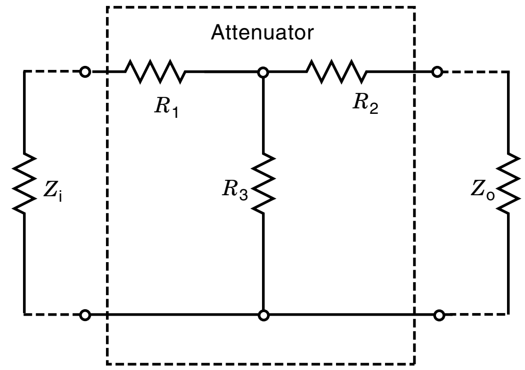


Figure 1. Resistor configuration for the "T" attenuator. Z_i and Z_o are the resistive impedances presented to the attenuator by external circuits.

ANTENNA ACCESSORIES

Accessories in antennas are devices for radiating or receiving electromagnetic waves. Their main purpose is to create means for the perfect operation of antennas. The accessories interconnect these antennas to the transmission and reception systems efficiently and safely. An electromagnetic coupling generally exists between the antennas and the accessories, and one can explore them to improve antenna performance. In this way, the accessories are as fundamental as the antenna itself, whose behavior depends on appropriate connections.

Among several kinds of accessories, the following are most commonly used: attenuators, baluns, chokes, coax lines, insulators, lightning arresters, planar structures, phase shifters, twin-lead. These components are described in the next sections.

ATTENUATORS

Attenuators are circuits designed to introduce a known loss between input and output ports (1). The power ratio, expressed in decibels, between the input and output represents the loss in these circuits. The main use of attenuators is in measuring standing wave ratio (SWR) in antennas and the transmission coefficient. In SWR measurement (2), one adjusts the attenuation value to maintain equal outputs in the stationary wave detector at the maximum and minimum points. Then the SWR in dB is equal to the difference between the readings of the attenuator.

The external circuits connected to the input and output ports of the attenuator should present purely resistive impedances. These are always matched to the input and output impedances of the component. Therefore, the resistors always constitute the attenuator circuit. In general, these resistive circuits use the "T" and "Π" circuit topologies. Figure 1 shows the "T" topology, where R_1, R_2, and R_3 form the circuit. The external circuitry presents purely resistive impedances (Z_i, Z_o) at the input and output of the attenuator. Figure 2 shows the configuration of the "Π" section. If the input and output impedances are the same ($Z_i = Z_o$), the circuits "T" and "Π" become symmetrical, that is, $R_1 = R_2$.

Design of an Attenuator

The designer has two concerns in designing an attenuator:

1. attenuation in dB
2. input and output impedances

The attenuation is given by Eq. (1):

$$\mathrm{ATT} = 10\log_{10} N \tag{1}$$

where ATT is attenuation in decibels and N is the ratio between the power absorbed by the circuit (from the generator) and the power delivered to the load.

The design starts from the desired input and output impedances and from the value of the ATT obtained in Eq. (1). Equations (2), (3), and (4) give the resistances of the "T" circuit:

$$R_3 = 2(NZ_iZ_o)^{1/2}/(N-1) \tag{2}$$

$$R_1 = Z_i[(N+1)/(N-1)] - R_3 \tag{3}$$

$$R_2 = Z_o[(N+1)/(N-1)] - R_3 \tag{4}$$

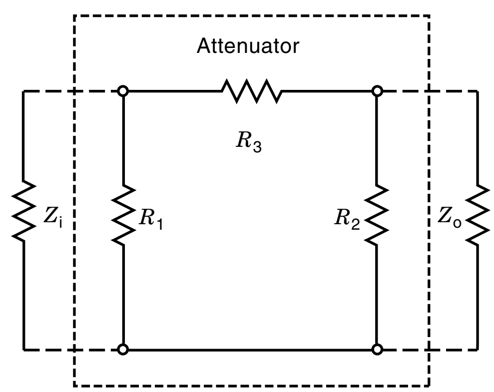


Figure 2. Resistor configuration for the "π" attenuator. Z_i and Z_o are resistive impedances presented to the attenuator by external circuits.

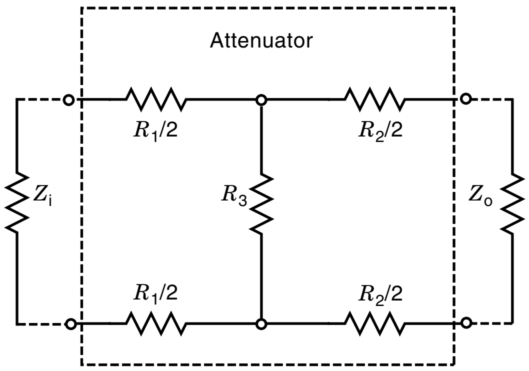


Figure 3. Balanced attenuator of the "H" type. The R_1 and R_2 resistors from the "T" circuit are distributed in the lower arms.

Equations (5), (6), and (7) give the resistances for a type "Π" circuit:

$$R_3 = 1/2(N-1)(Z_i Z_o/N)^{1/2} \quad (5)$$

$$1/R_1 = (1/Z_i)[(N+1)/(N-1)] - (1/R_3) \quad (6)$$

$$1/R_2 = (1/Z_o)[(N+1)/(N-1)] - (1/R_3) \quad (7)$$

Equations (2), (3), (4) and Eqs. (5), (6), (7) are also valid for symmetrical circuits (it is sufficient that $Z_i = Z_o$).

Attenuators "T" and "Π" are applied to unbalanced systems, such as coaxial cables. Bifilar lines feed many antennas that are intrinsically balanced, such as dipoles. In this case, the circuits "T" and "Π" change into their balanced versions and they are called "H" and "O" sections, respectively. Figure 3 shows the "H" circuit resistor configuration, and Fig. 4 illustrates the "O" circuit configuration. With respect to the circuits "T" and "Π," there is a distribution of the resistors R_3 and R_1, R_2 in the lower branch. All equations shown previously are valid for this case. However, it is necessary to take the half of the values of the respective resistors and distribute them in the upper and lower branches of their circuits.

Attenuators in Waveguides

Waveguides usually feed high-frequency antennas. In this case, (p. 262, Ref. 2) the attenuators are built starting from a very thin, tapered, resistive card. The attenuator device introduces this card in a section of the slotted guide. The adjustment of the penetration depth in the slot allows controlling the dissipation of power and the desired attenuation. This type of attenuator has a complex attenuation varying with penetration depth and frequency.

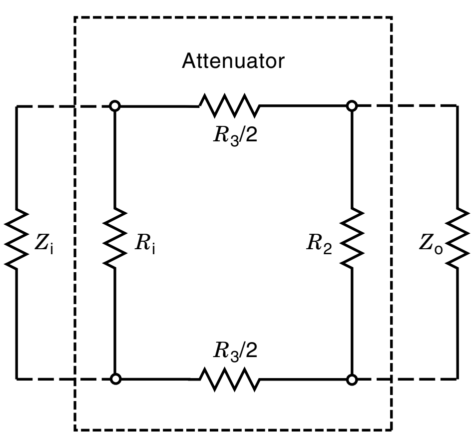


Figure 4. Balanced attenuator of the "O" type. The R_3 resistor from the "Π" circuit is distributed in the lower arm.

BALUNS

Although it is important, knowledge of the impedance value between two terminals is not enough to connect this impedance correctly to a transmission line because of couplings between the terminals and ground.

The equivalent circuit of Fig. 5 represents these couplings to ground. There are two special cases considered in detail (3). The first is when $Z_2 = Z_3$ in magnitude and phase. In this case, the impedance AB is balanced. Therefore, the voltages between A to ground and B to ground have the same magnitude and opposite phase.

The second case is when either Z_2 or Z_3 is zero. In this case, one of the sides is at ground potential, and the impedance is unbalanced. An example of a balanced line is a parallel wire line (twin-lead). The unbalanced lines are generally coaxial. Other types of lines exist, where the conductors present different couplings to the ground. An example of this is the twin lead which has conductors of different thickness. However, such a line is not common.

Because the conductors have different potentials with respect to the ground, the capacitance of the each conductor to the ground also differs. Therefore, the current in the two conductors can be different.

Antennas are generally designed for balanced or unbalanced input impedances. This simplifies the connection of the lines with the antenna. The connection to a symmetrical antenna requires a balanced line, whereas an antenna which has one of its feeding points at ground potential requires an unbalanced input.

A simple example of a balanced antenna is the dipole antenna shown in Fig. 6(a), and Fig. 6(b) illustrates a monopole, which is an unbalanced antenna. Figure 6(c) shows a dipole antenna with the feeding point between the center and the end (this is a procedure for achieving different input impedance values). Such an antenna has a current imbalance, which makes it impossible to feed it satisfactorily either by a balanced or unbalanced line.

Most frequently it is necessary to feed a balanced antenna with a coaxial cable and to feed (although less frequent) an unbalanced antenna with a balanced line. Such connections require special components to avoid problems with operation

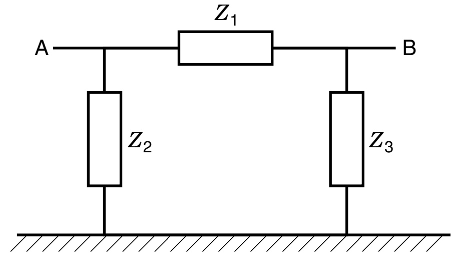


Figure 5. Equivalent circuit of an impedance AB.

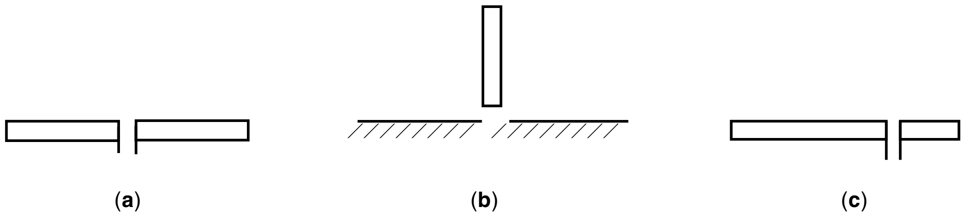


Figure 6. (a) Dipole antenna with symmetrical feeding. (b) Monopole antenna fed in the base. (c) Dipole antenna with asymmetrical feeding.

of the system (4). These devices are balanced-to-unbalanced converters, also known as baluns. The balun makes the voltages and/or currents in the two lines similar in magnitude.

If a coax cable is connected directly in a balanced antenna, currents are induced in the external part of the external mesh of the coax. This causes radiation of electromagnetic fields in unwanted directions. These currents cause an imbalance in the current distribution of the antenna. This affects the radiation diagram by altering the main lobe (sometimes drastically) and the gain of the antenna. For reception, interference signals can be induced in the external part of the coaxial cable and coupled inside the cable feeding the receiver.

The difficulties associated with the connection between an unbalanced and a balanced system can be understood by considering the coax line below a ground plane. Figure 7 illustrates the connection between a system and a parallel wire line. In this figure, Z_L is the load impedance and Z_S is the impedance associated with the support structures. The resulting currents in this system are equivalent to those created by ideal generator of Fig. 8.

The currents in lines A and B are not necessarily the same. The total current leaving point b flows into the line. However, the total current in b comes from line B and the connection with the ground. The main purpose of baluns is to ensure that the currents in lines A and B (Fig. 7) are similar.

Figure 9 illustrates a device capable of introducing symmetry in a line with respect to ground. Symmetry is essential in all kinds of baluns or balanced systems. Figure 10 shows the equivalent representation of this kind of balun. In this case $I_A = I_B$. Even so, if the length *FC* is small, the coax cable is almost short-circuited, and very little power is delivered to the load. To achieve satisfactory operation, the length *FC* should be of the order of a quarter wavelength ($\lambda/4$). Because of this restriction, the type of balun shown in Fig. 9 is inherently narrowband.

A possible solution to increasing the bandwidth is to wind the coaxial into a coil of length *FC*. This introduces a high impedance into the system. However, there are design limitations. One of them is in the lower frequency range. In this case, the impedance of the winding is small compared to the load (seen at the input of the twin-lead line). The upper limit is at the resonance point of the space of the windings. A technique used for increasing the inductive effect is to wind the cable in ferrite cores, which present a high impedance level on a wider frequency band. Careful designs of structures of this type allow their operation in frequency band ratios up to 10 : 1 (5).

The baluns most used are those based on quarter-wavelength short-circuited line sections. Such an arrangement is shown in Fig. 11. In this way, high impedance is obtained from point B to the external side of the outer mesh of the coaxial line. This prevents current flow into the outer surface of the external mesh of the cable.

In practice, the characteristic impedance of the transmission line is usually real, whereas that of the antenna element is complex. It is frequently desirable to operate the balun at a length other than a quarter-wavelength, in order to take advantage of the shunt reactance presented by the balun to the load, for impedance matching purposes.

Among many coupling matching networks that can be used to connect the transmission line to the antenna, we can introduce a quarter-wavelength transformer (see Fig. 23). If the impedance of the antenna is real, the transformer is attached directly to the load. However, if the antenna impedance is complex, the transformer is placed at a distance l away from the antenna. The distance l is chosen so that the impedance towards the load, at l, is real.

If the transformer is a 2-wire line section, and the line is a coaxial cable, obviously the coaxial line cannot be connected directly to the 2-wire line. In this case, a balun may be used to connect the transformer to the coaxial line.

RADIO-FREQUENCY CHOKES

A radio-frequency choke coil is an inductance designed to offer a high impedance to alternating currents over the frequency range for which the coil is to be used. This result is obtained by making the inductance of the coil high and the distributed capacitance low. The result is that the inductance

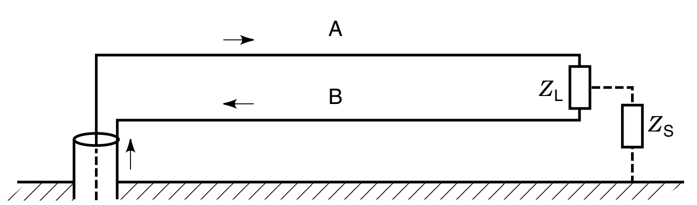


Figure 7. Unbalanced line connected to a balanced line.

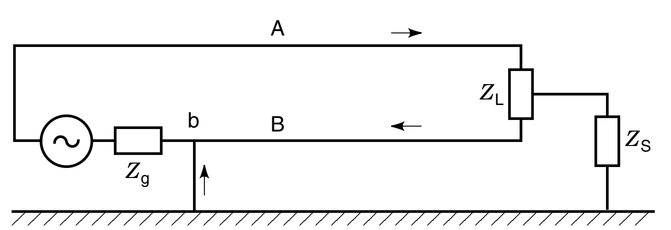


Figure 8. Idealized generator of the circuit of the Fig. 7.

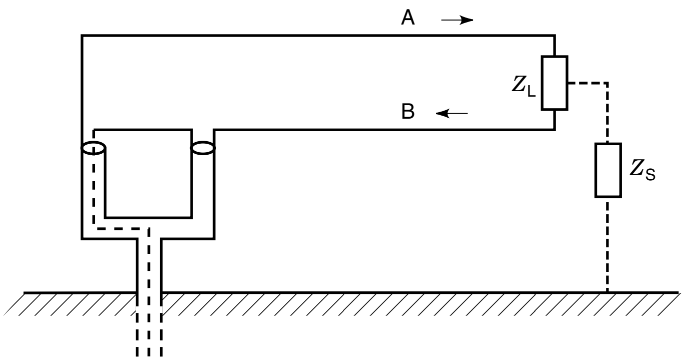


Figure 9. Connection by a balun.

is in parallel resonance with the distributed capacitance somewhere in the desired operating frequency range.

A typical radio-frequency choke coil consists of one or more universally wound coils mounted on an insulating rod or of a series of pies wound in deep narrow slots in a slotted bobbin. A long single-layer solenoid is sometimes used.

Proper use of slug-type magnetic cores improves the performance of radio-frequency chokes. These cores increase the inductance and hence the impedance of the coil without materially affecting the distributed capacitance.

COAXIAL CABLE

Coaxial cable is often used in antenna engineering to connect a transmitter to an antenna, particularly at high frequencies largely because of convenient construction and practically perfect shielding between fields inside and outside the line. The range of impedances obtained most conveniently by coaxial lines is from 30 to 100 Ω. However, because the cable is an unbalanced structure, its connection to a dipole needs a transformer called a "balun." Figure 12 shows its configuration with the dimensions of interest. "d" is the internal conductor diameter, and "b" the outer conductor diameter. Its characteristic impedance is given by

$$Z_0 = \left(60/\sqrt{\epsilon_r}\right) \ln(b/d)\,\Omega \quad (8)$$

where ϵ_r is the relative permittivity of the medium.

The attenuation α_c of a coaxial line due to ohmic losses in the conductor is given by

$$\alpha_c = (4.343 R_s/Z_0\pi)(1/b + 1/d) \quad \text{dB/m} \quad (9)$$

where $R_s = \sqrt{\pi f \mu/\sigma}\ \Omega$ where f is measured in Hz, μ in H/m, and σ is the conductivity in S/m.

The attenuation α_d of a coaxial line caused by losses in the dielectric is given by

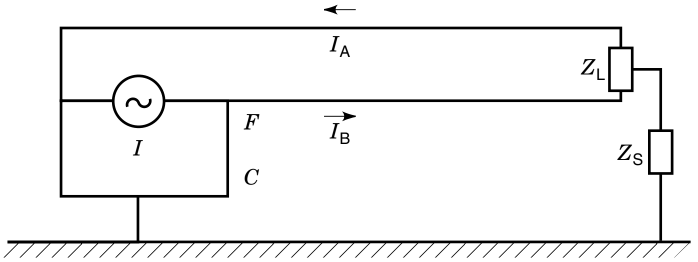


Figure 10. Equivalent circuit of Fig. 9.

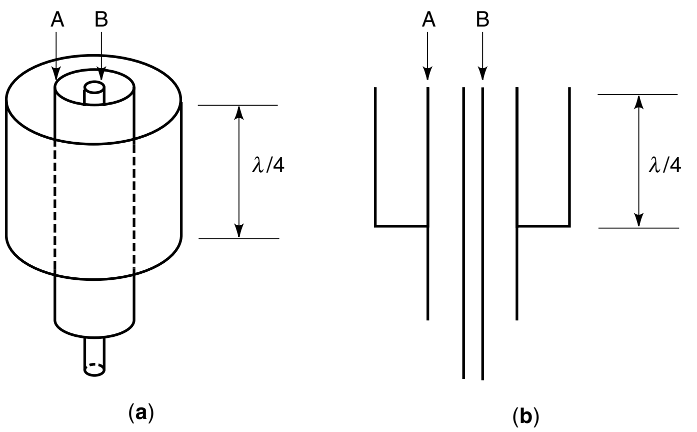


Figure 11. (a) Balun bazooka. (b) Cross section of the balun.

$$\alpha_d = 27.3\sigma Z_0/\ln(b/d) \quad \text{dB/m} \quad (10)$$

Breakdown occurs in an air-filled coaxial line at atmospheric pressure when the maximum electric field E_m reaches a value of approximately 2.9×10^6 V/m. The average power P that is transmitted in a matched coaxial line under these conditions is given by

$$P = (E_m^2/480) b^2 [\ln(b/d)/(b/d)^2] \quad \text{W} \quad (11)$$

The first high-order transversal electric (TE) mode that propagates in a coaxial cable has its cutoff when the average circumference is about equal to its wavelength, so

$$\lambda_c \approx \pi/2(b+d) \quad (12)$$

INSULATORS

A material is an insulator if this is its most dominant characteristic. This does not mean that other electric effects, such as conductivity and magnetism, are not present, but they are just less significant. From a macroscopic point of view, the constitutive relationship of the material, specifically in the case of insulators (dielectrics), is given by

$$\boldsymbol{D} = \epsilon \boldsymbol{E} \quad (13)$$

where $\boldsymbol{E}$ is the electric field intensity vector and $\boldsymbol{D}$ is the electric flux density vector.

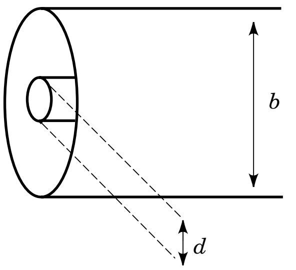


Figure 12. Geometry of a coaxial cable. A dielectric of relative permittivity ϵ_r fills the space between the inner and outer conductors.

From the microscopic point of view, it is better to use the relationship introduced by Lorentz (6), where two terms contribute to the vector density of electric flow. The first is related to the vector electric field by the vacuum permittivity, and the second is called vector polarization $\boldsymbol{P}$. Mathematically,

$$\boldsymbol{D} = \epsilon_0 \boldsymbol{E} + \boldsymbol{P} \tag{14}$$

where $\epsilon_0 = 8.856 \times 10^{-12}$ F/m.

The contributions of the material to the behavior of vector $\boldsymbol{P}$ are different for solids, liquids, and gases. These are very complicated to summarized here [for more details, see Ref. (6)]. For linear materials, however, $\boldsymbol{P}$ is directly proportional to $\boldsymbol{E}$. Therefore

$$\boldsymbol{P} = \epsilon_0 \chi_e \boldsymbol{E} \tag{15}$$

where χ_e is the electric susceptibility to the medium.

Substituting Eq. (15) in Eq. (14), one returns to Eq. (13), because

$$\epsilon = \epsilon_0 (1 + \chi_e) \tag{16}$$

Therefore there is no difference between this point of view and that of the constitutive relationship.

The term that defines the loss of energy in Poynting's theorem has the vector current density (which considers the conduction current) in phase with the intensity of the vector electric field. In the general characterization of dielectrics with losses, the complex permittivity is introduced and is given by

$$\epsilon = \epsilon' - j\epsilon'' \tag{17}$$

This is introduced in Maxwell's equations as the current $j\omega(\epsilon' - j\epsilon'')\boldsymbol{E}$, where the second term is in phase with the electric field. Both ϵ' and ϵ'' are, generally functions of frequency. Table 1 lists values of ϵ'/ϵ_0 and ϵ''/ϵ' for some representative materials used in antenna engineering. It also lists the breakdown field for some materials.

Table 1. Characteristics of Insulating Materials[a]

Material Composition	ϵ'/ϵ_0 at 10^8 Hz	ϵ''/ϵ' at 10^8 Hz	Breakdown field at 25°C (V/m)
	Ceramics		
Aluminum oxide (alumina)	8.8	0.00030	–
Porcelain (dry process)	5.04	0.0078	–
	Plastics		
Polyethylene	2.26	0.00020	47.2×10^6
Polytetrafluorethylene (Teflon)	2.1	<0.00020	39.4×10^6
	Adhesives		
Epoxy resin (Araldite CN-501)	3.35	0.034	15.9×10^6
	Glass		
Fused quartz	3.78	0.00020	16.1×10^6

[a] Data from Ref. (1).

LIGHTNING ARRESTERS

Lightning is an atmospheric phenomenon with potentially harmful consequences. It is caused by the accumulation of electric charges in a cloud and the consequent discharge to the terrestrial surface. This effect occurs mainly on structures that offer favorable conditions to discharge. The best protection against atmospheric discharges is the enclosure by a grounded conductive structure. Even so, such protection when not expensive, sometimes becomes impracticable.

Antennas must be immune to atmospheric discharges and maintain integrity as lightning current flows to the ground. Protection against the direct and induced effects of lightning has to be designed to avoid damage to operators, equipment, and structures. For antennas, several techniques can be used, such as

- protective conductors on a radome (7) (without degrading the performance of the antenna)
- grounding the towers (always positioning the antennas below the tower summit)
- lightning arresters (mainly in HF operation, where the antennas are relatively large structures)

Common types of lightning arresters are varistors, gas discharge devices, and semiconductors. No one type is suitable for all applications, and each may be combined with another into a hybrid device. A simple lightning arrester can be assembled by creating a gap between the structure to be protected and the ground. These gaps are adjusted to minimum width such that no arcing occurs when the transmitter is operating.

PHASE SHIFTERS

A phase shifter is a two-port component, which provides a fixed or variable change in the phase of the traveling wave. The shift is with respect to "reference" (the line without the component) and "test" (the line and the component) lengths. Therefore it is always understood as the phase difference between the two. The shift may be fixed or variable. The variable phase shift uses mechanical or electronic techniques to change the phase dynamically. The main uses of these devices are as testing systems, measurement systems, modulation devices, and phased array antennas. The use of phase shifters in antenna systems provides controllable steering of the main beam of the radiation pattern without moving the antenna. There are several ways to implement these components for communication systems. The type of phase shifter used in this process depends on the kind of antenna used, its costs, and power requirements.

Types of Phase Shifters

Shifters can provide fixed or variable phase shifts. Fixed-phase shifters are usually extra transmission line sections of certain lengths to shift the phase with respect to the reference line. Variable shifters use mechanical or electronic means to achieve a dynamical range of phase difference. In antenna systems, the main use of variable phase shifters is in beam steering, whereas fixed shifters are used more for beam forming (pattern synthesis).

A mechanically tuned phase shifter usually consists of variable shorts used with hybrids or, in the case of waveguide components, dielectric slabs with variable positions in the guide. A step motor moves the slab across the guide (from its center toward outer walls), therefore accomplishing maximum or minimum phase shift. Another way of obtaining the desired mechanically tuned phase shift is by combining variable shorts and hybrids. The movement of the short circuit along a transmission line results in the phase shift, therefore making it appear shorter or longer.

There are two classes of electronically tuned phase shifters depending on component cost, weight, and power handling capability: ferrite and switching circuit shifters. The ferrite shifter uses slabs of this magnetic material inside waveguides. Because the electrical properties of ferrite change depending on an applied bias magnetic field, this process yields the desired phase shift. Ferrite phase shifters are essentially low-cost, bulky, devices with high power handling capability. Switching-circuit shifters are essentially low-power *pin* diodes or FETs used as switching (ON–OFF states) elements. Depending on the applied bias current, these devices work either as short or open circuits. The phase shift is the result of an extra section of transmission line added to the switching element. Therefore, depending on the bias current, the wave traveling along the transmission line has an additional traveling path. Because these devices are binary switches, only discrete phase shifts are possible. The construction of a continuously variable shifter demands the use of several of these switching elements. The design of variable phase shifters depends on the expression of the variable shift with respect to a specific position or applied bias magnetic field or current. In all cases, the procedures depend on the topology of the device and its mathematical model.

Ferrite Phase Shifters

Ferrite phase shifters are common in power applications in waveguide technology. The ferrite phase shifter is the result of combining two properties of this medium: the dependency of the velocity of the propagating waves on the direction of propagation and the application of a bias magnetic field. Ferrite is an anisotropic medium. Therefore, the propagation constant (and the velocity of propagation) depends on the bias applied to the sample. The anisotropic properties of ferrites also depend on whether a magnetic field is applied. This is a macroscopic result of the microscopic interaction of the fields with atomic nuclei.

The phase shift in these devices is the difference between forward and backward traveling waves. This is a nonreciprocal phase shift, which depends on an applied external magnetic field. Accurate design of these devices is not straightforward because of the complexity of the mathematical models of ferrites. A detailed procedure is given by Pozar (8) and Bornemann (9).

Switching-Circuit Phase Shifters. The switching-circuit phase shifter uses solid state technology (*pin* diodes). In these kinds of applications, *pin* diodes behave as ON–OFF elements. The switching-circuit shifters are common to several manufacturing technologies, such as microstrip, coplanar guides and lines, slotline, waveguides, and finlines. The diodes act as binary switches along the line (Fig. 13). With the bias applied, the diode behaves as a short circuit for the incoming wave. Therefore, the phase-transmitted signal has two possible states, determined by which diode is polarized. A variable phase shift results from mounting several diodes on the line. The returning wave enters a hybrid and then the antenna.

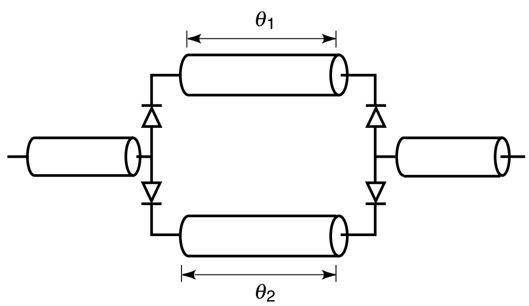


Figure 13. PIN switching-circuit phase-shifter representation. The application of the bias current results in a change of the electric length of the line. The resulting shift is the difference between the electrical angle in the direct and reverse polarized cases.

PLANAR TECHNOLOGY

Planar transmission lines are the basic transmission media for microwave integrated circuits when hybrid or monolithic technology is used. Figure 14 shows the cross sections of three basic planar transmission lines: stripline, microstrip, and slotline. Several other configurations are possible, such as listed in Ref. (10). All configurations are variations of the three main types. Each offers certain advantages, depending on the circuit type and its application. Some of these configurations are the suspended stripline, the suspended microstrip, the inverted microstrip, the coplanar guide, and the coplanar stripline. Planar designs using special substrates have been applied at frequencies up to 100 GHz.

A designer can produce several microwaves circuits and electric functions with planar transmission lines. Therefore, planar technology constitutes a very important technology for developing antenna accessories in the microwave range. These components, called distributed circuits, are generally of the order of a wavelength or more (in the propagation medium). Therefore, in the lower microwave range, these circuits can have large dimensions. In this case, one uses lumped components.

Several lumped components can be obtained starting from planar technology. These components are called microwave integrated circuits (MIC). Figure 15 illustrates lumped and distributed circuits in planar technology.

Following are the most important design parameters for planar transmission lines:

- effective dielectric constant
- characteristic impedance
- dispersion

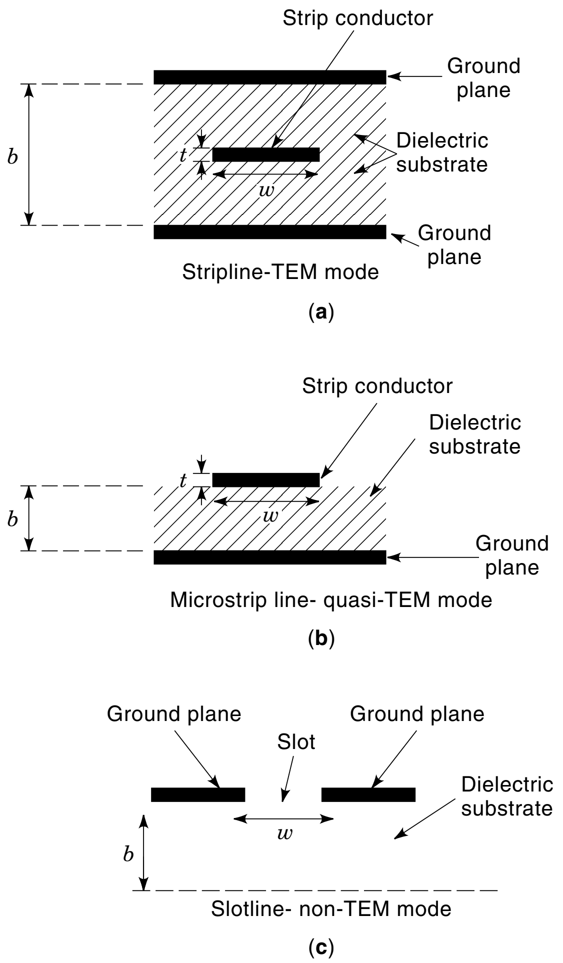


Figure 14. Basic planar transmission lines. (a) Stripling—TEM mode. (b) Microstrip line—quasi-TEM mode. (c) Slotline—non-TEM mode.

- losses
- working frequency limitations

Several antenna accessories are manufactured for planar technology. Examples are directional couplers, hybrid rings, power dividers, and impedance matchers (11,12). These components are used as matching, feeding, and sampling networks in antenna systems.

The effect of the electromagnetic coupling between two parallel transmission lines (sufficiently near each other) constitutes the design framework of several microwave devices (especially for antennas). An instance is the directional coupler, which has wide application in circuits and telecommunication systems. It performs several electric functions, such as separating and combining signal power (received or delivered from/to antennas). The directional coupler also allows sampling power and separating incident and reflected waves of a system. This separation is used for power control or performance measurements.

Couplers are either conductively or electromagnetically coupled transmission line circuits with three, four, or more ports. In general, a directional coupler is a four-port device. Two of its ports are mutually decoupled with respect to the other two. Figure 16 shows the block diagram of a directional coupler. From a theoretical point of view, if port 1 receives a microwave signal, this power is delivered to ports 2 and 3, but no power appears in port 4. Port 4 is decoupled from port 1. On the other hand, if port 2 receives a microwave signal, then this signal is divided between ports 1 and 4, and port 3 is decoupled relative to port 2.

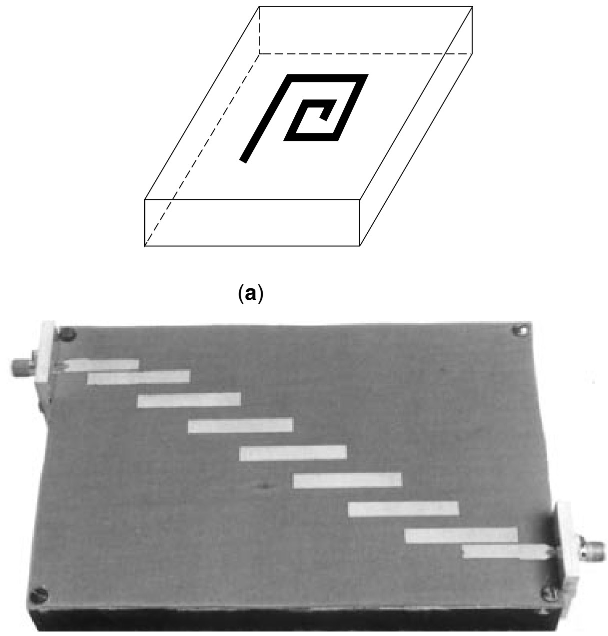


Figure 15. Lumped and distributed MICs. (a) Spiral inductor (lumped element). (b) Band-pass interdigital filter (distributed MIC) (courtesy of the Department of Electrical Engineering of the University of Brasília).

In the special case where power is equally divided between ports 2 and 3, the coupler is called a hybrid or 3 dB directional coupler. If the phase difference between ports 2 and 3 is 90°, the coupler is called a 90° hybrid or quadrature hybrid. Another hybrid design provides 180° between the output ports.

Different topologies of directional couplers exist (11,12). Figure 17 shows a directional coupler in the topology of parallel lines of $\lambda/4$ length. As in any directional coupler, a fraction of the incident power in the primary branch is coupled to the secondary branch. In fact, when a microwave signal is incident in port 1, port 2 receives part of its power. In this topology, another part of the power is coupled to port 3. This coupling happens through the gap, and practically no power is coupled to port 4, which is terminated by the load. The load has the same value as the characteristic impedance of the coupler lines. In parallel line couplers, the smaller the gap between branches of the coupler, the greater the coupling.

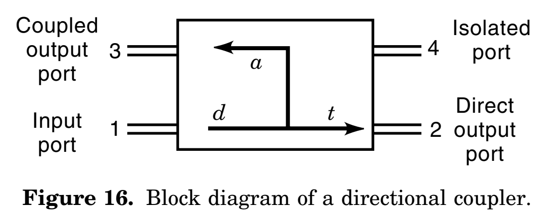


Figure 16. Block diagram of a directional coupler.

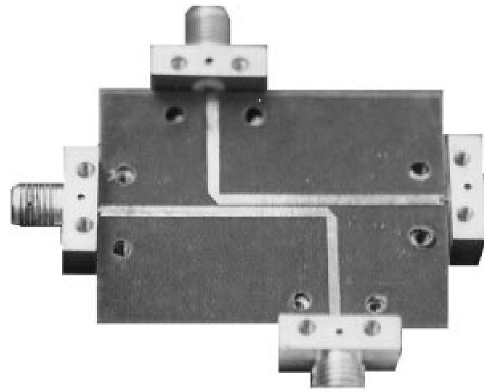


Figure 17. Directional couplers in λ/4 parallel lines in microstrip technology (courtesy of the Department of Electrical Engineering of the University of Brasília).

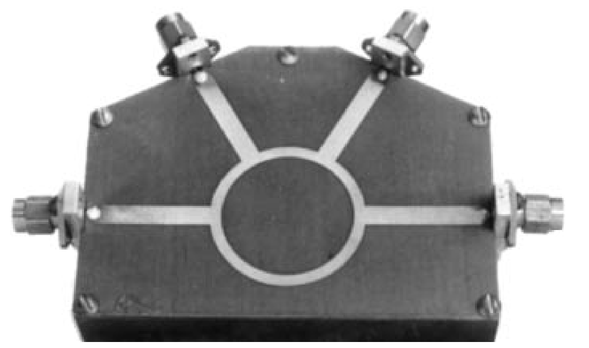
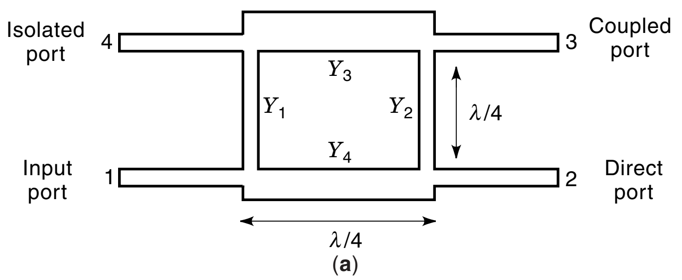


Figure 19. Hybrid ring (courtesy of the Department of Electrical Engineering of the University of Brasília).

The spacing between conductors can compromise high coupling designs.

The main parameters of a directional coupler are the coupling, directivity, isolation, and transmission factors. The coupling factor is defined by the relationship between the power in the coupled (port 3) and input (port 1) ports. This factor is a measurement of the coupling. The transmission factor is defined by the relationship between the power in ports 2 and 1. The directivity is the relationship between the power in the isolated (port 4) and coupled (port 3) ports. This measures the undesirable coupling. The isolation is the relationship between power in the isolated (port 4) and input (port 1) ports.



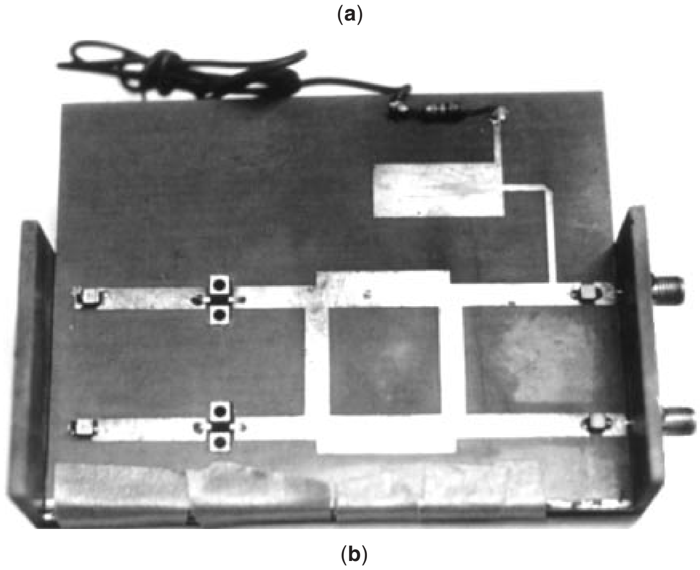


Figure 18. (a) Branch line coupler structure. (b) Branch line coupler in a phase-shifter circuit (courtesy of the Department of Electrical Engineering of the University of Brasília).

Figure 18(a) shows a coupler in branch line topology (11,12). A larger coupling factor can be obtained in this directional coupler. It also allows keeping dc continuity between the ports, which is an interesting feature for phase-shifter design [as shown in Fig. 18(b)]. It allows high power levels. The branch line coupler uses branching transmission lines to couple 3 dB to 9 dB. The lengths of all branches of this coupler are λ/4. However, the widths and consequently the impedances of these branches are different: Y_1, Y_2, Y_3, Y_4. The operating principle of this coupler can be presented in the following way: when the input port (port 1) receives a signal, the energy is divided and it propagates clockwise and counterclockwise through the branches. The power in each output (ports 2, 3, and 4) depends on the phase relationships between the propagating signals. In a 3 dB coupler, power is divided equally between the coupled (port 2) and matched (port 3) ports. Port 4 is the isolated port.

A special version of the branch line coupler is the hybrid ring (Fig. 19). The ring has a 1.5 λ circumference and all sections of the same impedance. It can be assembled on several technologies, such as coaxial line, stripline, microstrip line, or even waveguide. The operating principle is same as that of the branch line coupler. The signal injected in the input port is divided into two propagating waves with opposite directions. In the coupled output port, the two signals arrive in phase and reinforce each other, providing an output signal. In the direct output port, the signal that propagates counterclockwise travels a length of 1.25 λ, and the signal propagating clockwise travels 0.25 λ. The two signals arrive at this port in phase, providing an output signal at this port. For the same reason, there is an isolated port, which has no output

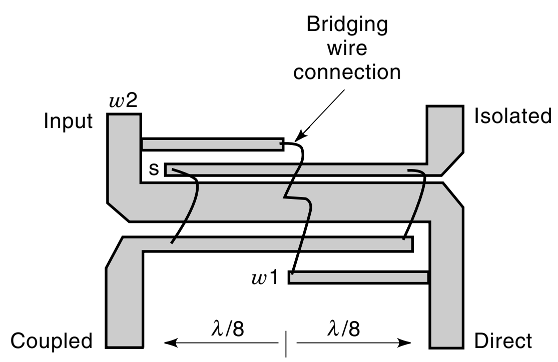


Figure 20. Lange coupler.

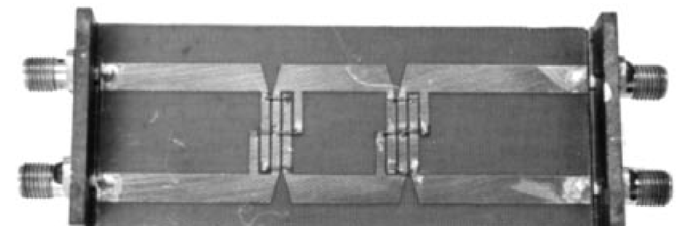


Figure 21. Lange coupler in tandem topology (courtesy of the Department of Electrical Engineering of the University of Brasília).

signal. The output signals in the direct and coupled ports are 180° out of phase.

Another topology of the directional coupler is the Lange coupler (Fig. 20) which is well adapted to microstrip technology (13). It is an interdigital structure with superior performance compared with the parallel line coupler. Its bandwidth can be greater than one octave. The Lange coupler associated with tandem topology (Fig. 21) allows obtaining power coupling near 3 dB with a combination of two lower order couplers, whose design dimensions are feasible.

Power dividers are used in feeds and performance measurements of antennas. Directional couplers and hybrid rings can be used as power dividers. The branch line coupler output signals are in phase quadrature (90° out of phase). Output signals of hybrid rings are in opposite phases. However, in several microwave applications, such as parallel feeds for phased array antennas, the input has to be divided into an arbitrary number of signals with equal power and in phase. A symmetrical power divider of n branches provides successive division of the input signal into n signals with the same power and phase at all frequencies. This power divider can also be used as a signal combiner by inverting the input and output ports.

There are several kinds of power dividers/combiners. One of them is the two-branch power divider, shown in Fig. 22. It is also possible to obtain output signals in phase but with different power levels.

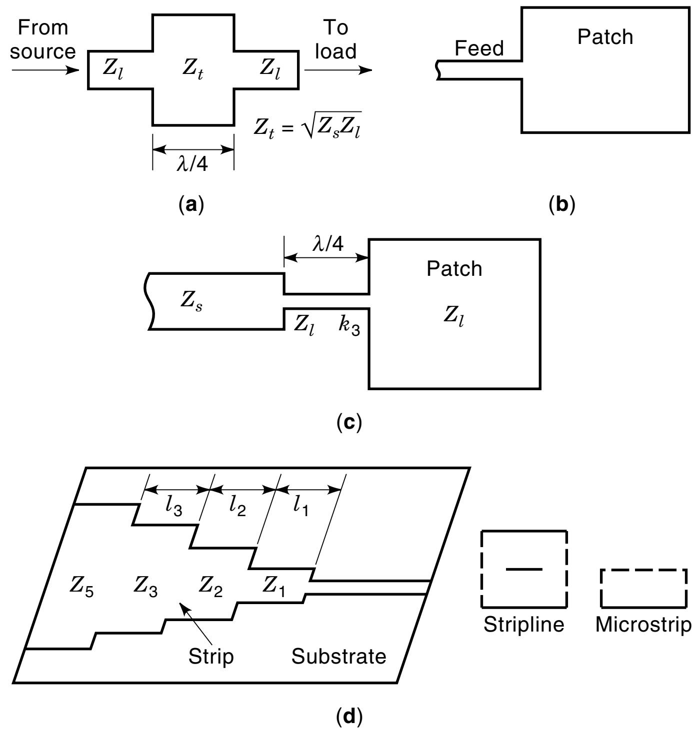


Figure 23. (a) Quarter-wave impedance transformer, (b) top view of a patch antenna, (c) microstrip edge feed with quarter-wave transformer, (d) multisection quarter-wave transformer.

Impedance matching circuits are another kind of antenna accessory. They are also manufactured in planar technology (2). These circuits are commonly used in antenna feeds, avoiding power losses caused by mismatch. In most cases, the load has an impedance different from the transmission line. This is the case in telecommunications systems, where the output impedance of the generator is usually different from the input impedance of the antenna. The problem is how to reduce or eliminate the resulting reflections and high SWRs.

One method is to cascade transmission line sections (planar structures) or lumped components with different impedances between the mismatched transmission line and its terminations (the system output and the antenna input). Then the antenna input impedance is transformed. This new impedance value seen at the output point (the end of the transmission line) can be matched to the system for maximum power transfer.

Different topologies and technologies are used in impedance matching circuits. Some of these are stub association and quarter-wave transformers. Figure 23(a) shows a schematic of a quarter-wave impedance transformer. The impedance Z_l can be matched to a transmission line of characteristic impedance Z_s with a section of transmission line that is a quarter-wave long based on the wavelength in the transmission line. This characteristic impedance of the matching section is shown as $Z_t = \sqrt{Z_s Z_l}$. The microstrip feed on Fig. 23(b) is planar, allowing the patch and the feed to be printed on a single metallization layer. The impedance of the edge-fed patch can be transformed by using a quarter-wave matching

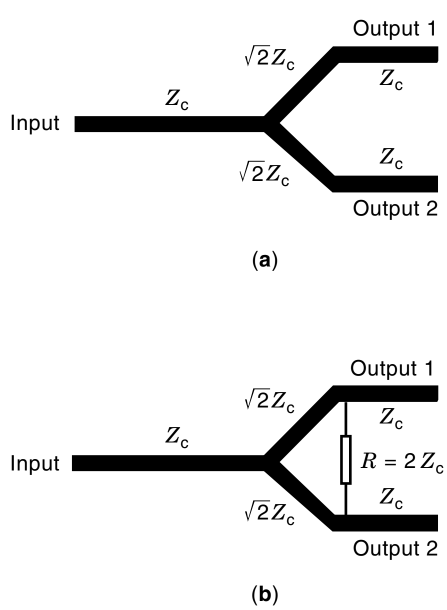


Figure 22. Unmatched (a) and matched (b) two-way power dividers.

section of microstrip transmission line, as shown in Figs. 23(a) and 23(c). Another kind of quarter-wave transformer is the multisection quarter-wave transformer, as shown in Fig. 23(d). This kind of transformer is commonly used in wideband systems.

TWIN LEAD

The bifilar line (twin-lead) is a two-wire parallel conductor line, which carries power from the generator to the antenna. Each wire carries equal and opposite currents (180° out of phase). Because the wires are spaced by a certain distance, however, their radiated fields do not cancel out completely. If the two wires were in the same place in space, there would be no radiated field. However, because this is not possible, there will be a certain amount of radiation loss. Keeping the distance between the parallel wires small (typically on the order of 1% of the wavelength of the radio wave) reduces this loss. The spacing between conductors is a constraint of the physical limitations of line construction. Parallel conductor lines, open-wire lines, two-wire lines, two-wire cable and two-wire ribbon cables are also bifilar lines.

Types of Bifilar Lines

The bifilar lines are usually supported a fixed distance apart. Insulating rods, called spacers, or molded plastics, such as polystyrene, provide the support. The spacers have to be placed at short intervals to prevent the wires from moving apart from each other. The lines are air-insulated, because this is the medium between them. The spacers have little effect on the impedance behavior of the line. Flexible dielectrically separated (molded) lines have several advantages over air-insulated types. In this type, a plastic coating similar to a ribbon surrounds the conductors. These lines maintain uniform spacing between conductors. They are also less bulky, weigh less, and are easier to install. One common example of a flexible dielectric bifilar lines is the television receiving cable (called ribbon cable or two-wire cable).

Uses of Bifilar Lines

In television reception applications, polyethylene molded ribbon bifilar lines are available with spaces of 2.54 mm (1 in.) and 1.27 mm (1/2 in.) and conductor sizes of AWG 18 (diameter of 2.053 mm). These lines have characteristic impedance of 450 Ω and 300 Ω, respectively. The attenuation is quite low for these receiving applications (typically under 0.03 dB/m for applications under 20 m). There is also a 75 Ω bifilar line, which has AWG 12 conductors (diameter of 1.024 mm) and close spacing between them (see Ref. 14 for a detailed description). The spacing keeps most of the field confined to the solid dielectric instead of the surrounding air.

Bifilar lines are also used in amateur radio applications and LF to VHF antennas, although for higher frequency applications, the coaxial line is used more because of its superior characteristics. The lines are simple to assemble. However, the designer should avoid sharp bends. Bends change the characteristic impedance of the lines, which causes reflections and power mismatch at each bend.

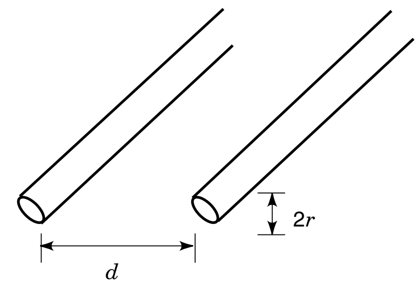


Figure 24. The geometry of bifilar lines. In the molded plastic casing, the dielectric surrounding the lines has the shape of a ribbon. Because most of the field is inside the ribbon, the line is less susceptible to external influences.

Design Equations of Bifilar Lines

The analysis of bifilar transmission lines uses Maxwell equations subject to the appropriate boundary conditions. Following are the main parameters of bifilar transmission lines:

Characteristic impedance

$$Z_0 = \frac{\eta}{\pi}\cosh^{-1}\left(\frac{d}{2r}\right) = \frac{120}{\sqrt{\epsilon_r}}\cosh^{-1}\left(\frac{d}{2r}\right)\,\Omega \tag{18}$$

where r is the radius of the conductors, d is the spacing between them, and ϵ_r is the relative dielectric of the surrounding medium (Fig. 24).

Attenuation

$$A(dB/m) = 8.686\left\{\frac{1}{\pi r}\sqrt{\frac{\omega\mu}{\sigma}}\left[\frac{d/2r}{\sqrt{(d/2r)^2-1}}\right] + \frac{\pi}{\lambda}\left(\frac{\epsilon''}{\epsilon'}\right)\right\} \tag{19}$$

where μ is the permeability of the surrounding medium, ω is the operating frequency, σ is the loss in the conductors, and ϵ''/ϵ' is the relative loss of the dielectric constant.

Propagation constant

$$\beta = \omega/v \tag{20}$$

where v is the phase velocity in the medium surrounding the conductors. In a ribbon transmission line, ϵ_r is 2.56 with $(\epsilon''/\epsilon') = 0.7$. If the wire is made of copper, then σ is 58 MS/m. In this case, because most of the field is inside the dielectric, the relative dielectric constant ϵ_r is approximately 2.56.

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ANTENNA, APERTURE. See APERTURE ANTENNAS.