source scanners. *SAL*

Ref.: Johnson (1984), p. 18.21.

A **Zeiss-cardioid scanner** is a parallel-plate scanner using two reflectors: one is a circle and another is a cardioid. *SAL*

Ref.: Johnson (1984), p. 18.14.

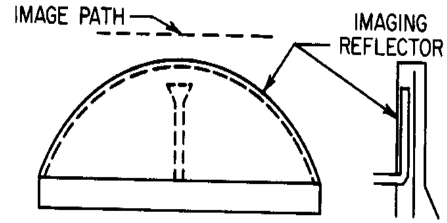


Figure S17 Virtual-source scanner (from Johnson, 1984, Fig. 18.22, p. 18.21, reprinted by permission of McGraw-Hill).

SCATTERER, SCATTERING. Scattering is "the process in which the energy of a traveling wave is dispersed due to interaction with inhomogeneities of the medium" (of which the radar target is one). Scattering characteristics are typically divided into integral and local scattering characteristics. The former describe the total energy scattered into space by the target as a whole. In the monostatic case, a single integral characteristic is used: the mean scattering cross section, equal to the average value $\bar{\sigma}$ of the monostatic effective scattering cross section $\sigma(\psi,\theta)$ over the entire solid angle 4π:

$$\bar{\sigma} = \frac{1}{4\pi}\int_0^{2\pi} d\psi \int_{-\pi/2}^{\pi/2} \sigma(\psi,\theta)\cos\theta\, d\theta$$

where the angles ψ,θ are azimuth and elevation or an equivalent pair of angles.

For ideal conductive convex bodies that are large in comparison with the wavelength, the value of σ is on the order of four times the surface area of the body.

In the bistatic case, two characteristics are used: the integral scattering cross section, equal to the average of the bistatic scattering cross section (bistatic RCS) over the entire solid angle and obtained as an average at the receiver position, and the average scattering cross section, obtained as an average including both the receiver and transmitter position.

The local scattering characteristics describe the scattering of an incident wave at separate elements of the object's surface. They are used when the target dimensions are significantly greater than the radar wavelength, when the signal reflection process consists of reflections from multiple scattering centers, each of which is associated with a different feature of the target's surface. The number of local characteristics is defined by the scattering cross section, pattern and scattering matrix of the individual scattering centers. The local characteristics are usually used in the analysis of high-resolution radar systems, and also in conventional radar problems, where the complete scattering characteristics of complex targets are obtained through superposition of reflections from individual specular points (see also **RADAR CROSS SECTION**; **REFLECTION**). *IAM*

Ref.: IEEE (1993), p. 1170; Kobak (1975), p. 66; Beckmann (1987); Tuchkov (1985), p. 60; Bancroft (1996).

Auroral scatter is caused by auroral disturbances that give radio reflections at HF and VHF. The main mechanisms of auroral scattering are weak scattering from randomly distributed gradients in electron density and strong scattering from randomly distributed clouds of ionization. Auroral reflections are aspect-sensitive and influence primarily over-the-horizon radars when the path of its propagation lies in northern latitudes (See also **CLUTTER, aurora.**) *SAL*

Ref.: Fink (1975), p. 18.115.

backscattering (see **BACKSCATTERING**).

Bragg scattering (see **BRAGG**).

The **scattering center** is the hypothetical point, which for a given trajectory, bistatic angle (angle between the directions to the transmitter and the receiver) and polarization may be considered to be the source of the scattered wave. For actual targets, the position of the scattering center wanders in various directions due to changes in the trajectory and (or) the bistatic angle. *IAM*

Ref.: Kobak (1975), p. 74.

coherent scattering (see **REFLECTION, specular**).

scattering cross section (see **RADAR CROSS SECTION**).

differential scattering cross section (see **CLUTTER reflectivity**).

diffuse [incoherent] scattering (see **REFLECTION, diffuse**).

End-region scattering is "the scattering from the end regions of finite bodies, which produces sidelobe scattering in directions away from specular." *SAL*

Ref.: Knott (1993), p. 90.

forward scattering (see **REFLECTION, diffuse** and **specular**).

Inverse scattering is the method for obtaining the target size and shape based on the measurement of backscattered field at different frequencies and for different aspect angles. The received data can be used for target recognition and requires the use of high-range-resolution and inverse synthetic aperture radar to get detailed information about recognition features. *SAL*

Ref.: Skolnik (1980), p. 437.

The **scattering matrix** is (1) a shortened term for the polarization scattering matrix, or (2) in circuit analysis, "an $n \times n$ (square) matrix used to relate incident waves and reflected waves for an n-port network."

Ref.: IEEE (1993), p. 1,170.

Polarimetric scattering is when the scattered electromagnetic field has a polarization that differs from that of incident field and is of interest in radar measurement. *SAL*

Ref.: Bhattacharyya (1991), p. 97.

A **Rayleigh scatterer** is one with dimensions significantly less than the radar wavelength. (See **scattering regions.**)

Scattering regions are the regions of normalized target dimensions expressed in wavelength (electric size) for which the wavelength dependence of scattering phenomena is the same. Typically, they are classified as the *Rayleigh scattering region* where the normalized dimension is much less than a wavelength and the RCS is proportional to the square of the volume of the body; the *resonant scattering region*, where the dimension is of the order λ/2 to 10λ wavelengths, and which is characterized by oscillating dependence of RCS on electrical size; and the *optical scattering region*, for dimensions more than 10λ in which geometric optics approximations work satisfactory and the RCS, at least for the sphere, closely approaches the projected area. Normalized RCS of conducting sphere and the scattering regions for it are shown in Fig. S18.

Sometimes, scattering regions are specifically defined for different illuminated surfaces. For example, for sea clutter, the scattering regions are defined depending on the grazing angle: the quasispecular region near vertical incidence, the interference region for low grazing angles (several degrees or less), and the intermediate diffuse or plateau region between the first two in which scattering is similar to that from a rough surface. *SAL*

Ref.: Skolnik (1980), p. 474; (1990), p. 11.5.

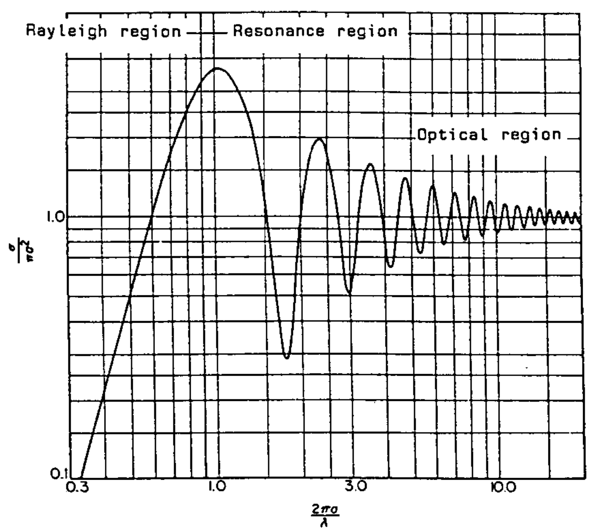


Figure S18 Normalized RCS of conducting sphere (from Barton, 1991, Fig. 5.2.3, p. 5-7).

Thomson scattering is the incoherent scatter from electrons in the ionosphere. *SAL*

Ref.: Fink (1982), p. 18.116.

SCATTEROMETER, SCATTEROMETRY. A scatterometer is "a radar calibrated to allow measurement of signal amplitudes," and scatterometry is the amplitude measurement technique. In the broad sense, any radar (e.g., an imaging or tracking radar) can be called a scatterometer if it is calibrated for amplitude measurement, but the term is normally applied to specialized radars used to measure the scattering coefficient of the earth's surface. A convenient approach to measuring the scattering coefficient at many angles simultaneously is to use a CW radar on a moving platform. In this case, doppler filters are used to separate returns from doppler frequencies corresponding to different grazing angles. A fan beam in elevation permits simultaneous measurement at points ahead of and behind the radar platform, which may be on a ground vehicle, helicopter, aircraft, or spacecraft.

When the scattering coefficient σ^0 is constant in the illuminated area, antenna gain is constant over the beamwidth and zero elsewhere, and range variation across the beam can be neglected, one can write:

$$\sigma^0 = \frac{P_r}{P_t} \cdot \frac{2vR^2}{\lambda^2 G_0^2 \psi_0 \Delta f_d}$$

where P_r is the received power, P_t is the transmitted power, v is the radial velocity, R is range to the illuminated area, λ is the wavelength, G_0 is the antenna gain, ψ_0 is the grazing angle, and Δf_d is the doppler filter bandwidth.

Pencil-beam scatterometers also can be used when higher gain is required than is available from fan beams. The SEASAT SASS and SCAT are examples of spaceborne doppler scatterometers. When the measurement can be made over a wide band of frequencies, the scatterometer can be termed a *spectrometer*. *SAL*

Ref.: Cantafio (1989), pp. 289–309; Skolnik (1990), pp. 12.18–12.28.

scintillation (see **TARGET fluctuation**; **LOSS, fluctuation**; **ERROR, scintillation**).

SEA EFFECT. The sea effect is a phenomenon related to the influence of the sea surface on the doppler spectrum of a signal reflected from it. In contrast to dry land, the reflectivity of the sea surface is extremely dependent on the angle of incidence. Since the receiver accepts signals incident on the surface at various angles in relation to the antenna beam, a redistribution of energy occurs in the doppler spectrum of the sea return compared to that from dry land. The envelope of the spectrum is distorted and its maximum is shifted toward lower frequencies.

In doppler navigators at the commonly used angles of incidence of 25° to 30° the error in velocity measurement due to the sea effect has a magnitude on the order of 1%. Narrowing the beam and reducing the angle of incidence reduce error. Reducing the angle of incidence, however, leads to broadening of the spectrum and to a secondary doppler effect. *IAM*

Ref.: Vinitskiy (1961), p. 397.

SEARCH, radar (see **COVERAGE, radar**).

SEARCHLIGHTING is "the process of projecting a radar beam continuously to a particular object or in a particular direction as contrasted to scanning". *SAL*

Ref.: IEEE (1993), p. 1,176.

SEEKER, radar. A radar seeker is a guided-missile-based, onboard radar used to extract guidance information from received signal for interception and destroying the victim targets. Typically, they are categorized on the basis of sensed radar energy into active seekers, semiactive seekers, and passive seekers. The main feature of radar seekers compared with conventional radars is the operation in severe missile environment: shock, vibration, humidity, and temperature spread in conjunction with significant limitations for weight and volume. In addition to these severe requirements the seeker hardware must be low cost, very reliable, and producible, making it one of the most sophisticated products of electronic technology. The main solutions for up-to-date systems lie in the use of digital large-scale integration technology (monolithic microwave integrated circuitry, MMIC, and very-high-speed integrated circuits, VHSIC), monopulse technique, and solid-state transmitters (in active seekers.) *SAL*

Ref.: Ivanov, in Skolnik (1990), pp. 19.1–19.43; Currie (1987), pp. 647–691.

An **active seeke**r is a radar seeker having its own transmitter giving it the capability to illuminate the target independently of a guidance radar. A typical block diagram of an active seeker is shown in Fig. S19. Typically, active seekers use either tube transmitters and solid-state transmitters depending on the operating frequencies and the range requirements, available antenna aperture (that depends on missile size), and required power/weight and power/volume considerations. For

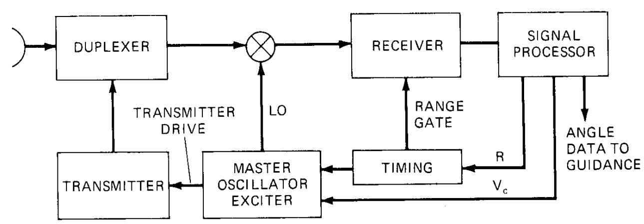


Figure S19 Active seeker block diagram (from Skolnik, 1990, Fig. 19.9, p. 19.16, reprinted by permission of McGraw-Hill).

surface targets with large RCS (like ships), solid-state transmitters are used, and for low-RCS air targets when several hundred watts of average power are required, the past preference has been for tube transmitters. Magnetrons, klystrons, and traveling-wave tubes have been used in active seeker vacuum-tube transmitters; field-effect transistors, Gunn diodes, and IMPATT diodes were the main types of microwave components used in solid-state transmitters. One of the main differences between the transmitter in an active seeker and that in conventional radar is the thermal design consideration, because conventional radar operates continuously and is designed for active cooling, while the active seeker operates for a short period (tens of seconds), permitting passive cooling.

Antennas can be of gimballed types (flat plate, Cassegrainian or inverse Cassegrainian antenna) or phased array. The latter might be the best choice for high-performance active seekers, but the hardware complexity and higher cost of a phased array makes gimballed types more competitive for all seeker applications. Because of very limited isolation achievable in the single onboard antenna, active seekers use noncoherent pulses or coherent waveforms rather than CW. Noncoherent waveforms have been widely used in seekers designed for attacking surface targets with large RCS, while coherent waveforms are required for air targets with low RCS.

Compared with semiactive seekers that can operate with high-power, large-antenna illuminators, the operating range of an active seeker is considerably less and is usually applied to short-range homing or to the terminal guidance mode of a multimode long-range system. *SAL*

Ref.: Skolnik (1990), p. 19.15.

A **lock-on-after-launch seeker** is designed to acquire and lock onto the target after launch of the host missile. Some radar homing seekers, such as the semiactive Raytheon Improved HAWK, while technically a lock-on-after-launch design, usually achieve seeker lock within a second or so of launch and can, under the right conditions (e.g., sufficient target RCS), lock on the target before missile launch. Because there is no midcourse guidance mode associated with the Improved HAWK, it is designated as a homing-all-the-way missile. Most modern surface-to-air radar guided missile systems (SAMs) incorporate one or more midcourse guidance modes and use the radar homing seeker, whether active or semiactive, for the terminal mode or "end-game." All such seekers are considered to be lock-on-after-launch radar seekers. (See also **active seeker**, **semiactive seeker**, **lock-on-before-launch seeker.**) *PCH*

A **lock-on-before-launch seeker** is designed to acquire, lock on, and establish target track while the host missile is still on the launcher. Several relatively short-range defensive missile systems, such as the passive infrared-guided Sidewinder air-to-air missile (AAM) and some radar-guided AAMs, including passive antiradiation homing missiles (ARMs), employ lock-on-before-launch seekers. The main advantages of this seeker design over the lock-on-after-launch type are (1) reduced cost and complexity and (2) higher overall probability of target intercept, since there is no transition between modes (i.e., the probability of target acquisition (given launch) is equal to unity). The disadvantage of the lock-on-before-launch seeker is that the seeker power-aperture product (in the case of an active radar seeker), or the combined illuminator radar-seeker power-aperture product (in the case of a semiactive radar homing system), is generally limited by the size and weight restrictions associated with tactical system. This limitation translates into a shorter effective range capability for most tactical defensive missile systems than would be possible with lock-on-after-launch, multimode guidance systems. The Russian SA-5 long-range strategic SAM system appears be an exception to the rule, in that the missile may be capable of target lock-on-before-launch. If this is the case, the designers have chosen to pay a high price to realize the power-aperture product required of the ground-based illuminator radar. *PCH*

A **passive seeker** is one using the energy radiated by the target for missile guidance without radiating its own energy in the target direction. There are three basic operating modes for passive seekers: antiradiation homing (ARH), home-on-jam (HOJ), and radiometric. The first mode typically is used in attacking hostile radars, usually in air-to-surface applications. Its main feature is the use of very wideband (octave bandwidth or wider) RF chains, making it similar to an electronic support measures (ESM) system. In the second, case jamming is used as a powerful source of radar energy providing angle information for homing. This is a necessary mode for semiactive and active seekers in which the target echo may be masked by jamming. The radiometric mode uses the natural thermal radiation for guidance and the sensitive receiver detects the difference in radiation between the target and the ambient background. In this mode seeker operates as a radiometer. *SAL*

Ref.: Skolnik (1990), p. 19.17.

The **seeker search volume** is the solid angle, in steradians, that a missile seeker must search in attempting to acquire (lock on) a designated target. The search volume required is a function of the seeker pointing error (SPE) that exists at the start of the search process, which in most systems will increase directly as time from launch and inversely with missile-to-target range. For a "no-scan" seeker acquisition process, the peak-to-peak value of SPE must not exceed the 3-dB beamwidth θ_3 of the seeker antenna, and the seeker acquisition range capability must be greater than the missile-to-target range for which the SPE > θ_3. If these conditions are not met, the seeker must scan. The size of the scan volume required is again a function of the SPE, which, in a multi-mode guidance system is a function of the midcourse-to-terminal "handover" errors that accrue during the midcourse flight phase. *PCH*

A **semiactive seeker** is a radar seeker that detects and tracks the signals reflected from the target that are illuminated by a separate radar (Fig. S20). The original semiactive seeker concept was CW semiactive homing, but pulsed doppler waveforms are now in common use. The local oscillator for a semiactive seeker can achieve coherence with the illumination in either of two ways: by using an on-board STALO of very high stability, or by receiving from the illuminator a reference signal from which the local oscillator is derived. In this latter case there are two antennas located on the missile: a front antenna that receives the signal reflected from the target and a rear (reference signal) antenna receiving a sample of the illumination (often through the sidelobes of the illuminator

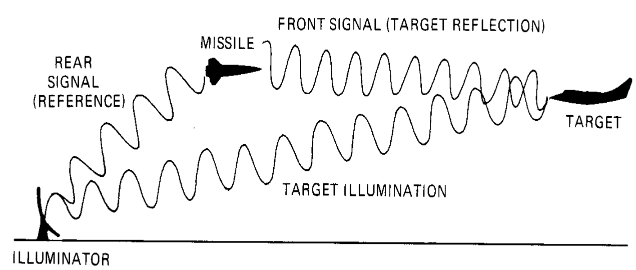


Figure S20 The concept of the semiactive homing (Skolnik, 1990, Fig. 19.1, p. 19.3).

antenna). The front receiver searches the doppler-shifted spectrum containing the target and clutter, locks on the target return, and extracts guidance information from it. In a CW system, the tracking illuminator uses a dual-antenna system, as sufficient receiver-transmitter isolation cannot be achieved in a single-antenna system. As an alternative to CW, coherent pulse-doppler waveforms can be used to achieve the benefits of CW operation without using two antennas. *SAL*

Ref.: Skolnik (1990), p. 19.3.

SELECTIVITY is the property of a system permitting the extraction of information from input signals based on specified criteria. In radar, one distinguishes four types of selectivity:

(1) Amplitude selectivity separates signals based on their levels, and is realized through use of limiters or clippers in the receiver.

(2) Frequency selectivity separates signals based on their spectra, through use of filters in the receiver.

(3) Time selectivity separates signals based on their times of arrival, through the use of gates or sampling strobes in the receiver or signal processor.

(4) Spatial selectivity separates signals based on their direction of arrival, through the pattern of the radar antenna.

Frequency selectivity is a major characteristic of receivers and is often termed simply *selectivity*. The more general selectivity capabilities of the radar are more usually discussed as resolution. *SAL*

Ref.: Skolnik (1970), p. 5.5.

SENSITIVITY is the minimum level of input signal required for target detection with specified probabilities that can be obtained at the receiver output. The minimum detectable signal power can be defined as the signal-to-noise ratio (S/N or SNR) required for reliable detection (which is the detectability factor D_x) times the receiver noise, that is:

$$S_{min} = kT_sB_n\left(\frac{S}{N}\right) = kT_sB_nD_x$$

where k is Boltzmann's constant, T_s is the system noise temperature, and B_n is the receiver noise bandwidth. The signal-to-noise required is a function of the probability of detection and the probability of false alarm. Sensitivity is often expressed in mW or decibels relative to 1 mW:

$$S_{\min}[\text{dBm}] = 10\log\frac{S_{\min}}{10^{-3}}$$

A receiver can be desensitized by raising the detection threshold (i.e., lowering the probability of false alarm).

Another definition, often used in ESM receivers, is the tangential sensitivity, defined as the signal level required to raise the lower level of visible noise at the receiver output to match the upper level of noise in the absence of a signal Fig. S21. Since the visible upper and lower levels are poorly defined, this criterion is not accurately specified, but it is normally taken to correspond to an SNR of about +8 dB. The concept of threshold sensitivity is introduced to define the

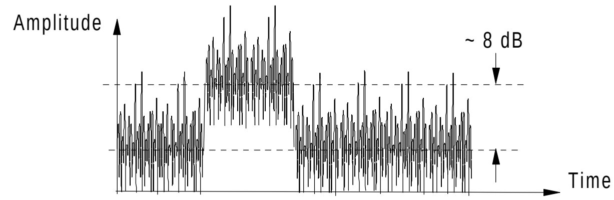


Figure S21 Tangential sensitivity of a receiver.

minimum level of the signal exceeding some specified threshold. (See also **DETECTABILITY FACTOR.**) *PCH, SAL*

Ref.: Druzhinin (1969), p. 88; Skolnik (1990), p. 1.7.

Sensitivity time control (STC) is a programmed variation in receiver gain designed to maintain constant echo signal strength as a target approaches the radar from some preset range. Since the signal power varies inversely as the fourth power of range, the common approach is to vary the receiver gain (using electronic gain control or inserted attenuation) as the fourth power of the range:

$$G(R) = G_0\left(\frac{R}{R_0}\right)^4 \tag{1}$$

where R_0 is the range beyond which full receiver gain, G_0, is used. To preserve linear operation of the receiver in the presence of clutter, the variation of gain with range may be modified to use and R^3 or other relationship. In practice, the continuous curve defined by (1) may be replaced by a series of steps of constant gain, which are sometimes referred to as an *STC map*.

The use of STC prevents the radar from detecting low-RCS clutter (e.g., birds or insects) at short range, but also prevents the detection of real targets of low RCS, and normal targets lying outside the antenna mainlobe (as in the upper coverage of a cosecant-squared pattern). STC is also called *swept gain*. *SAL*

Ref.: Skolnik (1980), p. 261; Barton (1988), p. 53.

SEQUENCE. A sequence is a mathematical term used to describe the set of symbols representing a phase-coded waveform. Binary sequences are most often used in pulse compression waveforms, the symbols 0 and 1 representing phase shifts of 0 and 180°.

An optimum binary sequence is one whose peak sidelobes are the minimum possible for a given code length. Computer simulation is generally used to find optimum binary sequences, the most frequent being Barker codes and allomorphic forms of binary codes. As the number of subpulses N increases, the computer search time becomes too long and it is more expedient to use pseudo-optimum codes that may not be strictly optimum but possess tolerable characteristics. The total number of optimum binary codes for N up to 40 and one code for each N are shown in Table S1.

The most commonly used type of sequence is the maximum length (m-)sequence, the set of N periodically varying symbol d_i, each can take values 0 or 1. They have the following qualities:

(1) The sidelobes of the periodic autocorrelation function of the waveforms coded with an *m*-sequence are equal to $1/N$.

(2) The multiplication of the elements of the original sequence d_i and a sequence shifted by k elements, d_{i+k}, after changing the product sign, gives the same sequence shifted by some number of elements.

Such a sequence is also called a *binary-shift-register sequence*, *linear recursive sequence*, *pseudonoise sequence*, or *pseudorandom sequence*.

Table S1
Optimum Binary Codes

Length of code, N	Magnitude of minimum peak sidelobe	No. of codes	Code (octal notation* for $N > 13$)
2	1	2	11, 10
3	1	1	110
4	1	2	1101, 1110
5	1	1	11101
6	2	8	110100
7	1	1	1110010
8	2	16	10110001
9	2	20	110101100
10	2	10	1110011010
11	1	1	11100010010
12	2	32	110100100011
13	1	1	1111100110101
14	2	18	366324
15	2	26	74665
16	2	20	141335
17	2	8	265014
18	2	4	467412
19	2	2	1610445
20	2	6	3731261
21	2	6	5204154
22	3	756	11273014
23	3	1021	32511437
24	3	1716	446650367
25	2	2	163402511
26	3	484	262704136
27	3	774	624213647
28	2	3	1111240347
29	3	561	30612440333
30	3	172	6162500266
31	3	502	16665201630
32	3	844	37233244307
33	3	278	55524037163
34	3	102	1447716045524
35	3	222	223352204341
36	3	322	526311337707
37	3	110	1232767305704
38	3	34	2251232160063
39	3	60	4516642774561
40	3	114	14727057244044

*Each octal digit represents three binary digits:

0	000	4	100
1	001	5	101
2	010	6	110
3	0011	7	111

(from Skolnik, 1990, Table 10.4, p. 10.18, reprinted by permission of McGraw-Hill)

A minimax sequence is one for which the peaks of the autocorrelation function are equal $1/N$, where N is the number of elements in the code. The parameters of some known minimax sequences are cited in Table S2.

Table S2
Minimax Sequences

Sequence	Period	Comments
M-sequence	$N = 2^n - 1$	n is integer
Legendre sequence	$N = 4i + 3$	i is integer
Hall sequence	$N = 4t^2 + 27$	t is integer
Jacobi sequence	$N = t(t + 2)$	t, $t + 2$ are simple figures

If different sequences correspond to the specified period N then these sequences can coincide. The Hall sequence with $N = 4t^2 + 27 = 4(t^2 + 6) + 3$ coincides with Legendre sequence for $i = t^2 + 6$. *AIL*, *SAL*

Ref.: Varakin (1970), p. 234-243; Skolnik (1990), p. 10.17.

SERVICE, radar

Servicing and maintenance is a complex set of operations to maintain the serviceability or only the operability of a radar during its preparation for and designated use, storage, and transportation. Radar servicing and maintenance includes the following components: technical inspection; preventive maintenance; supply; and collection and processing of operating results. A technical inspection is performed to evaluate the condition of the radar. The major element of servicing and maintenance is preventive maintenance, which envisions: equipment inspection and cleaning; adjustment operations; failure forecasting; seasonal lubrication and tightening operations; technical inspections. Inspection and adjustment operations and the failure forecasting work closely linked with them are the most labor-intensive part of servicing and maintenance.

Preventive operations are performed periodically based on radar operating time or calendar operating period. Preventive maintenance with its assigned periods and times is referred to as operational checks. Basic servicing and maintenance indicators include servicing and maintenance periodicity, labor intensity, and cost. Servicing and maintenance periodicity is the time between conclusion of the previous and beginning of the next preventive maintenance. Servicing and maintenance duration is the average time expended on one preventive maintenance session. Servicing and maintenance duration is measured in labor hours and is determined by the ratio of average servicing and maintenance duration to the overall number of maintenance personnel doing in the servicing and maintenance. Servicing and maintenance cost includes the cost of labor expenditures and of the materials expended on servicing and maintenance. *AIL*

Ref.: Aleksandrov (1976), p. 98.

SEXTANT, radar. A radar sextant is a passive radar used to measure the angular coordinates of astronomical sources: the sun, moon, or stars. *IAM*

Ref.: Rakov (1969), p. 395.

SHADOW, radar. A radar shadow is the "absence of radar illumination because of an intervening reflecting or absorbing object; the shadow is manifested on the display by the absence of blips from targets in the shadow area." For permanent configuration of obstructing objects within radar coverage such as terrain obstructions (hills, mountains, ridges, cliffs, forests, etc.) or man-made obstructions (buildings, towers, chimneys, etc.) the line-of-sight cutoff ranges can be plotted as the function of ground distance to obstructions, its height and the specified altitude cut. These plots are termed shadow diagrams and give the representation of shadow zones in azimuth coverage (Fig. S22).

In calculating the bistatic radar cross section of an object in the forward-scatter region, the shadow cast on the wavefront immediately beyond the target is considered an added source of radiation equal to that of the incident wave but with 180° phase shift. This contributes a large component of RCS given by

$$\sigma_f = \frac{4\pi A^2}{\lambda^2}$$

and having the pattern of a planar array in the shape of the shadow. Bistatic receivers in the vicinity of this forward scattered lobe can exploit this enhanced RCS for target detection. *DKB, SAL*

Ref.: IEEE (1990), p. 25; Barton (1988), p. 121; Willis (1991), pp. 150–155.

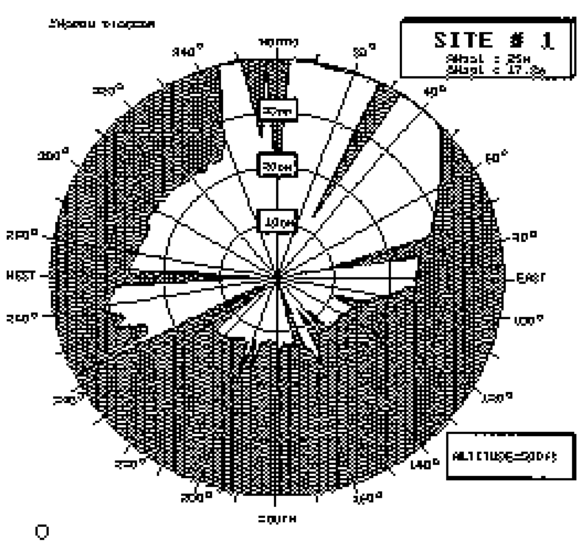


Figure S22 Shadow diagram from typical surveillance radar site.

SHAPING, low-reflection (see **STEALTH**).

SHIFT KEYING. Shift keying is amplitude, frequency, or phase modulation when the modulated parameter is changed discretely, typically assuming either of two constant values in accordance with a code for pulse compression. (See **WAVEFORM, phase-coded.**)

SIDELOBE

antenna sidelobe (see **PATTERN, antenna**).

A **sidelobe blanker** is "a device that employs an auxiliary wide-angle antenna and receivers to sense whether a received pulse originates in the sidelobe region of the main antenna and to gate it from the output signal is it does." *SAL*

Ref. IEEE (1993), p. 1217.

sidelobe canceler (see **CANCELER, CANCELLATION**).

Sidelobe masking is the ECCM technique in which sidelobe radiation is directed at an enemy jammer. If frequency agility is employed the jammer receives not only the real antenna mainlobe frequency, but also energy at other frequencies that forces the jammer to disperse the jamming power over a wider frequency band. To implement the masking signal a separate antenna or the same antenna with multibeam capability may be used. *SAL*

Ref.: Cantafio (1989), p. 451.

Range [time] sidelobes are the portions of the pulse-compression output waveform (responses) leading and trailing the principal response of compressed pulse. The view of the pulse-compression waveform at the output of the matched filter (especially for linear FM waveforms) resembles an antenna pattern with the main response (lobe) at the principal range and spurious responses (sidelobes) at other ranges; hence the name range sidelobes (Fig. S23). They can severely

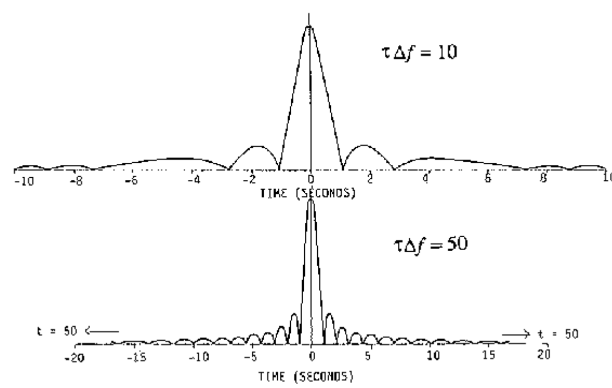


Figure S23 Linear FM matched-filter response for $\tau\Delta f = 10$ and $\tau\Delta f = 50$, Δf normalized to unity (from Barton, 1991, Fig. 7.2.4, p. 7.6).

limit the resolution and mask contiguous signals located close to each other in range, requiring range sidelobe suppression (reduction). For linear FM pulse-compression waveforms received with a matched filter, the sidelobe levels follow the $(\sin x)/x$ shape. Phase-coded waveforms have range sidelobes theoretically at a voltage level $N^{-1/2}$ relative to the mainlobe, for an N-element code. Low sidelobe levels can be achieved using a nonlinear FM waveform even without weighting, but at the cost of hardware and processing complexity. Range sidelobes are also termed *time sidelobes*.

Range sidelobes result primarily from the spectral components originating at the leading and trailing edges of the transmitted pulse, and at any discontinuities within the pulse. These components pass through the pulse-compression network with delays between zero and τ and thus appear at the output as small signals removed from the main response. In linear and nonlinear FM pulse compression, most sidelobes are caused by irregular errors in weighting of the frequency components in the pulse compression network. However, sidelobes due to reflections within the microwave system and anomalies within the waveform generator and pulse compression processing networks are not affected by weighting. When control of such sidelobes is necessary, a transversal equalizer may be used, adjusted to cancel the sidelobes out to $\pm\tau c/2$ each side of the main output lobe. This is accomplished at the expense of further spreading of low-amplitude lobes beyond the region originally affected, and increasing the sensitivity of the response to target doppler shift.

When linear FM waveforms are used, weighting of the spectrum can reduce the sidelobes, at the cost of signal-to-noise ratio reduction. The spectrum can be tapered by weighting in frequency or amplitude either the transmitted waveform, the matched filter, or both. The main technique is to use a mismatched receiving filter, instead of one strictly matched to the transmitted signal. Obviously, that results in reduction of signal-to-noise ratio, which is accepted as a necessary system loss to be paid for sidelobe reduction. The same illumination functions used in antenna design to reduce antenna sidelobes can be used to reduce range sidelobes of pulse-compression signals. Several weighting functions for a linear FM waveform are given in the Table S3. Analogously with antenna pattern considerations, weighting of the signal spectrum reduces the sidelobes but increases the mainlobe width and signal-to-noise ratio. For phase-coded waveforms, sidelobe reduction can be achieved by mismatched filtering using special weighting functions or by using complementary codes.

In the case of the nonlinear FM waveform there is no need to use special weighting, since the nonlinear variation of frequency in time can produce amplitude weighting of the signal spectrum to achieve range sidelobe reduction. Some weighting is required to reduce the spikelike, far-off sidelobes at each end of the sidelobe region, caused by the leading and trailing edges of the transmitted pulse. *SAL*

Ref.: Cook (1967), Ch. 7; Fink (1975), p. 25.76; Morris (1988), pp. 141–145; Skolnik (1990), p. 10.3.

SIGHT, radar. A radar sight is a system consisting of a radar range finder, an optical view finder, and a computer that observes the target, estimates of coordinates, and produces the data required for fire control. Typically, radar sights are classified for location as ground-based, airborne, and shipborne. *SAL*

Ref.: Popov (1980), p. 346.

Table S3
Weighting Function Data

Weighting function	Peak side-lobe level (dB)	Pulse widen-ing	Mis-match loss (dB)	Far sidelobe falloff rate
Dolph-Chebyshev	−40	1.35	-	1
Taylor, $\bar{n} = 6$	−40	1.41	−1.2	$1/t$
$k + (1 - k)\cos^n$				
Hamming ($k = 0.08, n = 2$)	−42.8	1.47	−1.34	$1/t$
Cosine-squared ($k = 0, n = 2$)	−32.2	1.62	−1.76	$1/t^3$
Cosine-cubed ($k = 0, n = 3$)	−39.1	1.87	−2.38	$1/t^4$
$n = 1, k = 0.04$	−23	1.31	−0.82	$1/t$
$n = 2, k = 0.16$	−34	1.41	−1.01	$1/t$
$n = 3, k = 0.02$	−40.8	1.79	−2.23	$1/t$
(from Cook, 1967, Table 7.1, p. 205)				

SIGNAL, radar. In radar applications, the term signal is used in the following senses:

(1) A received wave containing information about the observed environment, including targets of interest and interference.

(2) The useful portion of this wave containing information about the target.

(3) A radar waveform transmitted in the direction of the target.

In the last sense, the term *waveform* is preferable.

There are two ways of representing radar signals: in the time domain or in the frequency domain. (See **SPECTRUM**.) These are equivalent as sufficient descriptions of the signal. Most commonly used is the model representing the radar signal as the sum of a useful component, $u(\vec{\alpha}, t)$, and random noise, $n(t)$:

$$y(t) = A_0 \cdot u(\vec{\alpha}, t) + n(t) \qquad (1)$$

where A_0 is a detection weight ($A_0 = 1$ if the target is present, zero if not), and $\vec{\alpha}$ is a vector of parameters containing information about the target. The vector $\|\vec{a} = t_d, \phi, \theta, f_d, A\|$ represents information about the time delay t_d (range), azimuth ϕ, elevation θ, doppler frequency f_d (radial velocity), and signal amplitude. It should be noted that some confusion can arise since the term *signal* is sometimes used (as in *signal*-processing discussions) to denote $y(t)$, and other times (as in *signal*-to-noise ratio) to denote $u(t)$. To avoid this, $y(t)$ may be called the signal, and $u(t)$ the useful signal.

Since the noise component is always present in the input, it is a random signal. (See **FUNCTION, random.**) In theoretical analysis, the concept of a deterministic signal (all parameters of which are known) is sometimes used (see **DETECTION of a signal with known parameters**) as a simplified model for detection. For theoretical analysis and description of signals based on pulse-compression waveforms the concept of a complex waveform is used, where the useful signal is

$$u_c(t) = U_c(t)\exp(j2\pi f_0 t)$$

where $u_c(t)$ is a complex waveform, $U_c(t) = U(t)\exp(j\phi t)$ is a complex amplitude, $U(t)$ is the waveform amplitude (or envelope), $\phi(t)$ is the waveform phase, and f_0 is the carrier frequency. In this case the useful signal can be represented as

$$u(t) = \mathrm{Re}[u_c(t)] = U(t)\cos\phi(t) = U(t)\cos[2\pi f_0 t + \phi(t)]$$

The way in which the envelope $U(t)$ changes with time defines signal amplitude modulation (typically amplitude-pulse modulation), and $\phi(t)$ defines frequency (or phase) modulation.

If the waveform phase $\phi(t) = \phi$ is constant, there is no intrapulse modulation, and the resulting waveform:

$$u(t) = U(t)\cos(2\pi f_0 t + \phi)$$

is called a *simple* (uncoded) *waveform*. Variation in $\phi(t)$ with time introduces intrapulse modulation, as used in pulse compression, and the result is called a *complex waveform*.

The instantaneous frequency of the useful signal is defined as

$$\omega(t) = \frac{d}{dt}\phi(t)$$

Thus, for the simple waveform, $\omega = 2\pi f_0$, while for the complex waveform ω depends on the phase (frequency) modulation function $\phi(t)$. For linear FM, $\phi(t) = kt^2$, and

$$\omega(t) = 2\pi f_0 + 2kt$$

Depending on the value of f_0, signals are classified as RF, IF, or video (baseband). For the RF signal, f_0 is the transmitted carrier frequency, while for IF it can be any one of the intermediate frequencies used in the receiver. The video signal is a low-frequency wave at the output of the receiver or video amplifier, used as input to subsequent signal processing or to drive a display device such as a cathode-ray tube. If converted from IF in an envelope detector, it can be used for noncoherent processing or directly displayed, while if converted in a phase detector (bipolar video) it can preserve coherent information for use in MTI or pulsed doppler processing. Baseband signals are also generated for modulation of the transmitter carrier at IF or RF.

The term *coherent signal* is typically applied to a coherent pulse train, while the noncoherent signal implies a noncoherent pulse train. Analog, discrete, and digital signals are distinguished, based on the degree of discretization. Amplitude(-pulse) modulated, frequency-modulated, and phase-modulated (phase-coded) signals denote the modulation method (See **WAVEFORM.**) The wave entering the radar through the image channel is called the *image signal*. In more complicated theoretical models (e.g., in array antenna analysis), the signal in (1) is represented as a function of time and target coordinates $y(t, \vec{r})$, where $\vec{r} = (x, y, z)$, leading to the concept of the *space-time signal*. *SAL*

Ref.: Cook (1967); Mitchell (1976); Urkowitz (1983); Picinbonbo (1988).

signal bandwidth (see **BANDWIDTH**).

A **bipolar video signal** is “a radar video signal whose amplitude can have both positive and negative values.” It is derived from a phase-sensitive detector by multiplying the input signal by a sinusoidal reference (usually at IF). *SAL*

Ref.: IEEE (1993), p. 111.

Signal classification is the procedure used in electronic intelligence (ELINT) and electronic warfare to determine the type of the radar signal and select the possible ECM that must be used to jam the victim radar. The basic signal parameters used for classification are carrier frequency, bandwidth, intrapulse features (pulsewidth and modulation), and interpulse features (pulse repetition frequency, stagger parameters). *SAL*

Ref.: Wiley (1993); Goj (1993), p. 52.

The **signal-to-clutter ratio (SCR)** is the ratio of the received target signal power to the received clutter power as measured at a common point in the radar (e.g., at the radar antenna terminals). For a ground-based pulsed radar in volume clutter (rain and chaff), and a target at range R, the input signal-to-clutter ratio is

$$\frac{S}{C} = \frac{\sigma F^4 L^2{}_p R^2{}_c L_{\alpha c}}{R^4 \theta_a \theta_e \frac{\tau_n c}{2} L_\alpha \eta_v F^4{}_c}$$

where $F_c{}^4$ and $L_{\alpha c}$ are the pattern-propagation factor and atmospheric attenuation for the radar-to-clutter path, η_v is the volume clutter reflectivity (m^2/m^3), L_p is the beamshape loss, R_c is the range to the clutter, θ_a and θ_e are the radar antenna azimuth and elevation beamwidths, τ_n is the pulsewidth, c is the speed of light, and σ is the target RCS. When the radar has no range ambiguities in which clutter is present, $R_c = R$, $L_{\alpha c} = L_\alpha$. When clutter is present in multiple range ambiguities, the equation shown must be modified to sum the contribution of clutter in the multiple ambiguities.

For homogeneous, constant-gamma (γ) land clutter (see **CLUTTER**), which fills the radar azimuth beam out the horizon range, the signal-to-clutter ratio for a system with no range ambiguities containing clutter, can be expressed as

$$\frac{S}{C} = \frac{\sigma L_p F^4}{h_r \theta_a \frac{\tau_n c}{2} \gamma F^4{}_c}$$

where h_r is the height of the radar antenna above the clutter surface. In the region well within the horizon, $F = F_c = 1$. When there are range ambiguities that are occupied with clutter, the clutter contribution of the ambiguous cells must be summed, but as with volume clutter, it is usually the first ambiguity that dominates the result. *PCH*

Ref.: Barton (1988), pp. 40–43.

Signal discretization is the process of transforming analog signals into discrete signals. It involves two main operations: sampling (discretization in time or frequency domain), and quantization (discretization in amplitude). In radar this process is performed in analog-to-digital converters. *SAL*

Ref.: Wiley (1993), pp. 27–30.

Signal distortion is the change in shape or parameters (amplitude, frequency, phase, or polarization) of the radar signal between generation of the waveform and extraction of the needed information. Distortions can be introduced by reflection from a complex target (e.g., depolarization), by the propagation medium, and by radar circuits in the transmitter and receiver. The main sources of distortions are nonideal radar circuits, resulting in frequency change, interpulse phase change, pulse amplitude change, pulse width and timing jitter, generation of harmonics, and so forth. In modern radars with digital signal processing, the limitations of the analog-to-digital converter are significant sources of distortion. Distortions are classified as linear or nonlinear, the second leading to appearance of new spectral components. Signal distortion is an important issue in radar performance quality, especially for MTI and pulsed doppler radars, where they establish the stability of the system. *SAL*

Ref.: Skolnik (1980), p 130.

Signal energy is the signal parameter that characterizes the capability of the radar for detection and measurement of target coordinates. Given a signal $u(t)$, the signal energy E_s is determined by the equation:

$$E_s = \int_{-\infty}^{\infty} u^2(t)dt$$

For a rectangular pulse of width τ, the received energy is $E_s = S\tau$, where S is the signal power. This relationship applies regardless of the presence or absence of intrapulse phase or frequency modulation.

The energy is related to the signal spectrum $S(\omega)$ by the relationship:

$$E_s = \frac{1}{\pi}\int_0^{\infty} |S(\omega)|^2 d\omega$$

The energy of a received target echo signal, like its power, is inversely proportional to the fourth power of distance to the target. Signal energy is measured in joules. *AIL*

Ref.: Varakin (1970), p. 6; Shirman (1970), p. 236.

Signal extinction is the loss in strength of the signal due to its propagation in the atmosphere, considering two different physical processes: absorption of the signal (transformation into heat) and scattering of the signal (reradiation) from air molecules and particles (rain, snow, dust, etc.). It differs from signal attenuation (see **LOSS, atmospheric**) only in the sense that the latter usually does not differentiate between the two processes involved. *SAL*

Ref.: Currie (1992), p. 13.

The **signal-to-interference ratio** is the ratio of received signal power to the power of the combined interference. Sources of interference include thermal noise N, noise jamming J, and clutter C. If all three sources of interference are present, the signal-to-interference (S/I) ratio is

$$\frac{S}{I} = \left[\frac{1}{\left(\frac{S}{N}\right)} + \frac{1}{\left(\frac{S}{J}\right)} + \frac{1}{\left(\frac{S}{C}\right)}\right]^{-1}$$

Because the spectral and statistical properties of clutter may differ significantly from those of noise and jamming, calculation of radar performance must often use the ratio of signal energy to effective interference spectral densities:

$$\frac{E_1}{I_0} = \frac{E_1}{N_0 + J_0 + C_0} = \left[\frac{1}{\left(\frac{S}{N}\right)} + \frac{1}{\left(\frac{S}{J}\right)} + \frac{1}{\left(\frac{S}{C}\right)\left(\frac{D_x}{D_{xc}}\right)}\right]^{-1}$$

where D_x and D_{xc} are the detectability factors for noise and clutter, respectively. In cases where $D_{xc} > D_x$, the simple equation using signal and interference powers can lead to optimistic results. *DKB*

Ref.: Barton (1988), Ch. 1.

signal-to-jamming ratio (see **JAMMING**; **RANGE EQUATION with jamming**).

A **limited signal** is one "limited in amplitude by the dynamic range of the system." *SAL*

Ref.: IEEE (1993), p. 716.

The **minimum detectable signal** is "the minimum signal level that gives reliable detection in the presence of white Gaussian noise." The value of such a signal is limited by the presence of noise, and must be described in statistical terms, typically the probabilities of detection and false alarm. (See **SENSITIVITY.**) *SAL*

Ref.: IEEE (1993), p. 808; Skolnik (1980), p. 16; Sauvageot (1992), p. 16.

The **minimum discernible signal** is "the minimum detectable signal for a system using an operator, display, or aural device for detection." *SAL*

Ref.: IEEE (1993), p. 808.

A **narrowband signal** is one whose bandwidth is small relative to its carrier frequency (typically not more than about 10%). *SAL*

Ref.: Wehner (1987), p. 6.

Signal parameter estimation is the process of obtaining the estimator $\hat{\alpha}$ of the parameters extracted from a useful signal $u(\vec{\alpha}, t)$ (see **SIGNAL**). Since the received radar signal is a random process, $y(t)$, with some statistic (e.g., the pdf, W_y), the estimator of a scalar parameter α,

$$\hat{\alpha} = \hat{\alpha}[y(t)]$$

is a function of the received signal and can be described by the *a priori* statistic $W(\alpha)$ and the conditional statistic, $W(y|\alpha)$. The optimum estimator for minimum rms error:

$$\int_{-\infty}^{\infty} \{\hat{\alpha}[y(t)](-\alpha)\}^2 W(\alpha|y)d\alpha = \min$$

can be found as the mathematical expectation of the measured parameter

$$\hat{\alpha}_{opt}(y) = \int_{-\infty}^{\infty} \alpha W(\alpha|y)d\alpha$$

where

$$W(\alpha|y) = \frac{W(\alpha)W(y|\alpha)}{W(y)}$$

In practice, the likelihood ratio

$$l(\alpha) = \frac{W(y|\alpha)}{W(y)}$$

is used to describe the *a posteriori* pdf $W(\alpha|y) = W(\alpha)l(\alpha)$. In this case:

$$\hat{\alpha}_{opt}(y) = k\int_{-\infty}^{\infty} \alpha W(\alpha)l(\alpha)d\alpha$$

where

$$k = \left[\int_{-\infty}^{\infty} W(\alpha)l(\alpha)d\alpha\right]^{-1}$$

The estimation of a nonrandom parameter α is referred to as maximum likelihood estimation.When the parameter is random the Bayes estimation procedure is used. The process of signal parameter estimation using practical radar circuits is referred to as radar measurement. (See **MEASUREMENT**.) *SAL*

Ref/: Barkat (1991), Ch. 4; Shirman (1970), p. 173.

Signal selection is the discrimination of a useful signal against an interference background (e.g., jamming). The basic types of signal selection are shown in Table S4, which identifies the parameter used to discriminate between the useful signal S and interference I.

The selection techniques listed in Table S4 are typically referred to in Russian literature as *primary selection*. Secondary selection can use special coding of the transmitted waveform to increase interference immunity, in which case the corresponding received signal parameters are monitored to discriminate between useful signals and interference. Amplitude, frequency, phase, time, and other discrimination methods can also be used for secondary selection. When spatial selection is used by processing signals from several independent receivers to enhance angular resolution, it is called *functional selection*. *SAL*

Ref.: Maksimov (1979), pp. 108–114, Chs. 6, 8.

Table S4
Signal Selection

Type of selection	Discrimination parameter	Typical techniques and applications
Amplitude selection	Amplitude difference	Clipper and integration of the useful signal in a matched filter
Frequency selection	Difference in carrier frequency, pulse repetition frequency, scanning frequency, etc.	Frequency diversity, pulse-to-pulse, or scan-to-scan integration
Polarization selection	Polarization difference	Widely used to discriminate complex targets (e.g., aircraft) from rain clutter, using circular polarization
Phase selection	Phase	Phase detectors and phase tracking systems (phase-locked loops)
Spatial selection	Angle of arrival	Choice of aperture illumination, and use of nonlinear signal processing at antenna output.
Time selection	Time of arrival	Time position (range delay), pulsewidth selection
Structural selection	Signal structure (e.g., modulation function)	Signal pattern recognition
Combined selection	Combinations of previous parameters	Combinations of previous techniques (e.g., space-time, frequency-time, etc.

The **signal-to-noise ratio (SNR)** in a radar system can be expressed either in terms of the ratio of signal energy E at the input, to the noise spectral density N_0, or the ratio of signal power S to noise power N within the receiving system bandwidth. The relationship between the two measures, for an individual received pulse, is

$$\frac{S}{N} = \frac{E_1}{N_0 L_m}$$

where E_1 is the energy of a single received pulse and L_m is the receiver matching loss. When multiple pulses are coherently integrated, the SNR at the output becomes

$$\left(\frac{S}{N}\right)_o = \frac{E}{N_0 L_m L_{mf}}$$

where $E = nE_1$ is the total received energy of the pulse train and L_{mf} is the matching loss of the doppler filter.

The use of input energy ratio frees the radar analysis from difficulties in defining the "effective bandwidth" of complex signal processors. *DKB*

Ref.: Skolnik (1970) Ch. 2; Barton (1988) Ch. 1.

SIGNAL PROCESSING, SIGNAL PROCESSOR. Radar signal processing is an operation in which desired data about a target are extracted from the received radar signal. Signal processing is necessary in performance of all major radar operations: detection, measurement, and, if applicable, recognition and identification. The radar element performing this operation is termed a *signal processor*, and that using the data output of the signal processor is termed a *data processor*. The signal processor can be considered the final portion of the radar receiver, but it is more usually considered a separate radar subsystem due to the significance of its functions in the radar channel. The distinction between signal processing and data processing is not always clear, but in general the latter involves operations on angle coordinate data obtained from the antenna as well as range (and sometimes velocity) data from the signal processor, and in many cases the data processing is carried out over several scans of the antenna (for search radar), or periods in excess of a few seconds for tracking and guidance radar.

The top-level diagram of a typical modern signal processor is shown in Fig. S24. The matched filter is usually considered part of the radar receiver, although it is an important element in signal processing. If the basic signal-processing operations are performed in analog circuits, the system is termed an *analog signal processor*. The system diagrammed, while using an analog matched filter, is a *digital signal processor*. The plot extractor (tracker) is often considered part of the radar data processor, although it may be integrated into the signal processor as a final stage of filtering (operating over several scans) to remove clutter and other forms of interference.

Most modern radars used digital signal processors following the general outline of Fig. S24 and containing pulse compression, doppler (or MTI) filtering, CFAR, binary integration, and possibly the plot extractor function. The first four operations are sometimes called *primary signal processing*, since the information extracted from the received signal (usually a pulse train) at this stage is gathered in a single scan. The plot extraction operation can be called secondary (or data) processing because the data are derived from several scans (primary processing intervals). This function is often performed remotely from the radar site and is not included in the radar signal processor. In the article on filters, that term was defined in three senses, the first referring to operations typically performed in radar receivers, the second to primary signal processing operations, and the third to data processing: plot extraction or tracking (see **FILTER**).

Classifications of signal processors consider different properties and implementations, as shown in Table S5. *SAL*

Ref.: Oppenheim (1975); Lewis (1986), Ch. 1; Kunt (1986); Farina (1986, 1987, 1992); Vanderlugt (1992); Nitzberg (1992); Malvar (1992).

Table S5
Classification of Signal Processors

Property	Types
Nature of signal	Analog, digital
	Coherent, noncoherent
Algorithms or performance features	Adaptive, nonadaptive
	CFAR, MTI, pulsed doppler (MTD), FFT, robust, SAR
Number of channels	Single-channel, multichannel

SIGNATURE, radar (see **TARGET RECOGNITION AND IDENTIFICATION**).

SILENCE, radar. Radar silence is defined as "an imposed discipline prohibiting the transmission by radar of electromagnetic signals on some or all frequencies." It is usually an operational state of deployed land, sea, or air forces during which no radar transmissions are allowed, to avoid detection of the force by passive electronic support measures (ESM). Radar and radio silence are enforced as part of emission control (EMCON). *PCH*

Ref.: Johnston (1979), p. 65.

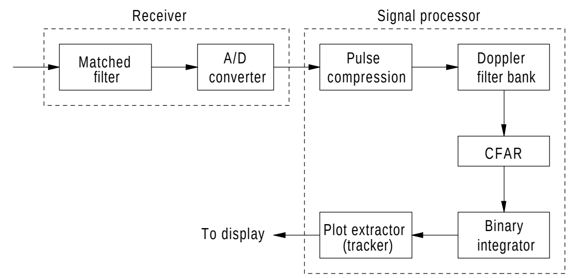


Figure S24 Block diagram of typical radar signal processor.

SKIATRON. A skiatron is a cathode-ray tube on which the signals are displayed as dark traces or dark blips against the bright face. In older radars it was used as a long-persistence display or storage tube. The device is now obsolete. Another term is *dark-trace CRT*. *SAL*

Ref.: IEEE (1993), p. 1,233.

SLANT-RANGE EFFECT. The slant-range effect is the effect of variation of the center frequency of moving radar return that corresponds to the change in antenna pointing angle with range. When the antenna is pointing ahead it is a predominant platform motion effect influencing doppler offset. When MTI is used it is desirable to center the clutter spectrum in the minimum response region (notch) of the MTI filter to obtain maximum clutter rejection. Since the clutter center frequency varies with range and azimuth, when the radar is moving, it is desirable for the filter notch to track the doppler-offset frequency. This can be done using an open- or closed-loop system; for example, a time-averaged-clutter coherent airborne radar (TACCAR). *SAL*

Ref.: Skolnik (1990), p. 16.4.

SLOW-WAVE STRUCTURE [CIRCUIT]. A slow-wave structure (or circuit) is used in certain types of microwave tubes to decrease the RF wave velocity, increasing the interaction time of the electron stream with the RF wave. To obtain efficient interaction the phase velocity of the RF wave should be approximately equal to the velocity of electrons. The crossed-field tube and traveling-wave tube are the major microwave devices where slow-wave circuits are employed. Slow-wave structures may be based on comb-quad circuits, meander lines, helices, helix-derived structures, bar and vane structures, and so forth. In general they are divided into circuits supporting electronic interaction with a forward-traveling wave, a backward-traveling wave, or a standing wave. The most common slow-wave circuits suitable for use in forward-wave interaction are helix-derived structures, bar and vane structures, meander lines, and capacitively-strapped bar circuits. Backward-wave circuits, usually derivatives of interdigital lines, strapped bar-and-vane circuits, or standing-wave circuits, are used in magnetrons and are resonant. Slow-wave circuits in TWTs are derived from the helix and coupled-cavities. Examples of typical slow-wave structures for crossed-field tubes are shown in Fig. S25. *SAL*

Ref.: Fink (1982), p. 9.48; Currie (1987), p. 460.

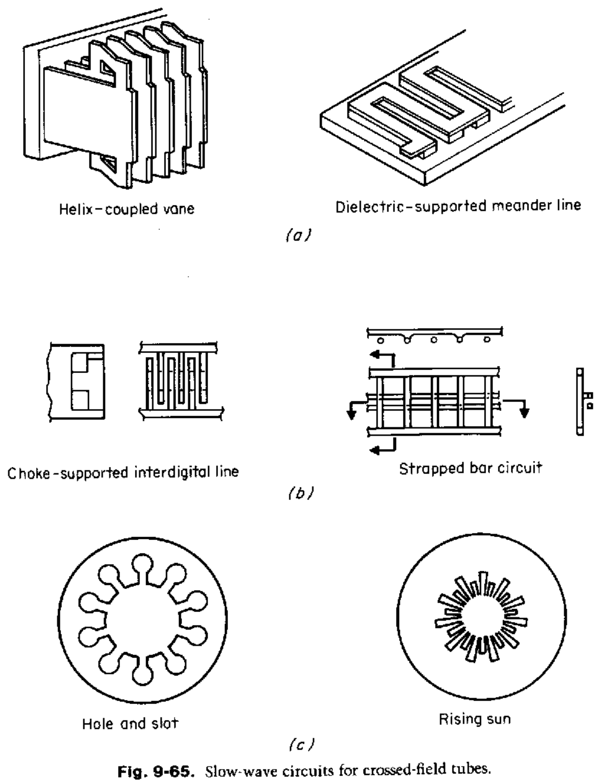


Figure S25 Slow-wave circuits for crossed-field tubes (from Fink, 1982, p. 9.52, reprinted by permission of McGraw-Hill).

SMOOTHING, data. Smoothing is a term describing the reduction of random variation of analog or digital data about its average value. In radar applications it is applied to filtering of measured data, as performed by filters of the α-β, Kalman, or other types. It is an effective technique when a series of data points are obtained in which errors are uncorrelated from point to point or over intervals of a few points (See also **ERROR**; **FILTER.**) *SAL*

Ref.: Kuz'min (1974), p. 399; Barton (1964), Ch. 13.

SPECKLE refers to the spots that form the random structure in a coherent image (speckle structure). The cause of the speckle is the interference of electromagnetic waves scattered from uneven surfaces of objects. The properties of speckle are associated with the nature of the image by the degree of roughness of the surface and by the parameters of the radar.

In radar, the speckle structure is usually manifested when a radar image of the earth's surface is obtained by a synthetic-aperture radar. The speckle structure is an informative feature for identification of surface-distributed objects and may be used for evaluating the characteristics of a radar. In actual tasks of imaging and detection of objects on a surface, the speckle structure is an interfering factor (speckle noise). In this case, smoothing of images is accomplished by noncoherent addition of two or more images (after envelope detection). The resulting improvement depends on the speckle model characterizing the image. A speckle model is a statistical description of the elements (pixels) comprising the radar image, including a description of the type of interaction of the noise and signal from the object of observation.

In terms of type of probability distribution, we distinguish between classical (Rayleigh) model and those with Rice, Weibull, log-normal, or K-distributions. In the classical model, the scattering source is represented by a large number of random independent point reflectors with similar scattering properties and uniformly distributed phases. Speckle statistics in the classical model follow the Rayleigh distribution.

For an analytical description of the interaction of noise and signal, we usually use the model of multiplicative noise, as well as other more complex ones: additive noise modulated by a signal, multiplicative plus additive noise, and others. *IAM*

Ref.: Bernstein R., *IEEE Trans.* ***CAS-34***, no. 11; Trevett (1986).

SPECTROMETER, SPECTROSCOPY, radar. Radar spectroscopy uses radar techniques to investigate the spectral properties of substances that emit or absorb specific frequencies as a result of their atomic or molecular structure. The radar spectrometer is an instrument that determines the distribution of emitted or absorbed energy in the radar band. Analysis of the fine structure of absorption is carried out with coherent signals.

Scatterometers capable of measuring signal amplitudes over a wide range of frequencies are also termed *spectrometers*. (See **SCATTEROMETER.**) *SAL*

Ref.: Currie (1987), p. 12; Popov (1980), p. 406; Skolnik (1990), p. 12.18.

SPECTRUM. A spectrum is "the distribution of the amplitude and phase of the components of a wave as a function of frequency." In radar applications the term is used to describe the radar signal. If the waveform in the time domain is a deterministic function $s(t)$ its spectrum is found as the Fourier transform

$$S(\omega) = \int_{-\infty}^{\infty} s(t)e^{-j\omega t}dt$$

For a random signal, when $s(t)$ is a random function of time (a random process), the concept of power spectrum (power spectral density), $G(\omega)$, is introduced, defined as the Fourier transform of the autocorrelation function of the signal $K_s(\tau)$:

$$G(\omega) = \frac{1}{2\pi}\int_{-\infty}^{\infty} K_s(\tau)e^{-j\omega\tau}d\tau$$

(the multiplying factor before the integral can be different for different cases, depending on the normalization used). The power spectrum describes only the average distribution of signal energy over frequency, without consideration of the phase structure. For a real signal $s(t)$ the function $Ks(\tau)$ is a real, even function of τ, leading to:

$$G(\omega) = \frac{1}{\pi}\int_0^\infty K_s(\tau)\cos(\omega\tau)d\tau$$

and the inverse Fourier transform gives the relationship between K_s(t) and $G(\omega)$:

$$K_s(\tau) = 2\int_0^\infty G(\omega)\cos(\omega t)d\omega \qquad (1)$$

It can be seen from (1) that the area in the power spectrum is equal to the average power $\sigma_s{}^2$ of the random signal, since

$$\sigma_s^2 = K_s(0) = 2\int_0^\infty G(\omega)d\omega$$

The concept of the spectrum is used extensively in the theory of radar signals and signal processing. (See also **TRANSFORM, Fourier**.) The spectra of typical pulse waveforms are illustrated in Table P5 under **PULSE**. *SAL*

Ref.: IEEE (1993), p. 1256; Barkat (1991), p. 89.

SPEED, blind. The blind speed is "the radial velocity of a target with respect to the radar for which the MTI response is approximately zero." In coherent MTI with uniform PRF, a blind speed is the velocity at which the target changes its range by one-half wavelength between pulses. It is a function of radar wavelength λ and pulse repetition frequency f_r:

$$v_b = f_r\frac{\lambda}{2} = \frac{\lambda}{2t_r}$$

When the blind speed is greater than the maximum velocity span of expected targets and clutter, the waveform is referred to as *high PRF*. For an airborne radar, the velocity span extends from the maximum closing velocity of targets approaching from the nose, to the maximum receding velocity of targets or clutter behind the aircraft (radar platform).

An important relationship that has a profound effect on the design of all types of radar is that between unambiguous velocity (blind speed) and unambiguous range. The product of the two is constant:

$$R_u v_b = \left(\frac{c}{2f_r}\right)\left(f_r\frac{\lambda}{2}\right) = \frac{c\lambda}{4}$$

Thus, for a given transmitted wavelength, a large unambiguous range can be obtained at the expense of a small blind speed and vice versa. The equation also shows that use of a sufficiently long wavelength λ would allow a radar to determine, for a given class of targets, both range and velocity unambiguously. Such a solution is seldom practical however, and most radars will be ambiguous in one coordinate or the other. (See also **MOVING-TARGET INDICATION**; **PULSE REPETITION FREQUENCY**.) *PCH*

Ref.: Barton (1988), pp. 234–236.

SPOOFING. Spoofing in ECM is "a type of deception by using an electronic device to transmit a 'target' echo. The spoofing transmitter must operate at the same frequency and PRF as the radar to be deceived." In ECCM it is a tactic in which a radar transmits signals that are not representative of the true radar operational waveform. Spoofing may also include the use of offsite transmitters that emit the actual waveform of the radar but are located at some distance from it in order to decoy antiradiation homing missiles (ARMs). In a tactical situation, proper timing is critical to the success of spoofing techniques, and for this reason the effectiveness of spoofing is often transitory in nature. *PCH*

Ref.: Johnston (1979), p. 67.

SQUINT (see **ANGLE, squint**).

SQUITTER. A squitter is a random reply by a transponder not triggered by an interrogator (i.e., by noise or by an internal triggering system). *SAL*

Ref.: IEEE (1993), p. 1268; Stevens (1988), p. 292.

STABILITRON. A stabilitron is a microwave oscillator in which the electron flux moving in crossed electrical and magnetic fields interacts with the reverse harmonic of the wave. It is distinguished by the combination of a nonclosed slow-wave system with an electron flux closed in a loop. Essentially it is an amplitron connected to an external circuit (Fig. S26). It is distinguished by frequency stability and less pulling of frequency than in the magnetron when there is a change in the nature of the load. *IAM*

Ref.: Dulin (1972), p. 130.

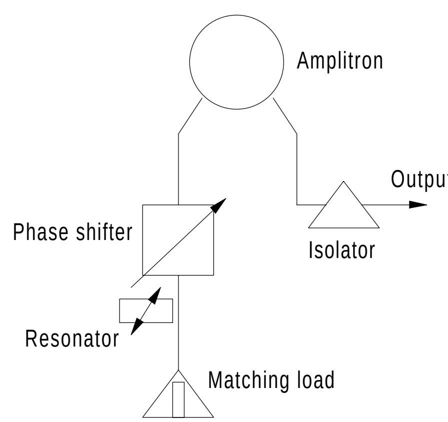


Figure S26 Stabilitron circuit (after Dulin, 1972, Fig. 6-19, p. 130).

STABILITY. Stability is constancy of radar circuits and waveforms. It is measured reciprocally by the level of instabilities that degrade radar performance and is of primary concern in systems using MTI or other doppler processing to reject clutter (See **MTI, limitations to performance.**) *SAL*

STAGGERED PRF (see **PRF**).

STATIONARY TARGET IDENTIFICATION (see **TARGET RECOGNITION AND IDENTIFICATION**).

STEALTH. Stealth technology is the means of reducing the reflections and emissions of vehicles to avoid detection by radar and other sensors. A vehicle to which this technology has been successfully applied is called a stealth vehicle, and when used in military applications it is expected to operate undetected within the normal coverage volume of defensive systems. The technology was developed primarily for stealth aircraft (Fig. S27), and is based on several fundamental principles:

(1) The use of passive (absorber) and active systems for RCS reduction.

(2) The use of shaping to produce minimum scattering, concentrated in a few directions away from the direction of incidence (e.g., fabrication of almost perfect joints between surfaces, minimally inclined surfaces).

(3) The avoidance of right-angle intersections that would support corner reflection effects.

(4) The use of synthetic materials transparent to microwaves (e.g., carbon fibers) in large surfaces like wings.

(5) Additional reduction and masking measures such as installation of gas exhausts on the top of the fuselage, suppression of infrared signatures, installation of ordinance and countermeasures inside the airframe, and so forth.

Figure S27 Photograph of a typical low-RCS aircraft (from Bhattacharyya, 1991, Fig. 4.84, p. 233).

The techniques applied to stealth aircraft are summarized in Fig. S28. The reduction in RCS achievable with stealth technology can be 30 dB or greater, corresponding to detection ranges ≤ 18% of those for normal target RCS, but this reduction is accompanied by considerable increase in cost of the vehicle. *SAL*

Ref.: Neri (1991), p. 249; Bhattacharyya (1991), p. 235; B. Sweatman, *Stealth Aircraft,* Motorbook International, 1986.

STEERING (see **BEAM**; **PATTERN, array**).

STEP FUNCTION. The step function changes its values only in a discrete sequence of points of a break, in which the values of the function can be either definite (given) or indefinite. Most often an unit step function is used in the form of an unit signal at the origin of the coordinates:

$$u(x) = \begin{cases} 0, & x < 0 \\ 1, & x \geq 0 \end{cases}$$

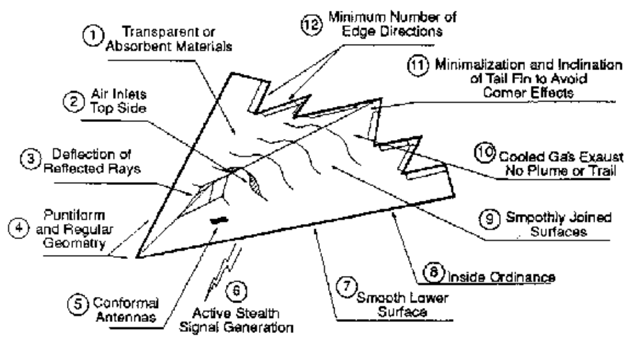


Figure S28 Stealth techniques applied to low-RCS aircraft (from Neri, 1991, Fig. 3.26, p. 250).

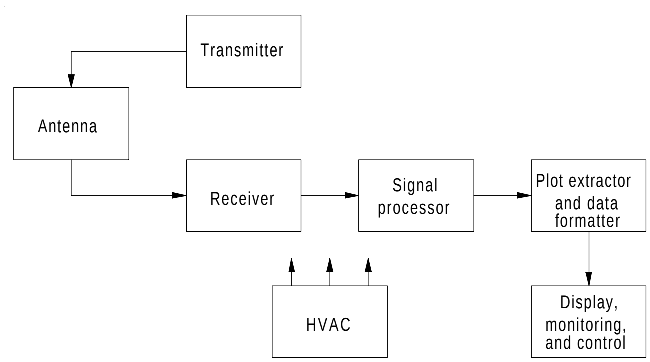
The reaction of a system to an input signal of the type of an individual step function is called the *unit step response* of the system. *IAM*

STRIPLINE (see **TRANSMISSION LINE**).

STROBE. A strobe is (1) an impulse applied to such devices as an analog-to-digital converter to control the sampling of the signal; or (2) a jam strobe, defined as an "indication of jammer azimuth bearing, one form being a marker on the radar PPI display." *SAL*

Ref.: IEEE (1993), p. 1300.

SUBSYSTEM, radar. A radar subsystem is a major constituent of a radar system, controlling its operational capabilities. The main radar subsystems are the antenna, transmitter, receiver, and signal processor. The last may be integrated with the receiver but is usually considered a separate subsystem, especially when implemented digitally. Other elements, such as a plot extractor (tracker) with output data formatter, display, monitoring, and control equipment are considered separate subsystems, along with special-purpose equipment (e.g., heating, ventilation, and air conditioning (HVAC)). A typical radar block diagram by subsystems is shown in Fig. S29. *SAL*



FigureS29 Block diagram of typical radar by subsystems.

SUPERHETERODYNE (see **RECEIVER, heterodyne**).

SUPERPOSITION. The superposition principle states that in any physical system, if the system is acted upon by a number of independent influences (causes), the resultant influence (effect) is the sum (vector or algebraic) of the individual influences (causes). The principle of superposition does not apply to nonlinear systems. *PCH*

Ref.: Van Nostrand (1983), p. 2,728.

SUPPRESSION

Defense suppression is the action of offensive forces in physically destroying a defensive system. It is also identified by the acronym SEAD (suppression of enemy air defenses). Applied to radar it often takes the form of antiradiation missile (ARM) attack. *SAL*

Ref.: Schleher (1987), p. 22.

sidelobe suppression (see **SIDELOBE, range**; **BLANKING, sidelobe**; **CANCELER, sidelobe**).

Small-signal suppression refers to the nonlinear effect of receiver circuits (limiters, detectors) in which the largest signal present acts to suppress smaller signals. When the signal-to-noise (SNR) ratio drops near or below unity, the effect is to reduce the SNR even further. The effect is also characterized as detector loss (see **LOSS**) or the capture effect. *DKB*

Ref.: Barton (1988), pp. 64, 467; Vaccaro (1993), pp. 92, 135.

SURVEILLANCE, radar. Radar surveillance is the periodic observation of a specified volume of space, typically to maintain cognizance of selected traffic, e.g., for air traffic control or air defense (See **RADAR, surveillance.**)

Adaptive surveillance can be used when some areas are found to have higher priority by virtue of greater numbers of target detections, requiring higher revisit rates.

Distributed surveillance is a concept based on sharing of data from different sensors operating in the same volume but in different regions of the electromagnetic spectrum (ultraviolet, optical, infrared, radar), from ground or moving platforms (spacecraft, aircraft, ships). A system using distributed surveillance can be flexible, using all or some of the available platforms and sensors depending on efficiency and cost trade-offs. This approach permits optimization of system coverage (from global to more localized regions), and of performance in such roles as low-RCS target (stealth target) detection. *SAL*

Ref.: IEEE (1993), p. 1,315; Long (1992), p. 379.

SWERLING MODEL (see **RCS fluctuation; LOSS, fluctuation**).

SWITCH, SWITCHING:

electronic feed switching (see **SCANNING, electronic**).

lobe switching (see **LOBING**).

A **microwave switch** is used to make or break microwave electrical circuits. The ideal switch would have zero voltage drop when closed, regardless of the current flow, and zero current when open, regardless of the applied voltage. Switches are classified as mechanical or electronic, the latter based on passive or active elements. Most commonly used are active electronic switches based on transistors or diodes as the switching elements. The significant parameters are switching speed (often specified as the time to reach the 3-dB response point when turning on, and to reach 90% of full attenuation when turning off), on-off ratio, insertion loss, and bandwidth. Electronic microwave switches can have switching speeds from microseconds to tens of nanoseconds, on-off ratios $\geq$ 60 dB, insertion loss of a few decibels, and bandwidths in excess of one octave. *SAL*

Ref.: Fink (1975), pp. 7.86, 16.10–16.18; Wiegand (1991), p. 79.

A **transmit-receive (T/R) switch** is "an automatic device employed in a radar for substantially preventing the transmitted energy from reaching the receiver but allowing the received energy to reach the receiver without appreciable loss." It is sometimes accompanied by an anti-T/R (ATR) switch that is "a radio-frequency switch that automatically decouples the transmitter from the antenna during the receiving period; it is employed when a common transmitting and receiving antenna is used." (See also **DUPLEXER**). *SAL*

Ref.: IEEE (1993), pp. 46, 1,406.

SYNCHRONIZER. A synchronizer is a device delivering timing pulses to various radar subsystems, ensuring synchronous operation of these subsystems. The radar synchronizer produces timing pulses (triggers) by introducing delay, by analog or digital means, in a starting trigger (pretrigger) which initiates the radar cycle. Many modern radars incorporate a digital synchronizer into the signal processor (Fig. S30). In a digital synchronizer the clock shifts the initial timing pulse through shift registers, the number of which is determined by the required delay. An additional analog delay may be introduced as necessary to provide fine adjustment to any point between clock pulses. *SAL*

Ref.: Fink (1982), p. 25.72.

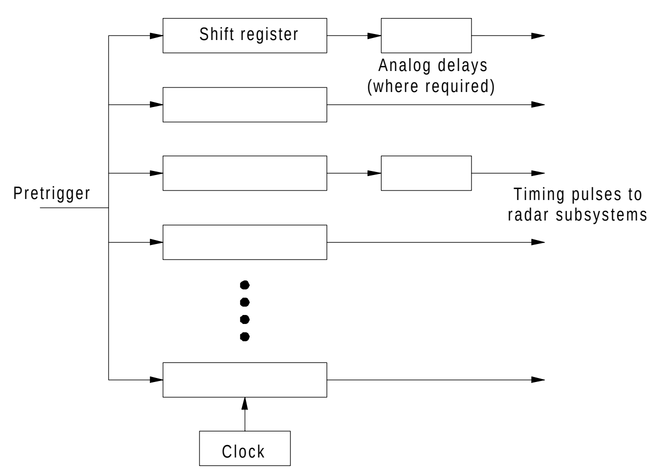


Figure S30 Block diagram of radar synchronizer (after Fink, 1982, Fig. 25.83, p. 25.73).

SYNTHESIZER

A **delay synthesizer** is an analog or digital device intended to generate a reference (tracking) pulse delayed relative to the transmitted pulse by a time proportional to the extrapolator inputs. In terms of method of delay generation, we distinguish between phase and pulse synthesizers. The former are used with a sinusoidal master oscillator and are based on a phase shift proportional to the required delay of initial sinusoidal oscillations, sharp pulses being formed from the shifted sinusoid. Pulse synthesizers are used both for sinusoidal and for pulse master oscillators. They contain an electronic delay circuit that operates on the basis of comparison of a linearly increasing voltage (generated, for example, by a fantastron, starting with the transmitted pulse at the input) with the range voltage, which arrives from the output of the extrapolator. Pulse analog synthesizers do not assure high precision and are used in radars when the range does not exceed several tens of kilometers. Phase analog synthesizers provide high precision over a wide range of velocities, but they have a bulky electromechanical system of the selsyn-transformer type as the phase shifter.

In comparison with the analog synthesizer, digital delay synthesizers make it possible to obtain high precision of the tracking measurement device. Pulse digital synthesizers in a counter/shaper operate like analog ones, by the principle of voltage comparison. Pulse digital synthesizers based on a controlled delay line contain switching circuits controlled from discharge flip-flops. The first type of synthesizer is easier to implement for large delays, the second for short ones, so a combination of synthesizers of both types is used, arranging them in series.

Besides digital synthesizers for simple signals, delay synthesizers for complex signals are also used. These ensure a given time shift while retaining the signal shape. A synthesizer of such a type is designed in a circuit of a controlled delay line and replaces a pulse delay synthesizer and functional oscillator which generates a functional signal of a given shape.

Delay synthesizers are widely used as one of the basic elements of tracking range measurement devices. *IAM*

Ref.: Dulevich (1978), pp. 327, 339.

A **frequency synthesizer** is a frequency generator that produces sinusoidal oscillations of the required frequencies or frequency, controlled by a specific program. The synthesizer is made with either digital or analog control. The latter is usually a voltage-control oscillator. The control of a digital synthesizer is by digital (binary) code. There are three basic types of digital frequency synthesizers: harmonic frequency synthesizers, and open- and closed-loop synthesizers based on controlled frequency division. The first type of synthesizer contains a harmonic oscillator and serial chain of stages consisting of a commutator, mixer and frequency divider.

A harmonic digital synthesizer is heavy and bulky, and its production technology is complex.

The open-loop digital frequency synthesizer based on a controlled divider has a reference frequency oscillator and digital frequency divider with variable division factor, based on counters/dividers. The drawbacks of this type of synthesizer include rounding errors with conversion of the controlling number to the division factor. The closed-loop synthesizer, which constitutes a conventional phase-locked circuit with tunable divider in a feedback circuit, does not have this drawback.

Frequency synthesizers are widely used in STALOs of modern radars. *IAM*

SYNTHETIC APERTURE RADAR (see **RADAR, synthetic aperture**).

T

TAPERING is the weighting used to adjust the aperture illumination, reducing it near and at the edges of the aperture to control sidelobes. (See **APERTURE illumination.**) *SAL*

Ref.: Silver (1951), pp. 179–199.

TARGET, radar. A radar target is “(1) specifically, an object of radar search or tracking; (2) broadly, any discrete object that scatters energy back to the radar.” Usually the object producing a desired radar echo is termed a target, while other objects are considered clutter. A given object may thus be a target for one radar (e.g., weather, for a meteorological radar) while it is clutter for another (e.g., an air traffic control radar). Radar targets are classified as simple and complex, and the latter as discrete (point) or distributed targets, depending on their extent relative to the radar resolution cell. Most targets of interest for conventional radars (aircraft, ships, satellites, etc.) are point targets, while clutter is distributed (over a surface for land or sea clutter, within a volume for precipitation or chaff). The main characteristics of targets are

(1) Radar cross section (RCS).

(2) Fluctuation parameters (pdf, correlation time and frequency).

(3) Trajectory parameters.

From the viewpoint of location or trajectory, they can be airborne, surface based (land or marine targets), or space targets. (See also **RADAR CROSS SECTION.**) SAL

Ref.: IEEE (1993), p. 1,338.

A **calibration target** is used for calibrating and pointing a radar. Most used are reflectors such as flat plates, spheres, dihedrals, trihedrals, and Luneburg lenses. These and other shapes are shown in Fig. T1. (See also **CALIBRATION.**)

target capacity (see **THROUGHPUT CAPACITY**).

A **complex target** is “a target composed of more than one scatterer within a single radar resolution cell.” The distinction between this term and distributed target should be noted (a target may be both complex and distributed).

SAL

Ref.: IEEE (1993). p. 226.

A **distributed target** is “a target composed of a number of scatterers, where the target extent in any dimension is greater than the radar resolution in that dimension.” Different clutter sources (ground, sea, rain, etc.) are examples of distributed targets for radars seeking to observe them. Distributed targets are further classified as surface-distributed (land, sea) or volume-distributed (precipitation or chaff). The term *extended target* is also used to describe the distributed target. *SAL*

Ref.: IEEE (1993), p. 375; Ostrovityanov (1985).

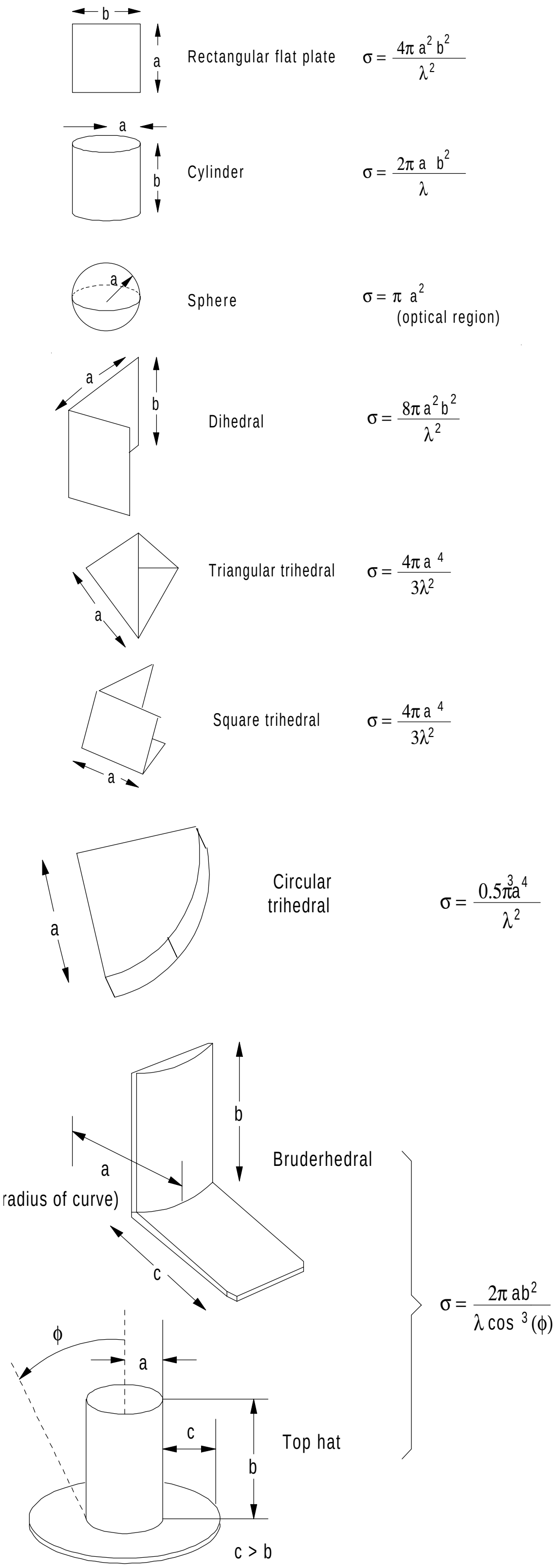


Figure T1 Characteristics of several typical reference reflectors (after Currie, 1989, Fig. 4.10, p. 107).

dumbbell target (see **two-point target**).

An **elementary target** is the target of the simplest configuration (see **RCS of simple shapes**). *AIL*

Ref.: Finkel'shteyn (1983), pp. 129–146.

extended target (see **distributed target**).

A **false target** is a detection resulting from interference, rather than from an object on which detection is desired.

Target fluctuation is "variation in the amplitude of an echo from a complex target caused by a change in target aspect angle, vibration of target scattering sources, or changes in radar wavelength; the amplitude component of target noise. The terms *scintillation* and *amplitude noise* have been used in the past as synonymous for target fluctuation and also to denote location errors caused by target fluctuation, and should be avoided because of their ambiguity." If the target produces fluctuations in its radar return, it is called a *fluctuating target*, and, if not, a *nonfluctuating* or *steady target*. The latter is hardly ever observed in real radar operation and is used mainly in theoretical analysis. Target fluctuation is usually described by Swerling models. (See **RCS fluctuations.**) *SAL*

Ref.: IEEE (1993), p. 1,338.

ghost target (see **GHOST**).

target glint (see **GLINT**).

target identification (see **TARGET RECOGNITION AND IDENTIFICATION**).

Target lock-on is the stage of acquisition of radar information from a detected target in which the trajectory parameters of the target are observed to a level sufficient for switching to the tracking mode. Several lock-on cycles are used, in each of which a pulse or pulse train (usually of the same type as the tracking signal) is sent. The number of pulses in a train is that necessary to obtain a sufficiently high probability of detection ($P_d > 0.9$) with the false-alarm probability in the resolution cell on the order of 0.01 to 0.1. In the lock-on cycle, a target search is conducted in a region whose dimensions are determined on the basis of designation data or results of the initial detection or preceding lock-on cycles) and on the timing of the train. If several detections appear in the lock-on region, then one is selected which is closest to the trajectory of the target (usually in the range coordinate). If there are no detections in the lock-on region, then the initial detection is considered unvalidated, and the target is false. More complex criteria of the *m*-out-of-*n* type, and rejection of a false trajectory based on the lack of detections in several lock-on cycles, are also possible. Usually a total of 4 to 5 lock-on cycles are used. *IAM*

Ref.: Savresov (1985), p. 70.

A **meteorological target** is an atmospheric phenomenon which the radar seeks to observe: cloud, precipitation, thunderstorm, tornado, clear-air turbulence, and so forth. (See **RADAR, meteorological.**) *SAL*

Target missing is the situation when the radar detector makes a decision that there is no target in the searched volume of space when the target is actually present. The sum of the probability of detection and the probability of target missing is equal to unity. *SAL*

Ref.: Leonov (1988), p. 11.

Target noise is "random variations in observed amplitude, location, and doppler of a target, caused by changes in target aspect angle, rotation, or vibration of target scattering sources, or changes in radar wavelength." (See **RCS fluctuation**; **GLINT**; **ERROR, scintillation.**) *SAL*

Ref.: IEEE (1993), p. 1,339; Skolnik (1970), Ch. 28; Barton (1969), Ch. 6.

A **nonfluctuating target** is one containing a single scatterer of constant RCS. Only a sphere, a point target, or a constant-amplitude repeater or transponder with a smooth antenna pattern qualifies as a nonfluctuating target. (See **target fluctuation.**) *DKB*

A **phantom target** is "(1) an echo box or other reflection device that produces a particular blip on the radar indicator, or (2) a condition, maladjustment, or phenomenon (such as a temperature inversion) that produces a blip resembling those of targets for which the system is being operated." *SAL*

Ref.: Johnston (1979), p. 64.

A **point target** is one occupying a space much smaller than the radar resolution cell. The concept of a point target is a convenient idealization having the following properties:

(1) The locus of phase- (wave-)fronts is a sphere with its center at the target position.

(2) The amplitude of the scattered electric field is uniform over the wave front.

(3) The normal to the phase-front passes through the target.

The ideal point target is nonfluctuating and does not broaden the width or the spectrum of the radar transmitted pulse. *AIL*

Ref.: Vasin (1977), p. 84.

Rayleigh target (see **RCS fluctuation**).

target recognition (see **TARGET RECOGNITION AND IDENTIFICATION**).

target scintillation (see **target fluctuation**).

Target signatures are the target features depicting the information about physical parameters of the target, the type of the target, scattering characteristics, and so forth. Typically, they are used for target recognition. (See **TARGET RECOGNITION AND IDENTIFICATION.**)

A **simple target** is one "from which the reflected signal is not resolvable in terms of interfering echoes from two or more discrete echoing elements." Examples are the sphere and the half-wave dipole.(See **RCS of simple shapes.**) *SAL*

Ref.: IEEE (1993), p. 1,225.

Target suppression is the loss in detectability of targets arising from other interfering target or clutter signals in the reference cells of a CFAR system. It can be minimized by removing large returns from the calculation of the threshold or by reducing effects of such returns with limiters or nonlinear (e.g., logarithmic) weighting of CFAR inputs. *SAL*

Ref.: Skolnik (1990), p. 8.17.

The **two-point target** is a simple model for a complex target consisting of two scattering centers connected by a nonreflecting, rigid rod. The average RCS of the two-point target is the sum of the individual scatterer RCS values. The glint characteristics of this target form the basis for theoretical models of more complex objects, and have been used extensively to model measurement errors. *SAL*

Ref.: Barton (1975), pp. 13–22; Skolnik (1970), p. 28.8; Vasin (1977), p. 85; Ostrovityanov (1985), p. 5; Barton (1988), pp. 115–120.

TARGET RECOGNITION AND IDENTIFICATION. An important feature of modern radars, in addition to the ability to detect and measure target position, is the ability to classify the type of target and if possible to identify the particular target being observed. There are many levels of classification, recognition, discrimination, and identification that have been discussed in the literature and attempted, with greater or lesser success, in deployed radars. The definitions of these levels is important in understanding radar performance in this area:

(1) Target identification is "the knowledge that a particular radar return is from a specific target. This knowledge may be obtained by determining size, shape, timing, position, maneuvers, rate of change of any of these parameters, or by means of coded responses through secondary radar." The implication in this definition is that the target is not only determined as specific type (e.g., F-16 aircraft), but that the aircraft tail number (flight number) or name of a ship target has been determined. It should be apparent that this can only be expected through active emissions from the target, such as a coded transponder response or emissions in other bands that carry detailed information limiting the target to a specific vehicle. In theory, given an extensive data base on vibrations, accidental modulations, and details of shape of all potential targets, it might be possible to distinguish among several aircraft, ships, or other vehicles of the same class. However, it requires a considerable stretch of imagination to conclude that identification in this sense is operationally realizable from radar echo signals.

(2) Discrimination is "separation or identification of the differences between nonsimilar signals." The term has been applied to processes intended to select threatening targets (e.g., reentry vehicles) while rejecting (discriminating against) accompanying objects such as rocket fragments, decoy, and chaff. Obviously if the signals are sufficiently nonsimilar, radar processing should be able to accomplish

this objective. A challenge to radar designers is to find waveforms and processes that will expose subtle differences (e.g., when the spread of RCS and trajectory parameters of RVs and decoys overlap).

(3) Classification has no precise definition, but is generally understood to imply that the type or model of the target is determined (e.g., F-16 aircraft vs. MiG-29; tank vs. personnel carrier; frigate vs. cruiser). It can be interpreted as coarsely as distinguishing fixed-wing from rotary-wing aircraft, or aircraft from surface vehicles. Again, this capability has been demonstrated and should be expected in properly designed radar systems if they conduct detailed measurements on the target. Classification during the brief dwells characteristic of surveillance radars will be of the coarser varieties (e.g., large vs. small aircraft). It is sometimes assumed that the finer levels of target classification will answer the "friend or foe" question, but in an era when many forces fly the same aircraft types and operate the same models of surface vehicles that may not be a realistic expectation.

(4) Recognition is another general term lacking precise definition, but it has been defined by leading Russian experts in this field as "determination of the type or class of an object."

These several terms and the methods of implementing them with radar (as opposed to ESM, ELINT, and infrared or optical imaging) are discussed in the following articles. *DKB*

Ref.: IEEE (1993), pp. 367, 614; Gorelik (1990); Nebabin (1995).

Target classification requires that the radar measure with sufficient accuracy a set of target parameters that will permit it to be accepted as a member of (or rejected as not belonging to) a class of objects the system is intended to detect, and to which it is intended to react. The classes may be broad or narrow, but will generally fall within those shown in Fig. T2.

Target discrimination is the term usually applied to processes in an antiballistic-missile (ABM) radar to select the threatening reentry vehicle from the potential cloud of arriving objects, most of which will be fragments of the booster or final missile stage, or decoys intentionally added to the arriving traffic. The methods used for this include measurement of

(1) Ballistic coefficient (weight-to-drag ratio) and its variation with altitude, by observing the deceleration of the target in the upper atmosphere.

(2) RCS and its fluctuations.

(3) Presence, velocity, and velocity spread of the ionized wake behind the target.

(4) Range profile of the target, using wideband waveforms.

(5) Two-dimensional profile using a combination of high resolution in range and inverse synthetic aperture (ISAR) processing in cross-range.

(6) Inherent resonances in target response, by observation over a broad bands in the VHF region.

(7) Polarization ratio of the RCS or individual scattering elements on the target.

These and other methods can be classified as shown in Fig. T3.

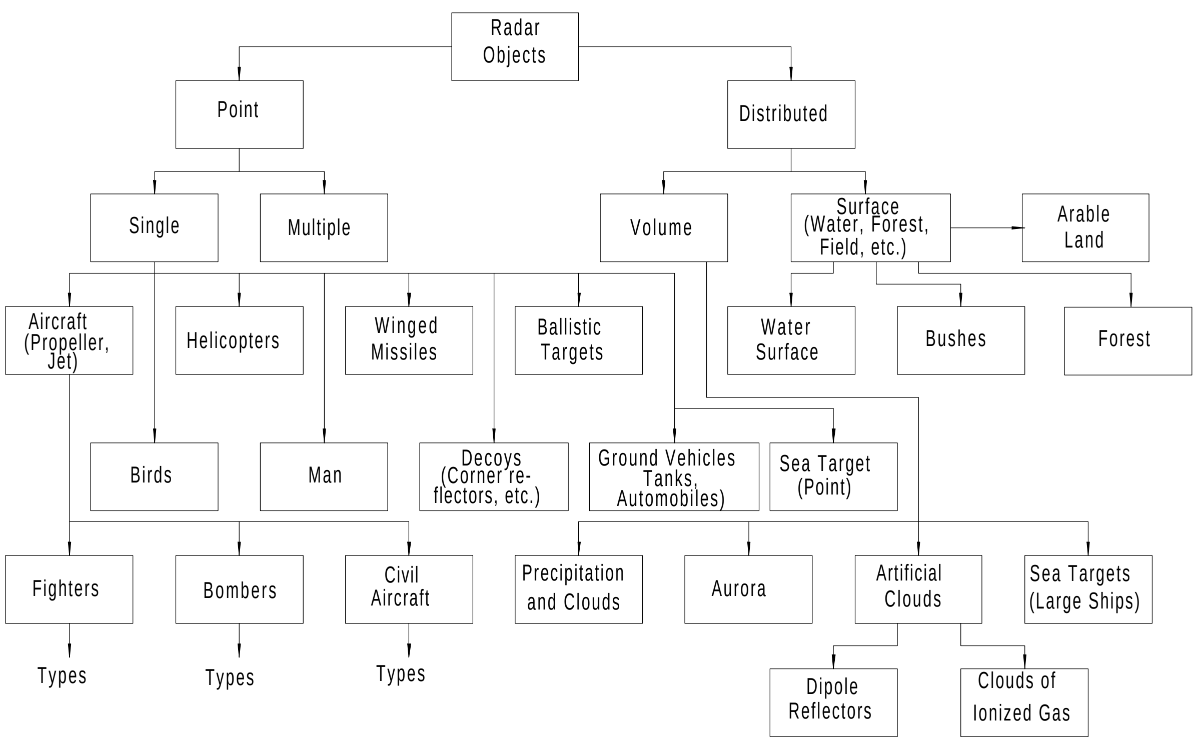


Figure T2 Classification of radar targets (from Nebabin, 1995, Fig. 16, p. 12).

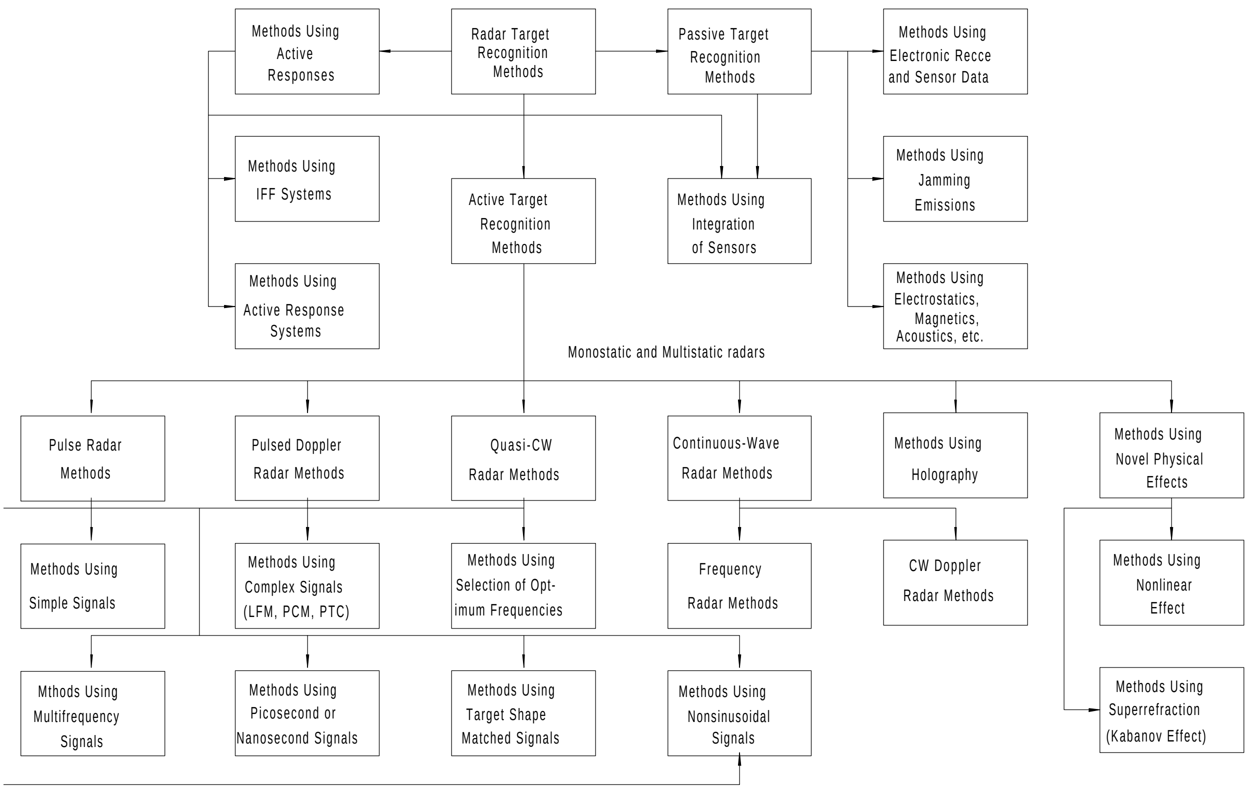


Figure T3 Radar recognition methods (from Nebabin, 1995, Fig. 1.21, p. 22).

The radar designs to accomplish these measurements at adequate range on multiple targets with low RCS are complex and expensive and have challenged designers in the major nations concerned with ballistic missile defense. The radar (and associated system software) problems have been regarded as so severe that only limited deployments have been made, and even these have been limited by treaty terms. (See **RADAR, instrumentation**.) *DKB*

Ref.: Fagen (1978), pp. 455; Freeman (1995), Ch. 6.

Target identification is a procedure or process used by a radar to positively identify a target that has been detected within its search volume. Cooperative target identification relies on the Identification Friend or Foe (IFF) technique developed during World War II, in which each "friendly" aircraft is equipped with a transponder. The search radar, through its IFF transmitter and antenna system, interrogates the target and the transponder aboard the target is triggered to respond with an encoded signal at the correct frequency. The response is usually visually displayed as an IFF symbol. The IFF system is used in most air forces of the world. All commercial airline operations use a system of continuous beacon tracking based on the IFF concept, referred to as secondary radar.

IFF systems can be classified as cooperative (designed to respond via a cooperative transponder that indicates that they are friendly to the interrogation equipment), or noncooperative (designed to use the radar echo itself as the indication). IFF is fundamental to military operations. Sometimes these systems are called identification, friend, foe, or neutral (IFFN). To some extent acronym IFF (or IFFN) is a misnomer, as the system can identify friendly targets but cannot positively identify enemy or even neutral platforms. Examples of U.S. IFF configurations are MK (Mark) X IFF, MK XII IFF, and MK XV IFF.

Noncooperative target identification applies in cases for which there is no communication between the radar and the target. Here the radar and its operator attempt to correlate detectable target parameters such as speed, altitude, and target signal characteristics such as amplitude and modulation, with those of known (stored) target types. The success of noncooperative target identification depends greatly upon the complexity of the radar, the sophistication of the signal processing system, and the time available for target observation. For example, a coherent CW or pulsed doppler radar may be capable of detecting doppler sidebands due to propeller or jet engine modulation (JEM lines) that may identify the target. Also, 2D or 3D radar imaging of the target may be possible with high resolution radars using inverse synthetic-aperture radar (ISAR) techniques. These and other noncooperative target identification techniques have been under study for decades, and despite some positive experimental results, there is little evidence that these techniques have achieved the reliability required for day-to-day radar operation in the field. *PCH, SAL*

Ref.: IEEE (1993), p. 614; Stevens (1988); Long (1992), p. 357; Nebabin (1995).

Target recognition requires at least classification of general target type (e.g., bomber, fighter, or civil aircraft; fixed or moving surface target), and possibly a finer classification by vehicle type or nomenclature (e.g., F-16, MiG-29; DDG, DLG; tank, personnel carrier). The many approaches to this capability are shown in Fig. T3. All these methods except the active response options in the upper left of the figure come under the general term noncooperative target recognition (NCTR), and all those in the lower portion of the figure use echo response to active radar.

The presence of recognizable lines in the target doppler spectrum has been the basis for NCTR on aircraft. These lines come primarily from rotating propellers or turbines, and in some cases from scanning antennas, and they provide information about the type of aircraft, including the propeller blade rate or the type of jet engine. The spectrum from a helicopter would appear (over a period longer than the reciprocal of the blade rate) as shown in Fig. T4. Typical measured spectra from a propeller aircraft and a jet aircraft are shown in Fig. T5.

The use of doppler spectra to recognize ground targets depends on phenomena similar to those used for aircraft, except that the observed spectrum may result from vibration or moving wheels or tracks.

For ground targets that are not moving, a series of stationary target recognition (STR) techniques have been developed (referred to by some authors as stationary target identification, or STI). These are based more on geometrical considerations, using high resolution radar and polarization measurements. Table T1 shows the division of the problem into three phases: detection, discrimination, and recognition (note that the terminology differs from that defined above).

STR is based on standard amplitude threshold techniques, using a CFAR processor. Difficulties can occur when clutter in the reference cells contains target-like objects or a discontinuity in clutter reflectivity (e.g., a transition from open field to woodland, when the target is located close to a treeline to escape detection by radar or visual means). Additional discriminants may then be needed, as listed in Table T2.

Stationary target recognition typically involves two stages: initial classification into generic classes, followed by recognition within the class (as listed in Table T2). The procedures involve target feature extraction, preclassification (discrimination), and classification, in which the input features are assigned to one or several classes with corresponding probabilities. The most important stage of target recognition is feature extraction, as performance is determined to a great extent by the quality of the features used, and even the best classifier cannot compensate for poor or incorrect features.

The main techniques in stationary target recognition are high resolution (profiling) and full polarization reception (measuring the polarization scattering matrix), or a combination of the two. Millimeter-wave radars and SAR techniques are the most suitable for both range and cross-range resolution. The results of a high-range-resolution millimeter-wave radar (with 0.5m resolution) are shown in Fig. T6.

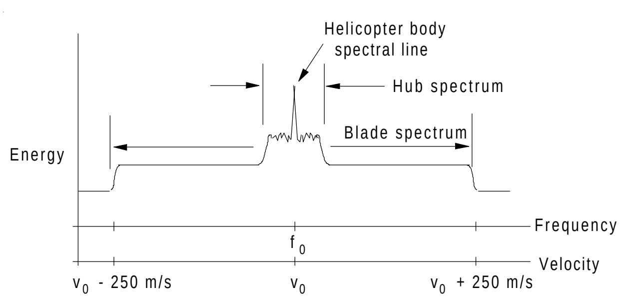


Figure T4 Idealized helicopter power spectrum (after Currie, 1987, Fig. 6.32, p. 273).

Table T1
Stationary Target Identification

Phase	Purpose
Detection	Separation of targets from noise and benign clutter
Discrimination	Elimination of strong target-like clutter returns
Recognition	Recognize targets
Classification	Determines generic target classes (e.g., trucks, tanks)
Identification	Specifies targets within generic classes (e.g., M-48, T-62)
(from Currie, 1987, Table 6.2, p. 279).	

Table T2
Target-Clutter Discrimination

Three classes of target-clutter discrimination algorithms
Scalar discriminants (not using any polarization data)
Clutter decorrelation techniques (frequency-induced and natural)
Spatial feature algorithms
High resolution techniques
Vector Discriminants (using limited polarization data)
Depolarization techniques
Depolarization techniques with frequency agility
Matrix discriminants
Null-polarization techniques
Matrix polarization techniques
Mueller and density matrix decomposition techniques
(after Currie, 1987, Table 6.3, p. 281)

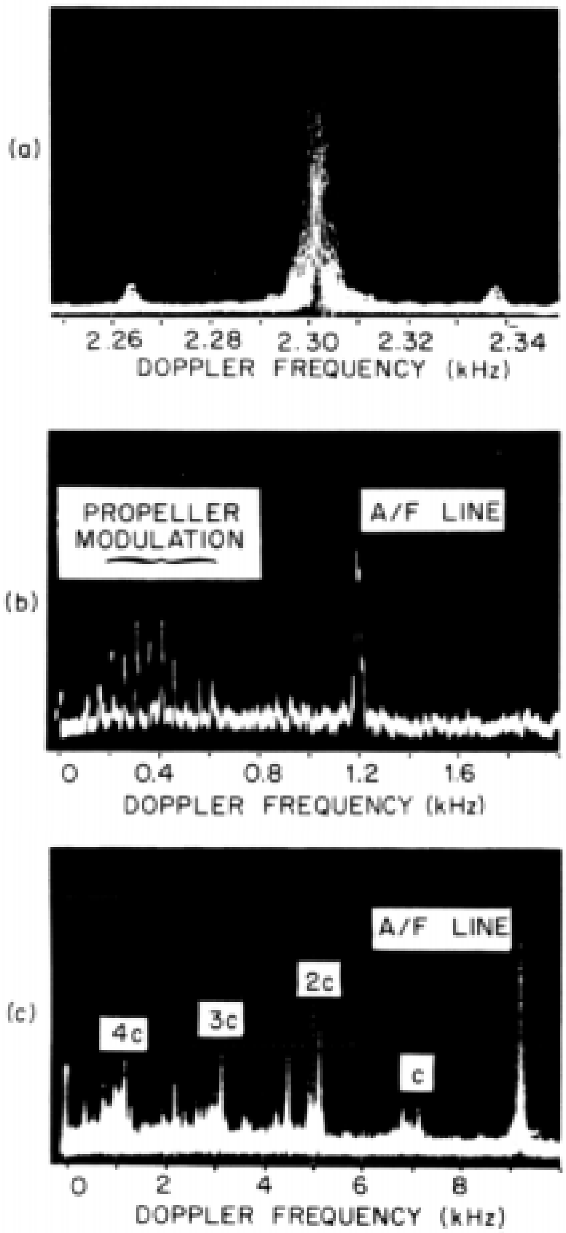


Figure T5 Pulse-doppler spectrum measurements showing (a) an AM sideband pair caused by a conical-scan radar antenna; (b) FM modulation from propellers of a DC-7 aircraft; and (c) jet engine modulation at the compressor rotation rate and its harmonics (from Skolnik, 1970, Fig. 17, p. 28.22).

Recent results in processing and interpretation of SAR data have shown that it is necessary to retain both amplitude and phase data if interpretation if to be reliable. The difficulties in reliance on conventional intensity-based methods are attributed to the presence of creeping waves and multiple reflections, which distort the profiles of many man-made targets. *DKB, SAL*

Skolnik (1970), pp. 28.17–28.23; Currie (1987), pp. 269–310; Nebabin (1995); Rihaczek (1996).

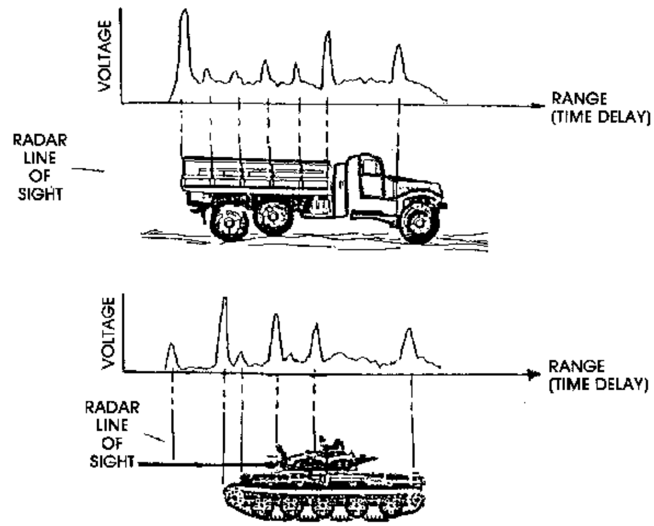


Figure T6 Ultra-high-resolution profiles of stationary targets (from Currie, 1987, Fig. 6.32, p. 273).

TEE

A **coaxial tee** is a power divider constructed as shown in Fig. T7. The width of the center conductor is increased to compensate for capacitive reactance at the splitting point.

Ref.: Lavrov (1974), p. 348.

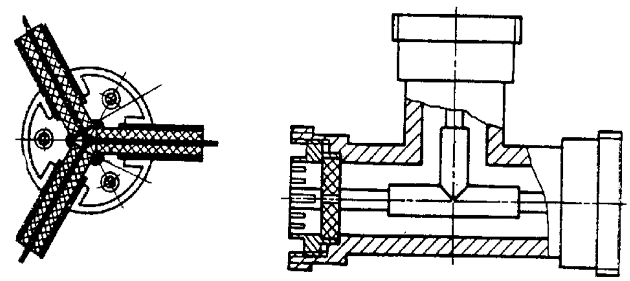


Figure T7 Coaxial tee configurations: (a) (after Lavrov, 1974, Fig. 18.43, p. 348.

A **waveguide tee** is a power divider constructed of waveguide, as shown in Fig. T8. Splitting in the H-plane is shown in Fig. T8(a) and (b), and in the E-plane in Fig. T8(c) and (d). The H-plane tee has higher power-handling capacity, while the E-plane tee has wider bandwidth, although both are narrowband devices.

Ref.: Lavrov (1974), p. 348.

TEMPERATURE, noise. Noise temperature is the temperature of a fictitious input termination resistor that would account for the noise actually observed at the output of a network. The spectral density of available noise from such a termination at temperature T is

$$N_0 = kT$$

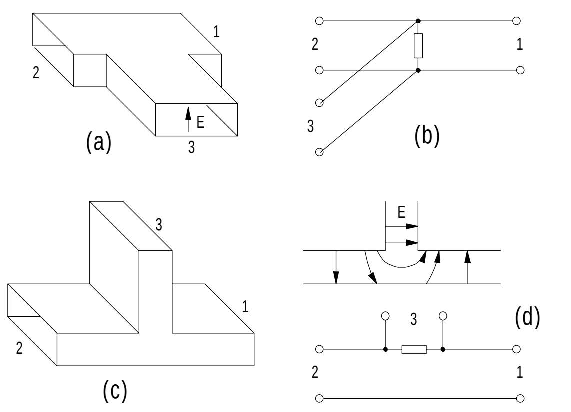


Figure T8 Waveguide tee configurations: (a) H-plane tee, (b) equivalent circuit of H-plane tee, (c) E-plane tee, (d) equivalent circuit and fields in E-plane tee (after Lavrov, 1974, Fig. 18.43, p. 348.

The concept of noise temperature is derived from Nyquist's theorem, which relates the thermal noise voltage to the temperature of a resistive circuit element. The noise temperature of a simple resistor is its physical temperature, but that of a complex network can differ from its physical temperature, and is sometimes called the effective noise temperature. The word *effective* is generally omitted discussions of radar and will be in the following articles.

The use of noise temperature to describe the performance of radar receiving systems can be traced to the early work at the MIT Radiation Laboratory (Lawson, 1950, pp. 103–108), and has been systematized by Blake in his Naval research Laboratory reports and textbook (1980). Those procedures are now widely used in radar analysis and can be summarized in the equation for (effective) system noise temperature T_s:

$$T_s = T_a + T_r + L_r T_e$$

where T_a is the antenna (noise) temperature component; T_r the component due to loss, L_r, in the transmission line connecting the antenna terminal to the receiver; and T_e the temperature of the receiver itself, multiplied by L_r to refer it back to the antenna terminal. *SAL, DKB*

Ref.: Lawson (1950), pp. 103–108; Blake (1980), Ch. 4; Skolnik (1980), p. 345, (1990), p. 2.26.

Antenna noise temperature is "the temperature of a black body which, when placed around a matched antenna that is similar to the actual antenna, but loss-free, produces from the antenna the same available power, in a specified frequency range, as the normal antenna in its normal environment." It is the component of system noise temperature passed by the antenna to the subsequent receiving components, normally denoted by T_a, and arises from three sources:

(1) The sky temperature T_a' reduced by ohmic loss L_a within the antenna structure.

(2) The temperature T_0 of the surrounding surface, picked up in earthwards-directed sidelobes of the antenna pattern.

(3) The noise temperature of the antenna loss L_a, assumed also to be at a physical temperature T_0.

Blake modeled the antenna by assuming that 88% of its integrated power pattern would be directed skyward (terminating at T_a'), the remaining 12% directed earth-ward (terminating at T_0 = 290K). The result is

$$T_a = \frac{0.88 T_a' - 254}{L_a} + 290$$

DKB, SAL

Ref.: IEEE (1993), p. 45; Blake (1980), p. 172.

The **receiver noise temperature** characterizes the noise component generated within the receiver in terms of the temperature T_e of a fictitious input termination resistor that would account for that noise component actually observed at the receiver output. It is related to receiver noise factor (figure) F_n:

$$T_e = T_0(F_n - 1)$$

where T_0 is the standard temperature, 290K. *SAL*

Ref.: Blake (1980), p. 147; Skolnik (1990), p. 2.31.

The **receiving (transmission-)line noise temperature** is the component of system, noise temperature originating in the passive RF components connecting the antenna to the receiver. It depends on the physical temperature T_{tr},of these components and their loss L_r:

$$T_r = T_{tr}(L_r - 1)$$

SAL

Ref.: Blake (1980), p. 152; Skolnik (1990), p. 2.30.

The **sky temperature** is the component of antenna noise temperature that would result from an antenna pattern entirely sky-directed (no earth-directed sidelobes). It is the result of interaction with atmospheric gases (atmospheric attenuation), galactic radiation, and interference generated in the atmosphere and by human activities (and transmitted into the antenna through atmospheric paths), and depends on radar frequency and one-way loss through the entire atmosphere (and thus on beam elevation):

$$T_a' = T_g + T_{pa}$$

where T_g is a galactic component dependent on frequency f_0 and T_{pa} is an atmospheric component depending on one-way total attenuation $L_{\alpha 1}$. Approximations for these two components are

$$T_g = \frac{k_g}{f_0^{2.5}}$$

where $k_g = 3 \times 10^8$ for normal solar and galactic noise conditions, 5×10^7 for maximum conditions, and 1.8×10^9 for minimum conditions;

$$T_{pa} = T_0\left(1 - \frac{1}{L_{\alpha 1}}\right)$$

The variation of sky temperature with frequency and elevation angle for clear air is shown in Fig. T9. *SAL*

Ref.: Blake (1980), p. 170; Skolnik (1990), p. 2.29; Barton (1993), p. 142.

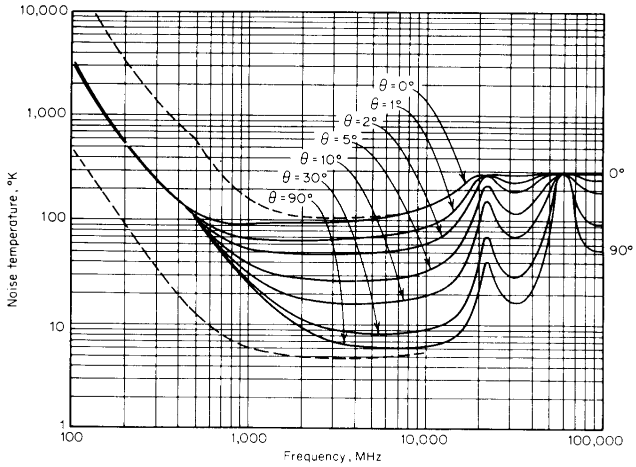


Figure T9 Sky temperature vs. frequency for different beam elevation angles (from Blake, 1980, Fig. 4.8, p. 170).

The **standard temperature** for noise calculations is $T_0 = 290$K, a reference specified in the IEEE standard for noise factor measurement and calculations. It represents an average room temperature. *SAL*

Ref.: Skolnik (1980), p. 344.

TEST, TESTING, radar

Antenna testing [measurement] is the process used to determine basic antenna performance parameters: radiation pattern, polarization, directivity, and gain. Two approaches are available:

(1) The far-field range. An unobstructed view over a large area is required.

(2) The compact range (Fig. T10).

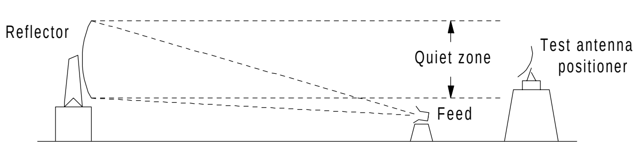


Figure T10 Compact antenna range configuration (after Barton, 1991, Fig. 13.5.1, p. 13.9).

Both types of test range operate under computer control to orient the antenna or test probe and to generate plots of antenna patterns. *SAL*

Ref.: Barton (1993), p. 13.9.

Computer-aided testing (CAT) involves use of a computer to simulate input data, control test procedures, and analyze output data in test of radar hardware or software. A typical block diagram of a computer-aided test configuration is shown in Fig.T11. *SAL*

Ref.: Evans (1990); Slater (1991); Barton (1991), p. 13.13.

Hypothesis testing is the procedure of making a decision in radar detection problems: whether the target is present or

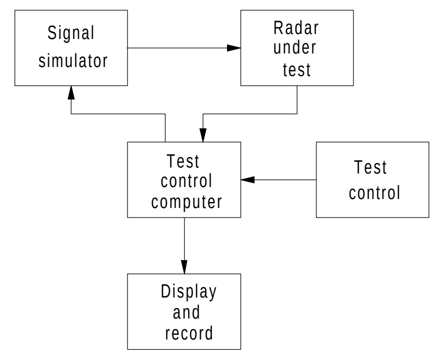


Figure T11 Laboratory testing under computer control (after Barton, 1991, Fig. 13.6.1, p. 13.15).

absent. In most such procedures there are two possible outputs at a given instant of time, one corresponding to the null hypothesis, H_0 (target absent) and the other to hypothesis H_1 (target is present). This is the binary hypothesis testing problem, in which the decision is made between only two possible outcomes, H_0 and H_1, and there are four possible outcomes: true detection, true nondetection, false alarm, and target miss. Other procedures consider additional alternatives. The M-ary hypothesis test requires a decision among M hypotheses, H_0, H_1, ... , H_M, with M^2 possible outcomes for each test. Different criteria may be used in making decisions as to which hypothesis is accepted (see **DETECTION**). *SAL*

Ref.: Barkat (1991), pp. 115–174.

Laboratory testing is performed on a radar system or its subsystems in a laboratory environment, or at special test facilities such as antenna ranges. Important factors are the designs of special simulators to generate signals representing targets, interference, clutter, noise, and so forth, with parameters sufficiently matching those of the real radar inputs. Examples of system-level parameters best evaluated in the laboratory are resolution in angle, range, and doppler, velocity response, target acquisition time, and multiple-target capability. CAT is the most appropriate in this environment (Fig. T12). Laboratory testing is, in principle, more cost effective than field testing, but cannot alone provide sufficient confidence or results covering all aspects of radar performance. *SAL*

Ref.: Barton (1991), p. 13.15.

Radar test is a set of procedures to evaluate and verify radar operability, to obtain estimates of radar performance, and to validate radar performance against specification requirements. Typical steps in a radar test program are

(1) Analysis and simulation.

(2) Subsystem test.

(3) Laboratory test.

(4) Field test.

(5) Performance evaluation based on data from the previous steps.

The flow of radar system design and testing is shown in Fig. 3. Detailed descriptions of performance evaluation methods are given in Leonov (1990). *AIL, SAL*

Ref.: Leonov (1990); Barton (1991), Ch. 13.

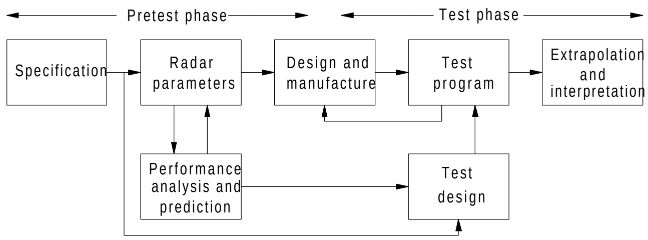


Figure T12 Flow of radar system design and testing program (after Barton, 1991, Fig. 13.1, p. 13.2).

Receiver test is performed to evaluate the main characteristics of the radar receiver: sensitivity, selectivity, dynamic range, interference immunity, and so forth, and the performance of the major receiver circuits: MTI canceler, pulse compression circuits, and so forth. The general test configuration is shown in Fig. T13. *SAL*

Ref.: Barton (1991), p. 13.12.

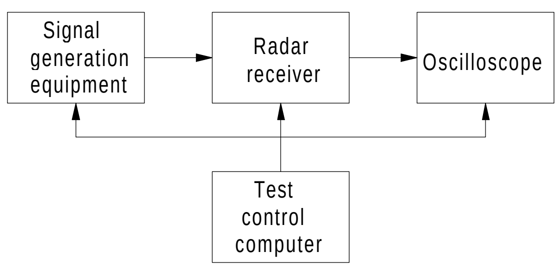


Figure T13 Receiver test configuration (after Barton, 1991, Fig. 13.5.4, p. 13.12).

A **signal-processor test** is performed to evaluate the performance of this radar subsystem. A typical test configuration is shown in Fig. T14, in which the test equipment includes a waveform synthesizer, an analog-to-digital converter similar to that used in the radar itself, a test data processor and control computer (often in the form of a personal computer), and display and recording equipment. To obtain additional data on subsystem interfaces and performance, the signal processor and receiver may be combined for testing. *SAL*

Ref.: Barton (1991), p. 13.12.

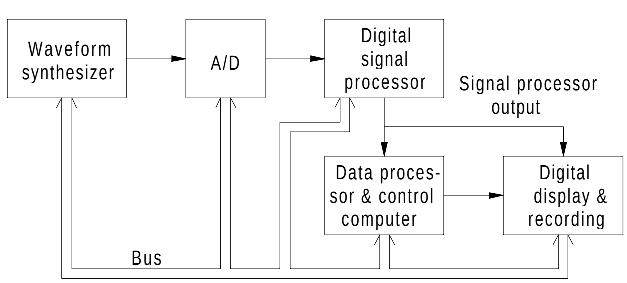


Figure T14 Signal-processor test configuration (after Barton, 1991, Fig. 13.5.5, p. 13.13).

Transmitter test is performed to evaluate the major performance characteristics of the radar transmitter: output power, bandwidth, frequency stability, noise performance, and waveform fidelity. A typical test configuration is shown in Fig. T15. *SAL*

Ref.: Barton (1991), p. 13.10; Sivan (1994), Ch. 15.

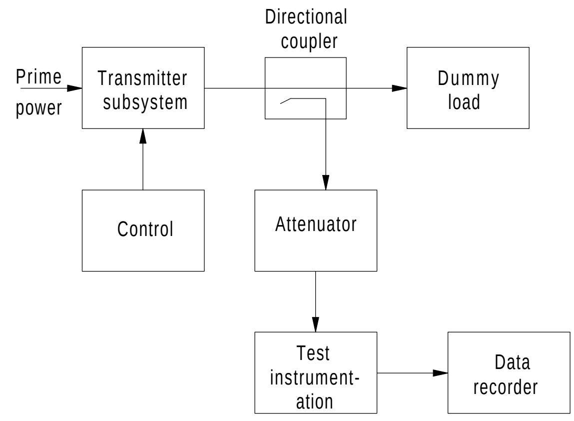


Figure T15 Transmitter test configuration (after Barton, 1991, Fig. 13.5.3, p. 13.11).

TETRODE. A tetrode is a four-electrode semiconductor or vacuum-tube device. In the latter, a second grid (screen grid) is included between the control grid and the anode, permitting increased gain. Tetrode tubes are used in radars operating in the VHF and low UHF bands. *SAL*

Ref.: Zherebtsov (1989), p. 251.

THRESHOLD. A threshold is "a value of voltage or other measure that a signal must exceed in order to be detected or retained for further processing." In radar detection, the threshold is set relative to noise or other interference to establish the required probability of false alarm. (See **DETECTION probability**; **FALSE ALARM probability**; **CFAR**.)

An adaptive threshold is controlled by factors considered important for the process of detection or signal processing. An example is the cell-averaging CFAR technique, in which the threshold for the test cell is based on the average output of surrounding range or doppler cells, which are assumed to provide an adequate estimate of the interference in the test cell. This is one form of adaptive video threshold or automatic threshold. *SAL*

Ref.: IEEE (1993), p. 1370; Skolnik (1980), p. 488, (1990), p. 8.13.

THROUGHPUT CAPABILITY. Radar throughput capability is the number of targets that can be processed by a radar with the specified performance. Here, the term *processed* includes the full set of operations performed by the radar according to its functional tasks: detection only, or detection and tracking, or detection, tracking, and recognition, and so forth. To determine throughput capability, the radar typically is represented as a server in which input targets are served sequentially at several different stages. for example, for three stages the throughput capability TC is equal

$$\mathrm{TC} = \sum_{i=1}^{L_1} \lambda_{1i} Q_{1i} Q_{2i} Q_{3i}$$

where L_1 is the number of channels for target search; λ_{1i} is the intensity of the target stream for the ith channel; Q_{1i} is the relative throughput capability of the first stage (detection of target at the boundary of the coverage sector); Q_{2i} is the relative throughput capability of the second stage (detection of tracks); and Q_{3i} is the relative throughput capability of the third stage (target tracking). In some simplified cases, analytical solutions for throughput capability are possible, typically for stationary input target streams and when processing capabilities do not impose restrictions upon throughput capability. The throughput capability is often called *target capacity. IAM*

Ref.: Shishkov (1987), pp. 85, 104.

THYRISTOR. A thyristor (or semiconductor thyratron) is a controlled switching diode with a PNPN structure, having three terminals. Because of the opposing connection of PN junctions there is insignificant reverse current through the thyristor at any polarity. However, when some quite high voltage is applied to the control electrode connected to the internal N-layer, the thyristor switches to a conducting state, and large current flow takes place with only a small voltage drop (e.g., 1V). In radars, thyristors are used as pulse modulator switches. *SAL*

Ref.: Zherebtsov (1989), p. 123.

TIME

Time of arrival (TOA) is the moment of time when a signal appears at the input of the receiver. The TOA location method is "a process by which the position of a radiating transmitter can be located by means of the relative time delay between its signals as received by multiple receivers of known relative position." *SAL*

Ref.: IEEE (1993), p. 1377.

The **time-bandwidth product** is the product of waveform duration and bandwidth. In a conventional (simple) pulse, the product is approximately unity. In pulse-compression waveforms, it refers to the product of the pulsewidth and the width of the transmitted spectrum τB and determines the pulse-compression ratio. (See **PULSE COMPRESSION**.) In continuous-wave signals and coherent pulse trains, it is the product of the coherent signal (burst) duration and the bandwidth $t_f B$. *DKB*

Ref.: Nathanson (1969), pp. 293, 297, 517, 518.

correlation time (see **CLUTTER correlation function**).

Dead time is "the time interval in an equipment's cycle of operating during which the equipment is prevented from providing normal response." In pulsed radar, the dead time occurs during and immediately following transmission of a pulse, until the T/R switch and receiver recover their normal ability to pass received signals. The loss of signals during this time is called *eclipsing loss*.

In a scanning radar, the time required to change the direction of the scan is also called dead time. For example, in an airborne intercept (AI) radar using a raster scan or n-bar scan, dead time occurs at the edges of the scan. It is called dead time because during that brief interval, no new space is being searched. *PCH*

Ref.: IEEE (1993), p. 313.

Delay time (see **DELAY**).

Dwell time is the time during which the radar beam illuminates a given angular position. In a phased-array radar, the time is usually defined quite precisely by the stepped motion of the beam from one position to the next. In mechanically scanned radars, the conventional definition is the time required to move through the half-power, one-way beamwidth. The term *observation time* is also used. *DKB*

Ref.: Billetter (1989), p. 80.

False-alarm time, in a radar receiver with a given detection threshold, is the average time between false alarms. The false-alarm time is the reciprocal of the average false-alarm rate:

$$t_{fa} = \frac{1}{\nu P_{fa}}$$

where ν is the rate at which detection decisions are made and P_{fa} is the probability of false alarm. In general this rate is

$$\nu = \frac{n_t n_f}{t_o}$$

where n_t is the number of range (time) cells tested for target presence and n_f is the number of doppler (frequency) cells per range cell. When video integration is used in a system without doppler filters, $n_f = 1$, and when there is no integration the rate is equal to the receiver noise bandwidth B_n. (See also **FALSE ALARM probability**.) *PCH, DKB*

Ref.: Barton (1988), p. 76.

The **fast time constant (FTC)** circuit is an ECCM technique employing a differentiating circuit matched to the length of the radar pulse to suppress the later portions of signals whose duration exceeds the pulsewidth. The disadvantage of this technique is that in the presence of a strong interference signal, the sensitivity of the receiver is reduced considerably (See **CFAR, log-FTC**.) *SAL*

Ref.: Neri (1991), p. 420.

Frame time, in a scanning antenna radar system, is the time required to perform one complete scan over the assigned volume of coverage. This is the data interval or the time between successive data readings from the same target, or "frame time" by analogy to television picture scanning. It is sometimes called *scan time*, but that term allows ambiguity when more than one rotation of an antenna is used to cover different elevations. *PCH*

Ref.: Barton (1964), p. 235.

Mean time between failures (MTBF) is a measure of the reliability of a complex mechanical or electronic mechanism, component, subsystem or system. Simply stated, the MTBF is the reciprocal of the failure rate λ in the expression for reliability:

$$R(t) = e^{-\lambda t}$$

where $R(t)$ is the probability that the item will operate without failure for time t under the stated operating conditions, $e = 2.7183$ is the base of natural logarithms. The failure rate λ is usually expressed in terms of failures per hour and is treated as constant for any given set of stress, temperature, and quality-level conditions. *PCH*

Ref.: Fink (1989), p. 28.3.

Mean time to repair (MTTR), in a complex mechanical or electronic system, is the average time required to return the system to operational condition after the occurrence of a failure. In reliability calculations for large systems, it is a component of the system availability calculation

$$A_o = \frac{MTBF}{MTBF + MTTR}$$

See also **mean time between failure.** *PCH*

Pulse rise time is the time taken by a pulse to begin measuring between same specified points relative to pulse amplitude (typically 10% and 90% points). *SAL*

Ref.: Stevens (1988), p. 292.

Recovery time is a term applied to receivers under jamming conditions and to T/R and ATR tubes. In the receiver case, it is "the time for a part of a receiver to recover to a zero-signal condition after receiving a jamming pulse of saturation intensity." For a T/R tube it is "the time required for a fired tube to deionize to such a level that the attenuation of a low-level radio-frequency signal transmitted through the tube is decreased to a specified value." For an ATR tube it is "the time required for a fired tube to deionize to such a level that the normalized conductance and susceptance of the tube in its mount are within specified ranges." *SAL*

Ref.: IEEE (1993), pp. 1,088–1,089; Johnston (1979), p. 66.

Resolving time is "the minimum interval by which two elements must be separated to be distinguishable by the time measurement alone." *SAL*

Ref.: IEEE (1993), p. 1,134.

scan time (see **frame time**).

search time (see **frame time**).

The **time-on-target (TOT)** is the time during which a target is continuously in the main lobe of a radar's antenna on any one scan of the antenna. Time-on-target can be expressed for a continuous scan as

$$\text{TOT} = t_o = \frac{\theta_3}{\omega_s}$$

where θ_3 is the antenna 3-dB, one-way beamwidth and ω_s is the antenna scan rate. In discontinuous scan (e.g., using a phased-array antenna with step scanning) the term *dwell time* is more commonly used. *PCH*

Ref.: Stimson (1983), p. 310.

TOMOGRAPHY, microwave. Microwave tomography is a radar imaging technique in which the final image is formed from the set of projections received in a microwave band. The underlying process of obtaining a radar image is based on scattering characteristics or (and) geometric shapes of objects. The restoration of the image is based on the reverse Radon transform. Both wideband and narrowband waveforms can be used in microwave tomography. Tomographic imaging based on wideband waveforms typically uses the linkage between the radar profile and the target scattering features. (See **IMAGING, 2D; IMAGING, 3D.**) When a narrowband signal is used, 2D and 3D images can be received through the observation of an amplitude of radar return for different aspect angles. The resolution of the synthesized image depends on the aspect angle sector, the width of the spectrum (for wideband waveforms) and the carrier frequency (for narrowband waveforms). The considerable volumes of the database and computation throughput make the requirements on signal processors very stringent. *IAM*

Ref.: Li, H. J., et al., *IEEE Trans.* ***AP-35***, 1987, no. 5.

TRACK, TRACKER, TRACKING. Tracking is "the process of following a moving object or a variable input quantity. This process may be carried out manually or automatically. In radar, target tracking in angle, range, or doppler frequency is accomplished by keeping a beam or angle cursor on the target angle, etc." Different methods and problems in tracking are discussed in the following articles. (See also **MONOPULSE**; **RADAR: fire control**; **gunfire control**; **instrumentation**; **tracking**; **track-while-scan**.) *DKB*

Ref.: IEEE (1993), p. 1,389.

The **α-β(-γ) tracker** is a sampled-data tracking technique used in surveillance radar and multitarget tracking radar to form track files from measurements of target position on a series of scans or dwells. (See **FILTER, α-β**.) In a track-while-scan surveillance radar, the outputs of the α-β tracking loops control the positions of correlation gates in range-azimuth space, within which a detected target report on the next scan will be accepted for track update. In a multitarget tracking radar they control positions to which the beam, range gate (and, where applicable, the doppler filter) is positioned on the next tracking radar dwell assigned to a particular target. The number of tracking channels in either case can be numbered in the hundreds, limited only by computational capacity and by time allocations in a multitarget tracker. *DKB*

Ref.: Blackman (1986), pp. 21–25.

Aided tracking is "a tracking technique in which the manual correction of the tracking error automatically corrects the rate of motion of the tracking mechanism." When the ratio of velocity increment to position input is properly adjusted, the operator can match the tracker output to the position and rate of a constant velocity target within a few seconds, after which only corrections for target accelerations need be applied. The rate memory of the aided tracker provides coasting through target fades. Use of aided tracking permits the operator to track even on accelerating targets with greater accuracy and less workload.

While manual tracking systems have given way to automatic, computer-controlled trackers in most applications, there are still situations (e.g., ECM environments and electro-optical tracking adjuncts for radar) in which the intervention of an operator can be effective. *DKB*

Ref.: IEEE (1993), p. 22.

Altitude tracking is carried out in radar altimeters to smooth the data derived from the receiver. In an FMCW altimeter the tracking loop is a frequency tracker or phase-locked loop, following the beat frequency between transmitted and received waveforms (which is proportional to altitude). In some cases, the frequency excursion of the waveform may be controlled by the loop to maintain a constant beat frequency. In pulsed altimeters the altitude tracker is implemented in the same manner as the conventional range tracker.

The degree of data smoothing depends on the tracking loop bandwidth, and if this is made narrow to minimize random error and irregular surface effects it may be impossible for the altimeter to follow abrupt changes in terrain height. Loss of track due to such abrupt changes may necessitate switching to a search mode to reacquire the ground echo. *DKB*

Ref.: Hovanessian (1984), pp. 330–334.

Angle tracking is the process of maintaining the radar beam (in tracking radar), or an electronic gate or cursor (in track-while-scan), on the target angle. In tracking radar, the techniques used are conical scan, lobe switching, and monopulse, all of which are methods of sensing an offset of the target from the beam axis and feeding back any error to keep the radar beam pointed at the target. Multitarget (phased-array) trackers use monopulse measurements during brief dwells on each target as inputs to α-β or Kalman filters, outputs of which are target coordinates and velocity components.

An angle tracking loop is designed to accept input data at the radar PRF (for single-target trackers), or at the target revisit rate (for multitarget trackers), to smooth these data, derive velocity, and extrapolate forward to a future target location. The bandwidth of the loop is optimized to minimize total error from random interference and dynamic lag. (See also **CONICAL SCAN**; **CONOPULSE**; **ERROR**; **LOBING**; **MONOPULSE**; **track-while-scan**.) *DKB*

Ref.: Skolnik (1962), Ch. 5, (1990), Ch. 8; Barton (1994), Ch. 9, (1988), Ch. 8; Neri (1991), pp. 135–175.

Tracking with angle-only measurements is possible from a single station if the target is constrained to follow a certain type of trajectory (e.g., satellite orbit). With multiple stations, triangulation methods may be used on targets with any type of motion. *DKB*

Ref.: Long (1992), p. 313.

Automatic tracking is "tracking in which a system employs some feedback mechanism, for example a servo or computer, to follow automatically some characteristic of a signal or target, such as range, angle, doppler frequency, or phase." In conventional tracking radar, the antenna is positioned by servomechanisms, and the range gate (and doppler filter if applicable) are controlled electronically. Multitarget phased-array trackers use the output of sampled-data tracking loops to control the electronic steering of the beam. The term automatic tracking is also applicable to the α-β tracking loops of track-while-scan surveillance radars, where correlation gates in range and angle, for each target, are controlled electronically. Automatic tracking in frequency is carried out using a voltage-controlled oscillator that converts the target signal to the center frequency of a fixed filter, or by interpolation between adjacent filters in a filter bank (e.g., FFT output) to generate a digital estimate of target doppler frequency. *DKB*

Ref.: IEEE (1993), p. 72.

Tracking (loop) bandwidth is defined as that of a low-pass filter: a one-sided bandwidth, measured from zero frequency to the frequency of half-power response B_3, or to the equivalent noise bandwidth β_n. The bandwidth can be no more than half the sampling rate of the system (the PRF, for conventional monopulse tracking radars; the lobe-switching rate for sequential scanning radars; the revisit rate for phased arrays; or the scan rate for search radars), but in most cases it is constrained in angle tracking by the mechanical inertia of the antenna. An optimum loop bandwidth can be found in any particular case which minimizes the total variance of random and (dynamic) lag error. *DKB*

Ref.: IEEE (1993), p. 72; Barton (1988), p. 464.

Coherent tracking radars use MTI, pulsed doppler, or CW techniques to reject clutter. In MTI radars the tracking loops accept envelope detected signals as input, and perform integration and averaging of data over the tracking loop time constant in the same way as in noncoherent radars. Pulsed doppler and CW systems take advantage of coherent integration for both initial detection and tracking, and are advantageous is several respects:

(1) They obtain accurate measurements of radial velocity through tracking and measurement of doppler frequency shift.

(2) They can reject clutter and other interference through doppler frequency resolution.

(3) They can maintain track as long as the signal-to-noise energy ratio within the doppler filter is well above unity, in contrast to noncoherent trackers in which the loop bandwidth is reduced sharply when the single-pulse SNR is near or below unity. *DKB*

Ref.: Barton (1988), pp. 567–472.

Tracking in clutter becomes a problem in track-while-scan radars when the density of clutter false alarms exceeds some allowable value that depends on the size of the correlation gate. In monopulse trackers, uncanceled clutter produces additive noise errors, and if not canceled below the signal level tends to capture the tracking loops, suppressing the desired signal. (See **ERROR, clutter.**) *DKB*

Ref.: Long (1992), pp. 308–313; Barton (1988), pp. 531–533.

Conical-scan tracking is a method in which a single pencil beam is rotated about the tracking axis, the beam axis forming a narrow cone. The method was used extensively during and immediately after World War II but has more recently been replaced in most applications by monopulse tracking. The major reason for this is the susceptibility of conical scan to amplitude modulation (scintillation) of the target (see **ERROR, scintillation**), exaggerated by intentional ECM using AM at the scan frequency. (See **ECM, angle measurement.**) Other disadvantages of conical scan include the limitation of tracking loop bandwidth to 1/4 the scan rate, greater error from random noise and interference, and mechanical vibration from the moving feed. *DKB*

Ref.: Barton (1988), pp. 388–390, (1988), p. 493.

Doppler tracking is possible in both CW and pulsed coherent radars. The usual technique is to include a fixed narrowband filter and discriminator at intermediate frequency, into which the target signal is converted by mixing with a signal from a voltage-controlled oscillator (VCO). The VCO is initially set to the correct value, using external designation or differentiated range data. Once the signal appears within the passband of the discriminator, the output of that circuit is used to control the VCO, closing the tracking loop. In pulsed doppler radars, there will be ambiguities in the data, and it is necessary to resolve these in order to force the loop to lock on the correct (central) spectral line. This can be done if accurate designation data are available, or by changing the PRF and observing the apparent (ambiguous) target doppler frequency, which will not depend on PRF when the central line has been acquired. *DKB*

Ref.: Barton (1988), pp. 439–448, (1969), Ch. 4; Chrzanowski (1990), pp. 122–127.

Track-before-detect is a process that uses large numbers of tracking channels to integrate signals (usually noncoherently) over long periods. The channels are designed to follow all possible trajectories of targets of interest, in range and angle, producing an output if an actual target matches one of the channels. Integration can be performed over periods such that the target moves across many resolution cells, over many scans of a search radar antenna. The sensitivity of such a system can be very high because of the integration, and because false alarms occur only when interference integrates in the moving channel sufficiently to exceed the threshold. The technique is also referred to as retrospective detection. *DKB*

Ref.: Long (1992), p. 458; Bar Shalom (1990), Ch. 4.

Group tracking is a process in which a number of closely spaced targets are tracked within one channel of a search radar or multitarget phased array. The bounds of the group are defined (usually within one beam position), and the midpoint is tracked in each coordinate. Any target leaving the group is assigned to an individual track file, while files on targets joining the group are merged with the group file. The result is to save radar resources that might otherwise be expended in scheduling separate dwells for individual targets within the same beam position. *DKB*

Ref.: Billetter (1989), p. 98.

Track initiation is the process of starting a track file on a new target after it is detected. The process starts with detection of a target whose reported position does not qualify as an update to an existing track. A tentative new track file is established using the initial reported coordinates, and a correlation gate is centered on that position with extent sufficient to encompass the motion of any target, plus position errors on the initial and subsequent scans (when doppler data are available, that coordinate may also be included in the correlation process). The tentative track is confirmed if *m* detections are received within the correlation gate over the next *n* scans (e.g., three out of four or five scans). Target location associated with each detection is used to refine the position and size of the correlation gate for subsequent scans. Once confirmation has occurred, the tentative track file parameters become those of the new track file. *DKB*

Ref.: Blackman (1986), pp. 6, 10, 151.

Track-on-jam is an electronic counter-countermeasures (ECCM) operating mode available to most fire control radars and radar homing seekers that allows the radar being jammed to transfer track from the target "skin" return to the jamming signal itself.

The track-on-jam capability of air defense missile systems serves as a strong deterrent to the use, by an attacking aircraft, of simple self-screening noise jamming (SSJ), especially when the target is within intercept range of the defensive missile. Track-on-jam strategies involve determination of the nature of the jamming (e.g., jammer-to-signal ratio (J/S), jammer duty cycle, noise bandwidth and modulation characteristics), and the implementation, via programmed logic, of the ECCM response. Depending on the jammer duty cycle, the appropriate response might be to ignore the jamming, command track-on-jam in synchronism with the jamming signal, or command track-on-jam as long as the jamming signal is present. A primary danger in track-on-jam ECCM is the potential for the defensive missile system to be deceived into launching its limited missile resources at targets that lie outside the lethal range of the missile. Although noise jamming usually denies target range data to the victim radar (and radar homing seeker), a coarser, but adequate, form of target range information may be available via a local search radar or the air defense net. Target range data may also be determined via triangulation using the angle tracking data from two adjacent tracking radars.

Track-on-jam techniques can be effective against coherent repeater jammers as well, provided that the jammer does not successfully negate the radar or missile seeker's angle tracking capability. Much of the modern tracking radar and missile-seeker-directed repeater jamming is designed to defeat the angle-tracking circuits of these systems, however, and track-on-jam ECCM is not generally useful against these techniques, unless, for some reason, the high *J/S* ratios required for jammer effectiveness cannot be maintained. (See also **HOME-ON-JAM**; **JAMMING.**) *PCH*

Leading-edge tracking is "a radar range tracking technique in which the range error signal is based on the range delay of the leading edge of the received echo. This provides ability to reject delayed interference, chaff, and more distant sources." The technique is applicable to simple, rectangular pulse waveforms, and requires that the receiver bandwidth be broad enough to reproduce the rise time of the pulse leading edge, typically about 10% of the full pulsewidth. Considerable loss in signal-to-noise ratio is incurred, but this is often acceptable once a track has been initiated. *DKB*

Ref.: IEEE (1993), p. 708; Barton (1988), pp. 436–438, (1969), pp. 78–81.

low-altitude target tracking (see **ERROR, multipath**).

Tracking at low signal-to-noise ratio is possible in a radar that observes the target continuously but is subject to nonlinear effects (small-signal suppression, or detector loss) in the envelope detector and error detector circuits of the receiver. These losses reduce the gain of the tracking loop, and hence its bandwidth. The usual expression for monopulse thermal noise error σ_θ, applicable for $S/N >> 1$, is modified to

$$\sigma_\theta = \frac{\theta_3\sqrt{1+S/N}}{k_m(S/N)\sqrt{2n'}} = \frac{\theta_3}{k_m\sqrt{(1+S/N)(f_r/\beta_{no})}}$$

where k_m is the monopulse error slope, n' is the number of pulses integrated in the actual tracking loop time constant, f_r is the pulse repetition frequency, and β_{no} is the design bandwidth of the tracking loop (for $S/N >> 1$). Similar expressions apply to range and doppler tracking error. It can be seen that the resulting random noise at the tracker output is finite even for $S/N \to 0$. However, the reduced loop bandwidth causes an increase in dynamic lag error on targets with even slight accelerations, and the track will generally be lost when $S/N <$ 0.5 (–3 dB) for more than a brief interval. This can occur even though the integrated ratio, $n'S/N >> 1$. *DKB*

Ref.: Barton (1988), pp. 467–472.

manual tracking (see **aided tracking**).

monopulse tracking (see **angle tracking**; **MONOPULSE**).

Multiple target tracking can be carried out by surveillance radars in the track-while-scan mode, and by phased array radars (and those with rapid electromechanical scan) that sequence their beams rapidly among many targets. The tracking loops operate on a sampled-data basis (see **α-β tracker**), with update rates dependent on the scan rate for surveillance radars and the tracking time budget allocation for multifunction and multitarget phased-array radars. *DKB*

Ref.: Blackman (1986); Billetter (1989); Bar-Shalom (1990, 1992); Sabatini (1994)

Off-axis tracking is a technique for avoiding surface effects (multipath and clutter) in low-altitude target tracking by biasing the beam position above that of the tracked target. The target elevation angle is then estimated by adding the (negative) off-axis angle sensed by the error detector to the angle of the beam axis. The extra elevation of the beam, combined with the steep slope of the antenna mainlobe pattern below the target elevation, reduce the input from surface reflections. (See **ERROR, multipath**; **SIGNAL-to-clutter ratio.**) *DKB*

Ref.: Barton (1975), pp. 343–366; Skolnik (1980), p. 174; Barton (1988), pp. 529–530.

Passive angle tracking is "a radar tracking technique that uses a received signal other than the backscattered radar emissions with which to track an object, jammer, or other signal source. Passive homing, home-on-jam (HOJ), and track-on-jam (TOJ) are examples of passive angle tracking." In some system applications, the angle data are sufficient even without accompanying range data. Most modern antiair missile systems are effective when seekers are used in the HOJ mode, and with command-guided missiles using the so-called three-point (command to line-of-sight) guidance method. Manned interceptors can also be vectored adequately to intercept a jamming aircraft using azimuth-only data. In other cases, angle-only data from two or more sites can be combined to solve for position of the passively tracked target. *DKB*

Ref.: IEEE (1990), p. 22; Hovanessian (1988), p. 259.

Range tracking is carried out in pulsed radar by direct matching of a range gate position to the delayed echo pulse. The usual technique if the split-gate range tracker, which is "a form of range tracker using a pair of time gates called an 'early gate' and a 'late gate,' contiguous or partly overlapping in time. When tracking is established, the pair of gates straddles the received pulse that is being tracked. The position of the pair of gates then gives a measure of the time of arrival of the pulse; that is, the range of the target from which the echo is received. Deviation of the pair of gates from the proper tracking position increases the signal energy in one gate and decreases it in the other, producing an error signal, which moves the pair of gates so as to establish equilibrium." The range difference channel is formed by subtracting the late gate output from the early gate output and integrating the result, to form a time-discriminator response. (See **DISCRIMINATOR.**) The error signal from the discriminator provides the input to an electronic tracking loop that controls the timing of the gates.

In CW radar, range tracking is possible when modulation of some type is applied to the transmission (e.g., sinusoidal or triangular FM). With sinusoidal FM, the phase shift between transmitted and received modulations provides the delay (and hence the range) estimate. With triangular modulation, the frequency difference between the transmitter output and the received echo is proportional to delay (and range, see **MEASUREMENT, range**).

In modern radar, the range tracker is implemented digitally. Figure T16 shows such a system, in which the time discriminator response formed by an early-late gate pair is used to control a variable-frequency oscillator that feeds a target range register. The use of gates followed by an integrator is necessary in most cases to form the error signal, because the signal-to-noise ratio (SNR) of the individual pulse may not be sufficient to provide a useful error signal. In phased array trackers providing a single pulse per dwell, with high SNR, the digital tracker may consist of a counter, which is started with the transmission and stopped by the threshold crossing of the strong echo pulse.

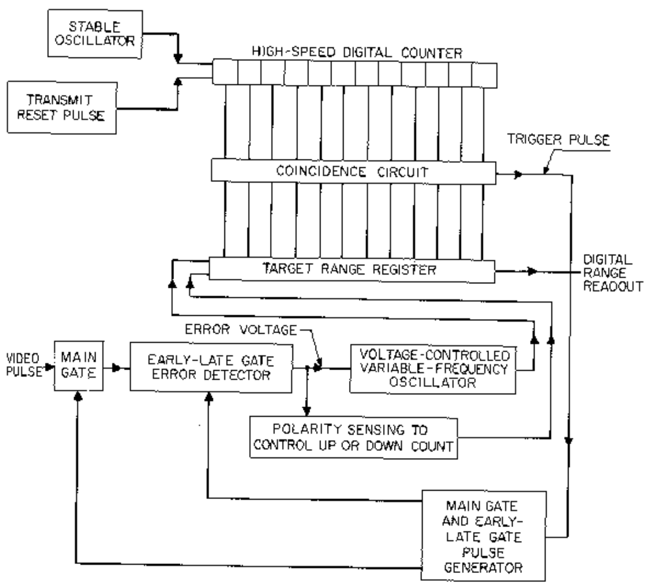


Figure T16 Block diagram of digital range tracker (from Skolnik, 1990, Fig. 18.26, p. 18.29, reprinted by permission of McGraw-Hill).

In a pulsed doppler tracking radar, the split-gate circuit can take the form of a balanced modulator at IF, to which the gating signal is applied in such a way as to reverse the phase of the signal pulse at its midpoint. A subsequent bandpass filter then provides the integration, not only over the pulse but over a train of coherent pulses, recovering the error signal without detector loss. In such a system, the single-pulse SNR can drop well below unity, and operation continues as long as the total energy ratio of the pulse train is well above unity.

Other methods of range tracking in pulsed doppler radars interpolate between the filter-bank (FFT) outputs of fixed, contiguous range gates to derive range estimates on one or more detected targets within the beam, or slow sinusoidal modulation of the position of a single gate to provide a sequential sensing of range error. *DKB*

Ref.: IEEE (1993), p. 1,263; Skolnik (1980), p. 176, (1990), p. 18.27; Barton (1988), Ch. 9, (1969), Ch. 3; Chrzanowski (1990), p. 120.

sequential-lobing tracking (see **SCAN, conical**; **LOBING**).

Track-while-scan (TWS) is "an automatic target tracking process in which the radar antenna and receiver provide periodic video data from the search scan, together with interpolation measurements, as inputs to computer channels which follow individual targets." Target detections (reports) from the search signal processor are applied to correlation gates that follow each tracked target, and those that correlate with existing tracks are used to update those track files, using an α-β or Kalman filter process. (See **α-β tracker**; **FILTER.**)

The TWS technique is used in surveillance radars, and also in sector-scanning trackers (e.g., many precision approach radars, and older fire control radars such as the Russian Fan Song).

The distinction between TWS and other tracking methods is that the tracking loop operates at the output of the signal processor and is not fed back to control the position of the antenna beam. Hence the regular search scan is not interrupted, and any number of detected targets may be tracked, up to the limit of the track processor. The limitations of TWS are

(1) The sample rate is low, being the search scan (frame) rate (this rate may be fairly high in sector-scanning radars using electromechanical or electronic scan).

(2) The correlation gates must be wide enough to encompass target maneuver and errors in reporting a detection, and hence, as a result of low sample rate, they may expand enough to accept reports on other targets or on uncanceled clutter, resulting in multiple split tracks or loss of track on the desired target.

(3) The signal-to-noise energy ratio is limited to that received on a scan past the target, and hence the inherent accuracy of the individual reports is low, compared with trackers which dedicate more time to an individual target.

(4) The position errors on a given scan are inherently greater, because of target scintillation, than with on-axis tracking radar using such techniques as monopulse.

(5) The vulnerability to angle jamming techniques is greater than with dedicated trackers, both because of sensitivity to amplitude modulation and lower signal-to-noise energy ratio resulting from spreading of radar energy into many beam positions. *DKB*

Ref.: IEEE (1993), p. 1,389; Skolnik (1980), p. 183, (1990), pp. 8.23–8.40; Blackman (1986); Barton (1988), pp. 390–397.

velocity tracking (see **doppler tracking**).

TRADEOFF. A tradeoff in radar design is the process of identifying and evaluating options, primarily during the conceptual and preliminary design phases of a radar development program. The result of tradeoff studies generally determine the manner in which the radar performance requirements are met through the radar configuration and design implementation. To be useful, such studies should take into account not only potential technical approaches (at the system and each major subsystem level), but the technical risk and life-cycle cost associated with each approach and its development time line.

Typical candidates for tradeoff studies include power-aperture versus frequency, scanning technique (e.g., mechanical scan versus phased array), active versus passive phased-array antenna, central versus distributed processing, MTI versus pulsed doppler waveforms, and the number and type of ECCM specific features. Figure T17 shows an example of a power-aperture tradeoff in terms of detection range for an S-band search radar. The steep slope due to increase in receiving area as compared to average power is clearly seen.

Despite the necessity for and apparent objectivity of engineering tradeoff studies such as those described here, the radar design often has much to do with the radar "culture" built up over time at a particular design agency or firm. *PCH*

Ref.: Barton (1991), p. K-5.

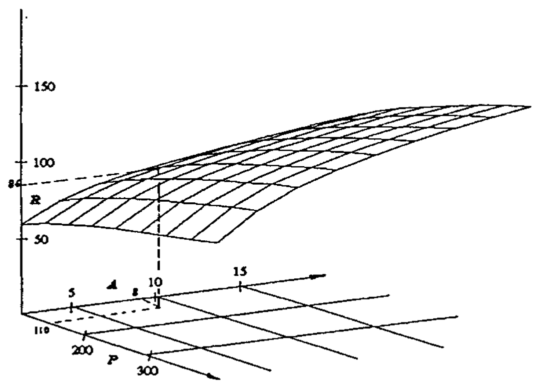


Figure T17 Average power vs. antenna aperture, S-band search radar (from Barton, 1991, Fig. K.1, p. K-6).

TRAINER, radar. A radar trainer is the equipment for professional training of radar operators. One of the possible configurations for a computer-aided trainer makes it possible to simulate any desired target environment, to simplify it or to complicate it, to vary the parameters of target trajectories in real time, and to get the estimates of operator training level in an automatic mode. *AIL*

Ref.: Romanov (1980), pp. 27–30.

TRANSCEIVER. A transceiver is a subsystem performing functions of both transmitter and receiver. In modern radar applications, individual transceiver modules are used in solid-state phased-array radar systems to feed antenna elements or subarrays on the array face. The module has the following functions: to switch between transmit and states, to provide phase shift for beam steering on both transmit and receive, to provide gain and power output in the transmit mode, and to provide gain and low noise figure in the receive mode. Phase shifting can be realized at a low power level, considerably reducing the dissipated power levels. A block diagram of typical transceiver module is shown in Fig. T18. *SAL*

Ref.: Skolnik (1990), p. 5.16; Leonov (1988), p. 41.

TRANSDUCER. A transducer is "a device by means of which energy can flow from one or more transmission systems or media to one or more other transmission systems or media." The energy at the input and output of the transducer usually are in different forms (e.g., mechanical and electrical, or acoustical and electrical). In radar applications, the most widely used are electro-acoustic transducers, such as the interdigital transducers in surface-acoustic-wave devices, used for pulse-compression networks (Fig. T19). These consist of two sets of interleaved metal electrodes (so-called fingers) deposited on the piezoelectric substrate. The acoustical wave is generated by application of an RF voltage across the adjacent sets of electrodes, which are spaced by one-half wavelength at the transducer design frequency. *SAL*

Ref.: IEEE (1993), p. 1,392.

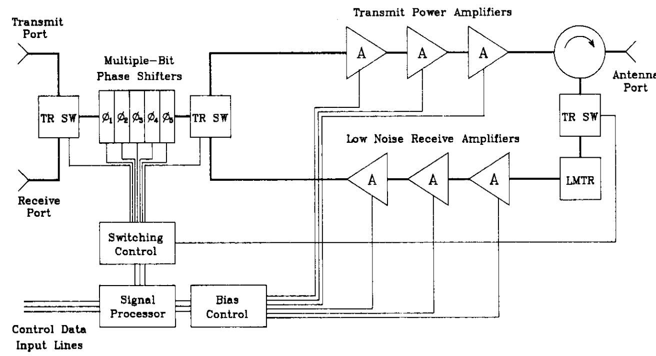


Figure T18 Block-diagram of transceiver module for phased array (from Skolnik, 1990, Fig. 5.9, p. 5.17, reprinted by permission of McGraw-Hill).

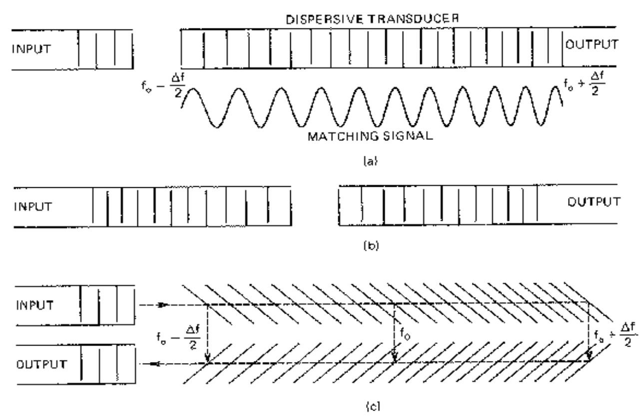


Figure T19 SAW transducer types (from Skolnik, 1990, Fig. 10.5, p. 10.11, reprinted by permission of McGraw-Hill).

TRANSFER FUNCTION. The transfer function is the characteristic of a linear system that describes its properties as a function of frequency. It is defined as the ratio of the Laplace transform of the output coordinate (voltage, current of the system) to the transform of the input coordinate for neutral initial conditions. The transfer function of a linear system with constant parameters is a rational function of the parameter of the Laplace transform p. Frequently the transfer function is introduced in the form of a ratio of z-transforms of output and input coordinates. (See **FILTER.**) The transfer functions introduced in this way are associated with the replacement of the variables $z = e^p$. *IAM*

Ref.: Gonorovskiy (1986), p. 152; Gol'denberg (1985), p. 50.

TRANSFORM

Cooley-Tukey transform algorithm (see **ALGORITHM**).

A **Fourier transform** is the transform relating the function $f(u)$ and its image $F(v)$ through the formulas

$$F(v) = \int_{-\infty}^{\infty} f(u)e^{-jvu}du \qquad (1)$$

$$f(u) = \frac{1}{2\pi}\int_{-\infty}^{\infty} F(v)e^{jvu}dv \qquad (2)$$

The first formula is called the *direct* Fourier transform and the second the *inverse* Fourier transform. The physical content of functions $f(u)$ and $f(v)$, as used in radar, are shown in Table T3.

Equations (1) and (2) are the one-dimensional Fourier transform. Two-dimensional transforms are also used; for example, to relate the aperture illumination function $g(x,y)$ for an aperture area to its antenna pattern $f(\theta,\phi)$. (See Barton, 1969, App. A.)

In digital signal processing, the discrete Fourier transform (DFT) is used to relate a discrete time function $s(kT)$ to its spectrum $S(n\omega_1)$:

$$S(n\omega_1) = \sum_{k=0}^{N-1} s(kT)\exp\left(-j\frac{2\pi}{N}nk\right) \qquad (3)$$

$$s(kT) = \frac{1}{N}\sum_{k=0}^{N-1} S(n\omega_1)\exp\left(j\frac{2\pi}{N}nk\right) \qquad (4)$$

$$n = 0, 1, \ldots N-1$$

where T is the sampling interval and N is the number of samples. (See also **SAMPLING.**) Equation (3) is called the *direct* DFT, and (4) is the *inverse* DFT.

Another form of DFT representation in digital filtering is a process applied to n successive signal samples and described by

$$F(p,k) = \sum_{p=1}^{n} s(p)w^{(p-1)(k-1)}$$

where $s(p) = x + jy$ is the input signal at sample point $p = 1, 2, \ldots, n$, $w = \exp(-j2p/n)$ is the complex weight kernel, and $k = 1, 2, \ldots, n$ is the output transform point (filter number). The frequency response of the DFT on unweighted input data sampled at repetition frequency f_r is

$$|H(f)| = \frac{\sin\left[\pi\left(\frac{nf}{f_r} - k + 1\right)\right]}{n\sin\left[\pi n\left(\frac{nf}{f_r} - k + 1\right)\right]}$$

Weighting may be applied to the n samples before the DFT operation to reduce sidelobe response. Typical filter responses for a DFT with and without low-sidelobe weighting of input data are shown in Fig. T20.

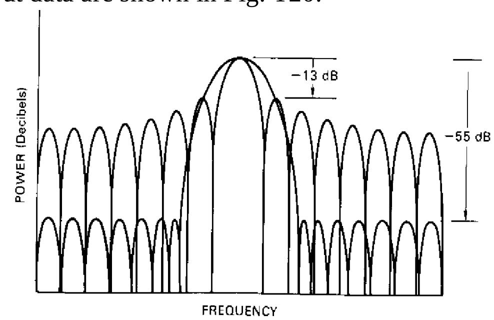


Figure T20 Transfer function of a discrete Fourier transform with uniform and weighted inputs (from Schleher, 1991, Fig. 2.34, p. 93).

To reduce the number of operations required to perform the DFT with digital computers, the fast Fourier transform (FFT) algorithm is used. The principle of the FFT is to reduce the calculation of the initial N-point DFT to calculation of several N_1-point DFTs with $N_1 << N$. Usually a value of N in the form of whole-number powers ν, in base 2, 4, or 8 is used, making it possible to repeatedly divide the series to be transformed into subseries of 1/2, 1/4 or 1/8 the length. Algorithms whose realization requires permutation of the readings of the input series $s(kT)$ are called *algorithms with decimation in time,* and algorithms with permutation of the readings $S(n\omega_1)$ are called *algorithms with decimation in frequency.*

FFTs reduce the number of required arithmetic operations by roughly N/ν compared with the methods of computing DFTs.

Table T3
Direct and Inverse Fourier Transform Pairs

Function, $f(u)$		Image, $F(v)$		Comments
Argument, u	**Function**	**Argument, v**	**Function**	
t = time	$s(t)$	ω = angular frequency	$S(\omega)$	Transform between time-domain deterministic waveform, $s(t)$ and its frequency-domain image (spectrum) $S(\omega)$
τ = time shift	$K(\tau)$	ω = angular frequency	$G(\omega)$	Transform between time-domain random signal correlation function, $K(\tau)$ and its frequency-domain image $G(\omega)$
x = coordinate in aperture plane	$g(x)$	θ = angular coordinate of antenna pattern	$f(\theta)$	Transform between linear aperture illumination function $g(x)$ and antennas pattern $f(\theta)$

The primary application of the FFT is as a digital filter-bank process in which $n = 2^m$ input samples, usually taken from a large number of contiguous range strobes within each pulse repetition interval, are processed to form n filter outputs for each range sample. If the input samples are unweighted, the output filter response has the form $H(f) = (\sin\pi fT)/\pi fT$, where T is the duration of the n-sample input. These filter responses overlap at the –4-dB level. To reduce sidelobe levels, time weighting is applied to the input samples before the FFT, using functions such as $W(t) = \cos^x(\pi t/T)$, where $t = 0$ is at the center of the sampled burst and x is a parameter that determines the sidelobe levels, or more complicated weighting functions of the Taylor or Dolph-Chebyshev type. Filters formed from weighted data overlap at levels above –4 dB, reducing the straddling loss for targets displaced from the center of a filter.

The performance parameters of FFT filters with different types of weighting are analogous to those of linear array antennas with illumination taper. Data on many types of weighted response may be found in Barton and Ward (1969).

The advantage of the FFT process, as compared with the DFT, is that the number of computations, for large n, increases as $(n/2)\log_2 n$ rather than as n^2 for the DFT. When the number of available signal samples $n \neq 2^m$, it is possible to add strings of zeroes at each end of the processed burst to fill the burst to the next value of 2^m, generating additional filter channels with greater overlap and preserving the processing throughput advantage of the FFT. *DKB, IAM, SAL*

Ref.: Barton (1969), App. A, B; Brigham (1974); Shirman (1981), p. 160; Oppenheim (1983), pp. 186–212; Gonorovskiy (1986), p. 29; Gol'denberg (1985), pp. 11, 14; Schleher (1991) pp. 94–103.

The **Hartly transform** is a transform of real discrete waveforms using the formulas:

$$X(k) = \sum_{n=0}^{N-1} x(n)\left[\cos\frac{2\pi nk}{N} + \sin\frac{2\pi nk}{N}\right]$$

$$x(n) = \frac{1}{N}\sum_{k=0}^{N-1} X(k)\left[\cos\frac{2\pi nk}{N} + \sin\frac{2\pi nk}{N}\right]$$

An advantage of the Hartly transform is the agreement of the direct and inverse transforms with an accuracy up to the constant multiplier $1/N$. One variety is the fast Hartly transform, whose algorithms may be used to perform base operations of digital filtration of both complex and real signal, but in the first case they do not have advantages over the fast Fourier transform. The advantages of the fast Hartly transform in processing real signals boil down basically to structural simplicity. *IAM*

Ref.: Vlasenko, V. A., *Radioelektronika* ***32***, no. 12, 1989, pp. 5–12, in Russian.

The **Hilbert transform** relates the signal $u(t)$ and its image $\hat{u}(t)$ (conjugate function) using the formula:

$$\hat{u}(t) = \frac{1}{J_1}\int_{-\infty}^{\infty}\frac{u(z)}{t-z}dz$$

$$u(t) = \frac{1}{J_1}\int_{-\infty}^{\infty}\frac{\hat{u}(z)}{z-t}dz$$

The Hilbert function is used to form a complex signal on the basis of a real signal $u(t)$ in the form $u(t) + j\hat{u}(t)$, and also to determine the inverse Radon transform. *IAM*

Ref.: DiFranco (1968), p. 58; Sloka (1970), p 13.

The **Laplace transform** of the time-domain function $f(t)$ is given by the formula:

$$F(p) = \int_0^{\infty} f(t)e^{-pt}dp$$

where p is a complex variable used to convert from the real variable ω to the complex frequency $p + j\omega$. The inverse transform, if it exists, is defined by the expression:

$$f(t) = \frac{1}{j2\pi}\int_{\sigma_1 - j\infty}^{\sigma_1 + j\infty} F(s)e^{st}ds$$

where the main value of the integral along the path of integration to the right of all singularities in $F(s)$ is taken.

The Laplace transform is widely used to describe automatic control devices, filters, and as the basis for the z-transform. *IAM*

Ref.: Korn (1961), pp. 8.6.1–8.6.4; Gonorovskiy (1986), p. 55; Jordan (1985), p. 46.34.

The **numerical transform** is an unrounded transform of a sequence of numbers of a given word length into a sequence of numbers of the same word length using the operations of modular comparison (addition module, multiplication module, raising to a power). Numerical transforms have no rounding noise. They can be used to obtain convolutions in digital form and make it possible to save time in obtaining convolution sums with a high accuracy of computation. *IAM*

Ref.: Shirman (1981), p. 164.

A **Radon transform** defines the projection of an arbitrary function $f(\vec{r})$ of the spatial variable $\vec{r}$ on the plane S, given by:

$$\vec{\xi}\vec{r} = p$$

where $\vec{\xi}$ is the vector of the normal to plane and p is the distance from the plane to the origin of the reference plane. This transform is used as the basic mathematical approach in microwave tomography. *IAM*

Ref.: Nebabin (1995), pp. 65–67.

The **Walsh transform** is a discrete transform of a series of N real numbers, $s(n)$, $n = \overline{0, N-1}$:

$$g(k) = \sum_{n=0}^{N-1} \mathrm{Wal}(k, n) s(n), k = \overline{0, N-1}$$

$$s(n) = \frac{1}{N} \sum_{k=0}^{N-1} \mathrm{Wal}(k, n) g(k)$$

on the basis of discrete Walsh functions, Wal(k,n).

Several varieties of Walsh functions are used, which differ in the order of numbering. Thus the discrete Walsh-Hadamard functions are determined by the relationship

$$\mathrm{Had}(k, n) = (-1)^{\vec{k}^{\mathrm{T}} \vec{n}}$$

where $\vec{n}, \vec{k}$ are vectors comprised of binary bits of the number n (the argument of the function), and the number k (the magnitude of the function).

The Walsh transform and its variety known as the fast Walsh transform are used in digital signal processing. *IAM*

Ref.: Shirman (1981), p. 160.

The **wavelet transform** is a mathematical tool developed to deal with analysis and synthesis of nonstationary signals in a manner analogous to the conventional Fourier transform for stationary signals. It is based on the concept of the wavelet, which is an elementary wave of small duration and amplitude. It is used for analysis of signals in radar signal processing and imaging. *SAL*

Ref.: Gregoris, D. J., and Yu, S., "Introduction to the Theory and Application of Wavelet Transforms," Spar J. of Eng. and Technology ***3***, July 1994, pp. 20–32.

The **Winograd-Fourier transform (WFT)** is a special algorithm for the discrete Fourier transform that minimizes the number of multiplications required for a given n-point fast Fourier transform. In this algorithm the weight matrix is factorized into three matrices, one of which is diagonal with elements that can be either real or imaginary. The total number of multiplications required is of order n, corresponding to the number of points in the transform, while the number of additions is comparable to that of the conventional FFT. *SAL*

Ref.: Schleher (1991), p. 102.

The ***z*-transform** is a transform of a series $x(n)$, determined by the formula

$$X(z) = \sum_{n=0}^{\infty} x(n) z^{-n}$$

where $X(z)$ is the z-transform of the series $x(n)$. The transform makes sense for those values of the complex variable z at which the series converges. The reverse transform is determined by integration with respect to the closed path C in the z-plane, covering all the specifics of the integral function

$$x(n) = \frac{1}{j2\pi} \oint_C X(z) z^{n-1} dz$$

The z-transform is a form of the Laplace transform, with $z = e^{pT}$.

Using the z-transform, it is convenient to write various forms of expressions for transfer functions and thus obtain different forms of realization of digital filters. The z-transform is the basic method of calculating the output signals of discrete and digital filters under complex input signals. *IAM*

Ref.: Oppenheim (1983), Ch. 10; Gol'denberg (1985), p. 7.

TRANSFORMER, microwave. A microwave transformer is a device used to convert microwave voltages in transmission lines. Basically microwave transformers are used for conversion of load resistances (see **matching transformer**) used to connect transmission lines of various types and for matching of active loads within a given frequency band. *IAM*

Ref.: Johnson (1984), Ch. 43; Sazonov (1988), p. 39; Druzhinin (1967), p. 325.

balun transformer (see **quarter-wave transformer**).

An **exponential-line transformer** is a section of an irregular transmission line whose wave resistance in the field changed by an exponential law. An exponential transformer is built in the form of a smooth adapter of a transmission line. It is used for wide-band matching of antenna loads. Good matching using these transformers is achieved in all frequencies above a certain boundary frequency of the transformer. *IAM*

Ref.: Johnson (1984), p. 43.18; Sazonov (1988), p. 145.

A **half-wave transformer** is a section of a transmission line with electrical length equal to π radians, or λ/2. A half-wave section at the rated frequency does not transform the load resistance. At other frequencies, there is a change in the input resistance, which increases with the frequency deviation. This property determines the use of the transformer for correction of the frequency characteristic of the load resistance. Normalized voltages at the input and output of the transformer are opposite in phase, which is used in the devices for strict antiphase supply of a symmetrical load. *IAM*

Ref.: Sazonov (1988), p. 41.

A **matching transformer** is a microwave transformer intended for transmission line matching. The simplest matching transformer is a quarter-wave transformer, but it has limited wide-band capacity. To expand the band of passed frequencies with simultaneous improvement in matching quality, stepped and exponential transformers are used. Matching of transmission lines with shapes of different cross section and transmission lines of various types requires wave-type transformers. Other types of transformers can also perform matching functions: half-wave transformer; pulse transformer. *IAM*

Ref.: Johnson (1984), Ch. 43; Sazonov (1988), p. 140.

A **mode transformer** is a microwave transformer for connecting sections of transmission lines that use different types of waves. In practice, the seven most common waves used are:

$H_{10}^{\square}$, $H_{20}^{\square}$, $H_{11}^{\square}$ in rectangular waveguides,

E_{01}^{O}, H_{01}^{O}, H_{11}^{O} in round waveguides,

and a T-wave in coaxial and strip transmission lines. Corresponding to this, there are 21 types of transformers for different types of waves. In practice, an even larger number of transformer types are used.

A transformer is an adapter between lines, creating an electromagnetic field with the same components as the field of the necessary wave type, and not creating (or creating with low intensity) components not contained in the necessary type of wave.

Mode transformers are used for connection of waveguide, waveguide and coaxial, and waveguide and strip transmission lines. For example, a transformer for connecting a coaxial line to a rectangular waveguide is a probe in the form of a section of the central strand of a coaxial cable and is the antenna in a waveguide (type $T - H_{10}^{\square}$ transformer). A wave-type transformer $H_{10}^{\square} - H_{11}^{\square}$ is an adapter with smooth change of configuration of cross section from rectangular waveguide to round. Mode transformers with various types of waves have bandwidths of 30 to 50%. *IAM*

Ref.: Lavrov (1974), p. 331.

A **pulse transformer** transmits pulses without distortion of their shape. To reduce the scattering inductance and capacitance of the pulse transformer, the volume of the transformer core, made of ferrite or high-quality steel with a high magnetic constant, is increased, the number of turns in the primary and secondary windings is limited (no more than a few dozen), special connection circuits of the windings are used, and the transformation factor of the pulse transformers is limited usually to values of 3 to 10, less often up to a 100. Pulse transformers can have a power over more than 100 kW and are designed for a voltage of more than 6 kV.

Pulse transformers are basically used to increase the amplitude of pulses, change their polarity, match resistance, and isolate dc sources of pulses and the load target, and so forth. They are widely used in pulse modulators of radars. *IAM*

Ref.: Glasoe (1948), Chs. 12-14; Druzhinin (1967), p. 325; Jordan (1983), p. 13.17.

A **quarter-wave transformer** is a section of a transmission line with an electrical length equal to $\pi/2$ radians, or $\lambda/4$. The quarter-wave transformer converts the load resistance into a resistance that is numerically equal to the conductivity of the load. It is used for connecting transmission lines to various wave resistances and for narrowband matching of random active loads. The quarter-wave transformer is also called a *balun (transformer)*. *IAM*

Ref.: Sazonov (1988), p. 41; Johnson (1984), pp. 43.12, 43.23–43.27.

TRANSISTOR, microwave. A microwave transistor is "an active semiconductor device with three or more terminals" operating at microwave frequency. Transistors are subdivided into two large classes: bipolar transistors and field-effect transistors (FETs). The basic parameters of transistors include power gain, noise factor, power, efficiency of the collector circuit, maximum generating frequency, maximum frequency, as well as the *S*-parameters of the transistor, the complex components of its scattering matrix.

Transistors are also classified in accordance with the type of semiconductor (silicon, germanium, etc.),and the specific features of the physical processes and manufacture (drift, diffusion, diffused).

In the microwave range, FETs (with Schottky barrier) are widely used. Owing to the simpler and more advanced technology of manufacture, they have a smaller spread of electrical parameters than bipolar transistors. The current in them does not flow through the PN junctions, but between ohmic contacts in a uniform channel medium, so FETs possess a higher linearity of volt-ampere characteristic, they have no current distribution noise, and the current density can be greater. Owing to the greater mobility of electrons in gallium arsenide than in silicon, and the low capacitance of the reverse bias of the Schottky barrier in the gate, FETs operate at high frequencies, up to 90 to 120 GHz. FETs surpass bipolar ones in their output power at frequencies of 4 to 5 GHz, and in noise factor at frequencies above 1 to 1.5 GHz. *IAM*

Ref.: Jordan (1983), pp. 18.14–18.26.

A **bipolar (junction) transistor** is a semiconductor device with two interacting PN junctions and three leads, whose amplifying properties come from the phenomena of injection from emitter to base of secondary carriers and their extraction from the base to the collector. In the operating (active) mode, the collector-base junction is closed by a strong reverse bias, and the emitter-base junction is open by a small direct bias. The input resistance of the transistor is therefore small. The output resistance is much greater than the input, since the collector current is determined by the current of the emitter, which depends hardly at all on the voltage at the collector.

Modern microwave bipolar transistors are made by planar-epitaxial technology. In low-power (low-noise) bipolar transistors, germanium PNP and silicon NPN structures are used, and in high-power ones, only silicon NPN structures, which have better parameters owing to the greater (about double) the mobility of electrons in comparison with holes.

Frequency properties are usually characterized by the maximum frequency f_{max} of generation, which is practically invariant with respect to the connection scheme of the transistor. The values f_{max} can reach 10 to 20 GHz but bipolar transistors are used at frequencies 1.5 to 2 times lower, where their gain exceeds 3 to 6 dB. Noise properties of transistors are described by the noise factor, which depends on the design of the transistor (see **low-noise transistor**), the frequency, voltage, and current.

To describe the microwave properties of a bipolar transistor, we use its *S*-parameters, complex coefficients of the scattering matrix, measured at operating frequencies at low and high amplitude levels. These define the gain and the input and output resistances of the transistor. *IAM*

Ref.: Jordan (1983), pp. 18.18–18.22; Gassanov (1988), p. 84; Liou (1996).

A **diffused transistor** can be either of two types:

(1) A transistor whose PN junction is made by the diffusion method. In this case drift transistors, which in the microwave range are made by planar-epitaxial technology, are counted as diffused transistors. This technology consists of coating (by epitaxial buildup) of a collector plate of semiconductor with a low specific resistance and a layer of the same semiconductor with a high specific resistance. The PN junctions are formed by diffusion of the admixture through an opening in a specially applied protective layer.

(2) A diffusion transistor, with movement of secondary carriers in the base that is basically of the diffusion type. It operates at lower frequency than the drift transistor. *IAM*

Ref.: Zherebtsov (1989), pp. 108–111.

A **drift transistor** is a bipolar transistor in which the transfer of secondary charge carriers through the base occurs chiefly through drift (i.e., under the action of an accelerating electrical field). In the base of the drift transistor, through nonuniform concentration of the impurities, an electrical field is created which accelerates the secondary charge carriers as they move to the collector. This helps raise the maximum frequency of the transistor and its gain in comparison with diffusion transistors. In the latter, a field does not arise in the base (with a low injection from the emitter), and movement of carriers in it is basically the diffusion type. A drift transistor is made by diffusion methods. (See **diffused transistor**.) *IAM*

Ref.: Zherebtsov (1989), p. 108.

A **field-effect transistor (FET)** is a semiconductor device with three leads, in which the output circuit does not contain PN junctions and the current in this circuit is controlled by an electrical field.

Microwave FETs are made of gallium arsenide (GaAs) of the N-type with planar-epitaxial technology with a gate that uses a Schottky barrier. The source and drain electrodes have ohmic contacts with the semiconductor. The supply voltage (U_{si}) is connected in such a way that electrons in the thin epitaxial N-layer channel move from the source to the drain. The current through the load is determined by the resistance of the channel, which depends on the voltage at the gate due to the change in thickness of the impoverished region of the Schottky barrier under the gate.

In contrast to the bipolar transistor, the FET has a high input resistance, is controlled by a voltage, and is characterized by the curve $s = \Delta I_s / \Delta U_{zi}$.

Thanks to the greater speed of the electrons in GaAs, than in silicon (Si), and the comparatively low capacity of the Schottky barrier, FETs can operate at frequencies up to 90 to 120 GHz, which vitally exceeds the frequency of use of bipolar transistors. This is why they are preferred, compared with bipolar transistors, for use as low-noise transistors at frequencies higher than 1 to 1.5 GHz, and as high-power transistors at frequencies above 4 to 5 GHz. The FET is also called a *unipolar transistor. IAM*

Ref.: DiLorenzo (1982); Gassanov (1988), p. 93.

A **germanium transistor** is a transistor that is manufactured on the basis of germanium (Ge). The addition of insignificant quantities of elements of a third group (indium, In, gallium, Ga, etc.) imparts a hole-type character to the conductivity of Ge, while elements of a fifty group (antimony, arsenic, etc.) give it an electron nature.

Ge microwave transistors include basically the low-noise bipolar PNP transistors. In the microwave band, silicon transistors and transistors based on GaAs are the most widely used. (See **field-effect transistor**.) *IAM*

Ref.: Gassanov (1988), p. 84.

A **heterojunction transistor** is a promising type of bipolar transistor with junctions formed at the boundary of two conductors having different-width restriction zones. In a bipolar transistor with a heterojunction, there is a nearly ideal unilateral injection of carriers into the base. Heterojunctions make it possible to vitally raise efficiency and the frequency boundary of operation of the device.

Models of transistors have been developed with a maximum frequency of 60 GHz, and it is expected that a maximum frequency of 100 to 200 GHz will be achieved. *IAM*

Ref.: Gassanov (1988), p. 211.

A **high-power transistor** is distinguished by its high output power. High-power bipolar and FETs are used (Table T4). High-power bipolar transistors are made of silicon, whose thermal conductivity is three times greater than that of gallium arsenide. At comparatively low frequencies in the microwave band (less than 2 to 3 GHz), the output power of bipolar transistors is limited by the difficulty of cooling the low-doping area of the collector and drops proportionally to the frequency. At higher frequencies, it is limited by the danger of avalanche breakdown of the collector junction and by the speed of the carriers, and drops proportionally to the square of the frequency.

Table T4
Parameters of High-Power Transistors

Type of transistor	Output power, 10 log P_o (dBW) at frequencies			Typical supply mode	
	2 GHz	5 GHz	35 GHz	Voltage, V	Current, mA
Bipolar	20	8	–	15-28 (collector)	units of amps
Field-effect	–	15	10	8-10 (drain)	units of amps

High-power FETs are made of GaAs because of its three-times greater mobility of electrons in the channel compared with Si, despite the worse thermal conductivity of the GaAs.

High-power transistors, as a rule, have several crystals in one housing, with multigate cells, which increases the total width of the channel. To improve cooling, a substrate is used that is coated with a gold heat sink, connected through small openings in the substrate to drain contacts.

With an increase in frequency, the efficiency of high-power transistors can drop to 15 to 20%, and the gain can be 3 to 6 dB. *IAM*

Ref.: Gassanov (1988), pp. 90, 97.

A **low-noise transistor** is marked by a low noise factor. As a rule it has low power due to the small dimensions and thickness of the junctions, and a high gain factor (no less than 6 to 10 dB). Bipolar and field-effect transistors may be used as low-noise transistors. The noise factor of bipolar transistors increases with the frequency at a rate of 1 to 3 dB/octave and with collector voltage and emitter current (Table T5). In FETs, the noise is basically of thermal origin and decreases with cooling. However at frequencies less than 1 GHz, the noise of FETs varies inversely with frequency. In this regard, FETs usually have advantages over bipolar ones in noise factor at frequencies higher than 1 to 1.5 GHz. *IAM*

Ref.: Gassanov (1988), pp. 89, 97.

Table T5
Parameters of Low-Noise Transistors

Type of transistor	Noise factor (dB) at frequencies			Typical supply mode	
	2 GHz	5 GHz	35 GHz	Voltage, V	Current, mA
Bipolar	1.5	2.5	-	5-10 (collector)	1-2 (emitter)
FET	0.6, 0.2 (*T* = 20K)	0.8, 0.2 (*T* = 20K)	2.5, 1.8 (*T* = 20K)	3 - 4 (drain)	10 (drain)

A **metal-insulator-semiconductor (MIS) transistor** is an FET with the structure of a metal-dielectric semiconductor whose metal gate is separated from the semiconductor by a thin layer of dielectric material. In the microwave region up to 30 GHz and above, type D-MIS and V-MIS are used. The first are made by the method of double diffusion. The channel is formed along the thickness of the thin layer of semiconductor, making it possible shorten the channel and to raise the frequency boundary.

In V-MIS transistors, a V-shaped groove is etched in the silicon NPN structure, with a dioxide fill formed on its surface. Metallization is performed for the gate. *IAM*

Ref.: Zherebtsov (1989), p. 123.

A **metal-oxide-semiconductor (MOS) transistor** is an MIS transistor whose dielectric is a layer of oxide, usually silicon dioxide (SiO_2). *IAM*

Ref.: Zherebtsov (1989), p. 120.

A **multiplier transistor** is a frequency multiplier that uses nonlinear resistance or nonlinear capacitance of a transistor junction. A multiplier based on a bipolar transistor with the use of a nonlinear resistance in the circuit is analogous to an amplifier with collector output circuit tuned to the *n*th harmonic. A transistor operates with cutoff of the collector current. The output resonant systems suppress the secondary components by 30 to 40 dB. The multiplication factor does not exceed 3 or 4. In comparison with the diode multiplier, they have good separation of input and output. The basic drawback is the low bandwidth of the processes, which limits use to one-tenth of the limit frequency of the transistor, f_t. A transistor-parametric multiplier operates with a nonlinear capacitance of the PN junction of the collector-base to a frequency of (2 to 3) f_t, but is quite difficult to tune due to the danger of oscillation.

A multiplier based on FETs is marked by a higher frequency limit, up to tens of gigahertz, better separation of input and output, and the absence of a modulating capacitance at the output, which leads to parametric effects. Multipliers use nonlinear resistance of the FET, connected in a circuit with a common source in a mode close to current cutoff. A frequency doubler at 8 GHz has an output power of 10 mW with gain of around 3 dB. The use of double-gated Schottky FETs, with the first transistor of a tetrode in a mode of linear amplification is promising. A doubler with a frequency of 18 GHz has an output power of 1 mW with a gain of 4 dB. When the input power is increased to 10 dB, the efficiency of the multiplier drops to 5 to 6 dB. *IAM*

Ref.: Gassanov (1988), p. 204.

A **silicon transistor** is a transistor made on the basis of silicon (Si). They are widely used as NPN low-noise and high-power bipolar microwave transistors, and also as MIS transistors with a V-type Schottky at frequencies below 1 GHz. At higher frequencies, FETs based on GaAs have better characteristics. *IAM*

Ref.: Gassanov (1988), pp. 84, 97.

A **tunnel transistor** is one in which, just as in a tunnel diode, the tunnel effect at the emitter junction is used. A tunnel transistor has good controllability and provides high amplification of voltage at the output. It eliminates the influence of the output on the input, which makes it possible to use a tunnel transistor as a switch. *IAM*

Ref.: Popov (1980), p. 439.

unipolar transistor (see **field-effect transistor**).

TRANSMISSION LINE, microwave. A microwave transmission line is a device that limits the range of propagation of electromagnetic oscillations and direct the flow of electromagnetic energy to the load. The choice of the line type (Table T6) is determined by the purpose and operating conditions of the system and depends considerably on the wave band and the transmitted power.

Depending on purpose, regular or irregular transmission lines are used. Transmission lines are used to transmit power from generators to loads, to form resonant systems (three-dimensional cavities and oscillating circuits with distributed parameters), and to form full load resistances. Sections of transmission lines are used also for combining individual microwave devices into a single circuit.

The basic properties of transmission lines are attenuation, reflection coefficient, efficiency, electrical length, and power-handling capacity.

Table T6
Microwave Transmission Lines

Type of line	Wave band
Wire, coaxial (coaxial cable)	VHF, meter, decimeter
Waveguide, coaxial (coaxial waveguide)	Decimeter, centimeter, millimeter
Strip, surface wave (single-wire)	Meter, decimeter, centimeter
Dielectric	Millimeter
Fiberoptic	Optical

Attenuation in a transmission line reduces the power P of a wave traveling along the line. It is characterized by the attenuation constant, which is equal to the length of the line in which power drops by a factor of e^{-2} and determines the dependence of the power on the distance l of the transmission line in the direction of travel of the wave:

$$P(l) = P(0)e^{-2\alpha l}$$

The attenuation constant usually increases with frequency, in dielectrics directly proportionally to f, in wires directly proportional to the square root of f, and in hollow waveguides depending on the ratio of operating to critical frequency (tending toward infinity as it approaches the latter).

The reflection coefficient of a transmission line is the ratio of the transverse components of an electrical field for reflected and incident waves in at one and the same point of the transverse section of the line. The coefficient of reflection in the field coincides with the coefficient of reflection for normalized voltages of reflected and incident waves and depends on the longitudinal coordinate l, which is figured from the load by the formula

$$\rho(l) = \rho(0)e^{j2\gamma l}$$

where $\gamma = \beta - j\alpha$ is the coefficient of propagation for the applicable type of wave, β and α are the phase and attenuation constants. With an increase in l, the reflection coefficient acquires a phase delay of $2\beta l$ and reduces its magnitude due to attenuation of the wave in the transmission line.

Transmission line efficiency is a coefficient equal to the ratio of the power delivered to the load to the power of the incident wave input to the line. The difference between the efficiency η and unity is due to the power losses from attenuation by a factor of $e^{-2\alpha l}$, and losses due to reflection of the incident wave from the load by a factor of $1 - |\rho_h|^2$, where $|\rho|$ is the magnitude of the reflection factor in the line section near the load (for $l = 0$):

$$\eta = e^{-2\alpha l}(1 - |\rho_h|^2) = \frac{4K_h e^{-2al}}{(1 + K_h)^2}$$

where K_h is the traveling-wave ratio.

The most favorable conditions for transmission of power occur when $K_h = 0$ or $|\rho_h| = 0$, which corresponds to a matched load.

The electrical length of transmission line characterizes the amount of phase delay of the reflection coefficient in the section, equal to

$$L = \beta l$$

where l is the length of the section and β is the phase constant.

The power-handling capacity of a transmission line is limited by electrical breakdown or overheating of conductors and insulators in the line. Usually the acceptable power is taken as 25 to 30% of the maximum rating.

The basic wave modes used in transmission line are T-waves, E-waves, H-waves, and EH- or HE-waves. (See **WAVE**.) *IAM*

Ref.: Jordan (1983), Chs. 29, 30; Sazonov (1988), pp. 12–40.

A **coaxial transmission line** is either a rigid line made of metal tubes secured one within the other by dielectric washers or metal stubs, or a flexible coaxial cable. In the coaxial transmission line, the main T-type wave is propagated (transverse wave). The attenuation factor comes from losses in the conductors and dielectric, and at a frequency of 1 GHz amounts to 0.1 to 0.7 dB/m. The standard impedances of 50 and 75 ohms are selected in the region where the parameters of the coaxial line are close to optimum. The diameters of the conductors are: internal, 0.06 to 3 mm; external, 4 to 20 mm; maximum operating voltage, 1 to 8.5 kV.

The bands in which such lines are used includes wavelengths from 3 to 5 cm up to 10m. *IAM*

Ref.: Jordan (1983), p. 29.28; Rakov (1970), vol. 2, p. 231; Sazonov (1988), p. 18.

Transmission-line components are sections of transmission line that implement various operations on the propagating waves within them: attenuation, power combining or dividing, phase shifting, and so forth. There are many such components, the most important of which are listed in Table T7. (See also **ATTENUATOR**; **BRIDGE**; **CIRCULATOR**; **COUPLER, DIRECTIONAL**; **DIVIDER**; **JOINT**; **PHASE SHIFTER**; **POLARIZER**; **SWITCH**.) *SAL*

Ref.: Lavrov (1974), Ch. 18.

A **coplanar transmission line** is a conducting strip transmission line formed by parallel, closely arranged slots in a metallic layer on one side of a dielectric plate with a high dielectric constant (> 10). The slow H-wave is its main type of wave. *IAM*

Ref.: Sazonov (1988), p. 22.

Table T7
Transmission-Line Components

Component Type	Function
1. Transmission-line sections: (a) coaxial line, (b) stripline, (c) waveguide	Transfer of RF along a specified path
2.Joints: (a) stationary, (b) rotary, (c) adapters, (d) transformers	Joining of two transmission line sections
3. Dividers: (a) coaxial tee (b) waveguide tee	Splitting of power from a single to multiple channels and combining of power from several to one channel
4. Switches	Time-dependent connection of channels
5. Decoupling devices: (a) attenuators, (b) directional couplers, (c) circulators	Reduction of power transferred from one channel to another or complete decoupling of the channels
6. Polarizers	Change in polarization of the transmitted wave
7. Phase shifters	Change in phase of the transmitted wave
8. Bridges	Addition, subtraction, and calibrated splitting of RF waves in 4-channel junctions
9. Protectors	Protection of a load from excessive power
10. Matching devices	Matching of transmission-line sections to obtain specified reflection coefficient
11. Balance-to-unbalance devices	Transfer from asymmetrical to symmetrical components and vice versa

A **dielectric transmission line** is one that uses slowed surface waves, the basic power of which is propagated over the surface of the dielectric. Dielectric transmission lines are made as continuous or hollow dielectric rods of rectangular or elliptical cross section, dielectric lines on the surface of metal plates (mirror dielectric waveguides), or between two parallel plates.

In the millimeter waveband, dielectric transmission lines have an attenuation coefficient of 0.02 to 0.2 dB/m, which is an order of magnitude less than in a copper rectangular waveguide. *IAM*

Ref.: Sazonov (1988), p. 25.

A **fiber-optical transmission line** is usually a multimode dielectric waveguide in the form of threads around 150 μm thick made of an especially pure quartz, which possesses insignificant losses in the optical wave range.

The operation of fiber-optical lines is based on the propagation of light as a result of complete internal reflection. To eliminate losses of light in irregularities of the surface, the boundary of full internal reflection is shifted inside the fiber through the creation of a stepped or smooth radial irregularity of the optical coefficient of refraction. Fibers coated with lacquer are used to form a multistrand fiber-optic cable.

Attenuation in these lines is determined only by the dielectric losses and amounts to 3 to 5 dB/km. They provide very wide operating frequency bands, ideal interference immunity, and practically complete separation between channels. They are used as multichannel connectors, directional couplers, commutators, filters, and so forth, in the submillimeter and optical wave bands. *IAM*

Ref.: Sazonov (1988), p. 28.

homogeneous transmission line (see **regular transmission line**).

An **irregular transmission line** has irregularities: specially constructed elements connected in transmission lines that support the given function of the line. Irregularities also may arise due to imprecise adherence to the given identical geometric dimensions and electrical parameters of the metals and dielectric materials in the process of manufacture and/or operation.

We distinguish between concentrated irregularities, whose extent along the transmission line is much less than the wavelength, and distributed ones, whose length amounts to a marked fraction of the wavelength or exceeds it. An example of a concentrated irregularity is a jump in the size of the wall of a rectangular waveguide, used in stepped microwave junctions. Distributed irregularities are characterized by the continuous reflection of electromagnetic waves over the entire length of the irregularity. One example of such an irregularity is the conical junction between coaxial transmission lines and various wave resistors, or the smooth junction between waveguides of different cross-section.

The basic waves of the transmission line are reflected from the irregularities. Other types of waves exist near the irregularity in the form of reactive fields that do not transmit energy. Reactive fields provoke scattering of energy into heat in the region of the irregularity, and this, as a rule, is greater than the losses in the same region from a passing main wave. Microwave junctions are examples of irregular transmission line. *IAM*

Ref.: Lavrov (1974), pp. 230, 284, 301.

An **irregularly connected transmission line** is a section of transmission line, usually two-wire, tape, multistrip, or coaxial line connected to the circuit in such a way that currents at the poles of a section are different. This is achieved through short-circuiting of the pair of poles either in the diagonal or in the horizontal, or through insulation of one pole from the outside circuit, and also through a combination of these techniques. The section of wire that causes the short must be maximally short, and for this purpose in practice line sections are rolled into a ring or wound on a toroid.

Irregularly connected lines are used as miniature functional devices in the decimeter and centimeter bands. A necessary feature of devices based on irregularly connected lines is a strong magnetic connection between wires of an irregularly connected line, which allows miniaturization and wideband characteristics. Matching devices based on irregularly connected lines can be simultaneously miniature and wideband, in contrast to stepped junctions. Low-frequency filters and high-frequency filters are easily made in the printed variant, and bandpass filters are made through cascade connections. *IAM*

Ref.: Zelakh (1989), pp. 3, 37, 53.

A **long transmission line** is one whose length is comparable to or longer than the wavelength. Propagation in such a line has a wave character, with a velocity less than that of free space, and the line can be regarded as a circuit with distributed parameters. If these parameters (capacitance, inductance, and resistance) are distributed uniformly, the line is called *homogeneous*, and if the resistance and loss are zero the line is called *ideal.*

When a long line of finite length is fed by an oscillator and is loaded with an impedance equal to the line impedance, the traveling-wave mode occurs. If the line is short-circuited and has no load or a reactive load, the standing wave mode occurs as a result of total reflection from the end. If the load is an impedance not equal to the line impedance, a hybrid mode occurs in which both standing and traveling waves are present. The main parameters describing long line performance are the traveling and standing wave ratios. (See **WAVE, standing.**). *SAL*

Ref.: Popov (1980), p. 119.

Transmission line matching provides a mode for a transmission line (matched mode) in which the traveling-wave ratio is equal to unity. For matching, reactive components are used (without ohmic losses) to raise the efficiency of the circuit. We distinguish between narrowband and wideband matching. In the former, the traveling-wave mode is achieved in a single frequency, and the matched bandwidth is determined by the condition of acceptable reduction in the traveling-wave coefficient to some level (for example, 0.8). In narrowband matching, compensation of reflection from the load is at a selected frequency by introducing an additional reflection to the line near the load using loops, irises, or other concentrated reactive components.

In wideband matching, a given value of the band (up to several octaves) is attained at an acceptable level of traveling-wave ratio by selecting the values of the matching components, multistage or smooth junctions. *IAM*

Ref.: Sazonov (1988), pp. 44, 140.

A **microstrip transmission line** is a trip transmission line with a substrate of material with a high dielectric constant, ε = 10 to 15, permitting the size to be reduced by roughly the square root of ε compared with an air line. Asymmetrical microstrip lines are the most common. *IAM*

Ref.: Sazonov (1988), p. 21; Gupta (1996).

A **regular transmission line** has identical geometric and electrical parameters along the entire length. In a regular line there are no irregularities of any kind, including junction and connecting devices, matching elements, and so forth.

Based on the type of propagated electromagnetic fields, we distinguish among: (1) transmission lines with transverse electromagnetic waves; (2) transmission lines with electric E- and magnetic H-waves; and (3) surface-wave transmission lines. *IAM*

Ref.: Lavrov (1974), p. 284; Sazonov (1988), p. 13.

A **slot(ted) transmission line** is a strip transmission line in the form of a slot in the conducting layer on one side of a dielectric sheet with high dielectric constant. In a slotted line, a slowed H-wave is propagated, with an electromagnetic field concentrated close to the slot. The critical frequency of this main wave is equal to zero, but there is significant dispersion. Slotted lines may be placed in rectangular screens. Such waveguide slotted lines are conveniently combined with circuits based on rectangular waveguides and also are often used in designs of waveguide strip radiating elements. *IAM*

Ref.: Sazonov (1988), p. 22.

Strip(-line) transmission line is formed from parallel metal conductors and dielectric plates. We distinguish between symmetrical lines, which have two perpendicular planes of symmetry in their cross section, and asymmetrical lines, which have one plane of symmetry. There are three varieties of striplines; rigid air lines; lines based on foil-covered dielectric plates; and lines based on dielectric plates made of ceramic or crystalline materials with a high dielectric constant. The last include microstrip, slotted, and coplanar lines.

Rigid air lines are used at high power, and most often are made to be symmetrical, in the form of a flat conductor secured by insulators between external conductors. Striplines of the second type are produced in symmetrical and asymmetrical forms.

Wave resistances of striplines are 20 to 100 ohms and are easily realized by selection of the width of the wires. Calculations of parameters of lines is done on a digital computer with special programs.

Striplines are widely used in decimeter and centimeter waves, basically for the formation of complex, ramified circuits combining many elements into a single microwave device. *IAM*

Ref.: Sazonov (1988), p. 19.

Surface-wave transmission line uses the slowed surface waves moving in free space over the slow-wave structure. Usually a dielectric material is used as the slow-wave structure (see **dielectric transmission line**), as well as a structure in the form of transverse grooves made on a wire, or a coaxial dielectric layer. The latter structure is used in single-wire transmission lines when the wave length is from several meters to 3 to 5 cm. Through the choice of the slow-wave coefficient in a single-wire line, an attenuation coefficient several times less than that of a rectangular waveguide can be obtained. However, the characteristics of a single-wire line degrade when it is bent. *IAM*

Ref.: Sazonov (1988), p. 210.

Transverse-wave transmission lines are those with T-type waves (T-waves) that do not contain longitudinal components of an electromagnetic field. They exist only in transmission lines which have no less than two insulated conductors (wire, coaxial transmission line, certain types of striplines). The critical frequency for T-waves is equal to zero. (See **WAVEGUIDE.**) *IAM*

Ref.: Lavrov (1974), p. 284; Jordan (1983), pp. 29.25–29.37.

Transmission lines with electrical and magnetic waves do not have longitudinal components of the magnetic and electrical field, respectively (waveguides, including with homogeneous dielectric filling, coaxial lines with higher types of waves). The critical frequencies of H- and E-waves differ from zero and depend on the shape and dimensions of the transverse section, and also on the parameters of the filling dielectric. *IAM*

Ref.: Lavrov (1974), p. 284; Sazonov (1988), p. 23.

A **waveguide transmission line** is based on metal waveguides of various cross-sections: rectangular, round, or H- and Pi-shaped waveguides. Most often, rectangular waveguides with a basic wave of type H_{10} are used for transmission of microwave power in the decimeter, centimeter, and millimeter-wave bands. The dimensions of their cross section $a \times b$ are selected on the basis of contradictory requirements on the transmission-line parameters (Table T8).

Table T8
Characteristics of Rectangular Waveguide Transmission Line

Size, $a \times b$ (mm)	Critical wavelength (cm)	λ_{min} – λ_{max} (cm)	Minimum attenuation, dB/m	Rated power (kW)
72×34	14.4	7.7–13	0.02	3000
23×10	4.6	2.3–4.1	0.12	300
7.2×3.4	1.44	0.75–1.2	0.51	40

N- and Pi-shaped waveguides have smaller transverse section and an expanded passband (to several octaves), but lower power-handling capacity and higher attenuation coefficient. Round metal waveguides are used chiefly for creating various circuit components, and less often for transmitting power over significant distances. Three types of waves are used most widely: H_{11}, H_{01}, and E_{01}. The level of transmitted power over a waveguide line may be increases several times over when the waveguide circuit is filled with gas such as SF_6, and the pressure is raised to (3 to 5) $\times 10^5$ Pa. (See also **WAVEGUIDE.**) *IAM*

Ref.: Jordan (1983), Ch. 30; Sazonov (1988) p. 23; Rakov (1970) Vol 2, pp. 232–238.

Wire transmission lines consist of two or four identical parallel wires. Open lines or lines in the form of flexible cables (two-wire cables) may be used. The wave resistance of these lines is 200 to 600 ohms. Four-wire transmission lines are formed from paired connected wires and are marked by lower parasitic emission, higher wave resistance, and better power-handling capacity.

Open wire transmission lines with T-waves are used in the hectometer and meter wavebands for connection of antennas to receiver and transmitter devices. *IAM*

Ref.: Sazonov (1988), p. 16.

TRANSMISSIVITY is a measure of an object's transparency to RF energy. An object that is completely transparent to RF energy (i.e., allows all incident RF to pass through), has a transmissivity of 1.0. Conversely, if the object allows no incident RF energy to pass through, its transmissivity is 0. *PCH*

Ref.: Currie (1987), p. 818.

TRANSMITTER, radar. A radar transmitter is a subsystem for generating and forming RF signals to be radiated in the target direction. There are two main classes of radar transmitters: vacuum-tube transmitters, and solid-state transmitters; and two basic transmitter configurations: power-oscillator transmitters and power-amplifier (amplifier chain) transmitters. A transmitter is a major part of any radar system, and its performance, size, cost, maintainability, and reliability significantly affects the corresponding parameters of the whole radar system. Moreover, the choice of transmitter type (tube or solid-state) defines the image of radar and approach to the system design in general.

The requirements on a transmitter evolve from a set of system requirements, principally the following: peak and average power, operating frequency range, bandwidth, waveform type, gain, pulse repetition frequency, MTI or doppler performance, reliability, maintainability and service factors, size and weight, and cost. There is no one universal transmitter that meets all the requirements equally well. Each type of transmitter and its configuration has its own advantages and limitations, and the final design typically is a series of compromises to reach required system performance. *SAL*

Ref.: Ewell (1981); Skolnik (1980), pp.190–221; Skolnik (1990), pp. 4.1–4.41; Gilmour (1986); Ostroff (1985); Fink (1975), p. 25.58; Brookner (1988), pp. 41, 266–268, 279, 326, 327, 329; Currie (1987), pp. 447–482; Leonov (1988), pp. 43–53; Sivan (1994).

An **amplifier-chain transmitter** is a transmitter in which the signal is generated at low level and then is amplified to required power level. The main advantage of such a configuration compared with a power-oscillator transmitter (usually based on magnetrons) is that such systems can usually ensure full coherence from pulse to pulse, compatibility with sophisticated pulse-compression waveforms, true frequency agility, combining, and arraying. Certainly, all these advantages are obtained at the expense of the higher complexity and cost. Many radars requiring high-power operation combined with the necessity of coherence and pulse-compression waveforms employ amplifier chain transmitters. The combination of a traveling-wave tube in the medium stage and output crossed-field amplifier is common that ensures good gain in the medium stages and high resultant power and efficiency in the output stage. In Russian radar literature, an amplifier chain transmitter is usually called a *multi-stage transmitter*. *SAL*

Ref.: Skolnik (1990), p. 4.9; Leonov (1988), p. 43.

An **astronomical radar transmitter** is one with high average coherent power. At frequencies higher than 400 MHz, it is generally based on klystrons due to their greater gain (40 to 60 dB per stage), high average output power (up to 400 kW per tube), and simplicity of changing the operating frequency. A klystron transmitter requires a high dc operating voltage and removal of a large amount of heat due to the low efficiency (around 40%). At lower frequencies, powerful triodes and tetrodes are used, which are controlled in a grid.

At frequencies above 3 GHz, radar astronomy transmitters are often used in the continuous mode, ensuring a high average power in a given mode. *IAM*

Ref.: Skolnik (1970), p. 33.20.

Transmitter efficiency is the ratio of the RF power available from the transmitter to the total power needed to operate the transmitter. The transmitter efficiency is less than corresponding tube efficiency. For example, if the efficiency of the tube employed in transmitter is 40 to 50%, the transmitter efficiency might be 20 to 25% depending on the tube type and condition of its application. *SAL*

Ref.: Skolnik (1980), p. 191.

A **laser radar transmitter** can be a continuous or pulsed light source. Three types are most widely used: a rubber laser with emission in the wavelength $\lambda = 0.69$ μm, a glass and neodymium laser with $\lambda = 1.06$ μm, and garnet laser, activated with neodymium, with the same wavelength. Pulse powers realized by these lasers reach 10^9W with a pulse duration of 10^{-8} sec and a repetition frequency of up to 10 Hz or more. These types of lasers, as a rule, operate in the pulsed mode with optical pumping. The luminous energy of pulsed or continuous tubes is used as the pump. Gas CO_2 lasers operating in a wavelength of 10.6 μm are marked by the monochromatic nature of their emissions. They have sufficient power (10W in the continuous mode), can operate both in continuous and in mode modes, and are economical (efficiency up to 10%). They are used to make various laser doppler systems with receiver devices of the heterodyne type.

To increase the emission power of lasers in the monopulse mode, multicomponent lasers (cascade lasers) are used in which one element performs the functions of the laser oscillator and the others the functions of amplifiers. To obtain a higher pulse repetition frequency at the output of the system, exceeding the repetition frequency of an individual laser, an array of several parallel lasers is used, whose beams are merged into a single beam using optical means. *IAM*

Ref.: Ustinov (1984), p. 158.

power-amplifier transmitter (see **amplifier-chain transmitter**).

A **power-oscillator transmitter** generates RF signals at the required power level using a single microwave device without additional stages of amplification. Typically, in radar applications, this device is a magnetron. In this case the configuration of the transmitter is more simple, and transmitting system usually smaller, lighter, and less expensive, but it is impossible to overcome the limitations introduced by the inherent oscillator properties. (See **MAGNETRON.**) In Russian radar literature this type of transmitter is usually called a *single-stage transmitter*. *SAL*

Ref.: Skolnik (1990), p. 4.9; Leonov (1988), p. 43.

A **solid-state transmitter** is a transmitter using solid-state devices as the principal active components. There are three main groups of solid-state devices that can be used in radar transmitters: microwave transistors, microwave diodes, and electron-bombardment semiconductor devices.

The main representatives of microwave transistors that can be a choice for solid-state transmitters are bipolar transistors and field-effect transistors (FETs). The are three main disadvantages that have power transistors from the point of view of their use in radar transmitters:

(1) The CW power that can be obtained from a single transistor is small compared with the power that can be obtained with a tube, typically not exceeding tens of watts. To increase the power output, transistors may be operated in parallel, but too large number can not be used because of losses occurring in combining. Typically two to eight transistors can be combined in single power module.

(2) Transistors in principle are average-power devices (while tubes typically are peak-power devices), their peak power outputs being limited by electrical characteristics, and as a result they are usually operated with relatively long pulsewidths and high duty cycles.

(3) The power output of microwave transistor decreases inversely as the square of frequency (approximately 6 dB/octave), so conventional silicon bipolar transistors with appreciable power output can be used only up to S-band. To increase operating frequency, frequency multipliers can be used (e.g., varactor frequency multiplier).

Diode components operating as low-power oscillators or negative-resistance amplifiers are usually considered for millimeter-wave solid-state power sources. The most promising results have been obtained for IMPATT and Gunn diodes. The output power of tens watts and efficiency up to 20% for these devices has been reported. A hybrid approach that combines semiconductor and vacuum-tube technology resulted in electron-bombardment semiconductor device containing a heated cathode that generates an electron beam striking a semiconductor diode at high energy. Such a device can be broadband, have a high gain (25 to 35 dB), high efficiency (50%), but at the expense of lacking of some primary advantages of solid-state devices as it requires high-power supply and cathode heating.

The choice between vacuum-tube and solid-state components is of primary importance for radar designer when the type of transmitter is under consideration, as it affects the image of the whole facility.

The main assets of solid-state transmitters as compared with tube transmitters are the following:

(1) Considerable simplification of operational, service, and maintenance requirements as the transmitter operates at much lower voltages.

(2) There is no need in hot cathodes, high-power modulators and other complex high-power elements.

(3) Air-cooling may be employed.

(4) The risk of dangerous x-rays and electric shock is eliminated.

(5) No pulse modulator is required, as solid-state microwave devices generally are self-pulsing.

(6) Wide bandwidth (up to 50% compared with 10 to 20% in tubes).

(7) Long, failure-free life and graceful degradation of system performance as the system usually consists of large number of modules.

(8) Improved and compact technology (e.g. employing active transceiver modules for phased-array radars), that makes it possible to develop inexpensive, compact, light, and reliable entirely solid-state radars that can operate without personnel attendance.

The main problem restricting usage of entirely solid-state transmitters arises from the fact that in many radar applications the requirement for high average power is fundamental, while the output power of the basic power-generating, solid-state unit is rather small. The solution is to combine large numbers of solid-state components to achieve required power levels, requiring careful design to minimize combining losses and ensure good transmitter efficiency. There are three basic configurations to combine lower power amplifiers:

(1) Corporate-combined structure (Fig. T21a), an approach suitable, for example, for a parabolic reflector antenna illuminated from a single feedhorn).

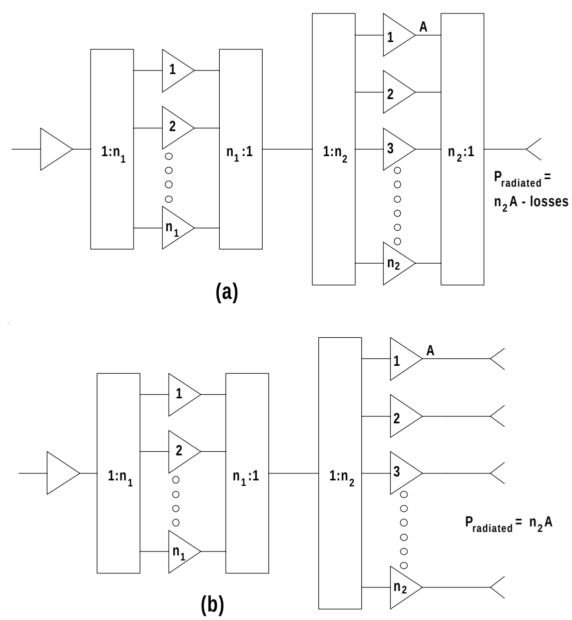


Figure T21 Block diagram of (a) a corporate-combined power amplifier; (b) a space-combined power amplifier (after Skolnik, 1990, Fig. 5.6, p. 5.12).

(2) Space-combined structure (Fig. T21b), an approach suitable for phased-array configuration in which each array element is fed by an amplifier module); and

(3) Hybrid approaches in which corporate-combined modules feed rows of a space-combined array. Development of individual transceiver module design based on solid-state technology for phased-array radars has proceeded rapidly in recent years.

The list of some fielded solid-state transmitters is given in Table T9. *SAL*

Ref.: Skolnik (1990), pp. 5.1–5.33; Skolnik (1980), pp. 216–220; Kaganov (1981); Ostroff (1985); Ewell (1981), p. 75; Leonov (1988), p. 51.

Table T9
Fielded Solid-State Transmitters

System	Contractor	Frequency (MHz)	Peak power (kW)	Duty cycle	Average power (kW)	No. of modules	Peak module power (W)	Year fielded
ROTHR	Raytheon	5–30	210	CW	210	84	3,000	1986
NAVSPASUR*	Raytheon	218	850	CW	850	2,666	320	1986
SPS-40*	Westinghouse	400–450	250	4%	10	112	2,500	1983
PAVE PAWS†	Raytheon	420–450	600	25%	150	1,792	340	1978
BMEWS*	Raytheon	420–450	850	30%	255	2,500	340	1986
TPS-59	GE	1,200–1,400	54	18%	9.7	1,080	50	1975
TPS-59‡	GE	1,200–1,400	54	18%	9.7	540	100	1982
SEEK IGLOO	GE	1,200–1,400	29	18%	5.2	292	100	1980
Martello	Marconi	1,250–1,350	132	3.75%	5	40	3,300	1985
RAMP	Raytheon	1,250–1,350	28	6.8%	1.9	14	2,000	1986
SOWRBALL	Westinghouse	1,250–1,350	30	4%	1.2	72	700	1987

*Solid-state replacement for tube transmitter; †Parameters per array face; ‡Upgraded with 100W modules (after Skolnik, 1990, Table 5.1, p. 5.3)

A **(vacuum-)tube transmitter** uses as the principal components microwave tube oscillators or amplifiers. The most common types are crossed-field tubes (see **MAGNETRON**; **CROSSED-FIELD AMPLIFIER**; **AMPLITRON**), linear-beam tubes (see **KLYSTRON**; **TRAVELING-WAVE TUBE**; **TWYSTRON**), and some other tubes used primarily at millimeter waves (see **GYROTRON**). Each of these components has assets and disadvantages (see corresponding entries), which have to be assessed when the choice is made for specific transmitter. The factors that most often dominate in tube selection are obtainable peak and average power, gain, bandwidth, operating voltage, spurious noise level, size, weight, cost, and availability. All these factors must be considered for the chosen frequency band of radar operation. In general, linear-beam tubes are less noisy and have higher gain compared with crossed-field tubes, but the latter are smaller, lighter, have lower operating voltages and are less costly. The main differences between the leading microwave tube types is summarized in the Table T10.

The main constituent parts of the tube transmitter are a self-excited oscillator (the exciter), amplifying tubes (if an amplifier chain configuration is used), the modulator, the power supply, cooling for the tube, safety interlocks, protection and monitoring devices, and x-ray shielding. A simplified block-diagram of a tube transmitter for phased-array radar is given in Fig. T22. The main advantages of tube transmitters are

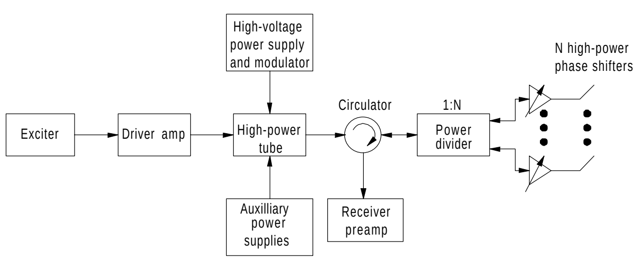


Figure T22 A simplified block-diagram of tube transmitter for phased-array radar (after Ostroff, et al, 1985, Fig. 8.3a, p. 208).

(1) Practically any desired requirement for radiated power can be satisfied with vacuum tubes.

(2) Tube technology has long traditions and is thoroughly developed.

(3) High peak powers, sufficient gain, good efficiency, practically all desired ranges of duty cycles and sophisticated pulse-compression waveforms can be obtained with this type of transmitter.

The main disadvantages that force radar designers to consider alternative solid-state transmitters are bulkiness, complexity, high operating voltages resulting in many unpleasant and dangerous effects, and insufficient lifetime, compared with solid-state components, which are inherent properties of vacuum tubes. *SAL*

Ref.: Skolnik (1990), pp. 4.5–4.31; Leonov (1988), pp. 44–50.

TRANSMITTER-RECEIVER (see **TRANSCEIVER**).

Table T10
High-Power Pulsed Amplifiers Compared for Same Frequency and Peak and Average Power Output

	Linear-beam tubes		Crossed-field tubes*	
	Klystron	TWT	Conventional	High-gain
Voltage	High voltage (1 MW requires approx. 90 kV)		Low voltage (1 MW requires approx. 40 kV)	
Gain	30–70 dB		8–15 dB	15–30 dB
Bandwidth	1–8%†	10–15%	10–15%	
X-rays	Severe, but lead is reliable		Not usually a problem	
Efficiency				
Basic	15–30%		35–45%	
With depressed collectors	40–60%		N.A.	
Ion pump	Required on large tubes		Self-pumping	
Weight	Higher		Lower	
Size	Larger		Smaller	
Cost	Medium	High	Medium	
Spurious noise‡	−90 dB		−55 dB	−70 dB
Spurious modes (typical)	None	π mode during rise and fall if cathode-pulsed; none if mod-anode- or grid-pulsed	π mode during rise and fall if cathode-pulsed; full-power noise output if turned on without RF drive	
Usable dynamic range	40–80 dB		A few decibels	
Control electrode	None, or mod-anode, or grid		None, or turnoff electrode	
Magnetic field	PPM up to 1 MW at S band; solenoid otherwise; barrel magnet rare (SLAC); none for ESFK		Permanent magnet	
Dynamic/static impedance	0.8		0.05–0.2	
Phase-modulation sensitivity	5–40° per 1% $\Delta E/E$		0.5–3.0° per 1% $\Delta I/I$	

*Distributed emission, reentrant, circular.
†Clustered-cavity klystron can achieve 10 to 15 percent bandwidth at higher cost.
‡In 1-MHz bandwidth

(from Skolnik, 1990, Table 4.1, p. 4.19, reprinted by permission of McGraw-Hill).

TRANSMIT-RECEIVE SWITCH (see **SWITCH**).

TRANSPONDER. A transponder is "a receiver-transmitter facility, the function of which is to transmit signals automatically when the proper interrogation is received from a radar. The main purpose of a transponder is to augment the power of the signal returned in the direction of the illuminating facility (interrogator). Typically, they are used in secondary radar applications, installed on aircraft, ships, missiles, and satellites to detect the interrogator transmission and respond with a coded reply stating vehicle identity, flight level (for aircraft), and so forth.

Transponders, consisting of an antenna, receiver, and transmitter, are usually low-power devices. They vary in complexity: some types employ various coded waveforms for relaying information on the status of the host platform, while others simply transmit an unmodulated beacon signal to facilitate tracking of the host platform by other (offsite) radars. Since the transmission path from a transponder is one way, the radar beacon equations apply. If the transponder is a repeater, the two-way equation for the repeater jammer applies.

Usually, transponders operate in the pulsed mode. They are classified as either active or passive. Passive transponders use radar reflectors focusing the illuminating signal with a parabolic antenna, sending it to the modulator for frequency shift, and retransmitting it via the same antenna. Active and passive transponders are used for ships, but for aircraft only active transponders are used. The block diagram of an aircraft transponder is cited in Fig. T23.

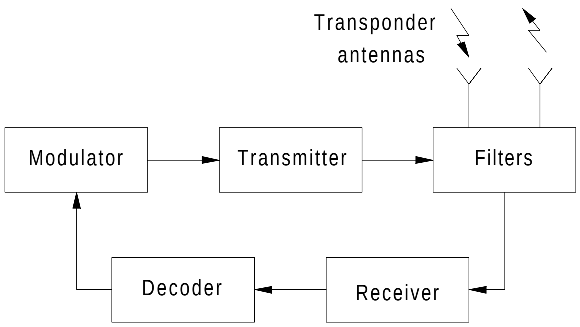


Figure T23 Block diagram of typical airborne transponder.

The interrogation comes from a secondary surveillance radar. The received signal from the transponder antenna passes through a filter to select the interrogating signal in the assigned band, and then to the receiver. The signals from the receiver go to the decoder that decodes received signals and codes replies, which pass to the transmitter modulator and then to the antenna for transmission. An example of aircraft transponder parameters is given in Table T11. A term used interchangeably with transponder is beacon. *PCH, SAL*

Ref.: IEEE (1992), p. 1,408; Stevens (1988), pp. 165–178; Stimson (1983), p. 610.

Table T11
Summary of Transponder Parameters

Parameter	Min.	Typical	Max.
Antenna gain (dB)	0	0	3
Sensitivity[a] (dBm)	−69	−71	−77
Power output[a] (dBW)	21	24	27
Reply delay (μs)	2.5	3.0	3.5
Transmitter recovery (μs)	25	50	125
Sidelobe suppression (μs)	25	35	45
Squitter replies (s^{-1})	0	0	30
DME suppression time (μs)	7	45	60
Reply rate threshold (s^{-1})	500	1,200	2,000

[a] Receiver sensitivity and transmitter power are measured at the antenna end of the RF cable. General aviation transponders can be operated with a reduced power output power specification if they are to be used in aircraft limited to flying below 15,000 ft. For this class of transponder, the output power is 18.5 dBW minimum.

(from Stevens, 1988, Table 8.1, p. 166).

TRAVELING-WAVE TUBE (TWT). A TWT is "an electron tube in which a stream of electrons interacts continuously or repeatedly with a guided electromagnetic wave moving substantially in synchronism with it, and in such a way that there is a net transfer of energy from the stream to the wave." The guided electromagnetic wave is created using a slow-wave spiral and propagates along its axis. In the process of this interaction, the electrons are slowed by the electrical field of the wave, giving up part of their energy, and thus providing amplification. We distinguish between TWTs with longitudinal magnetic fields (O-type TWTs) and ones with transverse magnetic fields (M-type TWTs). TWTs are used in input stages (low-noise TWTs) and in intermediate stages of microwave receivers, in exciters for wideband transmitters (see **oscillator TWT**), and in intermediate and final stages of wideband transmitters (see **AMPLIFIER, TWT**). In addition to RF amplification, other functions might also arise. (See **converter TWT**, **detector TWT**; **limiter TWT**, **phase-shifting TWT**, and **multifunctional TWT**.) *IAM*

Ref.: IEEE (1993), p. 1412; Popov (1980), p. 206; Andrushko (1981), p. 59; Sivan (1992), Ch. 4; Gilmour (1994).

A **converter TWT** is an O-type TWT whose output signal spectrum differs from that of the input signal. Multiplying and mixing TWTs are used. In the first, the frequency of the output signal is an integral multiple of the input frequency. The multiplier is a powerful TWT which operates in a nonlinear mode. The tube in which several frequencies are mixed is called a mixing tube, and it is used as a mixer. *IAM*

Ref.: Popov (1980), pp. 315, 401.

A **coupled-cavity TWT** is one using the slow-wave circuit with individual cells resembling the ordinary klystron resonant cavities (Fig. T24). Such a configuration offers increased power, and approximately 90% of all high-power TWTs employ the coupled-cavity approach for its good electrical and structural characteristics. So-called rabbit ears (e.g., oscillations appearing during the turn-on and turn-off portions of the pulse), is one of the spurious effects arising in such tubes and requiring special measures of suppression to be undertaken. *SAL*

Ref.: Ewell (1981), p. 64; Skolnik (1980), p. 207; Gilmour (1994), Ch. 13.

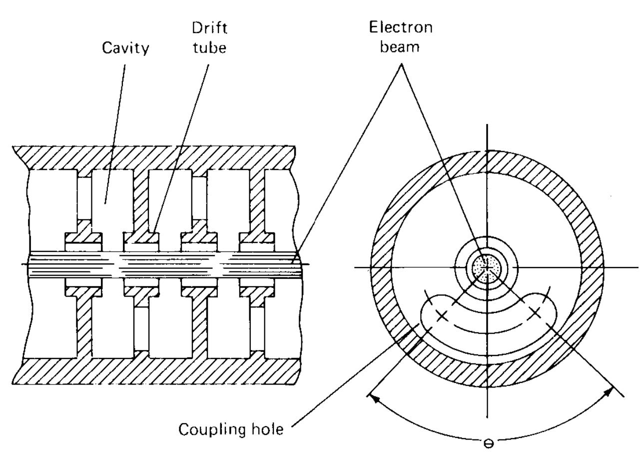


Figure T24 Coupled-cavity approach for slow-wave circuit in a TWT (from Ewell, 1981, Fig. 2.35c, p. 65).

A **depressed-collector TWT** uses depressed collectors (operating at voltages below body potential) to increase efficiency. Usually, multiple collector sections (for radar applications typical figure is three, see Fig. T25) are used that allows to catch each spent electron at the voltages near optimum. The efficiency of the tube can be increased up to 20%, but at the expense of equipment complexity. *SAL*

Ref.: Ewell (1981), p. 67; Skolnik (1990), p. 4.17; Gilmour (1994), Ch. 13.

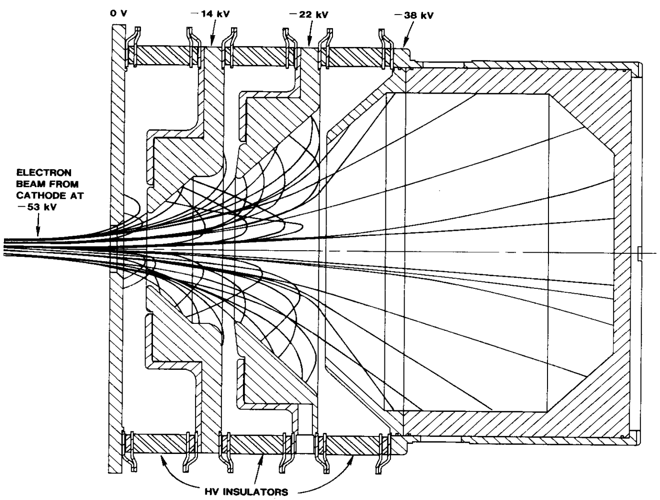


Figure T25 Multiple depressed collectors in TWT (from Skolnik, 1990, Fig. 4.11, p. 4.18, reprinted by permission of McGraw-Hill).

A **detector TWT** provides at its output the envelope of the modulated microwave signal appearing at its input. *IAM*

Ref.: Popov (1980), p. 105.

A **gyro-TWT** is a kind of gyrotron based on the interaction of the beam and RF field in a nonresonant structure, such as circular waveguide. Such devices have high output power and wide bandwidth and may have important applications in MMW radars. At 95-GHz peak output power of 20 kW, gain of 30 dB, efficiency of 8%, and bandwidth of 2% have been achieved. *SAL*

Ref.: Currie (1987), p. 470.

A **helix TWT** uses a helix or helix-derived structure as the slow-wave circuit. The simple helix slow-wave structure provides very broad bandwidth, more than an octave. However, it has restrictions on operation at high power levels as the resulting electron velocity is too fast to synchronize with the low velocity of the RF wave on the helix. To achieve higher power, modifications of helices known as ring-bar and contra-wound circuits (Fig. T26) are used. Often, when the tube operates at high average powers, the cooling liquid is passed through a helix constructed of copper tubing. *SAL*

Ref.: Ewell (1981), p. 64; Skolnik (1980), p. 206.

An **isophase TWT** maintains an optimum phase between the first harmonic of the grouped circuit and the traveling wave through a change in the phase speed of the electromagnetic wave. Here in the time of movement of the bunch of electrons along the entire line of the slow-wave system, it does not

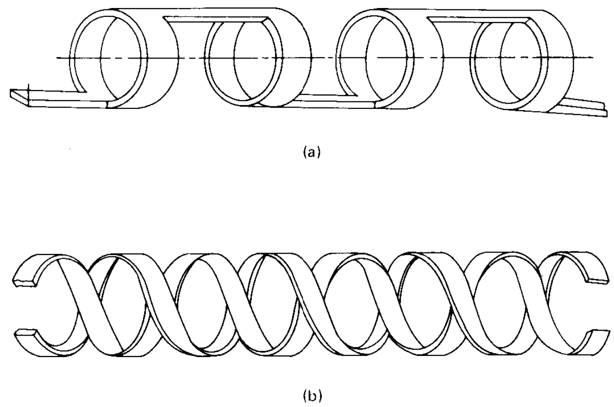


Figure T26 Helix-derived slow-wave circuits: (a) ring-bar circuit; (b) contra-wound circuit (from Ewell, 1981, Fig. 2.35, p. 65, reprinted by permission of McGraw-Hill).

leave the region of the slow-wave field. The isophase TWT belongs to the O-type devices. *IAM*

Ref.: Popov (1980), p. 143.

A **limiter TWT** is an amplifying TWT whose output power remains practically unchanged with a change in the level of the input power. It is an O-type TWT. *IAM*

Ref.: Popov (1980), p. 257.

A **low-noise TWT** is one whose noise factor is between 7 and 20 dB. The basic source of noises are the shot effect and losses to the resistance of the slow-wave circuit. To reduce the influence of shot noise, potential selection in the region of the electron gun is used to align the minimum density fluctuation of the electron flux with the start of the interaction space. The reduction in losses in the slow-wave circuit is achieved by selection of material and by precision in manufacture. Since these losses depend on frequency, the noise factor increases in the millimeter waveband. In this case, a spiral with a short step in its diameter is used, secured to a sapphire wedge to give it mechanical strength. This TWT is an O-type device. The low noise factor combined with the wideband properties of the TWT makes it widely used in input circuits of radar receivers. *IAM*

Ref.: Popov (1980), p. 219.

An **M-type TWT** is one with electron flux which moves in crossed electrical and magnetic fields. (See **CROSSED-FIELD AMPLIFER.**) The M-type traveling-wave tube is used as a high-power amplifier. (See **AMPLIFIER, magnetron.**) *IAM*

Ref.: Andrushko (1981), p. 75.

mixer TWT (see **converter TWT**).

A **multibeam TWT** is one in which the electron beam is the bunch of several electron beams interacting with an electromagnetic wave. This tube is an O-type TWT. *AIM*

Ref.: Popov (1980), p. 231.

A **multifunctional TWT** performs various functions: that of amplifier, mixer, amplifier, detector, oscillator, and level limiter. It is an O-type TWT. Dual-mode TWTs are used in radio countermeasure systems, where they may provide a variable-pulse or continuous operating mode, depending on the mode of the anode voltage. The use of such TWTs significantly reduces the weight and size of airborne equipment. *IAM*

Ref.: Popov (1980), p. 233; Andrushko (1981), p. 60.

multiplier TWT (see **converter TWT**).

An **O-type TWT** is one with a rectilinear electron beam from longitudinal electrical and magnetic fields. (See **TUBE, linear-beam.**) An O-type TWT is a glass tube whose bottom contains an electron gun for firing a narrowly directed beam of electrons. The slow-wave system is a spiral or chain of connected cavities (Fig. T27).

An O-type traveling-wave tube is used as an amplifier of medium and high power in a band of 0.96 GHz. (See **AMPLIFIER, TWT.**) *IAM*

Ref.: Dulin (1977), p. 68.

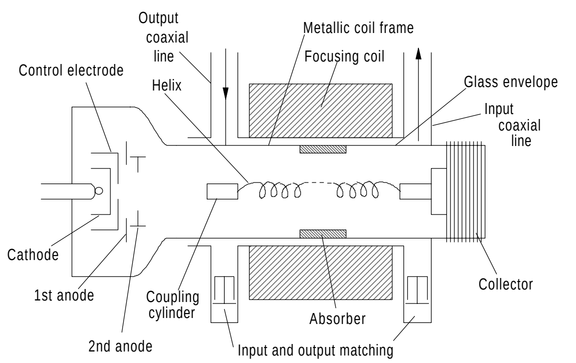


Figure T27 O-type TWT.

An **oscillator TWT** is one in which microwave oscillations are generated through a positive feedback circuit between the input and output. This circuit can be made either over a special channel or through internal reflections of the electromagnetic wave at discontinuities of the spiral. The tube is an O-type TWT. *IAM*

Ref.: Popov (1980), p. 82.

A **phase-shifting TWT** is one providing adjustable phase shift between the output and input signals by changing the mode or delivering modulated supply voltages to special phase-shifting elements that are built in. It is an O-type TWT. *IAM*

Ref.: Popov (1980), p. 453.

A **plasma TWT** is one whose operation is based on the use of prolonged interaction of the electron flux with a slow electromagnetic wave propagating along a plasma waveguide. The microwave energy is inserted directly into the plasma, whose use as a slow-wave medium makes possible broad electronic tuning in the operating band by changing the parameters of the plasma waveguide, to increase the amplification per unit of length (i.e. to reduce the length of the interactive space), and to increase the efficiency *IAM*

Ref.: Popov (1980), pp. 283, 453.

TRIGATRON. A trigatron (obsolete) "is an electronic switch in which conduction is initiated by the break-down in an auxiliary gap." *SAL*

Ref.: IEEE (1993), p. 1,414.

TRIGGER [FLIP-FLOP] circuit. A trigger circuit is "a circuit that has two conditions of stability, with means for passing from one to the other when certain conditions are satisfied, either spontaneously or through application of an external stimulus." Flip-flops are divided into symmetrical and asymmetrical. Symmetrical flip-flops are used as pulse repetition frequency dividers, for controlling difference devices, and as storage devices, counters, and switching devices. One possible circuit of a symmetrical flip-flop is given in Fig. T28(a), and its voltage time diagrams are given in Fig. T28(b).

The inputs of a flip-flop are video pulses, which convert the circuit from one stable state to another, which is maintained until arrival of the next pulse. Asymmetrical flip-flops are used chiefly to form a square wave from a sinusoid and as amplitude discriminators.

The circuits of an asymmetrical flip-flop can vary (tube, transistor, gas-discharge devices, with tunnel diodes, etc.). The asymmetrical flip-flop with a tunnel diode has a number of advantages: simplicity, high speed, and high stability with changing temperature. A disadvantage is the low output voltage that is eliminated by using transistors at the output.

In tunnel diode flip-flops, Fig. T28(c), one uses the descending section of a diode characteristic. To obtain two stable states the load line must intersect the diode characteristic at three points: in Fig. T28(d), points 1 and 3 are stable, point 2 is unstable. The circuit is triggered by pulses of alternating polarity. For the negative pulse a transition is made from point 3 to point 1, for positive pulse from point 1 to point 3. At the output one gets a square pulse, the duration of which depends on the repetition period of the triggering pulses, as in Fig. T28(e). *AIL*

Ref.: IEEE (1993), p. 1,414; Druzhinin (1967), pp. 238–244.

TRIODE, microwave. A microwave triode is a microwave electronic device with three elements. Triodes are divided into semiconductor triodes (or transistors) or vacuum-tube triodes: three-electrode tubes with electrostatic control of electron beam. Typically, vacuum-tube triodes are meant by the term *microwave triodes*. To reduce the influence of interelectrode capacities and pin inductances and decrease electron beam transit time, the microwave triode has the configuration of a lighthouse tower, with electrodes having the shape of flat disks with small distances between them and with circular pins (Fig. T29). They are used as oscillators and power amplifiers at lower frequencies, up to the UHF band. *AIL*

Ref.: Druzhinin (1967), p. 281; Betin (1972), p. 186.

TROPOSPHERE. The troposphere is the lower part of the atmosphere extending from the surface of the earth up to the heights of 8 to 10 km for polar latitudes, 10 to 12 km for medium latitudes, and up to 16 to 18 km in tropical regions.

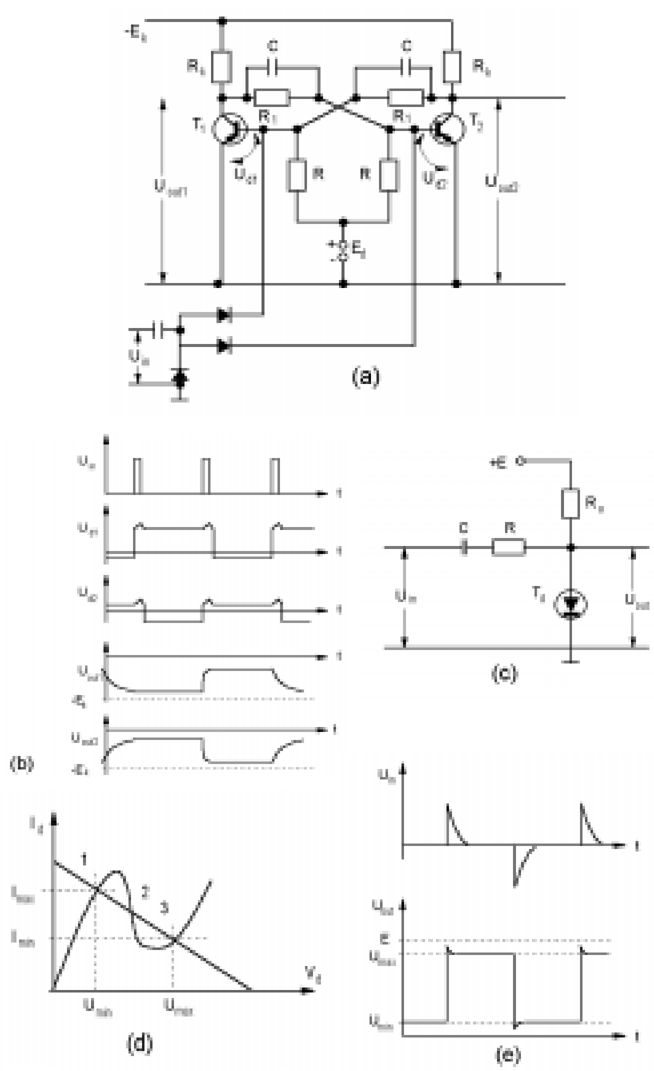


Figure T28 Flip-flops: (a) symmetrical, (b) voltages of symmetrical flip-flop, (c) tunnel diode type, (d) characteristic of a tunnel diode, (e) voltages of tunnel-diode flip-flop (after Druzhinin, 1967, Fig. 6.26, page 240 and Fig. 6.30, page 243).

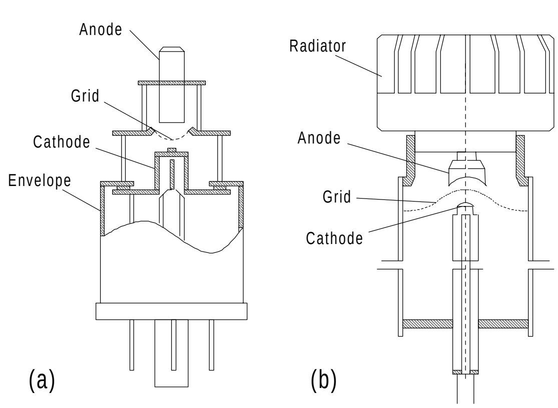


Figure T29 Triodes with disk outputs: (a) lighthouse tube; (b) metaloceramic.

The influence of the troposphere on wave propagation is mainly in tropospheric absorption (see **ATTENUATION**) and refraction. (See **ERROR, propagation**; **ATMOSPHERE, refraction.**) *AIL*

Ref.: Dolukhanov (1972), p. 120-136; Blake (1980), p. 205.

TUBE, microwave. A microwave tube is one in which the effects of interaction of the RF field and electrons are used to originate the process of oscillation or amplification of the microwave energy. From the point of view of the principles of electric and magnetic field structure used for their operation, all microwave tubes are divided into linear-beam tubes (O-type) and crossed-field tubes (M-type). The main representatives of the first class are klystrons, traveling-wave tubes, and twystrons; the primary types of devices falling into the second class are magnetrons and crossed-field amplifiers. *SAL*

Ref.: Gilmour (1986).

backward-wave tube (see **BACKWARD-WAVE TUBE**).

A **crossed-field tube** is a microwave tube in which the dc magnetic and electric fields accelerating the electrons are orthogonal to each other. The primary types of crossed-field tubes used in radar applications are crossed-field oscillators (magnetrons) and crossed-field amplifiers (magnetron amplifiers). These tubes are divided into two major groups as distributed-emission (emitting-sole) and injected-beam tubes. Subdivisions are based on whether linear or circular format; reentrant or nonreentrant electron stream; and forward-traveling wave, backward-traveling-wave, or a standing wave are used. (See **CROSSED-FIELD AMPLIFIER**) The family of crossed-field tubes is given in Fig. T30. Typically, these devices are characterized by high peak power, high efficiency, and relatively low (compared with linear-beam tubes) weight, size and operating voltages. Another term for this type of devices is *M-type tubes*. *SAL*

Ref.: Ewell (1981), p. 21; Leonov (1988), p. 44.

cathode-ray tube (see **CATHODE RAY TUBE**).

Tube efficiency is the ratio of the RF output available from the microwave tube to the dc power input in the electron stream. Sometimes this ratio is called *conversion efficiency*. *SAL*

Ref.: Skolnik (1980), p. 191.

A **grid-controlled tube** is a VHF or UHF microwave tube using a grid to control the electron beam. The process by which the electron beam is modulated to produce amplification is called *density modulation*. There main representatives of this class are microwave triodes and tetrodes, and they were the first RF components used in the transmitters of early radars during the 1930s, as there was no other alternative for large RF power sources. These devices are capable of operation at frequencies as high as 1 to 2 GHz, but they are not competitive with linear-beam and crossed-field tubes at frequencies much above 500 MHz. They are characterized by high power, high efficiency, wide bandwidth, low or moderate gain and inherent long life. The range of application is in

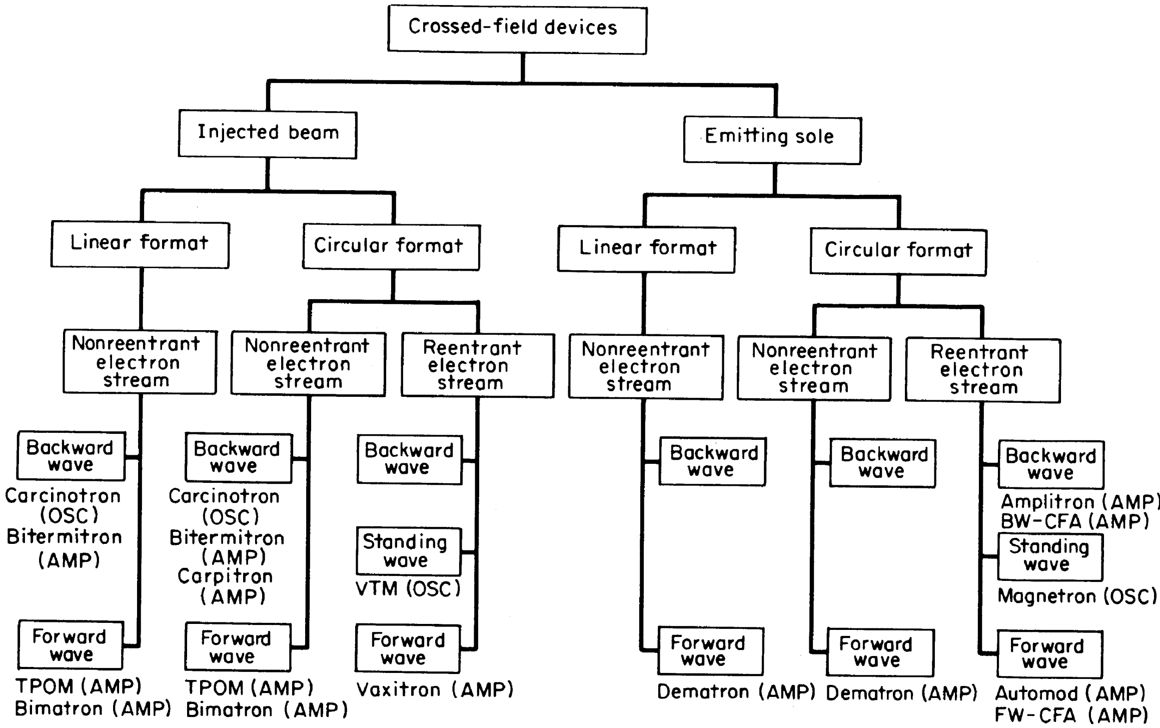


Figure T30 Family tree of crossed-field tubes (from Fink, 1982, Fig. 9.66, p. 9.54, reprinted by permission of McGraw-Hill).

high-power amplifier chains and transmitters in large radars operating in the VHF and lower UHF bands (e.g., early warning or satellite-surveillance radars). An example of grid-controlled tube suitable for high-power radar application at UHF is the coaxitron. *SAL*

Ref.: Skolnik (1980), p. 213.

A **linear-beam tube** is a microwave tube in which the electrons emitted from the cathode form a continuous beam traversing an interaction region and in which the dc electric field accelerating the electron beam aligned with the axis of the magnetic beam. In principle, the magnetic field cannot used at all, or used only for the purpose of beam focusing, and does not play the principal role in the process of the energy exchange. The primary types of linear-beam tubes used in radar applications are klystrons, traveling-wave tubes, and twystrons (a hybrid of the first two). They have high peak powers and high efficiency (the klystron can be called a champion of power and TWT a champion of bandwidth), but compared with crossed-field tubes, they are usually more complex and expensive and require higher operating voltages. Another term for this type of devices is *O-type tubes*. *SAL*

Ref.: Ewell (1981), p. 21; Leonov (1988), p. 44.

A **millimeter-wave tube** is usable in the millimeter-wave (MMW) portion of radar band (frequencies more than 30 GHz). They are classified as fast-wave or slow-wave devices, the latter including linear-beam and crossed-field tubes. The nomenclature of the slow-wave tubes is the same as used at lower frequencies: magnetrons, crossed-field amplifiers, klystrons, and traveling-wave and backward-wave tubes. The main difference is in design and fabrication of MMW tubes, as frequency scaling of a device imposes constraints on fabrication and power because of the smaller dimensions and associated lower voltage breakdown. Fast-wave devices such as gyrotrons (cyclotron resonance maser), gyroklystrons, gyro-TWTs, laddertrons, peniotrons represent new technology advancing the performance of MMW tubes. Examples of MMW klystrons and traveling-wave tube performance are given in the Tables T12 and T13. *SAL*

Ref.: Currie (1987), pp. 448–473.

Table T12
Typical EIA Klystron Characteristics

Frequency (GHz)	Power (W)	Duty (%)	Gain (dB)	Bandwidth (MHz)	Model
50–80	100	10	-	-	VKE2406
94–96	1,000	10	30	200	VKB2400T
95	30	200	30	150	-
220	60	1.7	-	-	-

(from Currie, 1987, Table 9.2, p. 456).

M-type tube (see **crossed-field tube**).

O-type tube (see **linear-beam tube**).

A **relativistic-electron-beam tube** is an SHF microwave tube using the effect of relativistic-beam traveling at the speed close to the speed of light to generate energy. An example is the tube operating at the submillimeter wavelength of 0.4 mm and generating pulsed waveforms with 1-MW power. *SAL*

Ref.: Brookner (1988), p. 334.

Table T13
Recent MMW TWT Developments

Vendor	Model	Freq. (GHz)	Pulsed or CW	Power (kW)	Power (W)	Gain (dB)	Struc-ture	Focusing	Weight (lbs)
EEV	N10043	18-40	CW	-	20	40	Helix	PPM	2.4
Stantel	W09MW1K	26.5-40	CW	-	10	30	Helix	PPM	5.5
Hughes	8900H	32-35	P	0.08	16	43	Helix	PPM	5.5
Hughes	921H	33.5-36.5	P	4	220	46	Cavity	PPM	17.5
Raytheon	QKW1995	Ka-band	CW	-	1000	43	Cavity	PPM	6
Varian	VTA-5700	34.5-35.5	P	30	3000	53	Cavity	Solenoid	350
Hughes	982H	93.75-95.75	P	0.1	50	60	Cavity	PPM	14
Varian	VTW-5795	94.5-95.5	P	5	500	46	Ladder	Solenoid	220
(from Currie, 1987, Table 9.4, p. 465).									

traveling-wave tube (see **TRAVELING-WAVE TUBE**).

TWYSTRON. A twystron is a linear-beam tube that is part traveling-wave tube and part klystron. It consists of a multi-cavity klystron input section coupled to an extended interaction traveling-wave output section. It was developed in 1963 by Varian, and its initial purpose was to produce a more efficient version of the broadband TWT, based on the more effective beam-bunching action of the cavities. The flexibility in tuning of the cavities combined with the broad power-bandwidth capability of the TWT output section not only improved the efficiency but obtained a significant improvement in bandwidth. The twystron appears capable of equally high power with broader bandwidth than klystrons (perhaps the clustered-cavity klystron is an exception), although at the expense of the cost and complexity. Typical Twystron amplifier characteristics are given in the Table T14. *SAL*

Ref.: Skolnik (1980), p. 208; Skolnik (1990), p. 4.17; Gilmour (1994), pp. 386–390.

Table T14
Typical Twystron Tube Characteristics

Tube type	VA-45	VA-146	VA-915
Frequency band (GHz)	2.7–2.9, 2.9–3.1, 3.0–3.2	5.4–5.9	3.4–3.6
Peak power (MW)	3.5	4.0	7.0
Average power (kW)	7.0	10.0	28.0
Pulsewidth (μs)	10.0	20.0	40.0
Efficiency (%)	35.0	30.0	30.0

U

UNCERTAINTY

The **uncertainty ellipse** is the locus of target position uncertainty at any time t due to radar measurement error. It can be represented by an ellipsoid of radar error, with semi-axes $R\sigma_e$, $R\sigma_a \cos E_t$, and σ_r, where R is the measured range to the target, σ_E and σ_A are the rms errors in target elevation angle and azimuth angle, respectively. If the range error $\sigma_r = 0$, the locus of target position errors is described by a target uncertainty, or error ellipse, as shown in Fig. U1.

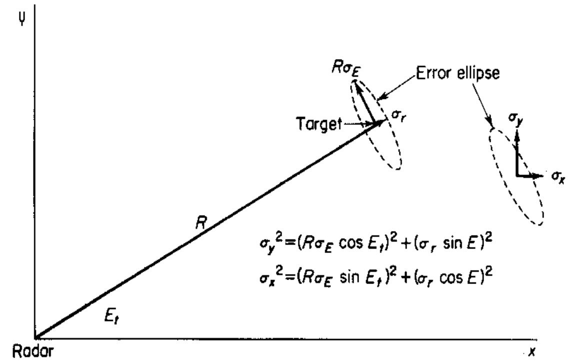


Figure U1 Target uncertainty ellipse (from Barton (1969), p. 241).

The **uncertainty principle** is the mathematical relationship between effective signal bandwidth β and effective time duration α:

$$\beta\alpha \geq \pi$$

Use of the term “uncertainty” is a misnomer, since there is nothing uncertain about the relationship expressed in the inequality shown. The relationship reflects the mathematical facts that (1) a narrowband waveform yields a wide frequency spectrum, (2) a wideband waveform yields a narrow frequency spectrum, and (3) signal bandwidth and frequency spectrum cannot simultaneously be made arbitrarily small. The utility of the relationship shown above lies in its predictive power in determining the accuracy with which time delay and frequency may be simultaneously measured. This can be summarized by the inequality:

$$\delta t_R \delta f \leq \frac{1}{\pi\left(2\frac{E}{N_0}\right)}$$

where δt_R and δf represent the rms error in time delay and frequency, respectively. This expression states that the time delay and frequency may be measured simultaneously with an accuracy determined only by the magnitude of the signal energy-to-noise power spectral density ratio. Substituting the first expression ($\beta\alpha > \pi$) into the second then states that for a fixed E/N_0, a large $\beta\alpha$-product achieves the same result.

It should be noted that the so-called uncertainty relation in radar bears no relationship to the uncertainty principle of quantum physics.

PCH

Ref.: Skolnik (1990), pp. 408–409.

V

VARACTOR. A varactor is a microwave diode with electrically controlled capacitance. This capacitance is on the order of tenths of a picofarad. Parametric, multiplying and tuning diodes are classed as varactors. A varactor uses a PN junction or Schottky barrier. (See **DIODE, PN** and **Schottky-barrier**.)

The basic parameter of the varactor is the maximum frequency, which determines its inertial properties, and Q-quality, which characterizes losses in the varactor. Other parameters include acceptable dispersed power, normalized reverse voltage (for a Schottky-barrier diode), or breakdown voltage (for a PN diode).

Varactors are produced in cermet housings or without housings. *IAM*

Ref.: Gassanov (1988), p. 80.

VARICAP. A varicap is a diode with electrically controlled capacitance in the range of units to tens of picofarads. It is used basically for electronic tuning of circuits in the HF and VHF bands. Varicaps in the microwave band are *varactor*s. *IAM*

Ref.: Gassanov (1988), p. 80.

VAXITRON. The vaxitron is an injected-beam, forward-wave, reentrant-stream, crossed-field amplifier tube using an interaction geometry similar to the voltage-tunable magnetron. The low-Q resonant structure is replaced in this tube by a broadband forward-wave circuit, and a multielement depressed collector may be used. Gain of 30 dB and efficiency of 50% have been reported. *SAL*

Ref.: Fink (1982), p. 9.60.

VEGETATION FACTOR. The vegetation factor (coefficient) is a factor in the surface reflection coefficient accounting for the absorption of incident electromagnetic waves by the vegetative layer on a land surface. This layer absorbs much of the incident wave, leading to a multiplicative vegetation factor approximated by

$$\rho_v = \exp\left(-\frac{k}{\lambda}\sin\psi\right)$$

where λ is the wavelength, ψ is the grazing angle, and the empirical constant $k = 1$ for thin grass, $k = 3$ for dense weeds or brush, and $k = 10$ for dense trees. *SAL*

Ref.: Barton (1988), p. 295, (1993), p. 148.

VELOCITY

The **critical velocity** for a phase-coded pulse compression system is that for which the doppler shift causes the phase error over the pulse to reach 180°, mismatching the waveform to the filter. For example, at X-band the critical velocity for a 13-element Barker code having 1-ms subpulses is 625 m/s. *SAL*

Ref.: Schleher (1991), p. 433.

doppler velocity (see radial velocity).

velocity gate pull-off (see **ECM, velocity measurement**).

Group velocity is the velocity with which the modulation envelope of the electromagnetic wave travels. It is the velocity at which energy is transported and is equal to or less than the velocity of light. *SAL*

Fink (1982), p. 18.94; Blake (1980), p. 193.

The **velocity of light** is $c = 2.99792458 \times 10^8$ m/s (see **WAVE, ELECTROMAGNETIC**).

Ref.: Jordan (1985), p. 3.14.

Phase velocity is the velocity of oscillations of an electromagnetic wave (a surface of constant phase) in a given medium. When the phase velocity in a medium is a function of frequency, that medium is called *dispersive*. The phase velocity is equal to or greater than the velocity of light. That fact does not violate the special theory of relativity because the transportation of energy is given by the group velocity. *SAL*

Ref.: Fink (1982), p. 18.94; Blake (1980), p. 193.

Radial velocity is the target velocity along the radar-target line that produces a doppler frequency shift in the echo signal. It is also called *doppler velocity*. *SAL*

velocity response (see **MTI improvement factor**).

VELODYNE. A velodyne is a motor with electrical adjustment of the rotation rate which has a toothed disk on its axle, connected to a magnetic core of a coil. When the disk turns, the action of the control voltage causes a change in the gap in the magnetic conductor of the coil, with the result that a variable voltage is induced with a frequency equal to the product of the number of teeth times the rotation rate of the disk.

Velodynes were used in a circuit for frequency tracking, which was used for range or velocity tracking in older radars with FMCW waveforms. The circuit used the property of the velodyne to generate an analog of the measured frequency in the form of an ac voltage whose frequency was proportional to the measured frequency. Voltage from the coil of the velodyne was applied to the input of a mixer, whose other input received the signal of a tracking local oscillator. The output signal of the mixer was used to generate the control voltage of the velodyne, this tuning the frequency of the toothed wheel of the velodyne to the measured frequency. *IAM*

Ref.: Vinitskiy (1961), p. 293.

VIDEO. The term video “refers to the signal after envelope or phase detection, which in early radar was the displayed signal. Contains the relevant radar information after the removal of the carrier frequency. *SAL*

Ref.: IEEE (1993), p. 1459.

Canceled video is the video output remaining in an MTI radar after cancellation is performed. *SAL*

Ref.: IEEE (1993), p. 153.

Coherent video is "bipolar video obtained from a synchronous (coherent) detector." *SAL*

Ref.: IEEE (1993), p. 206.

In-phase video is "one of a pair of coherent, bipolar video signals derived from the RF or IF signal by a pair of synchronous detectors with a 90° phase difference between the coherent oscillator (coho) reference inputs used for each. The other coherent video signal of the pair is designated quadrature video." *SAL*

Ref.: IEEE (1990), p. 18.

quadrature video (see **in-phase video**).

Raw video is the radar signal at the video output of the receiver or analog processor. This term is used with respect to displayed information, to distinguish signals sent in analog form (and preserving echo characteristics such as fluctuation and pulse widening) from those processed digitally. The latter, resulting from automatic detection and similar processes, generally appear as pulses of standard amplitude and width and are called *synthetic video.*

video integration (see **INTEGRATION, noncoherent**).

Video stretching is "the increasing of the duration of a video pulse." The procedure is used to improve visibility on a display, when the actual pulsewidth is a very small fraction of the sweep length. *SAL*

Ref.: IEEE (1993), p. 1459.

VISIBILITY

The **clutter visibility factor** in pulsed radar is "the predetection signal-to-clutter ratio that provides stated probabilities of detection and false alarm on a display; in moving-target-indicator systems, it is the ratio after cancellation or doppler filtering." In CW radar it is the corresponding single-look ratio, using a filter matched to the time on target. In automatic detection systems, a *clutter detectability factor* is similarly defined for the output of the automatic process. *DKB*

Ref.: IEEE (1993), p. 199.

The **visibility factor** is (1) in pulsed radar, "the ratio of single-pulse signal energy to noise power per unit bandwidth that provides stated probabilities of detection and false alarm on a display, measured in the intermediate-frequency portion of the receiver under conditions of optimum bandwidth and viewing environment"; (2) in continuous-wave radar, "the ratio of single-look signal energy to noise power per unit bandwidth using a filter matched to the time on target." This term is equivalent to the *detectability factor* for radar using automatic detection and the *clutter visibility factor* when the main source of interfering power is clutter. *SAL*

Ref.: IEEE (1993), p. 1,462; Blake (1980), p. 18.

Interclutter visibility is "the ability of a radar to detect moving targets that occur in resolution cells between points of strong clutter." (See **CLUTTER**.) *SAL*

Ref.: IEEE (1993), p. 665; Barton (1988), p. 247.

Subclutter visibility is "the ratio by which the target echo power may be weaker than the coincident clutter echo power and still be detected with specified detection and false-alarm probabilities. Target and clutter powers are measured on a single pulse return and all target radial velocities are assumed equally likely." (See **CLUTTER.**) Expressed in decibels, $SCV_{dB} = I_{db} - V_{xcdB}$, where SCV is subclutter visibility, I is the improvement factor, and V_{xc} is the clutter visibility factor. *SAL*

Ref. IEEE (1993), p. 1303; Schleher, (1991), p. 38.

VOLTAGE STANDING-WAVE RATIO (see **WAVE, standing**).

VULNERABILITY in electronic warfare is the susceptibility of a radar to electronic countermeasures (ECM). The degree to which the normal operation of a radar (or a radar homing seeker) is disrupted by ECM is a measure of its vulnerability. Quantitative determination of the vulnerability of a specific radar or seeker is a complex undertaking. Vulnerability to Gaussian (white) noise jamming and chaff often can be determined through analysis, but for most other forms of jamming, especially combinations of jamming techniques, the response of the victim radar and its electronic counter-countermeasures (ECCM) logic can be determined only through hardware-in-the-loop simulation and field tests. Such testing is necessary due to the influence of the total environment (jammers, natural clutter sources, reflections, etc.), as well as the nonlinear behavior of both jammer(s) and victim radar under some operational conditions.

For a search radar, vulnerability to noise jamming can be expressed in terms of the reduction in detection range caused by the jamming and the percentage of the radar coverage volume where target detection is denied by the jammer(s). In the case of sidelobe repeater jamming, a measure of vulnerability might include the number of false targets detected and false target track files established per scan. The major objectives of ECM used against tracking radars are to prevent target acquisition, break-lock if target acquisition has occurred, and provide false target position and speed data if target track cannot be prevented.

Over the years, specific ECCM techniques or "fixes" have been developed to counter specific forms of ECM, and many of these (e.g., the Dicke fix and sidelobe cancellers) remain effective techniques against unsophisticated jammers. A more robust way to reduce the vulnerability of a radar to ECM, however, is through the fundamental design approach adopted in the preliminary design stage.

Features of a radar design that provide inherent protection (but not total invulnerability) against ECM include high power-aperture product, ultra-low sidelobe antenna, adaptive antenna nulling, frequency agility, frequency diversity, wide instantaneous bandwidth, coherent coded waveforms, wide dynamic range in the receiver, and a spectral analyzer incorporated in the signal processor. *PCH*

W

WAVE, ELECTROMAGNETIC (EM). An EM wave is "a wave characterized by variations of electric and magnetic fields." The classical theory of electromagnetic phenomena was developed in the 19th century by James Clerk Maxwell (1831–1879), who built on the discoveries of Michael Faraday (1791–1867) and his predecessors, Coulomb, Ampere, and Gauss. Maxwell's theory may be summarized briefly as follows:

A stationary electric charge has associated with it an electric field **E**, which extends from the source to infinity, radially in all directions. If the charge moves, the **E**-field is altered in the vicinity of the charge, and the time-varying **E**-field induces a magnetic field **B**. If the motion of the charge is uniform, both fields are stationary, and there is no radiation. If, on the other hand, the motion of the charge is not uniform, that is, it accelerates or decelerates, the time-varying **B**-field (produced by the time-varying **E**-field) in turn generates an **E**-field, and the process continues, with **E** and **B** coupled in the form of a pulse. As one new field changes, it generates a new field that extends further, and the pulse moves out from one position in space to the next. The disturbance generated in the electromagnetic field takes the form of an electromagnetic wave that propagates beyond the original source and independent of it.

Unlike other forms of wave motion, water waves and sound, for example, electromagnetic waves can travel in a vacuum, that is, without a medium. This property of electromagnetic energy was hotly contested during the late 19th century, despite the negative results of several experiments. The most precise of these experiments, by the team of Albert A. Michelson (1852–1931) and Edward W. Morley (1838–1923), was designed to infer the presence of the "luminiferous aether" medium, that Maxwell and most other scientists had assumed, by comparing the speed of light along and perpendicular to the earth's motion around the sun. Albert Einstein's special theory of relativity, introduced in 1905, showed that all the experimental results could be explained within the context of his theory; there was no need of the aether. Light, Einstein held, and all electromagnetic radiation, travels in space at a constant speed of approximately 3×10^8 m/s, independent of the speed of the source.

With the advent of quantum mechanics, electromagnetic radiant energy is seen to be created, destroyed, and transported in discrete quanta or photons rather than through a continuous transfer of energy implied by electromagnetic waves in the classical representation of electrodynamics. Because the energy transported by large numbers of photons is, on the average, equivalent to the energy transferred in a classical electromagnetic wave, for macroscopic applications, including radar and communications, Maxwell's field equations are accurate and extremely useful tools.

In their original form, Maxwell's equations are amenable to solution only under a very restricted set of physical and geometric conditions. In word-form, these four equations are

(1) The total electric displacement passing through a closed surface is equal to the total charge within the volume enclosed.

(2) The net magnetic flux passing through any closed surface is zero.

(3) The magnetomotance (defined as the energy in watts per unit magnetic flux) taken around any closed path is equal to the sum of the conduction and convection current and to the time rate of change of electric displacement passing through the surface bounded by the closed path.

(4) The electromotance (voltage) taken around any closed path is equal to the time rate of change of the magnetic displacement passing through the surface bounded by the closed path.

For free-space conditions, Maxwell's equations can be expressed by two concise vector equations:

$$\nabla^2\mathbf{E} \approx \varepsilon_0\mu_0\frac{\partial^2\mathbf{E}}{\partial t^2} \quad \text{and} \quad \nabla^2\mathbf{B} \approx \varepsilon_0\mu_0\frac{\partial^2\mathbf{B}}{\partial t^2}$$

where ∇^2 is the Laplacian operator, ε_0 is the permittivity of free space, and μ_0 is the permeability of free space. These two vector equations represent a total of six scalar equations, in Cartesian coordinates, of the form:

$$\frac{\partial^2 E_x}{\partial x^2} + \frac{\partial^2 E_x}{\partial y^2} + \frac{\partial^2 E_x}{\partial z^2} = \varepsilon_0\mu_0\frac{\partial^2 E_x}{\partial t^2}$$

which take exactly the same form for E_y, E_z, and B_x, B_y, B_z. As Maxwell recognized, each component of the electromagnetic field (E_x, E_y, E_z, B_x, B_y, B_z) obeys the scalar differential wave equation

$$\frac{\partial^2 \psi}{\partial x^2} + \frac{\partial^2 \psi}{\partial y^2} + \frac{\partial^2 \psi}{\partial z^2} = \frac{1}{v^2}\frac{\partial^2 \psi}{\partial t^2}$$

provided that

$$v = \frac{1}{\sqrt{\varepsilon_0\mu_0}}$$

From experimentally determined values of μ_0 and ε_0, Maxwell predicted the speed of electromagnetic waves in free space to be approximately 3×10^8 m/s, in close agreement with the experimental measurement of Fizeau, who in 1849, had found a value of 3.153×10^8 m/s. In 1983, the 17th General Conference of Weights and Measures in Paris adopted a new definition of the meter, thereby fixing the speed of light in a vacuum as

$$c = 2.99792458 \times 10^8 \text{ m/s.}$$

Figure W1 represents a plane electromagnetic wave propagating in the x direction. The wavelength of the radiation is the distance traveled in one cycle, and is related to the frequency of the radiation, or the number of cycles per second (Hz) by the relationship $\lambda = c/f$, where c is the speed of light.

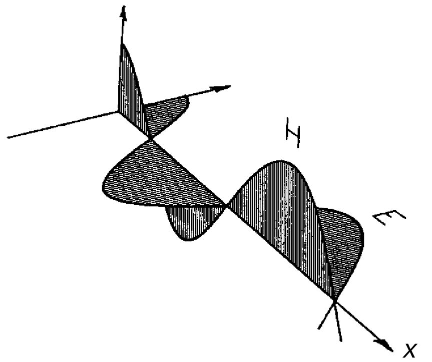


Figure W1 A plane electromagnetic wave.

The range of frequencies of electromagnetic radiation comprises the electromagnetic spectrum, represented here by Fig. W2. The upper end of the spectrum is shown here as approximately 10^{22} Hz, but in theory there is no limit. The period of the wave is the number of units of time per cycle, or $1/f$. The electromagnetic wave is a transverse wave, in that the variations of the electric and magnetic fields take place orthogonal to the direction of wave motion.

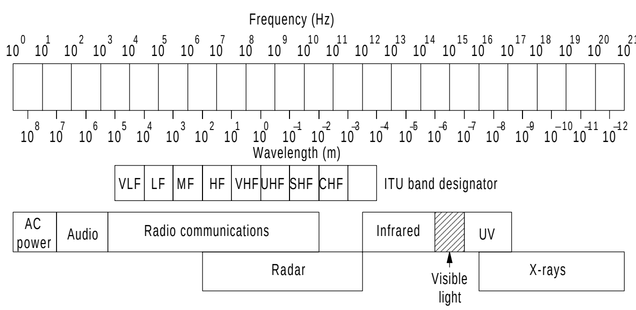


Figure W2 The electromagnetic spectrum.

The electric and magnetic fields are orthogonal to each other and are in phase at all points in space. The periodic wave may be described, as a function of time:

$$f(t) = A \sin(\omega t + \phi)$$

where A is the maximum amplitude of the electric field vector, $\omega = 2\pi f$, ϕ is the initial phase angle, and $(\omega t + \phi)$ is the total phase angle. The intensity of the wave, or the radiant flux, in watts, is proportional to the square of the amplitude. Electromagnetic waves obey the inverse-square law, in that the intensity, or power of the signal, is proportional to $1/R^2$. This is in accord with the law of the conservation of energy, since the total amount of energy flowing through a surface at range R_1 must equal the energy flowing through the surface at range R_2. The only way that this condition can be satisfied is for the radiant flux density (or power density) in W/m^2, to decrease with range. This principle is illustrated in Fig. W3.

By convention, the orientation of the electric field, shown here as vertical, determines the wave's polarization. Linear polarization (e.g., vertical, horizontal, or slant) is common in radar applications, and circular polarization, in which the **E** vector rotates 2π radians during one wave period, is sometimes used as well. Linear and circular polarizations are special cases of the more general elliptical polarization.

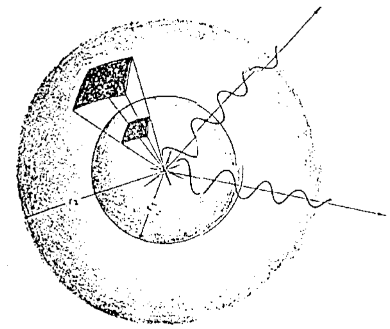


Figure W3: Geometry of the inverse-square law.

The propagating electromagnetic wave, in other than empty-space conditions, interacts with its environment. Effects due to the presence of the earth and its atmosphere, for example, include reflection, refraction (both diffuse and specular), diffraction, scattering, interference, and attenuation. To a first order, the magnitude of these effects are dependent on the frequency (wavelength) of the radiation, as well as the characteristics of the environment. (See **PROPAGATION.**) *PCH*

Ref.: Hecht (1990), Ch.3; Fink, pp. 1.38–1.45; Van Nostrand (1983), pp. 1,059–1,063, 2,991.

A **backward wave** is "a wave whose group velocity is opposite to the direction of electron-stream motion." The term is applied to traveling-wave tubes. *SAL*

Ref.: IEEE (1993), p. 86.

A **capillary wave** is a wave on the surface of a liquid for which surface tension is the dominant restoring force. Capillary waves are of interest to radar in that they make up the fine structure of the surface of large bodies of water (e.g., the sea) and as such contribute to the radar propagation effects and clutter backscatter characteristics. Capillary waves in sea water typically have wavelengths less than 0.02m. *PCH*

Ref. Skolnik (1990), p. 13.3.; Nathanson (1969), p. 231.

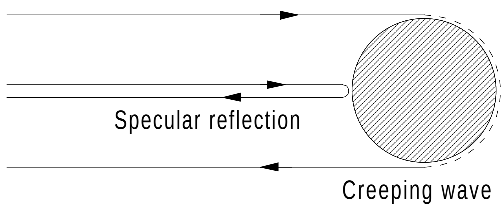
A **continuous wave** is a wave with no discontinuities or interruptions; that is, are everywhere defined as a function of a dependent variable. Continuous waves of interest to radar and communications are periodic (sinusoidal) in nature and can be described by the function

$$f(t) = A \sin \omega t$$

where the angular frequency $\omega = 2\pi ft$, A is the wave amplitude, f is the frequency, and t is time. Continuous waves form the basis of CW radar, which uses coherent, continuous wave transmission and reception to detect targets and measure their radial velocities using the doppler principle. (See **RADAR, CW.**) *PCH*

A **creeping wave** is an electromagnetic wave that forms in the shadowed region of a curved surface illuminated by radar energy. The creeping wave interacts with the specular reflection and both components determine the amount of energy reradiated back to the radar, i.e., the surface radar cross section (RCS).

If a curved surface, such as a sphere of radius r, is illuminated by electromagnetic radiation of wavelength λ, then the electrical circumference of the sphere, $kr = 2\pi r/\lambda$. The RCS of the sphere is a maximum for $kr = 1$, and its value oscillates about the geometric optics value (the area of a sphere πr^2), until $kr \approx 10$, at which point the oscillation is nearly damped out. The oscillation in the sphere's RCS with kr is due to the presence of a creeping wave of electromagnetic energy that circles the shadowed portion of the sphere and is launched back toward the source of illumination, i.e., the radar (Fig. W4).



FigureW4 Direction of travel of a creeping wave on a sphere.

The creeping wave loses energy in proportion to the distance traveled around the shadowed portion of the sphere; it thus becomes weaker as the sphere becomes larger, and the interference pattern decreases as the electrical size of the sphere increases. Similar creeping waves form in the shadowed regions of other curved surfaces, but their characterization and effect on the surface's RCS are more complex. (See **RADAR CROSS SECTION.**) *PCH*

Ref.: Knott (1985), pp. 161,162; Skolnik (1990), p. 11.5.

A **decameter [decametric] wave** is electromagnetic radiation having a wavelength between 10 to 100m. These wavelengths lie in what is termed the high frequency (HF) portion of the electromagnetic spectrum, 3 to 30 MHz in frequency (Fig. W2). (See **FREQUENCY, radar bands.**) *PCH*

A **direct wave** is propagated in a homogeneous or weakly heterogeneous medium, in space in particular, along approximately straight-line trajectories. All waves shorter than 10m are propagated as direct waves. *AIL*

Ref.: Dolukhanov (1972), p. 12.

An **electrical wave (E-wave)**, in electromagnetic radiation, is the propagation of the electrical field disturbance. The electrical wave is orthogonal to the magnetic wave. *PCH*

A **forward wave** is "a wave whose group velocity is in the same direction as the electron stream motion" in a traveling-wave tube. *SAL*

Ref.: IEEE (1993), p. 520.

A **gravity wave** is, in the context of radar, one formed on the surface of a body of water for which gravity is the dominant restoring force. Gravity waves are generally characterized as having wavelengths greater than about 2 cm. (See **CLUTTER, sea.**) *PCH*

Ref.: Skolnik (1990), p. 13.3.

A **ground wave** is one in which radio or radar energy propagates through interaction with the earth's surface, as opposed to free-space propagation or sky-wave propagation. At low frequencies, ground-wave propagation can travel thousands of miles, following the earth's curvature. Signal attenuation, which is a strong function of frequency, as well as the conductivity of the surface medium (soil or water), generally limits the usefulness of ground-wave propagation to frequencies less than 3 MHz, thereby making this mode of propagation mostly applicable to communications rather than radar.

At higher frequencies, those in the HF-band (3 to 30 MHz), radar propagation beyond the optical horizon can occur as a result of diffraction. Because signal attenuation increases exponentially with range at these frequencies, the range of over-the-horizon (OTH) radars using frequencies in the HF-band are generally limited to ranges of a few hundred kilometers. (See **PROPAGATION, DIFFRACTION, RADAR, OTH.**) *PCH*

Ref.: Van Nostrand (1983); Skolnik (1980), p. 536.

A **hybrid wave** is an EM wave for which presence of the longitudinal components of both magnetic and electrical fields is characteristic. They are designated as HE or EH waves. Hybrid waves as a rule exist in transmission lines with heterogeneous dielectric filling. The critical frequencies of hybrid waves depend complexly on the shape and dimensions of the transverse cross-section and on the parameters of the filled dielectric media. *AIL*

Ref.: Sazonov (1988), p. 13.

ionospheric wave (see **sky wave**).

A **magnetic wave (H-wave)** is a periodic variation of the magnetic field vector **H** in an electromagnetic wave not having a longitudinal component of electrical field, but having only the transverse component. A magnetic wave is sometimes referred to as a *transverse-electrical wave* (*TE-wave*). The magnetic wave is in phase with but orthogonal to the electric wave. *PCH, AIL*

Ref.: Sazonov (1988), p. 13.

A **meter [metric] wave** is one having a wavelength between 1 and 10m. These wavelengths lie in what is termed the very high frequency (VHF) portion of the electromagnetic spectrum, 30 to 300 MHz in frequency. (See **FREQUENCY, radar bands.**) *PCH*

A **millimeter (or millimetric) wave** is one having a wavelength between 1 mm and 1 cm. These wavelengths lie in what is termed the extremely high frequency (EHF) portion of the electromagnetic spectrum, 30 to 300 GHz in frequency (See **FREQUENCY, radar bands.**) *PCH*

A **monochromatic wave** is an electromagnetic wave with a single oscillation frequency. *AIL*

Ref.: Popov (1980), p. 238.

A **plane wave** is one for which the strengths of electrical and magnetic fields have identical phase and amplitude at all points lying in any plane perpendicular to the direction of its propagation at any moment (the wavefront is a plane). This wave sometimes is referred to as homogeneous. The term "homogeneous" wave reflects the fact that the amplitude of the wave does not depend upon coordinates; it is constant. In practice, the wave may be looked on as plane in homogeneous space at sufficiently great distances from the source. *AIL*

Ref.: Popov (1980), p. 291.

A **polarized wave** is a wave having different amplitudes of the vector of the magnetic and electrical fields in different directions perpendicular to the direction of propagation. (See **POLARIZATION.**)

Radio waves are EM waves occupying the radio frequency portion of the electromagnetic spectrum, nominally from a few kilohertz to about 1 THz. In addition to radio (FM, broadcast) and TV, radars and many other systems use this portion of the spectrum. (See **FREQUENCY, radar bands.**)

A **sky wave** is an EM wave in the HF portion of the electromagnetic spectrum that uses the refraction effects of the ionosphere to extend the range of transmission. Various segments of the HF spectrum are allocated by type of use (e.g., broadcast radio, maritime mobile, aeronautical mobile, amateur). Radars in the HF-band have been designed to provide OTH capability using ionospheric "bounce." This wave is also termed an ionospheric wave. (See **PROPAGATION**; **RADAR, OTH.**) *PCH*

A **spherical wave** is radiation from a point source, which travels from the source radially and uniformly in all directions. The source is said to be isotropic and the wavefronts created form concentric spheres that increase in size as they expand out into space. Most radiation occurs in the form of spherical waves, or approximately spherical waves. In radio and radar systems with directive gain antennas, the radiation wavefront can be represented as a segment of a sphere. As the wavefront expands, the radius of the spherical front increases so that at large distances from the source, the wavefront can be assumed to be essentially planar in nature.

The spherical wave of radius r can be described mathematically by the differential wave equation

$$\nabla^2\psi(r) = \frac{1}{r^2}\frac{\partial}{\partial r}\left(r^2\frac{\partial\psi}{\partial r}\right)$$

where ∇^2 is the Laplacian operator. *PCH*

Ref.: Hecht (1990).

A **standing wave** is the composite, nontraveling wave formed by two sinusoidal waves of the same frequency propagating along the same path in opposite direction. If E is the amplitude of the standing wave at any time t and distance x, and E_0 is the initial peak amplitude (assumed equal for the incident and reflected wave), then the wave may be described mathematically by:

$$E(x,t) = 2\,E_0 \sin kx \cos \omega t$$

At any point x the amplitude is constant. At certain points, $x = 0, \lambda/2, \lambda, 3\lambda/2, \ldots$, known as nodes, the amplitude will be zero at all times. Halfway between each adjacent node, at $x = \lambda/4, 3\lambda/4, 5\lambda/4, \ldots$, the amplitude has a maximum value of $2E_0$; these points are known as antinodes.

The ratio of the amplitude of a standing wave at an antinode to that at a node is called the standing-wave ratio, and the reciprocal is the traveling-wave ratio. The voltage standing-wave ratio (VSWR) is normally used. The VSWR is defined as the ratio of the maximum rms voltage at the peak of the standing wave in the transmission line E_{max} to the minimum rms voltage at the trough of the standing wave E_{min}. The typical notation is S or VSWR.

$$S = \frac{|E_{\max}|}{|E_{\min}|} = \frac{1+\rho}{1-\rho}$$

where ρ is transmission-line reflection coefficient. Since $0 \le \text{VSWR} \le \infty$, larger VSWR implies larger loss. *PCH, SAL*

Ref.: Hecht (1990); Jordan (1985), p. 29.8; Sazanov (1988), p. 36.

A **submillimeter [submillimetric] wave** is one having a wavelength between 0.1 mm and 1 mm. These wavelengths lie in what is termed the portion of the electromagnetic spectrum from 300 to 3000 GHz in frequency. (See **FREQUENCY, radar bands.**) *PCH*

A **surface wave** is a traveling wave of electrical current formed along the surface of a conducting body through the interaction of the surface with the electrical field of an incident wave of electromagnetic energy. For such a wave to form, there must be a component of the incident electric field tangential to the surface in the direction of propagation. This component induces surface currents that travel along the body in the direction of the incident energy (i.e., a traveling wave, that propagates at nearly the speed of light). Surface waves are induced on long surfaces (in comparison with the wavelength of radiation). The familiar end-fire radiation characteristic of long-wire antennas are due to this mechanism.

Surface traveling waves are of interest to radar in that they can contribute significantly to the radar cross section of a radar target. Typical radar targets, such as aircraft, missiles, and ships, usually present relatively long surfaces in the plane of the incident radar energy and thus are conducive to the formation of surface traveling waves. When the surface waves that propagate in the direction of flow of incident energy reach a discontinuity (e.g., a gap or termination of the physical structure) the wave is reflected. The reflected wave radiates in the end-fire mode, in the direction of the illuminating radar. The result can be an increase in effective radar cross section of the target that is many times that predicted by geometric optics. Much attention has been devoted, in so-called stealth technology, to reducing the surface wave effect through termination and other schemes. *PCH*

Ref.: Knott (1985).

A **transverse wave (T-wave)** is a wave not containing longitudinal components of electromagnetic field. They exist only in transmission lines with at least two insulated conductors. Thus, the critical frequency for T-waves equals zero. *AIL*

Ref.: Sazonov (1988), p. 13.

A **traveling wave** is an EM wave of a specific type being propagated in a transmission line only in one direction and characterizing the transfer of EM energy along the line. Traveling waves are categorized as *incident* or *reflected*. An incident wave is a traveling wave created by an oscillator and moving in the direction of the increase in a longitudinal coordinate. A reflected wave is a traveling wave originated by a load or by an irregularity and moving in a direction opposite that of an incident wave. *AIL*

Ref.: Sazonov (1988), p. 29.

A **tropospheric wave** is "a radio wave that propagates in the troposphere." In Russian literature, the term is used to denote waves propagated over long distances (up to 1,000 km) due to scattering or ducting in the troposphere. Only wavelengths shorter than about 10m can be propagated in this way. *AIL*

Ref.: IEEE (1993), p. 1,417; Dolukhanov (1972), p. 15.

A wavefront is a surface perpendicular to the direction of wave propagation at each point and passing through those points in space where the phase of the EM field is constant. Depending on the proximity to the source, the wavefront may be regarded as spherical, becoming essentially planar at great distances. Radiation or scattering from multiple points can produce irregularities in the wavefront. (See **GLINT.**) *SAL*

Ref.: IEEE (1993), p. 1453.

Wave impedance is "a complex factor relating the transverse component of the magnetic field to the transverse component of the electric field." When the sign is chosen so that the real part of impedance is positive, the wave impedance of a traveling wave is called *characteristic wave impedance*. The real part of wave impedance is wave resistance, showing the degree to which a medium or transmission line resists propagation of the traveling wave. For free space the wave resistance is 120π. *SAL*

Ref.: IEEE (1993), p. 1,484.

Wavelength is "the distance between points of corresponding phase of two consecutive cycles in the direction of the wave normal" (for a sinusoidal wave). The wavelength λ is related to phase velocity v and frequency f by $\lambda = v/f$. *SAL*

Ref.: IEEE (1993), p. 1,485.

Wave number is the reciprocal of the wavelength in a sinusoidal wave. It also denotes the magnitude of the wave vector (a vector that points in the direction of wave propagation). In radar applications, the wave number $k = 2\pi/\lambda$ is used instead of $1/\lambda$. *PCH*

Ref.: Van Nostrand (1983).

wave polarization (see **POLARIZATION**).

WAVEFORM, radar. A definition of the waveform is "the geometrical shape as obtained by displaying a characteristic of the wave as a function of some variable, usually, time." In radar, waveforms are typically divided into several groups; depending on the primary parameter chosen for classification:

(1) Waveforms are divided into continuous-wave waveforms and pulse waveforms depending on the character of their variation in time domain.

(2) From the standpoint whether they use intrapulse modulation, pulse waveforms are divided into simple (uncoded) pulses and pulse-compression waveforms (in Russian radar literature, these are usually termed *simple* and *complex signals*).

(3) If the waveform employs pulse compression, it is divided into frequency-coded (or frequency-modulated, FM) and phase-coded waveforms.

(4) Frequency-coded waveforms comprise linear FM waveforms, nonlinear FM waveforms, and time-frequency-coded (or frequency-stepped) waveforms.

(5) Phase-coded waveforms are divided into binary-coded waveforms and polyphase-coded waveforms, each having its own representatives depending on the code used for phase-shift keying.

(6) The most popular shapes of waveforms are rectangular, Gaussian, sawtooth, and triangular.

(7) From the viewpoint of discretization, waveforms are divided into analog waveforms (both in amplitude and time), discrete waveforms (analog amplitude and discrete time), and digital waveforms (digital code for amplitude at discrete moments in time).

The proper choice of waveform is crucial for radar operation as it defines all the main characteristics, including performance of target detection, resolution, measurement, and classification. The primary factors influencing the selection of a particular waveform are requirements of range and doppler coverage, probability of detection, resolving capability, measurement errors, interference rejection, the necessity of enhanced modes of operation (like target classification or imaging), and the complexity and cost of waveform generation and signal-processing systems. Most modern radars employ pulse-compression waveforms compatible with digital signal processing networks. *SAL*

Ref.: IEEE (1993), p. 1,481; Barton (1988), pp. 209–275, (1991), pp. 7.1–7.55; Skolnik (1980), pp. 411–434, (1990), pp. 10.1–10.39; Brookner (1977), pp. 121–198; Lewis (1986), pp. 17–108; Nathanson (1990), pp. 452–550.

The **Barker (phase)-coded waveform** is a phase-coded waveform using the Barker code. In Fig. W5, the waveform for a Barker code of length 13 is shown with a representation of the binary sequence (Barker code), the video signal as a frequency-modulated square-wave, and the corresponding phase reversals of a CW signal. *SAL*

Ref.: Hovanessian (1984), p. 241.

A **binary- (phase-)coded waveform** (Fig. W6) is a phase-coded waveform in which a carrier is divided in time into N segments (chips), and each is coded either "+" (corresponding to nominal carrier phase 0°) or "–" (corresponding to a 180° phase shift). *SAL*

Ref.: Skolnik (1990), p. 10.17.

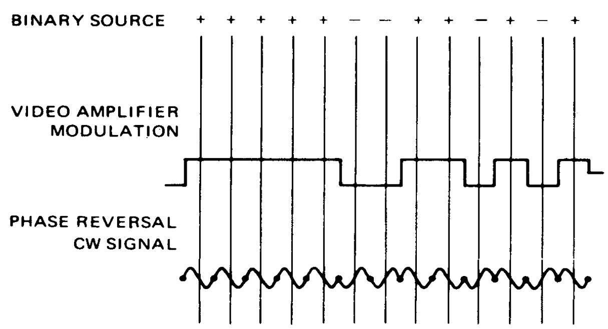


Figure W5 Waveforms for a Barker code of length 13 (from Hovanessian, 1984, Fig. 8-24, p. 241).

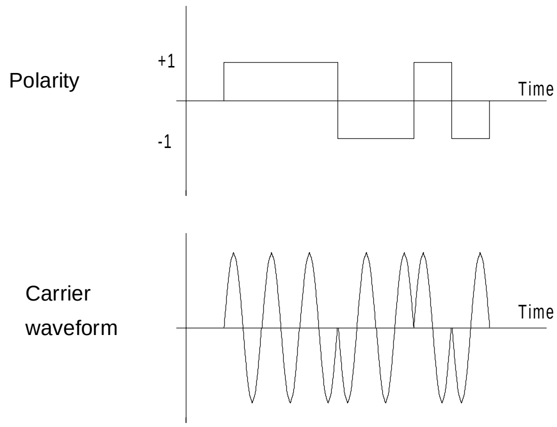


Figure W6 Binary-coded waveform (after Skolnik, 1990, Fig. 10.8, p. 10.17).

The **burst waveform** is a mode of radar operation in which bursts of N pulses (pulse trains) are transmitted at constant PRF and RF, for processing in a doppler-sensitive receiving system. (See **MOVING-TARGET DETECTOR.**) *SAL*

chirp waveform (see **linear frequency-modulated waveform**).

Classes of waveforms are defined:

Class A waveform refers to a simple pulse.

Class B waveform refers to a pulse-compression waveform.

Class C waveform refers to a coherent pulse train. *SAL*

Ref.: Bogush, 1989, p. 192.

A **complementary (phase)-coded waveform** is a phase-coded waveform using complementary codes to cancel range sidelobes. Two separate matched-filter outputs have equal range sidelobe amplitudes whose carrier frequencies are out of phase while the carriers of the main compressed pulse are in-phase (see Fig. W7). In the resultant compressed waveform, the sidelobes are canceled and the mainlobes are added coherently. Although theoretically this waveform represents a good solution to the sidelobe suppression problem, in practice there are difficulties in implementing such a technique (primarily related to doppler sensitivity) that curtail the range of its application. *SAL*

Ref.: Barton (1991), p. 7.31; Nathanson (1990), p. 491.

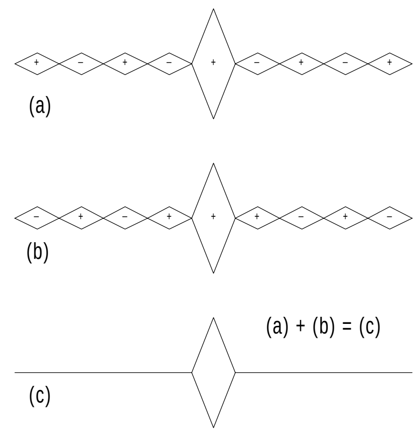


Figure W7 Range sidelobe cancellation with complementary sequences (after Barton, 1991, Fig. 7.2.30, p. 7-31).

A **continuous-wave waveform** is one that continues, theoretically, for an infinite time. In practice, such waveforms are limited in duration and are often coded in some manner to permit measurement of target range. This coding may be a frequency ramp (triangular wave, see **RADAR, CW**). If this waveform is interrupted periodically, it is *interrupted FMCW* (IFMCW).

Use of CW simplifies the radar design in some respects, providing high average power with lower operating voltages than in pulsed systems. The short-range performance can be greatly improved, since there is no dead time due to duplexing, but dual antennas are required when the transmitted power is more than a few watts. In addition, for high-power CW radars, extreme care is needed to isolate the receiver from the transmitter, and to minimize transmitted noise sidebands that would enter the doppler filters in which target detection is expected. FMCW waveforms must be generated with great linearity in the frequency deviation, if target energy is to be contained within a narrow doppler filter. *SAL*

Ref.: Currie (1987), p. 633; Barton (1978); Skolnik (1990), Ch. 14.

discrete-frequency-shift waveform (see **time-frequency coded waveform**).

fill-in sawtooth waveform (see **sawtooth waveform**).

The **Frank-coded waveform** is a polyphase-code waveform using the Frank code. It can be described considering a hypothetically sampled step-chirp waveform, like the four-frequency step-chirp waveform shown in Fig. W8, where F_i denotes the frequency of subpulse i. If the received signal using a Frank-coded waveform is processed through a matched filter, the compressed signal has lower range sidelobes than a linear FM waveform or a frequency-stepped waveform, but its performance deteriorates rapidly with doppler shift so it should not be used except under very-low-doppler-shift conditions. For large N, the peak range sidelobes for this waveform are approximately 20 log (πN), where N is the number of subpulses in the code. (See **CODE, Frank.**) *SAL*

Ref.: Barton (1991), p. 7.17; Lewis (1986), p. 10; Brookner (1977), p. 146.

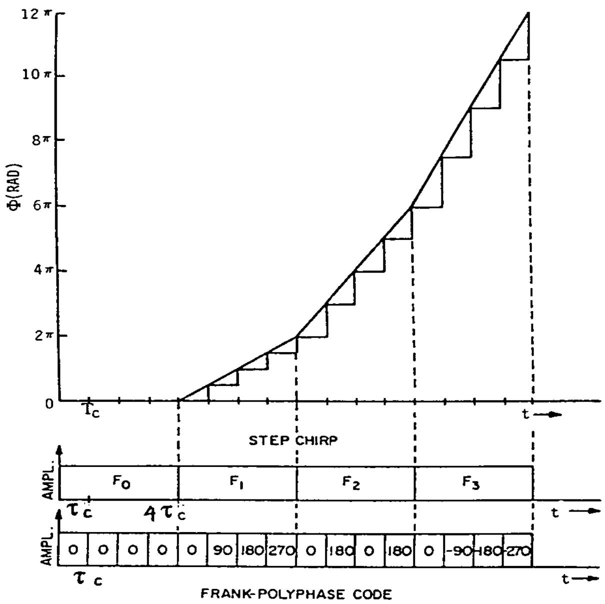


Figure W8 Step-chirp and Frank polyphase-code relationships (from Barton, 1991, Fig. 7.2.17, p. 7-18).

frequency-coded waveform (see **time-frequency-coded waveform**).

A **frequency-modulated waveform** is a waveform in which frequency modulation is applied in accordance with preset law. There are two main types of FM waveforms: linear and nonlinear FM waveforms. *SAL*

Ref.: Skolnik (1990), p. 10.3.

frequency-stepped waveforms (see **linear-FM waveform**).

Gaussian waveform (see **PULSE**).

A **Huffman-coded waveform** is a waveform in which a carrier is coded by a Huffman code to form amplitude modulation of the signal. At the output of a matched filter, such a signal has zero sidelobe levels except the beginning and the end of the waveform, for zero doppler. The disadvantages of such a waveform are that it does not stay well correlated under doppler shift conditions, and a large variation in the amplitude of the transmitted pulse envelope is required (29 dB in the example shown in Fig. W9). *SAL*

Ref.: Lewis (1986), p. 12; Brookner (1977), p. 146.

intrapulse-modulated waveform (see **pulse-compression waveform**).

large time-bandwidth product waveform (see **pulse-compression waveform**).

A **linear frequency-modulated waveform** is a frequency-modulated waveform in which the carrier frequency varies linearly in time for some specified period (for CW radar) or within the pulsewidth (for pulsed radar). To obtain this waveform, the phase must have a quadratic dependence on time. The waveform voltage can be written as

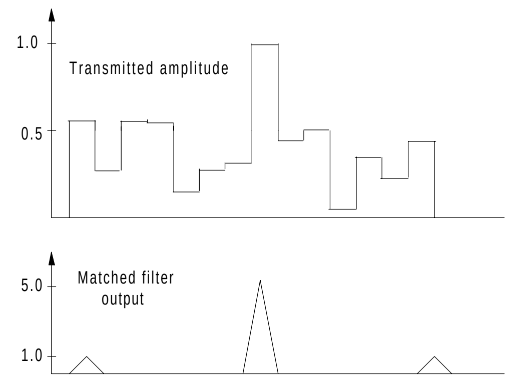


Figure W9 Huffman-coded waveform (after Cook, in Brookner, 1977, Fig. 6, p. 146).

$$v(t) = v_0 \cos\phi(t) = v_0 \cos[\omega_0 t + \theta(t)]$$

where v_0 is the amplitude, ω_0 is the carrier frequency, and $\theta(t)$ is the phase. When

$$\theta(t) = kt^2$$

the instantaneous frequency:

$$\omega(t) = \frac{d\phi(t)}{dt} = \omega_0 + 2kt$$

varies linearly (Fig. W10).

This is one of the oldest and most frequently used waveforms in both CW and pulse radar applications. In FMCW radars it makes possible the determination of target range as well as doppler. (See **RADAR, CW.**)

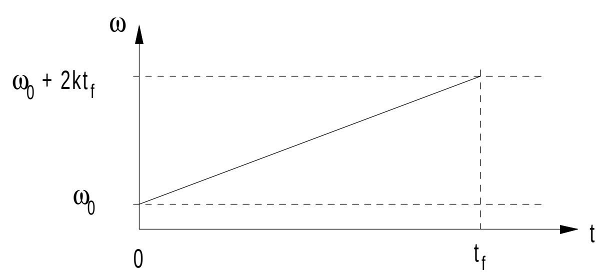


Figure W10 Frequency vs. time for linear FM waveform.

In pulse radar the frequency of this pulse-compression waveform increases (or decreases) linearly over the duration of the pulse (Fig. W11) providing intrapulse modulation. A variant of the linear FM waveform is the linear-step FM waveform, in which the frequency versus time function is not the continuous one, but comprises *N* equal steps as shown in Fig. W12. Such a waveform can approximate a linear FM pulse without using the dispersive delay line for generation.

There are two basic methods for generating linear FM waveforms: active and passive. Active methods generate the modulation by a special generator, for example, a voltage-controlled oscillator, with time-varying drive voltage resulting in the desired linear ramp. In the passive method, the

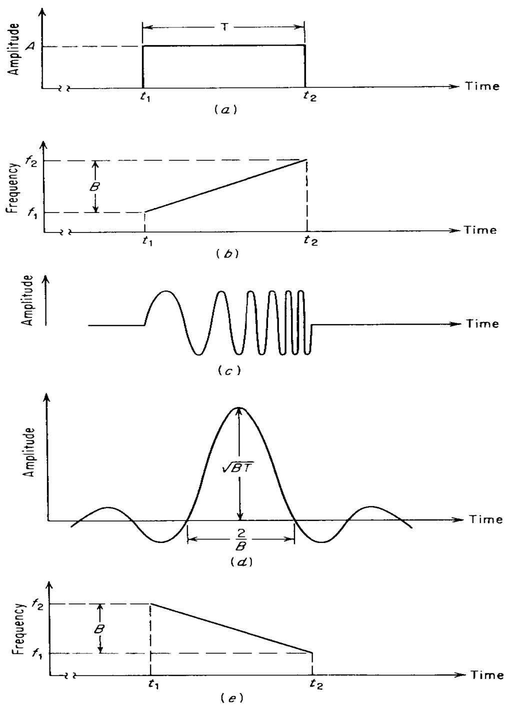


Figure W11 Linear FM waveform and its compression: (a) transmitted waveform; (b) frequency of the transmitted waveform; (c) representation of the time waveform; (d) output of the pulse-compression filter; and (e) same as (b) but with decreasing frequency (from Skolnik, 1980, Fig. 11.15, p. 423, reprinted by permission of McGraw-Hill).

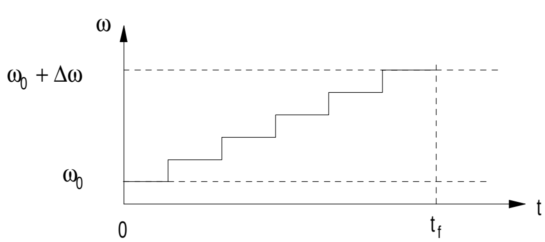


Figure W12 Frequency vs. time for stepped FM waveform.

waveform is generated by exciting a dispersive delay line with an impulse. The main advantages of the linear FM waveform are that it is quite insensitive to doppler shifts and is the easiest to generate, a variety of hardware being available to form and process it. The main disadvantages are

(1) It has range-doppler cross coupling, resulting in measurement errors unless one of the coordinates (range or doppler) is determined.

(2) Range sidelobes are high, compared with nonlinear FM and phase-coded waveforms. To reduce sidelobe level, weighting is usually required, resulting in a 1- to 2-dB loss in signal-to-noise ratio.

The linear FM pulsed waveform is often called *chirp*. *SAL*

Ref.: IEEE (1990), p. 8; Barton (1991), p. 7.14; Skolnik (1980), p. 422; Skolnik (1990), p. 10.4; Wehner (1987), p. 127; Brookner (1977), p. 136; Leonov (1988), p. 62.

A **noise-like waveform** is a sinusoidal or noise wave modulated by broadband low-frequency noise (amplitude, frequency, or phase modulated). For this type of waveform, the amplitude, frequency, and initial phase are random quantities even in the absence of any interference. This signal has a very broad band (ideal can be white noise with the infinite spectrum width), but it has strong limitations for using matched filtering to extract the useful signal based on its shape parameters. *AIL*

Ref.: Svistov (1977), p. 307.

A **nonlinear frequency-modulated waveform** is a frequency-modulated waveform in which the frequency of the carrier varies in time in accordance with any nonlinear law, either symmetrical or asymmetrical (Fig. W13). A symmetrical waveform usually has a frequency that varies in one direction during the first half of the pulse (e.g., increases) and varies in the opposite direction during the second half of the pulse (e.g., decreases). The phase variation for such a waveform can be written

$$\phi(t) = \omega_0 t + kt^2 + \omega_n(t)$$

where $\omega_n(t)$ is the nonlinear time function (e.g., sin(t), arctan(t), erf(t), or other).

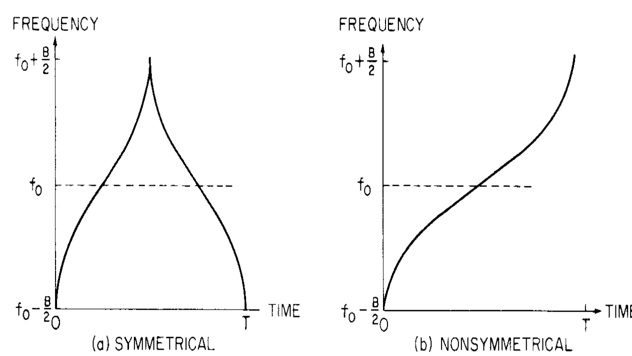


Figure W13 Nonlinear FM waveforms (from Skolnik, 1990, Fig. 10.2, p. 10.6, reprinted by permission of McGraw-Hill).

The asymmetrical waveform typically is one-half of the symmetrical waveform. In principle the nonlinear variation in time has the same effect as amplitude weighting of the transmitted signal spectrum, but a rectangular pulse shape, compatible with efficient transmitter operation, can be maintained. The nonlinear FM waveform has very low range sidelobes without necessitating the use of special weighting for their suppression, and hence has no signal-to-noise ratio loss as does the linear FM waveform. However, it is more sensitive to doppler frequency shifts, and the main factor affecting its acceptance is its complexity and the limited development of nonlinear FM generation techniques. *SAL*

Ref.: Skolnik (1980), p. 431, (1990), p. 10.4.

A **phase-coded waveform** is one in which intrapulse modulation is obtained by subdividing the pulse into subpulses of equal duration, each having a particular phase. The phase of each subpulse is set in accordance with a given code or code sequence. (See **CODE**; **SEQUENCE.**) Phase-coded waveforms may be binary- or polyphase-coded. First is a waveform in which there are only two levels of phase-shift keying: 0° and 180°, while the latter uses codes with the number of discrete phase values greater than two. The most popular binary-coded waveforms are the Barker codes and the allomorphic forms of binary codes. Polyphase-coded waveforms are represented by the Frank-coded waveform, quadriphase-coded waveform, and waveforms using P-codes. Compared with the pure linear FM waveform (without special weighting), phase-coded waveforms have lower range sidelobes. They are preferred when the radar operates in jamming conditions, as the coding of the transmitted signal (which can be changed) gives an additional degree of protection against ECM. But although the mainlobe of phase-coded waveform ambiguity diagram is rather narrow, it has a relatively wide sidelobe plateau, so resolution performance in a dense target environment or in presence of distributed clutter can be rather poor. In general, the implementation of phase-coded waveforms is more complex than that of the linear FM waveform. *SAL*

Ref.: Skolnik (1990), pp. 10.15–10.26.

A **polyphase-coded waveform** is a phase-coded waveform in which a carrier is divided into time segments coded by a polyphase code (as opposed to a binary code with two states of phase position). Historically, the first polyphase-coded waveform for radar applications was Frank-coded waveform, developed by R. L. Frank for Loran-C. Other waveforms of this type are waveforms using quadriphase codes and P-codes. The range sidelobes of polyphase-coded waveforms can be lower than those of the binary-coded waveforms of the same length, but the performance of this waveform deteriorates rapidly in the presence of doppler frequency shift. Generation and processing of polyphase-coded waveforms use technique similar to those of frequency-coded waveforms, but their range sidelobe parameters are much better than for unweighted FM waveforms. *SAL*

Ref.: Barton (1991), p. 7.17; Skolnik (1980), p. 432.

pulse waveforms (see **PULSE**)

A **pulse-compression waveform** uses intrapulse modulation to widen the signal bandwidth. It makes it possible to use long pulses to increase radiated energy and detection performance, but simultaneously having range resolution corresponding to short-pulse usage. (See **PULSE COMPRESSION.**) The most common types of intrapulse modulation are frequency and phase modulation (amplitude modulation is possible but seldom used). The basic types of pulse-compression waveforms are frequency-modulated waveforms and phase-coded waveforms. The comparative performance of these basic waveforms is given in Table W1. *SAL*

Ref.: Barton (1988) pp. 220–230, (1991), pp. 7.2–7.31; Skolnik (1980) pp. 420–34, (1990) p. 10.1–10.39; Brookner (1977), pp.143–148; Lewis (1986), pp. 7–116.

Table W1
Performance of frequency-modulated and phase-coded waveforms

	Linear FM		Nonlinear FM		Phase-coded	
	Active	Passive	Active	Passive	Active	Passive
Range coverage	Limited range coverage per active correlation processor.	Provides full range coverage.	Limited range coverage per active correlation processor.	Provides full range coverage.	Limited range coverage per active correlation processor.	Provides full range coverage.
Doppler coverage	Covers any doppler up to ± $B/10$, but a range error is introduced. *SNR* and time-sidelobe performance poor for larger doppler.		Multiple doppler channels required, spaced by $(1/T)$ Hz.			
Range sidelobe level	Requires weighting to reduce the range sidelobes below $(\sin x)/x$ falloff.		Good range sidelobes possible with no weighting. Sidelobes determined by waveform design.		Good range sidelobes. $N^{-1/2}$ for an N-element code.	
Waveform flexibility	Bandwidth and pulse width can be varied.	Limited to one bandwidth and pulse width per compression network.	Bandwidth and pulse width can be varied.	Limited to one bandwidth and pulse width per compression network.	Bandwidth, pulse width, and code can be varied.	
Interference rejection	Poor clutter rejection.		Fair clutter rejection.		Fair clutter rejection.	
SNR	Reduced by weighting and by ripple loss versus range.	Reduced by weighting.	Reduced by ripple loss versus range.	No *SNR* loss.	Reduced by ripple loss versus range.	No *SNR* loss.
Comments	1. Very popular with the advent of high-speed digital devices. 2. Extremely wide bandwidths achievable.	1. Widely used in past. 2. Well-developed technology.	1. Limited use. 2. Waveform generation by digital means most popular.	1. Limited use. 2. Extremely limited development.	1. Widely used. 2. Waveform very easy to generate.	1. Limited use. 2. Waveform moderately difficult to generate.

(from Farnett and Stevens, in Skolnik, 1990, Table 10.1, p. 10.5, reprinted by permission of McGraw-Hill).

A **quadriphase-coded waveform is** a polyphase-coded waveform in which carrier is divided in time into N segments and each is coded in accordance with quadriphase code sequence having 0-, 90-, 180-, or 270° values. *SAL*

Ref.: Morchin (1993), p. 379.

rectangular waveform (see **PULSE**).

A **sawtooth waveform** is a frequency-coded waveform in which frequency changes in time in a sawtooth fashion, having a long linear interval and a very short sweep recovery interval. This waveform is a practical realization of the linear FM waveform used in most continuous-wave radars, as it is impossible to realize unlimited linear variation of transmitter frequency without arranging periodic modulation relative to some preset carrier frequency. The combination of a frequency-stepped waveform with a sawtooth FM waveform provides tolerance to large doppler shifts and is known as a *fill-in sawtooth waveform*. *SAL*

Ref.: Nathanson (1990), p. 550; Leonov (1988), p. 23.

A **short-pulse waveform** is a pulse train consisting of a series of short simple pulses to achieve high range resolution. This waveform can be employed both in coherent and noncoherent radars. One way to form such a short pulse in a coherent system is to drive ringing filters with video or RF pulses that are even shorter than the desired one (Fig. W14). Very good range resolution can be achieved with this waveform (e.g., resolution of 8 cm has been achieved for RCS diagnostics), but since the pulsewidth is short, in some cases it requires extremely high peak power to ensure the tolerable *detection* performance at medium and long ranges that can restrict the application of this waveform primarily to the experimental purposes at short ranges. *SAL*

Ref.: Currie (1989), p. 375; Wehner (1987), p. 104.

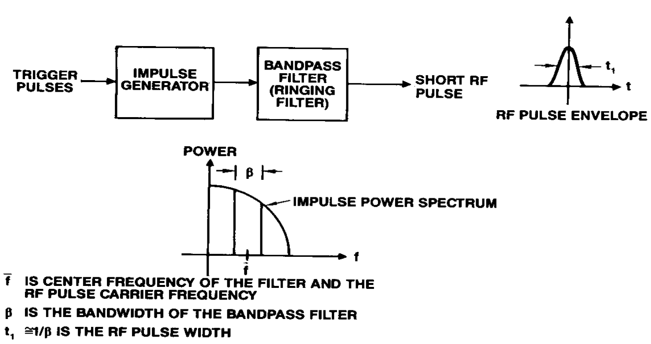


Figure W14 Generation of short pulses with ringing filter (from Wehner, 1987, Fig. 4.1, p. 104).

stepped-frequency (burst) waveform (see **linear-FM waveform**).

A **stretch waveform** actually refers to the technique using a linear FM waveform (chirp) on transmit, and down-converting on receive with a frequency-modulated local oscillator signal using a similar FM slope. For example, the transmitted signal can be a mixture of two chirps: one narrowband with bandwidth B_1 and the second wideband with bandwidth B_2. After reception, the radar return is mixed with the same wideband chirp B_2, resulting in a signal with bandwidth B_1 which goes to further processing. The range resolution corresponds to the signal with the bandwidth $B_1 + B_2$, while the bandwidth of signal processor may be equal to B_1.

In principle, this technique exchanges signal duration for signal bandwidth. It is done to obtain high range resolution by wideband transmitted signals while avoiding the use of the usual wideband processing circuitry. The main range of application is in the tasks where high resolution is required, such as radar mapping or imaging. However, keep in mind that only a portion of the whole range interval can be observed in this manner since the signal is stretched in time. This can be a problem for surveillance radars, but for tracking radar it may be tolerable. The method of using a stretch waveform to obtain high-range resolution signature of the target is shown in Fig. W15. *SAL*

Ref.: Skolnik (1980), p. 432; Wehner (1987), p. 115.

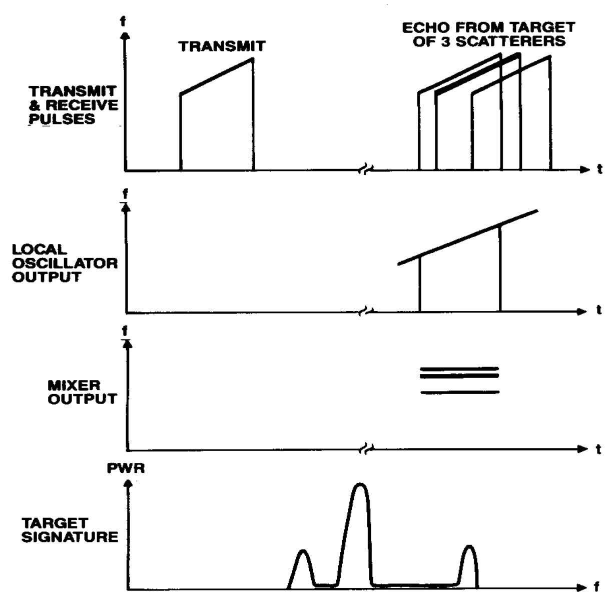


Figure W15 Stretch waveform and processing (from Wehner, 1987, Fig. 4.9, p. 116).

A **time-frequency-coded waveform** is a pulse train in which each pulse is transmitted at a different frequency (Fig. W16). In this case the range resolution of the compressed pulse is determined by the total bandwidth of all pulses, and the doppler resolution is determined by the waveform duration *T*.

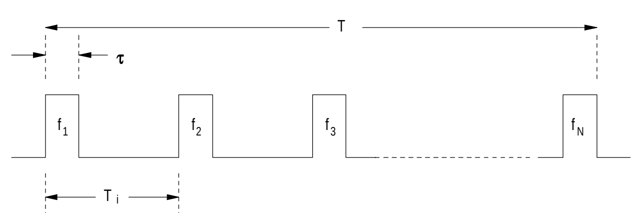


Figure W16 Time-frequency-coded waveform (after Skolnik, 1990, Fig. 10.14, p. 10.26).

Some relations for N contiguous pulses of width τ are given in the Table W2. Weighting of the pulse train, staggering of the pulse repetition interval, and intrapulse modulation of each pulse may be used to vary the basic waveform parameters. The time-frequency-coded waveform is preferred over the linear FM waveform for large signal bandwidths. It can be helpful under interference conditions if the order in which discrete frequencies are transmitted is varied. This kind of waveform sometimes is called the *discrete-frequency shift waveform*, *frequency-stepped waveform*, or *stepped-frequency waveform*. *SAL*

Ref.: Skolnik (1980), p. 431; Skolnik (1990), p. 10.26; Wehner (1987), p. 160; Currie (1987), p. 296.

Table W2
Waveform Relationships for *N* Pulses Contiguous in Time and Frequency

Waveform duration T	$N\tau$
Waveform bandwidth B	N/τ
Time-bandwidth product BT	N^2
Compressed pulsewidth $1/B$	$t/N = T/N^2$
(from Skolnik, 1990, Table 10.6, p. 10.26).	

triangular waveform (see **PULSE**).

A **wideband waveform** is a pulse-compression waveform with the spectrum width Δf up to the value of 0.1 of carrier frequency f_0 (e.g., $\Delta f/f_0 \le 0.1$). If the spectrum width is more than a half of the carrier frequency (i.e., $\Delta f/f_0 > 0.5$), the waveform is referred to as an *ultrawideband waveform*. *AIL*

Ref.: Dymova (1975), p. 97; Varakin (1970), pp. 122–250; Noel (1991); Taylor (1995).

waveform processing (see **SIGNAL PROCESSING**).

WAVEGUIDE. A waveguide is "(a) broadly, a system of material boundaries capable of guiding electromagnetic waves; (b) more specifically, a transmission line comprising a hollow conductive tube within which electromagnetic waves may be propagated or a solid dielectric or a dielectric-filled conductor for the same purpose." It serves as a channel over which energy of various types of oscillations can propagate: acoustic (see **ultrasonic waveguide**), RF (**dielectric**, **metal**, **stripline waveguide**); or optical (**beam**, **optical waveguide**). RF waveguides are used most widely in radar circuits. These are in the simplest case in the form of a metal tube of various cross-sections that transmit a microwave line. Other types of waveguides are also widely used: metal-dielectric, dielectric, and strip. Depending on the degree of screening, all waveguides are subdivided into two large classes of open and closed waveguides (Fig. W17).

A second conductor is not necessary for existence of waves in a waveguide, since the directed waveguide waves are formed through successive reflection from the walls or boundary of the dielectrics with different dielectric constants. The wavelength Λ of the result oscillations, propagating along the axis of the waveguide, is longer than the length λ of the wave in the dielectric (air):

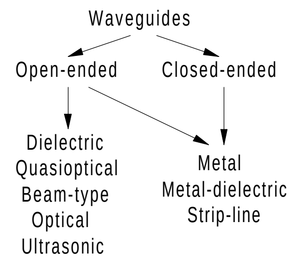


Figure W17 Basic types of waveguide.

$$\Lambda = \frac{\lambda}{\sqrt{1-(\lambda/\lambda_{cr})^2}}$$

where λ_{cr} is the critical wavelength. Along a waveguide of given transverse dimensions, it is impossible to transmit oscillations whose wavelength λ is greater than the critical length λ_{cr}. The critical wavelength depends on the geometric cross-section and type of propagated waves, and in size is comparable with the size of the transverse section (see formulas in the article **rectangular waveguide**). The value λ_{cr} depends on two integer indices m and n, which are used for indication of wave mode. Smaller values of λ_{cr} (higher wave modes) correspond to larger values of the indices. If conditions of propagation are satisfied ($\lambda < \lambda_{cr}$) for a given mode, then all lower mode waves can also be propagated. For suppression of the usually undesired lower wave modes, special filtration devices are used. (See **multimode waveguide**.)

Depending on the ratio of the operating and critical length of the wave, we distinguish between single-mode, multi-mode, and below-cutoff waveguides.

Waveguides are used most widely for channeling RF energy and for creation of various microwave elements and devices (filters, phase shifters, circulators, etc.) *IAM*

Ref.: Lavrov (1974), p. 295; Sazonov (1988), p. 23; Anderson, I. N., Microwave J ***25***, no. 12, 1982.

A **beam(-mode) waveguide** is an optical or quasioptical channel used to transmit the energy of EM waves in the optical, sub-millimeter, and shortest millimeter bands. Propagation of energy in beam waveguides is implemented in a wavebeam with a strictly localized cross section, as a result of which the effects of external emissions (e.g., of the sun and the scattered radiation of the sky), are eliminated. Lens and mirror beam waveguides are most used. These consist of a series of focusing lenses or mirrors. Losses to radiation in the waveguide are inversely proportional to the Fresnel parameter, which is equal to $2D^2/d$, where D is the diameter of the lenses or mirrors and d is the distance between them.

Since the dimensions of beam waveguides significantly exceed the dimensions of metal waveguides, they are not widely used in conventional radar.

The use of beam waveguides is most promising in the feeds of large-aperture antennas with mechanical scanning, since their use makes it possible to get along without rotating waveguide joints. *IAM*

A **below-cut-off waveguide** causes attenuation of waves with a wavelength λ_0 greater than the critical λ_{cr}. The below-cut-off waveguide has a purely reactive wave impedance and for this reason cannot be used directly as an energy transmission line. The optimum ratio of loss to volume for high-pass filters based on cut-off waveguides is obtained when the equation $\lambda_0/\lambda_{cr} \geq 1.6$ applies.

The use of below-cut-off waveguides in microwave generators makes it possible to reduce the dimensions and weight through a reduction in the size of the three-dimensional cavity. Sections of cut-off waveguides in combination with the nonuniformity introduced in them make it possible to create microwave assemblies for various purposes: below-cut-off attenuators, circulators, and bandpass filters. *IAM*

Ref.: Craven, G., and Mok, C., *IEEE Trans.* ***MTT-19***, Mar. 1971, pp. 295–308.

A **circular waveguide** has a circular cross section. Field patterns of some of the simpler waves in these waveguides are given in Fig. W18. For TE_{mn} waves, *m* denotes the number of axial planes along which the normal component of electrical field vanishes, and *n* is the number of cylinders including the boundary of the guide along which the tangential component of the electric field vanishes.

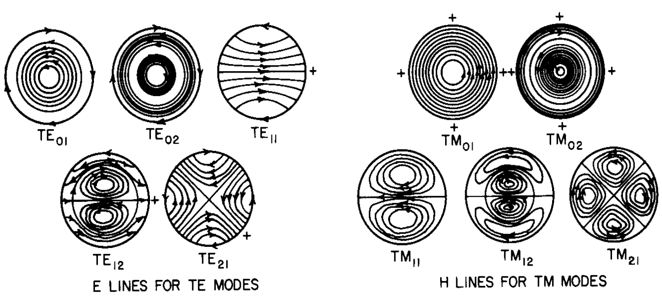


Figure W18 Field configurations for circular waveguide (from Johnson, 1984, Fig. 42-14, p. 42-33, reprinted by permission of McGraw-Hill).

Circular waveguides usually use H_{11} waves (critical length of wave $\lambda_{cr} = 3.42a$, where a is the radius), E_{01} waves ($\lambda_{cr} = 2.61a$), and H_{01} waves ($\lambda_{cr} = 1.64a$) (Fig. W19).

In properties, an H_{11} wave is analogous to an H_{10} wave in a rectangular waveguide. However, in a circular waveguide, two confluent (having identical λ_{cr}) H_{11} waves with orthogonal field polarization may exist. Any irregularities of a real waveguide leads to transition of the energy from one polarization to the other polarization.

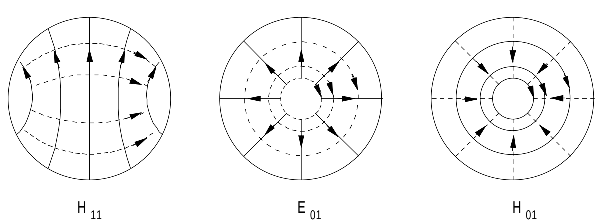


Figure W19 Field structure for main modes in circular waveguide (after Sazonov, 1988, Fig. 1.9, p. 25).

The axis-symmetrical wave E_{01} is used in rotating joints. In addition, an H_{11} wave may propagate in the waveguide. To prevent the appearance of an H_{11} wave, special measures are taken. The axisymmetric wave H_{01} has a relatively short critical wavelength and at least four types of waves can propagate simultaneously with it. However, thanks to the complete absence of longitudinal components of electrical currents on the walls of the waveguide, and the quick reduction in amplitudes of the transverse components of current on the walls with an increase in diameter, round waveguides with an H_{01} wave are used to form high-quality cavities and to channel millimeter waves. *IAM*

Ref.: Sazonov (1988), p. 24; Johnson (1993), p. 42.33.

A **closed waveguide** is a completely screened metal or metal-dielectric waveguide. It is used both as single-mode and as multi-mode closed waveguides, primarily with circular or rectangular cross-section. (See **circular waveguide**, **rectangular waveguide**.)

Closed waveguides are used for power circuits of radars in the decimeter, centimeter and millimeter wavebands. *IAM*

Ref.: Anderson, J. N., *Microw. J.* ***25***, no. 12, 1982.

A **dielectric waveguide** is an open waveguide in the form of a continuous or hollow dielectric rod of rectangular, circular, or elliptical cross section. For a dielectric cylindrical waveguide, joint existence of E- and H-waves that are nonhomogeneous in azimuth is characteristic. The main wave constitutes the simplest combination and is designed by the symbol EH_{11}. In contrast to the main wave of a hollow waveguide, it is propagated without cutoff at any frequencies (critical frequency is equal to λ_0). The real critical frequency of a dielectric waveguide is determined on the basis of a provisionally selected attenuation level of the wave due to radiation. The radius of the wave field, 10, in which 99.5% of the wave power is concentrated, corresponds to this level.

Dielectric waveguides are classed as decelerating systems. They are used in rod-type dielectric antennas, and as the component base of integrated circuits in the millimeter band. *IAM*

Ref.: Nikol'skiy (1964), p. 296.

An **elliptical waveguide** has characteristics resembling those of rectangular waveguide, but its bandwidth is narrower. Typically it is made in long lengths or is assembled in desired sections with an adaptive flange to rectangular guide. Elliptical waveguides find application in antenna feeds. *SAL*

Ref.: Johnson (1984), p. 42.42.

A **ferrite waveguide** use ferrite inserts to create various microwave components (assemblies). These use ferromagnetic properties and are associated with phase shift of propagating electromagnetic waves and the phenomenon of ferromagnetic resonance.

Based on such waveguides, circulators are developed with losses of less than 1 dB, an isolation between channels of more than 15 dB, and bandwidths of more than 10% (in the long-wave section of the millimeter band), tunable filters (see **FILTER, ferrite**), directional couplers, and phase shifters. (See **PHASE SHIFTER, ferrite**.) *IAM*

Ref.: Itoh, T., *Microw. J.* ***25***, no. 9, 1982.

H- and π(-type) waveguides have the form of tubes with a cross section in the form of an H or which differ from rectangular ones in the presence of internal projecting pieces. They use waves which are similar to the H_{10} waves of a rectangular waveguide. An increased concentration of the electrical field in H- and π-type waveguides near the projecting pieces is equivalent to the addition of some linear capacitance, which leads to a reduction in phase velocity and an increase in the critical wavelength λ_{cr}. At the same time, the λ_{cr} of the nearest higher type of wave is decreased. For this reason, H- and π-type waveguides can pass a larger frequency band than rectangular ones with the H_{10} wave. The transverse dimensions of H- and π-type waveguides are less than in rectangular guides for a given wavelength.

Drawbacks of H- and π-type waveguides include greater attenuation and lower power-handling capacity compared with rectangular ones. *IAM*

Ref.: Lavrov (1974), p. 300.

A **hollow(-tube) waveguide** is a hollow metal tube of rectangular, circular or other cross section in which the electromagnetic energy is propagated. A number of different types of waves can propagate through the waveguide, but typically it is designed for transmission of a single one; for example the dominant wave or the wave having the lowest cutoff frequency. *IAM*

Ref.: Lavrov (1974), p. 300.

Metal waveguide uses only metal limiting surfaces. Depending on the shape of the cross-section, we distinguish between rectangular, circular, and H- and π-shaped waveguides. In addition to smooth-sided waveguides, waveguides with corrugated walls are also used, as well as open waveguides in the form of two symmetrical slots in metal planes (low-power waveguides).

The wall thickness of waveguides is selected for mechanical strength. To increase the stability of oscillations, waveguides are covered on the inside with a thin film of lacquer, and to reduce losses they may be silver- or gold-plated. To transmit higher power, they are filled with a special gas with higher electrical strength. *IAM*

Ref.: Lavrov (1974), p. 295.

A **metal-dielectric waveguide** is made of dielectric materials and limiting metal surfaces. Closed waveguides are waveguides fully or partially filled with dielectric, with fully screening external metal surfaces. Filling the waveguide with material with a high dielectric constant significantly reduces the dimensions of the waveguide with an increase in the dissipative losses. Coating the metal walls with a dielectric layer on the order of $\lambda/3$ thick, or placing dielectric inserts in the central part, results in decreasing distribution of the field amplitude toward the walls and a reduction in losses. Single-mode rectangular waveguides, with a dielectric plate with a slot line (waveguide-slot line) perpendicularly mounted at the center, are used for waves in the millimeter band.

Among the open waveguides, a mirror line (see figure in article **open waveguide**) and insulated dielectric waveguide on a metal substrate are widely used. A layer of dielectric with lesser dielectric constant is used as the insulator, reducing the induced currents in the substrate, as well as ohmic losses.

At frequencies of 10 GHz and higher, metal-dielectric waveguides have significant advantages over microstrip lines: losses that are smaller by an order of magnitude, and acceptable power greater by an order of magnitude, with smaller mass and size characteristics.

Metal-dielectric waveguides are used to develop miniature multifunctional microwave devices, including those with a higher power level. *IAM*

Ref.: Koshevaya, S. V., et al., *Radioelektronika* **10**, 1990, p. 3, in Russian; Kanyushevskiy, Yu. Z., et al., *Radiotekhnika*, no. 1, 1982, p. 8, in Russian.

A **waveguide mode** is the term used to describe the electric and magnetic field pattern in a waveguide. To identify the mode the type of the wave and the number of relative maxima in the field configuration (usually with subscripts) are given. For example, TE_{mn} denotes that the electric field is transverse to the direction of propagation and it has m and n relative maxima occurring along the width and the height of the cross-section, respectively. *SAL*

Ref.: Fink (1975), p. 3.28.

A **multimode waveguide** can propagate EM waves of several modes in addition to the main mode. Multimode waveguides include quasioptical and optical waveguides (open waveguides), and closed waveguides used in the shortwave section of the millimeter band. The use of multimode closed waveguides is explained by the smaller losses of waveguides of increased section, whose dimensions violate the condition of single-mode propagation. (See **single-mode waveguide**.)

In contrast to the single-mode waveguide, forward and reflected parasitic modes are excited in the nonuniformities of a multimode waveguide. Filtering parasitic modes is a complex task, which is accomplished by various methods: breaking a multimode waveguide down into several single-mode waveguides, production of circuit components in the form of smooth junctions, or use of optical components (prisms, mirrors). Filtering modes with polarization different from the polarization of an operating mode is effected with the use of resistor elements arranged orthogonally to the electrical field of the operating mode, or with slots in the walls of a metal waveguide, oriented along the lines of current induced by the field of the operating mode.

Multimode waveguides are used beginning in the shortwave sector of the millimeter band, and in the submillimeter and optical bands. Usually circular metal waveguides (with a diameter > 0.8λ) are used as the circuits for the millimeter band, with an operating mode of H_{11} or H_{01}, and rectangular metal waveguides (with size of wide wall > λ) and operating mode H_{10}. *IAM*

Ref.: Anderson, I. N., *Microw. J.* ***25***, no. 12, 1982.

An **open waveguide** is partially screened or unscreened. Dielectric waveguides and partially screened metal and metal-dielectric waveguides are classed as open waveguides. Polarization of the electrical field usually is orthogonal to the metal wall, which has a dielectric layer on it (Fig. W20a–c), and parallel to the metal walls, forming edges (Fig. W20d).

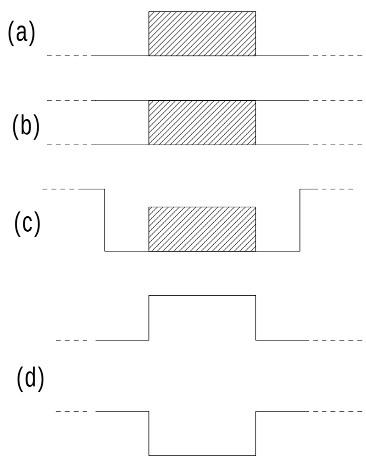


Figure W20 Basic types of open-ended waveguide.

Open waveguides, in contrast to closed types, are self-filtering. Their parameters are selected in such a way that all modes except one (the operating mode) are attenuated. The advantage of open waveguides is their lower optical losses due to the reduction in the number of metal walls and concentration of the field inside the dielectric (metal-dielectric waveguides). The disadvantage is the significant losses to radiation when there is any nonuniformity, for example of any elements of the circuit. In connection with the presence of the attenuation modes, compensation methods for reducing losses are not feasible, and elements of a circuit based on open waveguides are usually made in the form of smooth junctions (bends, smooth branches, etc.) with partial screening at the point of the nonuniformity.

Open waveguides, with the exception of tubular dielectric waveguides, are used for transmission of relatively low powers. They are used to make integrated circuits and planar antennas of the millimeter waveband.

Open waveguides also include microstrip lines whose use is limited to the long-wave portion of the millimeter band. *IAM*

Ref.: Wang, T., and Schwartz, S. E., *IEEE Trans.* ***MTT-31***, no. 2, 1983.

Optical waveguide is a multimode dielectric waveguide for transmission of waves in the submillimeter and optical bands. The cross-sections of optical waveguides amount to hundreds of wavelengths. Their operation is based on the phenomenon of full internal reflection. (See **TRANSMISSION LINE, fiber-optic**.) The path of propagation of beams inside the waveguide changes with a change in frequency, which leads to dispersion and distortions of transmitted wideband signals. Single-mode optical waveguides, which are fibers 3 to 5 μm in diameter, and are practically free of dispersion, are promising. *IAM*

Ref.: Adams (1981).

A **quasioptical waveguide** has a larger cross section compared with standard waveguides and is used for channeling millimeter or submillimeter waves. When waveguides for H_{10} waves are increased in size by a factor of 3 to 10, a corresponding 3- to 10-fold reduction in losses is achieved. Transmission of a pure H_{10} wave from a standard waveguide to a quasioptical waveguide is usually done using a conical adapter. *IAM*

Ref.: Harvey, *Proc. IEEE*, Mar. 1959, p. 154.

A **rectangular waveguide** is in the form of a tube of rectangular cross section $a \times b$ and is used for transmitting power in a basic H_{10} wave. Critical wavelengths in a rectangular waveguide are calculated from the formula

$$\lambda_{cr} = \frac{2}{\sqrt{(m/a)^2 + (n/b)^2}}$$

The long-wave boundary for the use of a waveguide λ_{max} is selected to be 10% below the critical length of the main wave H_{10} ($\lambda_{cr} = 2a$) to retain electrical strength. The short-wave boundary arises from the requirement for absence of propagation of higher mode waves, the nearest of which are the H_{20} ($\lambda_{cr} = a$), E_{11}, and H_{11} ($\lambda_{cr} = (2ab)/(\sqrt{a^2 + b^2})$). The structure of the field of wave H_{10} is shown in Fig. W21.

Rectangular waveguides are used most widely due to their simplicity of design and the stability of the modes in them. *IAM*

Ref.: Sazonov (1988), p. 23.

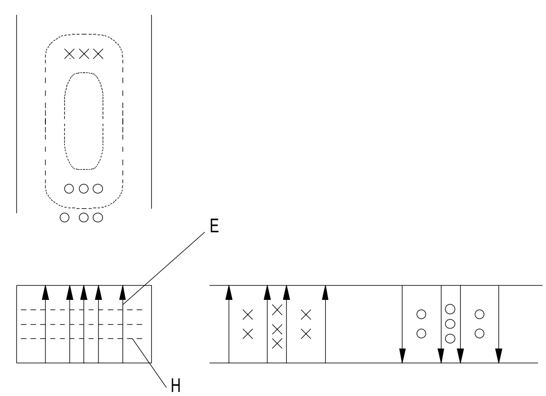


Figure W21 Structure of field of H_{10} wave in rectangular waveguide (after Nikol'skiy, 1964, Fig. 210, p. 275).

ribbed waveguide (see **H- and π-type waveguides**).

A **ridged waveguide** has single or double ridges inserted on the broad wall to obtain broader bandwidth in a single mode. The variations of cutoff wavelength for ridge waveguide are given in Fig. W22. *SAL*

Ref.: Fink (1982), p. 9.12.

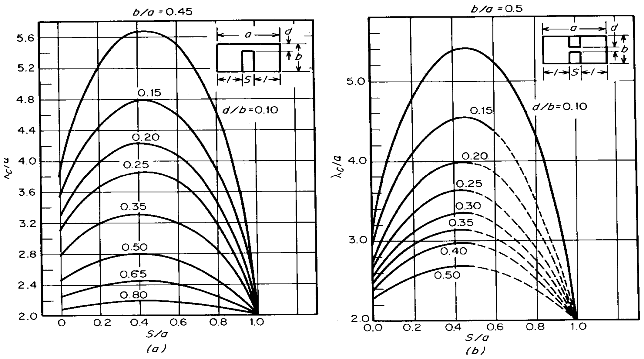


Figure W22 Cutoff wavelength of ridge waveguide: (a) single ridge; (b) double ridge (from Fink, 1982, Fig. 9-8, p. 9-13, reprinted by permission of McGraw-Hill).

Single-mode waveguide propagates waves in a single mode, eliminating those with $\lambda > \lambda_{cr}$. The minimum operating wavelength is at least 1% higher than the critical wavelength λ_{cr} of the next higher mode.

In contrast to closed waveguides, the parameters of open waveguides may be selected so that all modes except for the one operating mode are attenuated (self-filtering property). *IAM*

Ref.: Anderson, I. N., *Microw. J.* ***25***, no. 12, 1982.

A **spiral waveguide** is in the form of a conductor convoluted in the form of a spiral. The action of a spiral waveguide is represented in simplified form on the basis of motion of a T-wave along the spiral line of the conductor. The phase velocity of the wave along the axis will be less than the wave velocity in a vacuum by a factor of $2R/d$, where R is the radius of the spiral and d is the pitch of the spiral. Precise analysis shows the presence of dispersion in a spiral waveguide. With distance from a spiral waveguide, the field of the slow wave diminishes exponentially. Spiral waveguides are used as slow-wave structures. *IAM*

Ref.: Nikol'skiy (1964), p. 294.

Squeezable waveguide is "a variable-width waveguide for shifting the phase of RF wave traveling through it." *SAL*

Ref.: IEEE (1993), p. 1,267.

Strip(line) waveguide is in the form of a strip transmission line with solid dielectric or air filling. We distinguish between two basic types of strip waveguides: asymmetric (usually open) and symmetric (closed and open). The latter as a rule have smaller losses to radiation than asymmetric ones. Depending on the presence of dielectric, strip waveguides are classed as metal or metal-electric waveguides.

There are usually only transverse electromagnetic waves (T-waves) in a section of a strip waveguide. Strip waveguides in the millimeter band are inferior to completely filled metal waveguides in losses and throughput power. Ease of manufacture is an advantage of strip waveguides. *IAM*

Ref.: Coevolve (1972), p. 195.

A **trough waveguide** is a broadband waveguide having the cross section shown in Fig. W23. The cutoff wavelength of

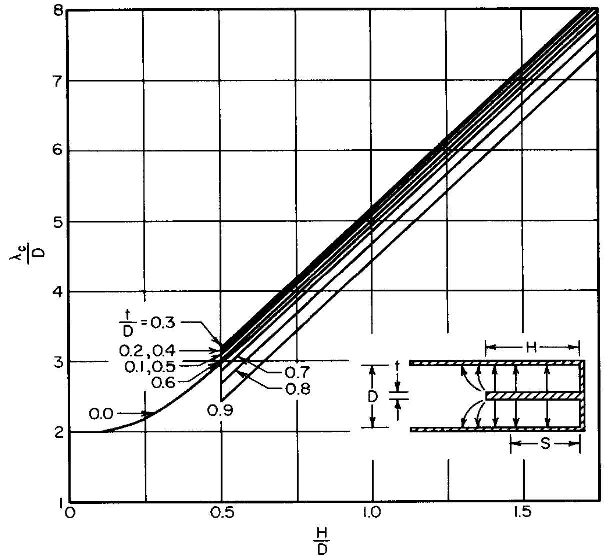


Figure W23 Cross section and cutoff wavelength of trough waveguide (from Johnson, 1993, Fig. 42.17, p. 42.40, reprinted by permission of McGraw-Hill).

the dominant mode may be determined from the graph in the figure. *SAL*

Ref.: Johnson (1993), p. 42.37.

Ultrasonic waveguide is intended for propagation of ultrasonic waves. Its operation is based on the propagation of longitudinal and transverse (shifting and twisting) waves in a sample of elastic material (sonic conductor). Depending on use, the sonic conductor may have various shapes: stepped, wedge-shaped, uniform, and so forth.

The waveguide may be nonpiezoelectric (fused quartz, sapphire) or piezoelectric (monocrystal quartz, lithium niobate), nonconducting (quartz, sapphire, lithium niobate) or conducting (aluminum, steel, etc.). Depending on the material of the waveguide, the connection to electrical circuits for excitation of the wave may be made using conducting electrodes attached directly to the piezoelectric sonic conductor, or to a layer of the piezoelectric material placed between the electrode and the nonpiezoelectric sonic conductor and also using electromagnets.

In ultrasonic and in radio waveguides, the effect of natural dispersion is manifested. Devices using this effect are often made in the form of metal bands or rods with a first longitudinal wave possessing the maximum length of the frequency segment of linear dispersion. In round waveguides in the form of wires or tubes, longitudinal waves at low frequencies have low dispersion, and this is used in nondispersive magnetostrictive delay lines. Torsional vibrations may be used as dispersion delay lines.

To obtain a linear dispersion characteristic, two-layer and multilayer waveguides are used. Their operation is based on the nonidentical penetration of oscillations of different frequency from one medium to another. *IAM*

Ref.: Shirman (1974), pp. 136, 142.

WAVE PROPAGATION (see **PROPAGATION**).

WEIGHTING is the process of multiplication of an input function to provide improvement in specified characteristics. In radar applications, weighting is applied to aperture illuminations (a process known as tapering) to reduce antenna sidelobes, to pulse-compression waveforms to reduce time sidelobes, or to coherent pulse trains to reduce doppler sidelobes. In all cases the reduction in sidelobe levels is achieved at the expense of broadening of the mainlobe and loss in gain or efficiency. The characteristics of different weighting functions are shown in Table W3. (See also **LOSS.**) *SAL*

Ref.: Barton (1969), App. A, B; Skolnik (1990), p. 10.27.

Table W3
Weighting Functions

Illumination function	Efficiency, η	Peak sidelobe (dB)	Beamwidth factor, *k* (deg)
Linear illumination functions:			
Uniform	1.00	–13.3	50.8
Cosine	0.81	–23	68.2
Cosine squared (Hanning)	0.67	–32	82.5
Cosine squared on 10 dB pedestal	0.88	–26	62
Cosine squared on 20 dB pedestal	0.75	–40	73.5
Hamming	0.73	–43	74.2
Dolph-Chebyshev	0.72	–50	76.2
Dolph-Chebyshev	0.66	–60	62.5
Taylor, $\bar{n} = 3$	0.9	–26	60.1
Taylor, $\bar{n} = 5$	0.8	–36	67.5
Taylor, $\bar{n} = 8$	0.73	–46	74.5
Circular illumination functions:			
Uniform	1.00	–17.6	58.2
Taylor, $\bar{n} = 3$	0.91	–26.2	64.2
Taylor, $\bar{n} = 5$	0.77	–36.6	70.7
Taylor, $\bar{n} = 8$	0.65	–45	76.4

Beamwidth = $k\lambda/a$ or $k\lambda/D$ where a is width of rectangular antenna, D is diameter of circular antenna (after Skolnik, 1990, Table 7.1, p. 7.38).

Bickmore-Spellmire weighting [distribution]. The Bickmore-Spellmire distribution is the distribution given by

$$f(r) = p\left[1-\left(\frac{2r}{D}\right)^2\right]^{p-1}\Lambda_{p-1}\left[jA\sqrt{1-\left(\frac{2r}{D}\right)^2}\right]$$

where p and A are the constants that determine the distribution, Λ is the lambda function, and D is the extent of the distribution.

In radar applications this distribution is typically used to describe the aperture distribution for antennas with continuous circular aperture (then D is the antenna diameter) in which it produces the Bickmore-Spellmire pattern. Detailed descriptions of the Bickmore-Spellmire distribution for the linear and circular apertures are given in Skolnik (1970). *SAL*

Ref.: Currie (1987), p. 532; Skolnik (1970), Ch. 9.

The **Chebyshev weighting [distribution]** is the distribution that produces the narrowest possible mainlobe for a given sidelobe level. This distribution is more theoretical than practical in antenna design, as it requires an impulse at both ends of the aperture that is rather difficult to realize, especially for relatively large apertures. *SAL*

Ref.: Currie (1987), p. 530; Skolnik (1970), Ch. 9.

Filter weighting is the process of shaping the filter response, typically to reduce range sidelobes in a pulse compression waveform. The term is applied more broadly to weighting of any waveform envelope (time weighting) or frequency spectrum (frequency weighting) in signal processing. Many different weighting functions (or windows) are used, including the classical Hamming function:

$$C_n = a_0 - a_1\cos\left(\frac{2\pi n}{N}\right)$$

where n is the sample index and N is the number of samples, $a_0 = 0.54$, and $a_1 = 1 - a_0 = 0.46$. For $a_0 = a_1 = 0.5$, the function is a $\cos^2$ weight (where the origin is taken as the center rather than the end of the window), also known as the Hann window. Multiple-term windows of the form

$$C = a_0 - a_1\cos\left(\frac{2\pi n}{N}\right) + a_2\cos\left(\frac{4\pi n}{N}\right)$$

are also used, where, for example, $a_0 = 0.423$, $a_1 = 0.498$, $a_2 = 0.075$ (known as Blackman-Harris windows). Other weighting functions are the Dolph-Chebyshev and the Taylor, the first giving the narrowest mainlobe width and the second the lowest loss, for a given sidelobe level, and Kaiser weighting, giving minimum integrated sidelobe energy. For characteristics of these functions see Table W3. The choice of appropriate weighting is based on a trade-off among sidelobe level, rate of sidelobe fall-off, mainlobe width, and weighting loss. *SAL*

Ref.: Barton (1969), App. A, B; Skolnik (1990), p. 10.27; Levanon (1988), p. 211.

WINDOWING (see **WEIGHTING, filter**).

X

(No entries)

Y

(No entries)

Z

ZONE, radar

A **Fresnel zone** is any one of the concentric surfaces between a transmitter and a receiver over which the difference between the indirect (reflected) path and the direct path is equal to a multiple of one-half of the wavelength. These zones of physical reflection are called Fresnel zones because they are a manifestation, at radio and radar frequencies, of Fresnel's optical diffraction theory. Fresnel zones are useful in describing the interference lobes produced by the interaction between a direct and a surface-reflected wave.

Figure Z1(a) illustrates the direct (r_1) and indirect (r_2) ray paths in the vertical plane between a transmitter at height h_1 and a receiver at height h_2 at some ground distance d away.

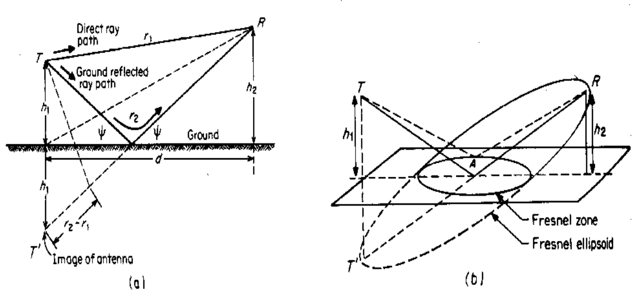


Figure Z1 (a) Ray paths for ground reflection; (b) Fresnel zone, plane earth geometry (from Fink, 1989, p.18.56, reprinted by permission of McGraw-Hill).

It can be shown that for grazing angles $\psi < 0.5$ radians, the phase difference, in radians, corresponding to the path difference (r_2-r_1) is

$$\Delta = \frac{4\pi h_1 h_2}{\lambda d}$$

and the resultant field strength at the receiver E_R is equal to $E_1\Delta$, where E_1 is the direct path field strength. For a given distance d, the received field strength will vary as the height of either antenna is altered. The grazing angles at which maxima and minima occur are determined by

$$\sin\psi = \frac{n\lambda}{4h_1}$$

which will result in maxima for n odd, minima for n even.

Figure W24(b) more clearly represents the actual situation, showing that the surface is illuminated over a broad region, which reradiates elementary wavelets in all directions. In the direction toward R is an elliptical zone on the ground near the point of ray reflection, where the wavelets add nearly in phase. From successive ring areas, or Fresnel zones, each bounded by larger ellipses, the wavelets alternately cancel and add. Most of the reflected energy comes from the first Fresnel zone, which is defined as that area from which all the reradiated elementary wavelets arrive within one-half wavelength of the phase of the direct path ray. Thus in Figure W24(b), the boundary of the first Fresnel zone is defined as the locus of points A such that the distance $TA + AR$ is different by a half-wavelength from the direct path distance TR. This locus is the Fresnel ellipsoid, an ellipsoid of revolution with foci at *T'* and *R*.

The Fresnel zone just described should not be confused with the Fresnel *region*, which is an interval in space between the *near field* of an antenna and the *far field* (see **ANTENNA radiation regions**). In the Fresnel region, ray paths from the antenna aperture to the observation point (target) are not parallel, and the antenna radiation pattern is not is not constant with distance. *PCH*

Ref: Van Nostrand (1983), p. 1,280.; Fink (1982), pp. 18.53–18.56.

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Radar Abbreviations and Acronyms

2D	two-dimensional
3D	three-dimensional
A	ampere
A/D	analog-to-digital
ABM	antiballistic missile
ac	alternating current
ACQ(N)	acquisition
ACR	approach-control radar
ADC	analog-to-digital converter
ADF	(1) automatic direction finder, (2) airborne direction finder
ADT	automatic detection and tracking
AEW	airborne early warning
AFC	automatic frequency control
AFR	acceptable failure rate
AG	available gain
AGC	automatic gain control
ARH	antiradiation homing
AI	airborne intercept
AIC	adaptive interference canceler
AJ	antijamming
AM	amplitude modulation
amp	(1) amplifier, (2) amplitude, (3) ampere
AMTI	(1) airborne MTI, (2) adaptive MTI
ANT	antenna
AOA	angle of arrival
AP	antenna pattern
APC	automatic phase control
ARM	antiradiation missile
ARR	antenna rotation rate
ARSR	air route surveillance radar
ART	automatic range tracking
ARTS	automated radar terminal system
ATCRBS	air traffic control radar beacon system
ASDE	airport surface detection equipment
ASK	amplitude-shift keying
ASR	airport surveillance radar
ASW	antisubmarine warfare
ATC	air traffic control
ATR	(1) antitransmit/receive (switch), (2) automatic target recognition
ATT	avalanche-transit-time
atten	attenuator
az	azimuth
BAM	bird activity modulation
BBD	bucket-brigade device
BEF	band-elimination filter
BF, BFP	bandpass filter
BLWN	band-limited white noise
BMWD	binary moving-window detector
BO	blocking oscillator
BPC	binary phase code
BRG, brg	bearing
BS	(1) backscattering, (2) beam-steering
BSF	bandstop filter
BSU	beam-steering unit
BW	(1) backward-wave, (2) bandwidth, (3) beamwidth
BWO	backward-wave oscillator
BWT	backward-wave tube
CA	(1) clutter attenuation, (2) circuit analog
CACFAR	cell-averaging constant-false-alarm-rate
CADF	commutated-antenna direction finder
CAT	computer-aided testing
CCD	charge-coupled device
CEP	circular error probability
CFA	crossed-field amplifier
CFAR	constant-false-alarm rate
CFT	continuous Fourier transform
CHILL	chaff, illuminated
CMF	coherent [circulating] memory filter
CMP	comparator
CO	crystal oscillator
coef	coefficient
COHO	coherent oscillator
colidar	coherent light detection and ranging
COMINT	communications intelligence
const	constant
CORA	coherent antenna array
CORAD	color radar
COSRO	conical scan on receive only
COTAR	correlation tracking and range (system)
CP	(1) circular polarization, (2) clock pulse, (3) control panel, (4) control point
CPACS	coded pulse anticlutter system
CPE	circular probable error
CPG	clock-pulse generator
CPI	coherent processing interval
CPSK	coherent phase-shift keying
CR	cancellation ratio
CRM	counter-radar measures
CS	(1) conical scanning, (2) control signal
CSR	clutter-to-signal ratio
CTD	charge-transfer device
cur	current
CW	continuous wave
CWDR	continuous-wave doppler radar
CWO	continuous-wave oscillator
D/A	digital-to-analog
DABS	discrete address beacon system
DAC	digital-to-analog converter
DAFC	digital automatic frequency control
DAGC	digital automatic gain control
dB	decibel
dBc	decibel wrt carrier
dBK	decibel wrt one kelvin
dBm	decibel wrt one milliwatt
dBm^2	decibel wrt one square meter
dBs	decibel wrt one second
dBsm	decibel wrt one square meter
dBW	decibel wrt one watt
DBF	digital beamforming
DBS	doppler beam sharpening
dc	direct current
DCON	downconverter
DD	(1) digital data, (2) digital display
DDP	distributed data processing
DECM	deceptive electronic countermeasures
decr	decrement
del	delay

dev	(1) deviation, (2) device
DEW	distant early warning
DF	(1) direction finder, (2) detectability factor, (3) doppler filter, (4) digital filter
DFT	discrete Fourier transform
deg	degree
DIC	digital integrated circuit
DL	(1) delay line, (2) data link
DME	distance-measuring equipment
DMF	digital matched filter
DOA	direction of arrival
DP	(1) data processing, (2) distributed processing
DPCA	displaced-phase-center antenna
DPLL	digital phase-locked loop
DPSK	differential-phase-shift keying
DR	(1) dynamic range, (2) detection radar
DRFM	digital radio-frequency memory
DSD	drop-size distribution
DSP	digital signal processor
DST	display storage tube
DVOR	doppler VHF omnirange
DWG	digital waveform generator
EAR	electronically agile radar
EAT	electronic angle tracking
EBS	electronic beam sharpening
ECCM	electronic counter-countermeasures
ECL	emitter-coupled logic
ECM	electronic countermeasures
EDC	electronic digital computer
eff	efficiency
EIA	extended interaction amplifier
EIKA	extended interaction klystron amplifier
ELINT	electronic intelligence
EMC	electromagnetic compatibility
EMCON	emission control
EMI	electromagnetic interference
ENR	energy-to-noise ratio
ENT	equivalent noise temperature
EOCCM	electro-optic counter-countermeasures
EPROM	electrically programmable read-only memory
ERP	effective radiated power
ERPD	effective radiated power density
ESJ	escort screening jammer
ESM	electronic (warfare) support measures
ESMR	electronically scanning microwave radiometer
EW	electronic warfare
EWR	early warning radar
FA	(1) frequency agility, (2) frequency adjustment
FB	feedback
FCS	feedback control system
FD	(1) frequency diversity, (2) field, (3) flux density, (4) focal distance, (5) frequency divided
fdbk	feedback
FET	field-effect transistor
FF	(1) far field, (2) flip-flop
FFH	fast frequency hopping
FFP	far-field pattern
FFT	fast Fourier transform
FH	frequency hopping
fil	filter
FIR	finite impulse response
FM	frequency modulation
FMCW	frequency-modulated CW
FOM	figure of merit
freq	frequency
FRUIT	false replies unsynchronized to interrogator transmission
FS	(1) forward scattering, (2) frequency shift
FSS	frequency-selective surface
FTC	fast time constant
FWHM	full width at half maximum
G	gain
GBP	gain-bandwidth product
GBR	ground-based radar
GC	gain control
GCA	ground-controlled approach
GCI	ground-controlled intercept
GEOSAR	geosynchronous synthetic-aperture radar
GHz	gigahertz
GMTI	ground moving-target detection
GN	Gaussian noise
GPR	Gaussian random process
GTA	gain-time control
H	(1) hardware, (2) homing, (3) horizon
HAR	harbor advisory radar
HDDR	high-density digital recording
hdw	hardware
HF	(1) high frequency, (2) height-finder
HFA	high-frequency amplifier
hgt	height
HIC	hybrid integrated circuit
HIPAR	high-power acquisition radar
HOJ	home-on-jam
HP	horizontal polarization
HPBW	half-power beamwidth
HPF	high-pass filter
HPRF	high pulse repetition frequency
HR	high-resolution
HRR	(1) high range resolution, (2) high-resolution radar
HTI	height-time indicator
HVAC	heating, ventilation and air conditioning
HVPS	high-voltage power supply
Hz	hertz
IAGC	instantaneous automatic gain control
IC	integrated circuit
ICR	integrated cancellation ratio
ICV	interclutter visibility
ID	identification
IF	intermediate frequency
IF amp	intermediate-frequency amplifier
IFF	identification, friend or foe
IFFN	identification, friend, foe or neutral
IFFT	inverse fast Fourier transform
IFM	instantaneous frequency measurement
IFMCW	interrupted FMCW
IFT	inverse Fourier transform
IIR	infinite impulse response
IISLS	improved interrogation sidelobe suppression
ISLS	interrogation sidelobe suppression
ILS	instrument landing system
IMPATT	impact avalanche-transit-time
int	interrogation
IPS	information processing system

I/Q	in-phase/quadrature
IR	(1) infrared, (2) impulse response, (3) interrogator-responder
ISAR	inverse synthetic aperture radar
ISL	integrated sidelobe level
ISR	interference-to-signal ratio
JAFF	jammer-illuminated chaff
JSR	jamming-to-signal ratio
KF	Kalman filter
KO	klystron oscillator
kW	kilowatt
LAS	low-altitude surveillance
LFM	linear frequency modulation
LHC	left-hand circular
LNA	(1) low-noise amplifier, (2) logarithmic narrowband amplifier
LNM	low-noise mixer
LO	local oscillator
log amp	logarithmic amplifier
LORAN	long-range navigation
LORO	lobe-on-receive-only
LP	linear polarization
LPAR	large phased-array radar
LPF	low-pass filter
LPI	low probability of intercept
LPRF	low pulse repetition frequency
LR	likelihood ratio
LRS	linear recursive sequence
LSB	least significant bit
LSIC	large-scale integrated circuit
LVA	(1) logarithmic video amplifier, (2) large vertical aperture (antenna)
MAG	maximum allowable gain
MAIR	modular airborne intercept radar
MBA	multibeam antenna
MC	microcircuit
MCG	master clock generator
MCW	modulated continuous wave
MDAC	multiplying digital-to-analog converter
MDL	minimum detectable level
MDS	minimum detectable (discernible) signal
mem	memory
MERA	molecular applications for radar electronics
MESFET	metal semiconductor field-effect transistor
MET-RRA	metal target reradiation
MEW	microwave early warning
MF	matched filter
MFAR	multifunction array radar
MFD	maximum frequency deviation
MFG	master frequency generator
MFMR	multifrequency microwave radiometer
MFR	multifunction radar
MFSK	multiple frequency-shift keying
MFWA	magnetron forward-wave amplifier
MHz	megahertz
MGC	manual gain control
MI	modulation index
MIC	microwave integrated circuit
MITATT	mixed tunneling and transit time
MLC	mainlobe clutter
MLD	maximum-likelihood detector
MLE	maximum-likelihood estimate
MLM	maximum-likelihood method
MLS	microwave landing system
MMW	millimeter waves
MO	master oscillator
MOD-FET	modulation-doped field-effect transistor
MOP	modulation on pulse
MOPA	master-oscillator-power-amplifier (transmitter)
MOSFET	metal oxide semiconductor field-effect transistor
MPRF	medium pulse repetition frequency
MS	main storage
MSA	(1) microstrip antenna, (2) Minimum-scattering antenna
MSLC	multiple sidelobe canceler
MSR	missile site radar
MSSR	monopulse secondary surveillance radar
MTBCF	mean time between critical failures
MTBF	mean time between failures
MTD	moving-target detector
MTI	moving-target indicator
MTR	missile track radar
MTTF	mean time to failure
MTTR	mean time to repair
mW	milliwatt
MW	megawatt
NDB	nondirectional (radio) beacon
NF	(1) noise factor [figure], (2) near field
NFP	near-field pattern
NPR	noise power ratio
NR	(1) noise ratio, (2) navigation radar
NSNR	normalized signal-to-noise ratio
O	oscillator
OA	omnirange antenna
OBA	off-boresight angle
OBC	onboard computer
ODL	optical delay line
op amp	operational amplifier
OPDAR	optical detection and ranging
OPINT	operational intelligence
OR	(1) operating range, (2) operational requirements
OSC, osc	oscillator
OSDR	oil slick detection radar
OTH	over-the-horizon
OWF	optimum working frequency
PA	(1) parametric amplifier, (2) power amplifier, (3) pulse amplifier
PANAR	panoramic radar
PAR	(1) precision approach radar, (2) phased array radar, (3) perimeter acquisition radar
par amp	parametric amplifier
PARROT	position adjustable range reference orientation transponder
PD	(pulsed doppler, (2) power divider
pdf	probability density function
PE	permanent echo
PLO	phase-locked oscillator
PM	(1) phase modulation, (2) preventive maintenance
PN	pseudonoise
PPF	pattern-propagation factor
PPI	plan position indicator

PPM	(1) pulse position modulation, (2) planned preventive maintenance, (3) parts per million
PR	(1) pseudorandom, (2) pulse rate
PFR	pulse repetition frequency
PRI	pulse repetition interval
PRN	pseudorandom noise
prob	probability
PRR	pulse repetition rate
PSD	(1) phase-sensitive detector, (2) power spectral density
PSK	phase-shift keying
PSNR	power signal-to-noise ratio
PSR	primary surveillance radar
PW	pulsewidth
PWD	pulsewidth discriminator
Q	quality factor
QAM	quadrature amplitude modulation
QF	quality factor
QI	quality index
QPD	quadrature phase detector
R&D	research and development
RADAM	radar detection of agitated metals
RADAN	radar doppler automatic navigation
radar	radio detection and ranging
RADINT	radiation intelligence
RAG	range-azimuth gate
RAM	radar-absorbent material
RAS	reliability, availability, serviceability
RASSR	reliable advanced solid-state radar
RATAN	radar and television aid to navigation
RBI	radar blip identification
RBW	receiver bandwidth
RCS	radar cross section
RCSR	radar cross-section reduction
RCVR	receiver
RDD	radar data display
RDF	radio direction finding
RECTENNA	rectifying antenna
ref	reference
REX	receiver-exciter
RF	(1) radio frequency, (2) range finder
RFA	radio-frequency amplifier
RFC	radio-frequency choke
RFI	radio-frequency interference
rg	range
RGPO	range-gate pull-off
RHC	right-hand circular
RHI	range-height indicator
RINT	radiation intelligence
RL	radiolocation
rms	root-mean-square
RMSE	root-mean-square error
RNF	receiver noise factor [figure]
RO	receive only
ROC	receiver operating characteristic
ROTHR	relocatable over-the-horizon radar
RPS	radar position symbol
RSR	(en)route surveillance radar
RT	real time
R/T	receive/transmit
RTC	receiver transfer characteristic
RTP	real-time processor
RTR	ready-to-receive
RV, rv	(1) random variable, (2) reentry vehicle
RWR	radar warning receiver
S	(1) search, (2) switch
SAR	synthetic-aperture radar
SAW	surface acoustic wave
SBD	Schottky-barrier diode
SBR	space-based radar
SC, sc	semiconductor
SCR	(1) signal-to-clutter ratio, (2) silicon controlled rectifier
SCV	subclutter visibility
seq	sequence
SDP	signal data processor
SHF	super-high frequency
SID	sudden ionospheric disturbance
sig	signal
SIGINT	signal intelligence
SIGSEC	signal security
sim	simulation, simulator
SINR	signal-to-interference noise ratio
SIR	signal-to-interference ratio
SJR	signal-to-jamming ratio
SLAR	side-looking airborne radar
SLB	sidelobe blanking
SLC	sidelobe cancellation
SLR	(1) sidelobe ratio, (2) sidelooking radar
SLS	sidelobe suppression
SNA	steerable null antenna
SNCR	signal-to-noise+clutter ratio
SNIR	signal-to-noise+interference ratio
SNJR	signal-to-noise+jamming ratio
SNR, S/N, snr	signal-to-noise ratio
SOJ	stand-off jammer
SPE	(1) seeker pointing error, (2) spherical probable error
spec	specification
SQNR	signal-to-quantization-noise ratio
sr	steradian
SS, ss	(1) spread spectrum (signal), (2) solid-state
SSJ	self-screening jammer
SSR	secondary surveillance radar
STALO	stable local oscillator
STC	sensitivity time control
STD	standard
STI	stationary target identification
STR	synchronous transmitter-receiver
SVA	small vertical aperture (antenna)
svc	service
SW	(1) software, (2) switch, (3) short wave
SWR	standing-wave ratio
SWS	search-while-scan
sync	synchronization, synchronizer
T	transformer, trigger
TACAN	tactical air navigation
TALAR	tactical landing approach radar
TAS	(1) track-and-search, (2) target acquisition system
TBWP	time-bandwidth product
TCR	target-to-clutter ratio
TD	(1) time delay, (2) tunnel diode
TDL	tapped delay line
TDWR	terminal doppler weather radar

TED	transferred-electron device
TEO	transferred-electron oscillator
TGT	target
TL	transmission line
TM	test mode
TOA	time of arrival
TOT	time-on-target
TR	(1) transmit/receive, (2) tracking radar
TRAP-ATT	trapped-plasma avalanche transit-time
TTG	test target generator
TTL	transistor-transistor logic
TUNNET	tunneling transit-time
TWA	traveling-wave amplifier
TWR	traveling-wave ratio
TWS	track-while-scan
TWT	traveling-wave tube
TWTA	traveling-wave-tube amplifier
TX	transmitter
UCON	upconverter
UHF	ultra-high frequency
ULSA	ultra-low-sidelobe antenna
UPS	uninterruptable power supply
UWB	ultrawideband
V	(1) volt, (2) voltage, (3) valve
VAD	velocity-azimuth display
var	variable
VCD	variable-capacitance diode
VCO	voltage-controlled oscillator
vel	velocity
VGPO	velocity-gate pull-off
VHF	very high frequency
VLA	very large array
VLB	very long baseline
VLDS	very large data store
VLSA	very-low-sidelobe antenna
VLSIC	very-large-scale integrated circuit
VOR	VHF omnidirectional range
VOR TAC	VOR-TACAN system
VOS	vertical obstacle sonar
VP	vertical polarization
VPS	variable phase shifter
VR	variable resistance
VSWR	voltage standing-wave ratio
VT	vacuum tube
VTM	voltage-tunable magnetron
WFM	waveform
W	(1) watt, (2) waveguide
WF	whitening filter
WFTA	Winograd-Fourier transform algorithm
WG, wg	waveguide
WL	wavelength
WPB	wide-pulse blanking
WV	(1) wave, (2) working voltage
WVL	wavelength
XCVR	transceiver
xform	transformer
xistor	transistor
XO	crystal oscillator
XPD	cross-polarization discrimination
X-POL	cross polarization
XSECT	cross section
XSTR	transistor
XTO	crystal oscillator
YIG	yttrium-iron-garnet
Z	(1) impedance, (2) azimuth
ZF	zero frequency