POWER COMBINERS AND DIVIDERS

Since the birth of radio over one hundred years ago (in 1895), there has been a very rapid growth of radio communication technology. To obtain higher information transmission rates over larger distances, there has been a steady demand for higher frequency bands and higher levels of transmitted power. The most significant progress in these two directions took place during World War II and afterward. This led to the emergence and expansion of what is now termed microwave (1 GHz to 30 GHz frequency band) and millimeter wave (30 GHz to 300 GHz frequency band) communications technologies. Important systems that operate now at microwave or millimeter waves include terrestrial links, satellite links, and radar. These systems use active and passive circuits, which fulfill a number of fundamental functions related to signal generation, modulation, launching, reception, and detection.

Prior to 1948, the year the transistor was invented, the only semiconductor element which was used in microwave communication circuits was a crystal detector diode. At that time, microwave signal generation and amplification were dominated by vacuum tubes including magnetron, klystron, and traveling wave tubes (TWTs) (1). These devices still remain in use at kW levels where solid state devices cannot yet compete. Since the late 1960s the situation with the dominance of microwave power generation by tubes has changed dramatically. This was due to the introduction of microwave semiconductor devices such as the Gunn diode, avalanche diodes, and bipolar and field-effect transistors (1), which were able to produce considerable power levels and gain at microwave frequencies. With the invention and introduction of these new devices, there has been growing interest in developing techniques to obtain higher power levels by combining power from individual modules. The motivation for such techniques has been due to a number of reasons. For example, in the mid-1970s it became clear that the combination of even a modest number of IMPATT (impact avalanche transit time) diodes could lead to the manufacture of a reasonable pulsed radar transmitter. This transmitter offered a reduction in size and weight and it required only tens of volts for its operation. This was in contrast to tubes that often required 1000 V supplies and considerable warm-up time. Since the inception of the concept of power combining, in the late 1960s, many power-combining schemes have been tried and developed. These schemes often include the two tasks of power division and combination and therefore the terms *power combining* and *power dividing* often appear simultaneously.

Power combining schemes involve different technologies and media. These range from metal waveguides and cavities, microstrip, and striplines for which the power is confined to a finite region. Alternatively, open space configurations are used. In that case, power is launched by a set of radiating elements, antennas, at the transmit side and is intercepted by another set of antennas at the receive side. The choice of technology and medium depends on the type of solid state device involved in the power combining scheme, operational frequency and bandwidth, and the type of application in which the combined power is required. The devices involved in power combining/dividing are known as power combiners/dividers. Because of their reciprocal character, the same circuits can alternatively be viewed as power combiners or power dividers. This depends on the nomination of their input and output ports.

The review of power combining/dividing structures is the subject of the presentation in this article. The presentation commences with an introduction of categories of power combiners followed by the fundamental characteristics of these structures. Methods of analysis as well as applications are also described.

MICROWAVE AND MILLIMETER-WAVE POWER COMBINING TECHNIQUES

Power Combining Principles

Power combining can be achieved in a variety of manners, but they can be generalized to two situations, which are shown in Fig. 1.

In Fig. 1(a), there are a number of individual oscillators and the task is to combine them into a more powerful synchronized signal source. This is achieved via a suitable combining circuit. This circuit with some modifications can be

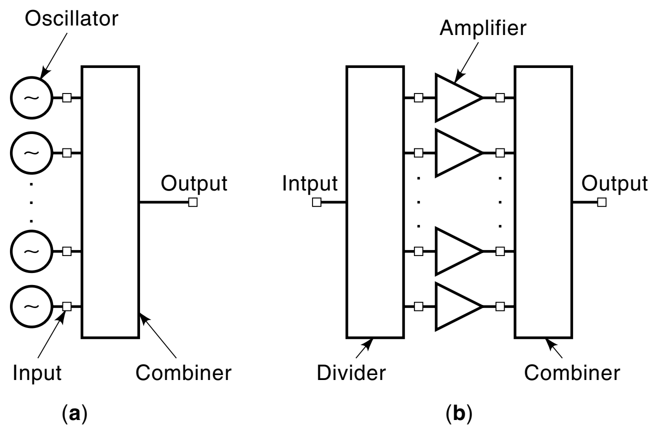

Figure 1. General configurations for power combining. (a) Oscillators. (b) Amplifiers.

used for injection-locked oscillators and reflection amplifiers. In that case, the output port is used as an input port for injection of an external signal.

In Fig. 1(b), there is a single low-power source that is split and applied to the inputs of amplifiers. In turn, the outputs of these amplifiers are combined into a common port to produce a more powerful source. In this case, the association of both dividing and combining circuits is used to obtain the power combination.

At this stage, the requirements on the combining structure have not yet been specified. In general, these requirements are related to the proper functioning of the entire combiner as well as to the efficient utilization of individual oscillators or amplifiers that appear in the combiner, which is explained as follows. First of all, the combined source has to function in a reliable manner. This means that it has to have stable and long lasting performance. The term *stability* may also refer to the purity of the frequency spectrum generated by the combined devices. The combining circuit should be able to synchronize individual sources and suppress any instability that can lead to multimode operation of multiple solid-state devices in the combiner. Reliability is often associated with graceful degradation of performance of the combiner. It means that if one of the oscillators or amplifiers fails, the remaining devices should not be affected. Consequently, the power produced by the combiner should be only slightly reduced due to the absence of the failed unit. Graceful degradation has always been regarded as the most important aspect of the power combining techniques in the competition of solid-state sources with tubes, for which failure is usually catastrophic. One additional aspect, in conjunction with graceful degradation, is feasibility of the “hot” replacement of a failed unit. A suitable provision in the power combining/dividing structure is necessary to accomplish this task.

Classification of Power Combiners/Dividers

A systematic grouping of power combiners has been accomplished by Russel (2) and Chang and Sun (3). As a result of this classification, three major categories of combiners have been introduced: chip-level, circuit-level, and spatial-level. In addition to these three major groups, multiple-level and other types of combining schemes have also been identified (3). This classification is not unique. However, it helps to recognize combiner/divider configurations in the context of their use. Chip-level combiners (4–7) can be viewed as a subcategory of circuit-level combiners. They utilize very small distances (a small fraction of a wavelength) between individual solid-state devices. As such, although they increase the generated power, they do not offer graceful degradation. This is because the failure of one chip usually leads to the destruction or prevention of operation of the remaining chips and thus to the catastrophic failure of the combined source. Other problems with this type of combiners are low effective impedance and thermal interactions. Standard circuit-level combiners utilize larger distances between individual solid-state devices and therefore they allow for the introduction of counter measures against a catastrophic failure of the combiner. Because of the present marginal importance of chip-level combiners, Fig. 2 shows the reduced classification of combiners, which include only circuit-level and space-level combiner categories.

Circuit-type combiners (2,3) confine power within a region often enclosed by a conducting surface such as a cavity or a waveguide. Radiation into a free space is considered an undesired phenomenon as it leads to power losses. In space combiners, power combining takes place in an unbounded (8) or partially bounded region (9). In the latter case, the combining region often forms a space resonator. The combining process is accomplished with radiative elements (antennas). The relative dimensions, in terms of wavelength, of the spatial combiner structure are usually much larger than of its circuit-type equivalent. Consequently, the quality factor which is achieved with space resonators is usually much larger than that obtained with circuit-type resonant cavities. Because of this feature, circuit type combiners are employed at the lower end of microwave and millimeter wave frequencies while the space-type combiners are aimed for use at the upper millimeter-wave frequency bands.

As can be seen in Fig. 2, circuit-level combiners can be divided into resonant and nonresonant combiners. Resonant combiners use resonant cavities which exhibit a high quality factor, Q and as a result feature small operational bandwidth. Due to this property, they serve the purpose of synchronizing multiple oscillators. Nonresonant combiners utilize structures with low quality factor, and therefore they exhibit an increased operational bandwidth.

At this stage it has to be mentioned that often similar-in-shape structures may appear in both resonant and nonresonant types of combiners (2,3). For example, a cylindrical cavity with N-coaxial ports including IMPATT or Gunn diode oscillators with a central port for power extraction is considered a resonant cavity combiner (10). On the other hand, the same cylindrical cavity, but with peripheral coaxial probes and a central probe, can also be considered a nonresonant, N-way radial combiner (11,12). The difference lies in the Q factor that the cylindrical cavities provide (13). The division into resonant and nonresonant cavity combiners may be less apparent for those structures for which it is difficult to identify whether or not they are resonant in operation. The proper identification of the Q factor may require solving an electromagnetic field problem to fully characterize a given structure. This is not always feasible.

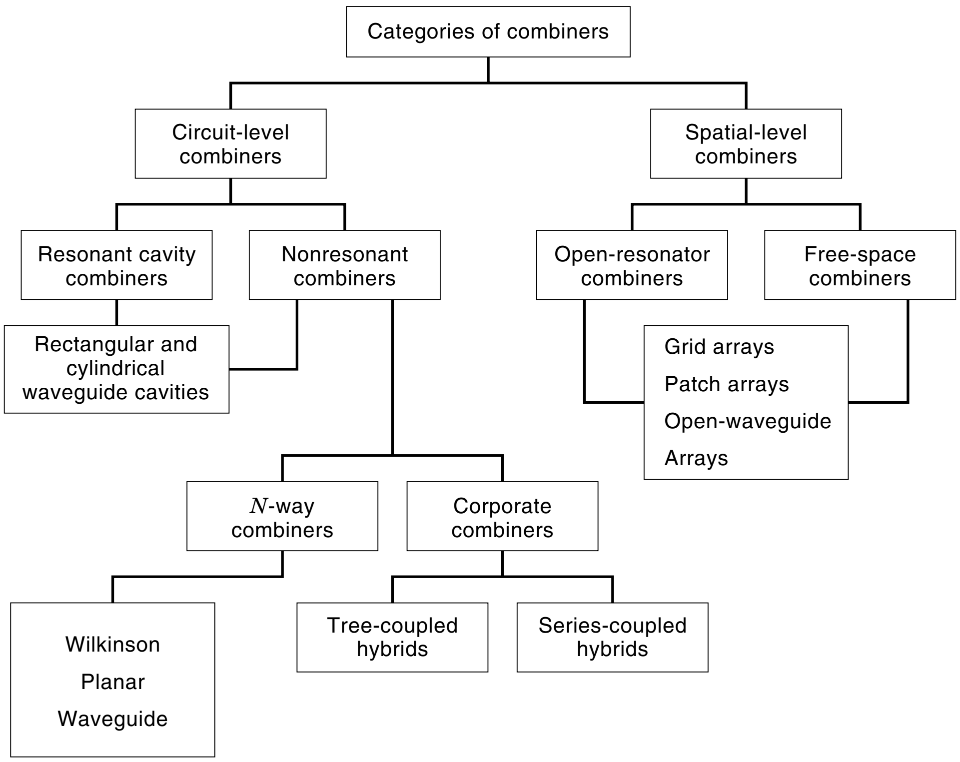

Figure 2. Categories of power combiners.

Nonresonant combiners are generally divided into N-way and corporate-type combiners. Again, this classification is not unique and conforms to some earlier accepted terminology (2,3). N-way combiners utilize a single structure with N input ports and one output port. Corporate combiners use a tree of M-way combiners, with M being a small number, to create a combiner with a large number of input ports. The case of $M = 2$ leads to the binary-type corporate combiner. As has been indicated earlier, all the nonresonant combiners discussed here can also be regarded as power dividers. This is accomplished by designating input ports of the N-way combiner as output ports of the divider, and by renaming the common output port of the combiner to the input port of the N-way divider.

Similar to circuit-level combiners, spatial combiners can be divided into resonant (open-resonator cavity) and nonresonant structures (14). Due to their similarities with Fabry–Perot cavity lasers (1), resonant-type spatial combiners using a space resonator with front and back reflection mirrors are also named quasi-optical power combiners (9,14).

RESONANT AND NONRESONANT CAVITY COMBINERS/DIVIDERS

For the purpose of power combining and dividing, rectangular and cylindrical cavities are often used. Both resonant and nonresonant power combining structures can employ such cavities. As has been explained earlier, the identification of the resonant or nonresonant nature of the application can be distinguished by the quality factor. This factor depends on the method of launching and extracting the power and therefore on the cavity loading conditions. Resonant and nonresonant combiners/dividers using cavities are described in the following sections.

Rectangular Cavity Combiners/Dividers

The most popular configurations of rectangular cavities, which have been used for the purpose of power combining, are shown in Fig. 3. The first structure, shown in Fig. 3(a), is

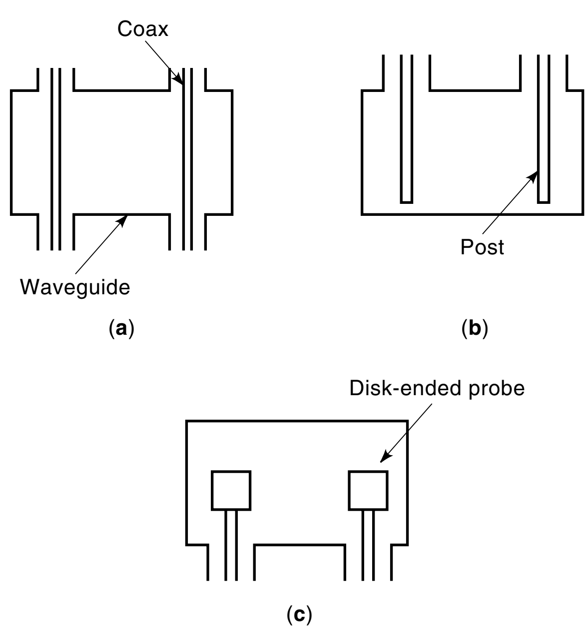

Figure 3. Rectangular waveguide cavity combiner/dividers. (a) cross-coupled coaxial lines. (b) Posts. (c) Disk-ended probes.

known as the Kurokawa rectangular waveguide cavity combiner (15,16). It is considered as a resonant-type of combiner. In this combiner, solid-state devices (IMPATT or Gunn diodes), operating as oscillators or amplifiers, are located in the lower part of the bottom coaxial lines. The upper coaxial lines are loaded with an absorber to stabilize the operation of the multiple-oscillator structure as well as to provide bias to solid state devices. Please not that for the proper operation, pairs of coaxial lines are located close to waveguide walls, at the maximum of the magnetic field, and are spaced approximately at a half-waveguide wavelength. The waveguide short circuit is located a quarter-wavelength from the first adjacent coaxial pair. The bottom coaxial lines often include impedance-transforming circuits to obtain proper impedance conditions for the solid-state devices.

Figure 3(b) shows one variation of the basic Kurokawa combiner (17,18). In this case, the bottom parts of the coaxial lines are removed and replaced by open-ended posts. In this configuration, solid-state devices are located in the gaps between the bottom waveguide floor and the tips of the posts. The rectangular waveguide in this configuration can be of a standard or oversized format.

Figure 3(c) shows a rectangular waveguide with probes (19). In contrast to the Kurokawa-type combiner, individual amplifiers including active devices (diodes or transistors) are connected externally to this structure via coaxial ports. These in turn couple power to the common rectangular cavity. This basic configuration with some modifications, depending on the method of launching and extracting power, is used in ladder-type (19,20) and traveling-wave type combiners/dividers (21). Coaxial probes or waveguide apertures can be used as combining ports. To combine power, reflective (single unit) and transmissive (using dividing and combining units) configurations are explored.

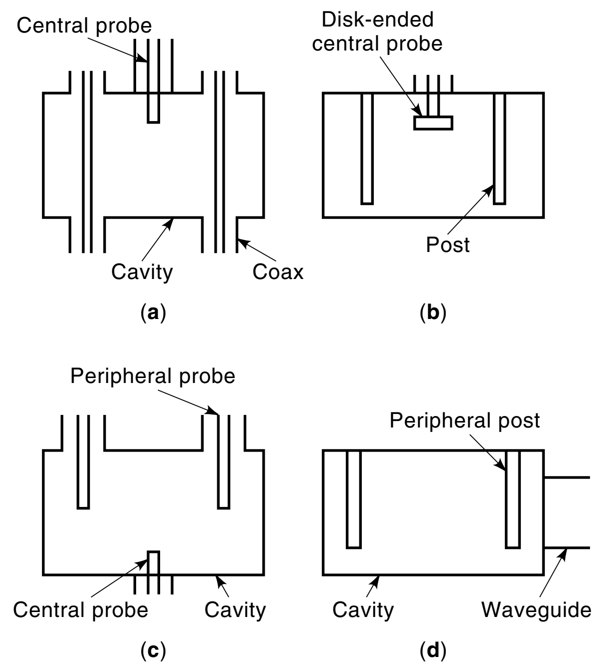

Figure 4. Cylindrical cavity combiner/divider configurations. (a) Cross-coupled coaxial lines (10,11,22) and a central coaxial probe. (b) Posts and disk-ended coaxial central probe (23,24). (c) Coaxial probes (25,26). (d) Posts and a waveguide aperture (27,28).

Cylindrical Cavity Combiners/Dividers

The principles of the use of cylindrical cavities for the purpose of power combining or dividing are analogous to those already presented for rectangular cavities. This is because both types belong to the same family of parallel-plate waveguide structures. The obvious differences are in the cross-sectional shapes, which consequently lead to their different use. Figure 4 shows typical configurations of combiners/dividers that use a cylindrical cavity as a combining structure.

For the purpose of combining, it is usually assumed that the cavity operates in its TM_{0N0} mode whose field is constant with the cavity height and axially (azimuthally) symmetric. Power is extracted or launched by the centrally located probe, Fig. 4(a) to (c), or by the rectangular waveguide port, Fig. 4(d). An important feature of the cylindrical cavity combiner/divider is that, since the mode of operation assumes fields azimuthally symmetric, in theory there is no minimum spacing between peripheral probes or coaxial lines. This is in contrast to the rectangular cavity combiner/divider in which the peripheral ports have to be spaced at half waveguide wavelengths for its proper operation.

Radial, Conical and Hemispherical Cavity Combiners/Dividers

The cylindrical cavity with symmetrically located peripheral coaxial probes or lines and a single central probe, already described above, can also be regarded as a radial cavity combiner/divider. This designation, instead of cylindrical cavity combiner, has been used in a number of references. Variations include shaping of the radial cavity as well as different types of central and peripheral ports. Standard designs concern uniform height radial cavity, which is fed by the central probe. However, some designs include a nonuniform height cavity (29,30). At the periphery, rectangular waveguide ports or coaxial probes are used. For the central port, full-waveguide-height coaxial probe (31,32) disk-ended (33,34), conical (35,36,37) and dielectric coated probes (38) are used. Similarly, peripheral probes can be formed by full-height coaxial probes (29,30), dielectric coated probes (37), conical probes (36), disk-ended probes (34), balun loops (39), continuous annular balun sections (35) and rectangular waveguide ports (33,38), instead of straight coaxial probes. The variations of radial-cavity combiner/dividers are shown in Fig. 5.

Conical (2,3) and hemispherical cavity combiner/dividers (40,41) shown in Fig. 6 exhibit axial symmetry and they utilize posts or coaxial probes for connecting solid-state sources or amplifiers and for power launching and extracting.

NONRESONANT *N*-WAY COMBINERS

Cavity-type Nonresonant *N*-way Combiners

N-way combiners form a single structure with *N* input ports and one output port. Some examples of these types of combiners have already been discussed and include radial, hemispherical and conical cavities. In these structures, an azimuthally symmetric field is used for the power combining/dividing purposes. In this case, the device operation is equivalent to *N* parallel transmission lines connected to one port. To support this mode of operation, input ports are excited in-

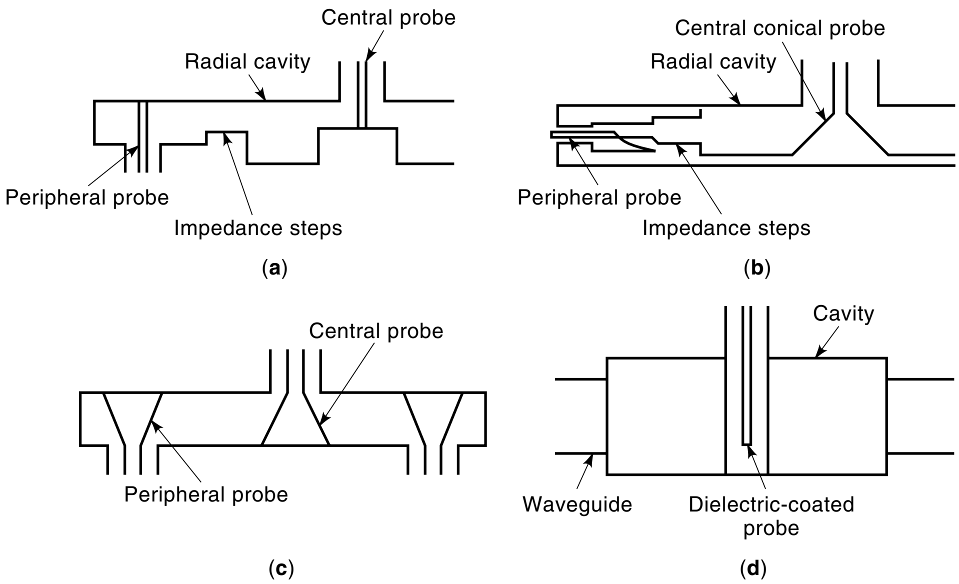

Figure 5. Variations of radial-cavity combiners/dividers.

phase. Since in practice, the arriving signals may have different phases, dielectric beads or rods (40) of variable length or height may be used for tuning purposes. Out-of-phase excitation of ports leads to the generation of higher order modes which feature azimuthal dependence. To suppress these higher order asymmetric modes, slits (24,26) and resistive absorbers (24,25) are used. These perturbation elements also help to increase the isolation between peripheral ports, which in turn increases the graceful degradation performance of the combiner, which is explained as follows.

In general, failure of a source, or an amplifier connected to a given port, results in a change of the load presented to this port. This new condition affects the performance of the remaining active devices. The extent of this influence depends on the isolation between the failed port and the remaining ports. Radially symmetric cavity combiners have a built-in isolation that usually increases with the number of ports, and is equal to $1/N$ on average. This value may be considered insufficient in some applications and this is why the resistive elements are introduced to increase isolation. When the slits or resistive vanes are located along the lines of the surface electric current produced by the dominant symmetric cavity mode their effect on the fundamental mode is negligible. However, they cut lines of the currents that are associated with higher-order azimuthally varying fields. This results in absorption of higher-order modes and consequently improves the isolation between the peripheral ports.

An equivalent circuit for the radial (also hemispherical and conical) combiner under the condition of the dominant (symmetric, in-phase) mode of operation is shown in Fig. 7.

As can be seen, N transmission lines of equal length (not necessarily quarter wavelength), which meet at a common port, form this circuit. These lines are nonuniform as their characteristic impedance varies with distance. In this case, they operate as impedance tapers and provide suitable impedance transformation between the peripheral ports and the central port. This is to counter the parallel loading effect at the common port. The peripheral ends are connected via resistors. These resistors represent mutual coupling between the lines. Also they may incorporate isolation resistors.

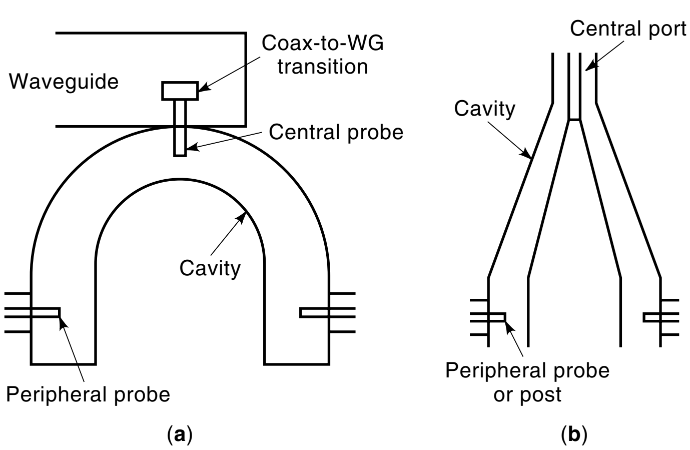

Figure 6. (a) Hemispherical. (b) Conical cavity combiner/divider configurations.

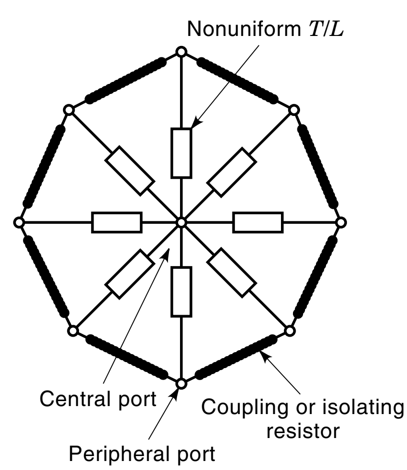

Figure 7. An equivalent circuit for a radial combiner under assumption of the dominant mode of operation.

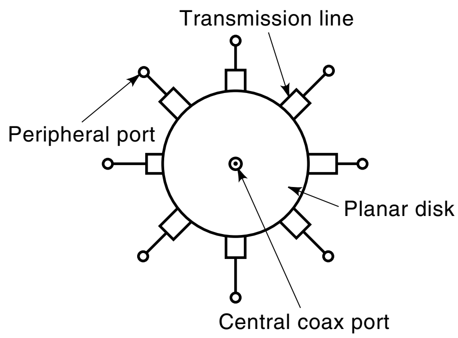

Figure 8. Planar version of a radial combiner with coaxial central port and microstrip transmission lines.

Nonresonant Planar Radial-Line Combiners

Radial combiners do not necessarily have to be realized in waveguide. Their planar equivalents have also been investigated (42–47). In these cases, a planar disk with a coaxial line at its center and N microstrip lines connected at its periphery form the combiner structure. Two varieties of the planar radial combiner are shown in Fig. 8 below.

As can be seen, depending on the design, nonuniform (tapered) or uniform microstrip lines are used. Additionally, resistors can be connected between the neighboring lines to improve isolation (42,43,46) and hence the graceful degradation performance of this device.

Wilkinson *N*-Way Combiners

Wilkinson's N-way combiner is one of the oldest power combining structures. It was introduced by Wilkinson in 1960 (48). The device resembles in its concept the radial-line combiner and is synthesized using M sections of N uncoupled transmission lines of equal length with isolation resistors. A similar concept was also followed by Yee et al. (49). The equivalent circuit of the N-way Wilkinson divider is shown in Fig. 9.

The difference, in comparison with an N-way radial line combiner, is that in the present case the lines are uniform and are quarter-wavelength long. To meet the impedance matching conditions, the characteristic impedance of the lines is equal to $\sqrt{N}\,Z_0$, where Z_0 is the characteristic impedance of input and output transmission lines. Isolation between peripheral ports is improved using resistors between adjacent lines. The most well known version of the Wilkinson's divider is the two-way divider, as shown in Fig. 9(b). It is used as a basic building block in a corporate type combiner, which is discussed later.

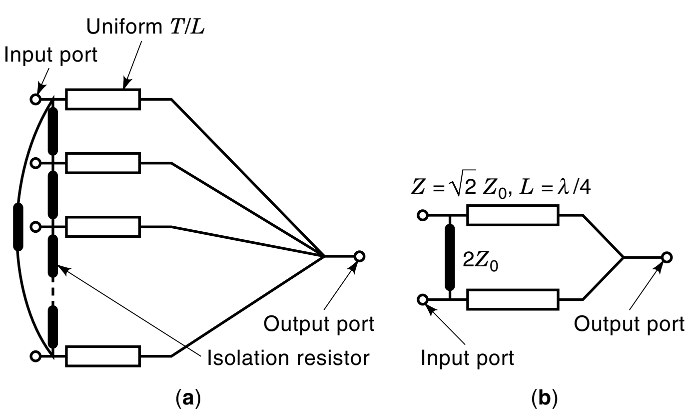

Figure 9. (a) An equivalent circuit of the Wilkinson's N-way power combiner. (b) Its two-way version.

Generalizations or variations of the standard Wilkinson divider were presented in Refs. 50, 51, 52, and 53. In particular, the synthesis of N-way dividers using M-sections of N-wire uncoupled or coupled transmission lines was described in Ref. 50. To avoid multilayer construction of the standard Wilkinson N-way divider, for $N > 2$, new planar solutions were presented in Ref. 51. Practical designs of four-way and seven-way dividers with isolation exceeding 20 dB between ports over a fractional bandwidth of 40% were demonstrated. The new structures presented in Ref. 52 feature an increased power handling capability. The work in Ref. 53 concerns the design procedures for unequal power-split tee junctions.

Sectorial Radial Combiners

One problem with the N-way planar radial combiner is that it uses hybrid (coaxial and planar) technology for its construction. This shortcoming can be avoided using a sectorial M-way planar combiner (54). Variations of this structure are shown in Fig. 10.

Used as dividers, these structures can provide approximately equal power split with equal phase distribution. To increase the isolation between the peripheral ports, lumped resistors, as in the case of Wilkinson's type combiner, can be used. In order to make output ports parallel as well as to

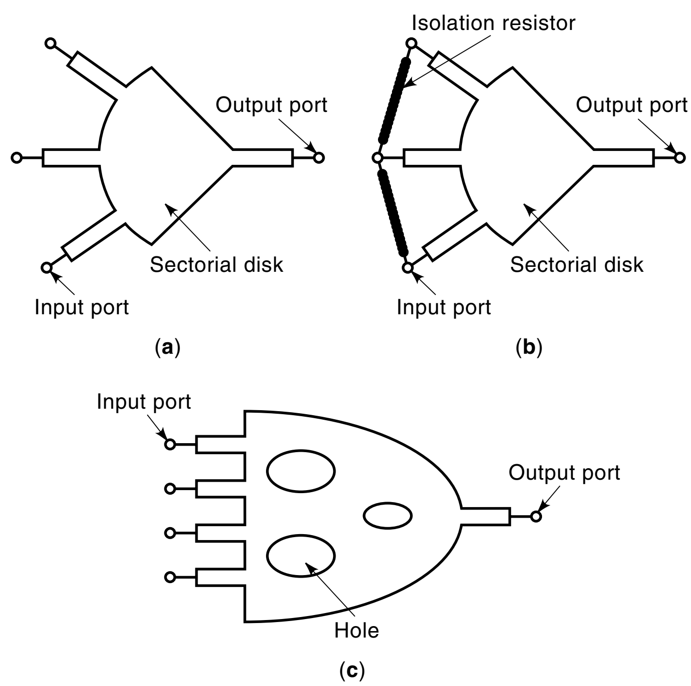

Figure 10. Sectorial planar combiners. (a) Without isolation resistors, (b) With isolation resistors. (c) With holes for phase and amplitude equalization.

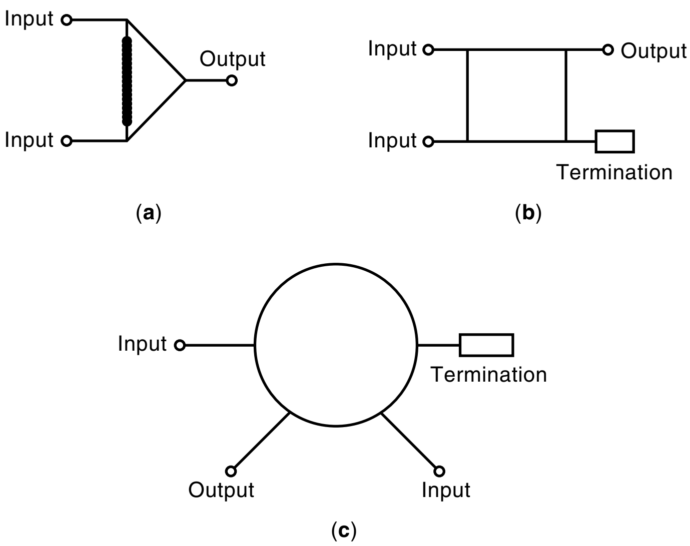

Figure 11. Basic blocks used in tree-type power combiners (a) Wilkinson 2-way divider. (b) 3 dB branch-line coupler. (c) Rat-race hybrid.

provide equal phase division, exponential taper and purposely created holes can be used (55,56,57).

Corporate Combiners

Combiners with a large number of ports can be built using a tree of identical or similar hybrid devices with a small number of input/output ports. These combiners can include binary or generally M-ary basic building blocks. As a result, the number of input ports in such combiners is not arbitrary but given by 2^L or M^L, where L is the number of section in the longitudinal direction. Basic building blocks for the binary tree-type combiners are shown in Fig. 11.

These basic structures include the two-way Wilkinson's combiner, the 3 dB quadrature hybrid and the rat-race hybrid (Ref. 58, Khoul or equiv.). Other possibilities include other types of couplers (for example coupled-line directional coupler) and the Gysel power combiner (52,59). These hybrids can be built in waveguide or planar (for example microstrip) technologies. These structures have in-built isolation between their input ports, exceeding 20 dB, and feature an increased operational bandwidth, typically 20% or more.

The common property of the Wilkinson's, 3 dB quadrature, and rat-race hybrids is that they can provide equal power division when they operate as dividers. Please note, however, that alternative designs of these dividers can also offer unequal power division (53). The differences between these building blocks concern the number of ports and phase relationshps. The Wilkinson's hybrid is a three-port device and provides equal in-phase power split. The rat-race hybrid is a four-port device and is capable of in-phase and out-of-phase (180° phase difference) equal power split. The 3 dB branch-coupler is a four-port device and provides equal power split with a 90° phase difference. Because of their four-port arrangement, the subtracting port of the rat-race hybrid and the isolated port of the 3 dB coupler need be match-terminated prior to their use in the tree-type combiners.

Figure 12 shows examples of corporate (tree-type) combiners with two-way Wilkinson's combiners and 3 dB couplers.

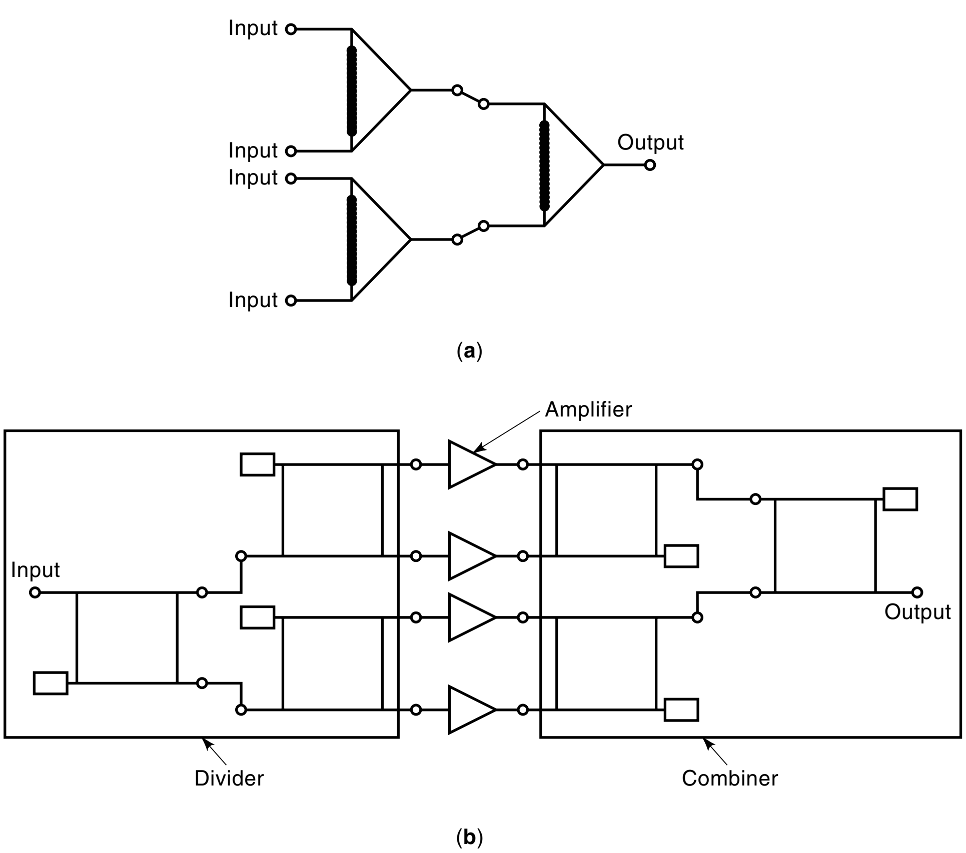

Figure 12. Typical configurations of corporate power combiners. (a) Wilkinson combiners. (b) 3 dB branch-line couplers.

As can be seen, the assembling of the binary corporate structure using two-way Wilkinson or rat-race hybrids is straightforward. The use of the 3 dB coupler requires an extra care to account for the 90° phase difference in its output ports. This is accomplished using suitable connection arrangements.

Major advantages of corporate combiners is that their analyses and designs are straightforward. Their disadvantages are due to losses. Heat sinking for isolation resistors also creates problems. Because each section includes quarter-wavelength transmission lines, the corporate combiner becomes less compact than the radial combiner does with a similar number of ports. Due to their losses and large size, corporate combiners with many branches can be inconvenient, especially at the lower end of microwave frequencies. An interesting discussion regarding the use of corporate and radial combiners has been presented in Refs. 59 and 39. The inconvenience of the corporate combiner due to its large dimension is diminished at upper microwave or millimeter-wave frequency bands, because its size is scaled.

Similar to binary corporate combiners, M-ary corporate combiners can be built using M-way combiners (54). However, they have not found practical application and therefore they are not discussed here.

As shown in Fig. 13, combiners can also be built using hybrid circuits in the chain or serial configuration (2). Using this approach, combiners with arbitrary number of input ports can be produced. These structures are easy to realize in practice. Usually, couplers with progressively increasing values of coupling are used in the chain. Because of the use of nonidentical couplers, phase shifters have to be included to compensate for unequal phases of input signals. As the phase equalization is not easy to accomplish at upper microwave frequencies, these circuits have not found wide acceptance at this particular frequency band.

SPATIAL POWER COMBINERS

General Considerations

Spatial combiners use arrays of radiating elements to combine power in an unbounded or partially bounded free space.

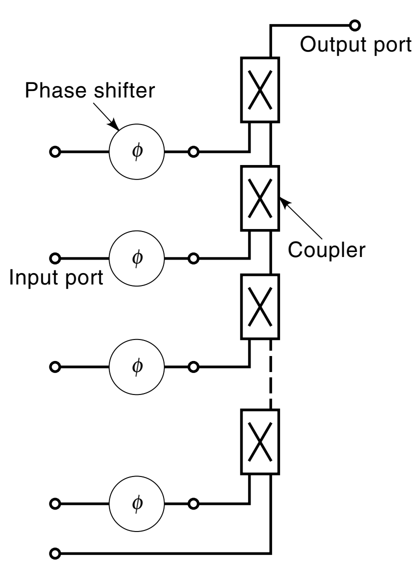

Figure 13. Chain-type combiner using couplers with unequal coupling coefficients.

The major difference with circuit-type combiners is that besides an active device and its biasing and impedance matching circuit, an antenna is an integral part. Because of the integration of an active element with an antenna, spatial power combiners are often regarded as an extension of what is known as an active array antenna. Different types of antennas including dipoles, patch antennas, notch antennas, and open-ended waveguides can be used for this purpose. This blending leads to a wide range of configurations of spatial combiners.

Similar to their circuit counterparts, spatial combiners can combine power due to individual oscillators or amplifiers. Thus the general principles for the combining schemes introduced earlier in Fig. 1 also apply to this type of combiners. The difference is that now the combining has to take place in a given direction in space. This creates an extra challenge as efficient oscillators or amplifiers as well as suitable radiating elements have to be simultaneously designed. It is apparent that the beam pointing at an undesired direction can be considered as a loss in combining efficiency. This is a new aspect in comparison with the circuit-level combiners.

Most of the reported work on spatial combiners has been related to oscillators (60). However, work concerning amplifiers, frequency multipliers, detectors, mixers, and modulators has also been accomplished (60).

Spatial oscillators, like their circuit counterparts, require some form of synchronizing mechanism. This can be achieved using internal or external means. A preferable option is to use a high Q resonator such as a spatial cavity resonator (61–65). In contrast to a circuit-type cavity resonator, this resonator, in addition to its synchronizing role, has to enable power radiation in a desired direction. Because of this requirement, one of its mirrors usually has to be semitransparent. Signals reflected from the resonator's walls (or mirrors) cause the self-injection of oscillators. Due to the high Q of the resonator, signals having high spectral purity can be established. The resulting frequency is closely related to the resonant frequency of a discrete field (mode) which can exist in the cavity. Note however that because the spatial cavity is partially enclosed, it can also support a continuous spectrum of modes. This is in contrast to the enclosed conducting cavities that exclusively support a discrete spectrum of modes. One inconvenience is that a spatial resonator occupies a considerable volume making the combiner's design less compact. Frequency synchronization is also achieved internally through the self signal-injection in the combiner structure (66–72). The mechanisms contributing to this phenomenon are due to mutual coupling of individual elements. This coupling can be accomplished due to free-space or surface waves (66–71), resulting in the weak coupling mechanism, or via specially arranged transmission-line-type connecting circuits (72) or due to close spacing of active devices in grid-type combiners (73,74), which produce strong coupling between individual oscillators. If successful, this synchronization approach makes the combiner's design compact and low profile. An alternative method for synchronizing individual oscillators is to inject an external high purity signal (74,75). An antenna can be used to illuminate the combiner array. Different mechanisms for synchronizing spatially combined oscillators are shown in Fig. 14.

The understanding of synchronization mechanism in the spatially combined oscillator array is not trivial. The general

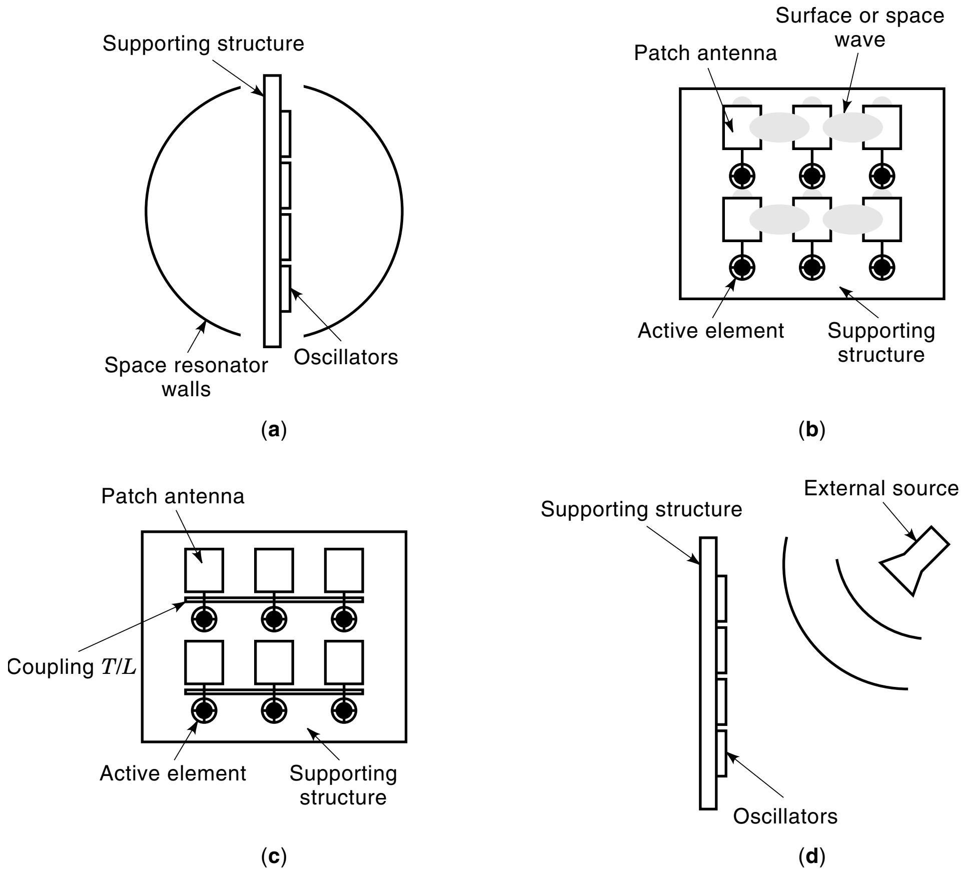

Figure 14. Methods of synchronizing spatially combined oscillators. (a) Spatial resonator. (b) Surface of space waves (weak coupling mechanism). (c) Connecting transmission lines (strong coupling mechanism). (d) External source.

answer to this problem is given by Adler's theory (76). This theory provides general conditions for synchronization, which are related to the amplitudes and frequencies of the free-running oscillators and the injected signal. In particular, this theory provides the information on the locking (synchronization) bandwidth of the externally injected oscillator. The extension of this theory in relation to linear arrays of spatially combined oscillators has been presented (66–72).

In general, two parameters are involved in the synchronization process. One is the frequency, the other is the phase. Due to a large number of competing mechanisms, whose parameters are not precisely known, the synchronization process may be unpredictable. The problem starts with a number of individual oscillators whose free-running frequencies may be different. This is due to differences in characteristics of individual solid-state devices as well as due to the different environments they are in. Middle elements in the array have a similar environment. However, this is not the case for end elements. Due to different environment and individual characteristics, the synchronized (in frequency) oscillators may feature different phases, which may generally have random values or result in a progressive shift. Random phase differences lead to power combining losses, as the signals, in a given direction, add out-of-phase. This is a similar phenomenon to that observed in circuit-level combiners using N-way Wilkinson or radial combiners. On the other hand, the constructive phase shift can result in an end-fire beam launching or beam squint. These phenomena were explained in Refs. 69, 70, and 71. It was shown that the phase shift could intentionally be controlled by end elements in the array (70,71). It was also shown that by varying the free-running frequency of the oscillator at one end of the linear array, beam scan could be obtained. Larger scan angles could be obtained for strongly coupled arrays of elements.

The formation of the beam in the broadside direction usually improves with the use of a space resonator (68). However, the thorough answer to the problem of synchronization of individual oscillators to achieve efficient directional combining has not yet been provided. Nevertheless, a number of relatively well synchronized spatial oscillators with good radiation characteristics have been demonstrated. Frequency tuning of such oscillators using mechanical and electronic means has also been shown (60). Features such as graceful degradation of spatially combined arrays have also been investigated (60).

The fundamental structures of spatially combined amplifiers are similar to those used in spatially combined oscillators. Their design often becomes less complicated than for oscillators. This is because solid-state devices usually operate at their small-signal mode, for which characteristics are better known. However, coupling mechanisms due to surface or space waves and circuit connections have to be taken into account to obtain a predictable design. Reflective and transmissive types of spatially combined amplifiers can be designed in the same way as for the circuit-type combiners.

Classification of Spatial Power Combiners

The spatial combining structures can be classified in a variety of ways. One approach is to group them into resonant and nonresonant structures. However, as has already been indicated for circuit-level combiners, this classification may not be an easy task. This is because the estimation of the Q-factor of a given structure can be difficult. A more convenient way is to classify spatial combiners in terms of their radiating elements. Using this approach, grid-type, patch (or slots), and open-ended rectangular waveguide types can be selected. These three classes are shown in Fig. 15.

Grid Combiners

The grid-type combiner is one of very early structures introduced for the purpose of spatial power combining. The term *grid* was introduced as early as the 1960s (77,78) and is related to wire antennas and selective periodic surfaces. The use of solid-state devices using grid structures was also reported at the same time (79,80). The first grid oscillators for quasi-optical power combining were introduced in 1988 (81,82).

As shown in Fig. 15, the grid combiner is formed by an array of identical units that are called cells. Unit cells are densely spaced (approximately quarter free-space-wavelength or less). This spacing creates a strong coupling mechanism between adjacent units. The vertical conducting strips in the grid structure act as dipole antennas. These can be fed by active elements such as diodes (81) or transistors (82,73,74). The horizontal conducting strips or bars serve the purpose of biasing as well as heat sinking.

For the case of in-phase operation, which results in broadside radiation, the considerations of an infinite array are often reduced to that of a single cell. Assuming that the array operates in its dominant mode, as established by the in-phase excitation of the identical unit cells, the structure can be regarded as an assembly of parallel rectangular waveguides with top and bottom walls formed by an electric conductor and sidewalls formed by magnetic conductors. In practice, the arrays are of finite size. However, an infinite array approach permits reasonable design rules.

As grid arrays are equivalent to densely packed dipoles, these arrays form inefficient radiators, as their equivalent radiation apertures are small, resulting in low directional gains. To obtain higher radiation efficiencies, antenna elements with a larger directional gain have to be used. This task can be accomplished with patch antennas that feature higher gain than the dipoles.

Spatial Combiners with Patch Antennas

One problem with grid-type combiners is that the design of the array has to be accomplished in a single step with all elements present. This is because there are almost no means for tuning or trimming. The available options for tuning include changing the bias conditions, use of transparent and nontransparent mirrors, and injection of an external signal. Consequently, the design has to be very predictable, as the design error can be costly. Although the design of the passive grid structure can be accomplished in a very accurate and repeatable manner, variations in performance of individual active devices may lead to design uncertainties in the active array.

Weakly coupled arrays of active patch antenna elements provide a better alternative to make the design more predictable. This is because single elements can be designed and tested before being assembled into the array. As has already been mentioned, a variety of antenna elements including different integration techniques with different types of active elements (diodes or transistors) can be used for this purpose. Figure 16 shows some typical arrangements that have already been explored.

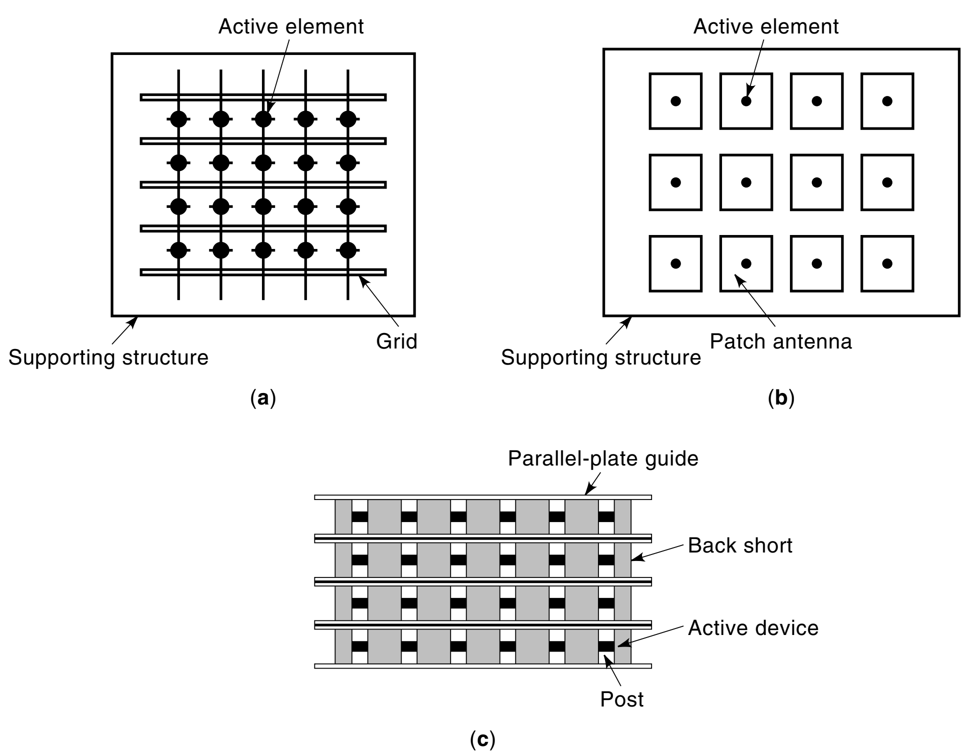

Figure 15. Basic configurations of spatial combiners. (a) Grid structure. (b) Microstrip patch antennas. (c) Open-ended parallel-plate waveguides.

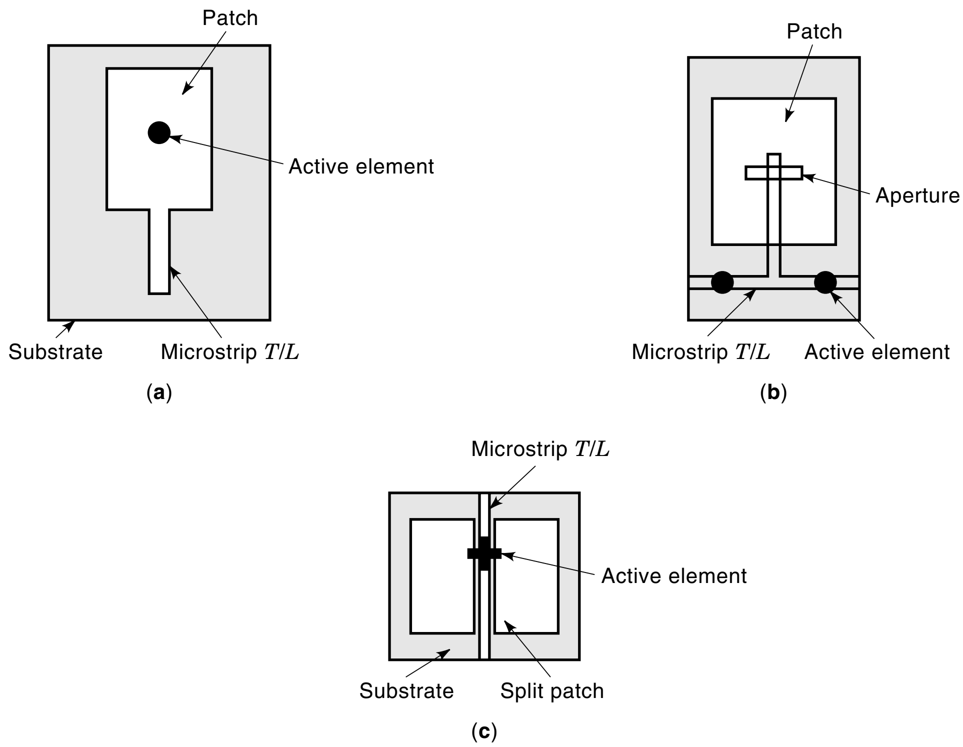

Figure 16. Different integration techniques used in spatially combined active array antennas. (a) Gunn or IMPATT diode beneath the patch (83,84). (b) Gunn diode integrated with the feedline of the aperture coupled patch (85). (c) FET transistor integrated with the patch (86,87).

As patch antennas and active devices take usually a larger space than dipoles in grid arrays, special care has to be taken to minimize their dimensions to avoid undesired grating lobes or nulls in the radiation pattern. This requires minimization of the interelement spacing to a value smaller than one wavelength. As active devices require biasing lines and pads, which take a considerable space, this task may become a challenge. In some cases, a triangular lattice instead of the usual rectangular lattice may provide a solution to this problem (88).

Spatial oscillators or amplifiers integrated with patches can also be strongly coupled. This can be arranged using circuit connections between active elements (85). To avoid multimode problems, resistors can be used (72).

Spatial Combiners Using Waveguides

Although spatial combiners with patch antennas seem to be advantageous versus grid-type combiners in terms of better integration with active elements, and superior qualities for heat sinking, they still have a number of deficiencies. As with grids, they feature a limited tuning range and wide power output deviations. At millimeter-wave frequencies patch dimensions may become too small to accommodate an active device. In this case, patches can be replaced by waveguides (64), rectangular horns, or their equivalents such as notch antennas (89,90).

The use of open-ended parallel-plate waveguides for spatial power combining has been demonstrated in (64,91). In this case, each parallel plate region consists of a number of equi-spaced Gunn diodes. The back shorts in the grooves are adjusted to provide the simultaneous frequency tuning. The parallel-plate guides are stacked in the vertical direction. This is equivalent to having a grooved mirror. For synchronization purposes, a spherical mirror with a waveguide aperture is positioned in the front of the grooved section, creating a Fabry–Perot resonator. The array of parallel-plate identical oscillators is analogous to the grid-type oscillator array. For the dominant mode of operation (established by its symmetric arrangement and in-phase excitation), an infinite array of these oscillators is equivalent to an infinite array of unit cells having electric walls at the top and the bottom walls and the magnetic walls on its sides. The difference with the planar grid is that this symmetry is enforced instead of being assumed. Perhaps because of this enforced mode of operation, the parallel-plate waveguide offers better stability in comparison with the planar grid.

A radial equivalent of the parallel-plate groove concept for the Fabry–Perot resonator operation has been presented in Ref. 92. In this case, instead of parallel-plate regions, Gunn diodes were located in a circular groove so that they could operate in an axially symmetric mode. As in the parallel-plate guide case, a ray-collimating window, which also served the purpose of frequency synchronization, was used. High combining efficiency close to 100% was obtained.

One of the very interesting alternative options to open-ended waveguides is a notch antenna, whose basic configuration is shown in Fig. 17.

This antenna is considered a planar equivalent of a horn antenna. Their variations include Vivaldi and antipodal antennas. These types of antennas feature an increased, often multiband, frequency operation. This is a significant advantage over the patch, which can barely obtain a 50% operational bandwidth. In contrast to the patch, which is used in the tile-type array, the notch antenna allows for the brick array construction, providing more space for inclusion of active devices. These are usually placed at the nonradiative end of the notch antenna. Examples of spatially combined oscillators and amplifiers using notch antennas have been presented in Refs. 90 and 93.

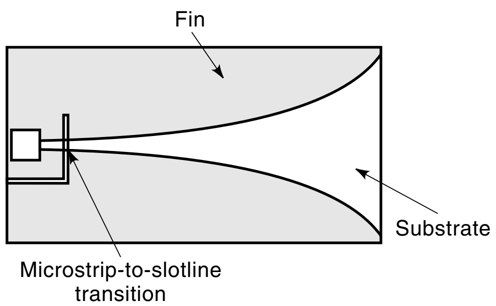

Figure 17. Configuration of a standard notch antenna fed by a microstrip coupled to the slot.

ANALYSIS METHODS FOR POWER COMBINERS/DIVIDERS

The knowledge of equivalent circuit parameters of power combiners/dividers is of paramount importance in the design process. This is often a nontrivial task, as it requires solving a three-dimensional electromagnetic field problem. In the past, many simplified methods were developed to overcome these difficulties. Using these simplified methods, ample design rules for varieties of combiners/dividers were obtained. This situation has changed dramatically in the 1990s with a number of commercial, general-purpose electromagnetic field solvers being developed. These solvers are based on integral or finite difference methods, in frequency or time domain, and are capable of accurately analyzing arbitrarily shaped passive structures. Some of these software packages are also able to simultaneously analyze assemblies of passive and active devices in real time. One problem with the general-purpose E-M field solvers is that they require large computing resources and long computational times. This may be inconvenient in cases when combining structures have to be optimized. This is because the optimization process usually requires much iteration time. Because of these shortcomings, it is advantageous to have alternative approximate ways for the analysis of combiners/dividers. In many cases, these methods can provide a relatively good approximation to the optimal design using less computing resources than the general-purpose commercial software. Nevertheless, the design process can be finalized using more accurate general-purpose software.

The following sections present examples of approximate methods for the analysis of selected power combining/dividing structures. The presentation commences with one-dimensional transmission-line type combiners and is continued with two- and three-dimensional structures.

Analysis of Transmission-Line Type Combiners

As has been demonstrated earlier, a number of basic power combining/dividing structures can be built using sections of transmission lines. For example, the Wilkinson power combiner/divider, 3 dB branch line coupler, and Gysel power combiner/divider can be built using microstrip or stripline transmission lines. The design equations for such structures using the transmission line approach are straightforward and can be found in standard microwave books (58,94,95). Thus they are not repeated here. In practice, these simple design equations are not accurate enough as they neglect discontinuities in the form of tee junctions, mitered bends, and impedance steps, which real combining structures incorporate. To obtain a more accurate design, these discontinuities have to be taken into account. The modern-day commercial software (96,97,98) not only provides this solution, but also it offers a necessary optimization before the etching of the circuit takes place. This software can also take into account the presence of active devices in these structures (96).

Analysis of Planar Two-Dimensional Power Combining Structures

Many planar power combiners cannot be represented by simple connections of transmission lines. Examples include microstrip planar circular disk combiners (45), annular-ring combiners (98), sectorial dividers (54,55), and planar dividers with holes (56,57). Also included in this list are microstrip disk and annular ring-type couplers (99,100,101).

The analysis of these structures can be handled by a number of commercial software packages, which have been developed for the purpose of analyzing arbitrarily shaped 2- and 2.5-D multilayer microstrip configurations (97). Full-wave analysis methods in real or spectral domains in association with the moment method are employed to accomplish this task. The use of these software packages may lead to time-consuming computations due to their general-purpose nature.

For single-layer microstrip structures, an alternative boundary-element method can be used (102). In this method, an arbitrarily shaped microstrip structure is modeled as a cavity, with top and bottom plates formed by a perfect electric conductor and their sidewalls formed by a perfect magnetic conductor. The electric field at the cavity's wall is assumed to be uniform with height but varying along the perimeter. In the combiner/divider, this cavity can incorporate a number of microstrip and coaxial ports. For coaxial ports, equivalent strips can be introduced to make the analysis uniform. To determine the equivalent circuit parameters, one of the ports is excited while the remaining ones are left open-circuited. At the excitation port the electric current is assumed to be known. The unknown is the electric field along the structure's perimeter, which has to be determined. As can be seen, using the magnetic wall model, a 2-D problem is replaced by a 1-D problem. This is in contrast to the usual full-wave analysis method in which the distribution of electric currents on an arbitrarily shaped conducting surface has to be determined.

For an irregularly shaped microstrip structure, the wall surface is divided into a number of sections, each with a constant value of the electric field. Using field equations for a parallel-plate guide, a set of linear algebraic equations is generated which can be solved using standard methods. Having determined the electric field distribution for the case of excitation of each single section of the boundary, an equivalent impedance matrix for the N-port formed by N-sections of the perimeter is determined (102). For some regular microstrip structures, such as a circular disk or an annular ring, an alternative method of determining the impedance matrix of the equivalent N-port can be employed. This time, the field in the cavity is expressed in terms of radial waves. The problem again reduces to a 1-D problem, in which field expansion coefficients are determined. This alternative approach has been demonstrated in Refs. 45, 46, 99, and 101.

One deficiency of the magnetic wall approach is the negligence of radiation. This leads to erroneous results at frequency points at which the structure becomes resonant and radiative. Nevertheless, this approach generates reasonably accurate results for equivalent circuit paremeters for many planar structures. The advantage of this approach is that these parameters are determined in a reasonably short time using an ordinary personal computer (PC), unlike general-purpose commercial software packages where the computational time becomes longer even using more powerful computing resources.

Analysis of Waveguide-Cavity-Type Combiners

As has been shown in a previous section, many power combining/dividing structures are formed by parallel-plate waveguides with coaxial probes or cylindrical posts perpendicular to the plates.

Rectangular cavity, cylindrical, and radial cavities belong to this category. Equivalent circuit parameters of these structures can be derived assuming that rectangular waveguide entries, coaxial entries, or gaps in the posts form input/output ports of these structures. The usual assumption about these ports is that they support only single-mode operation. This includes the TEM mode for the coaxial line, the TE_{10} mode for the rectangular waveguide, and the TEM mode, having uniform dependence with height and axial symmetry, in the cylindrical gap in the post. Equivalent circuits of these structures, as viewed from their ports, can be determined by solving an electromagnetic field problem in which single ports are excited while the remaining ports are terminated. Depending on the type of (admittance, impedance, or scattering matrix) parameters to be determined, a short circuit, an open circuit, or a match load is used as a termination. The choice depends on the configuration of the analyzed structure.

A major difficulty in finding a solution to this problem are arbitrarily positioned cylindrical posts and coaxial entries with respect to the walls of a rectangular or a cylindrical cavity. The analysis of these structures can be handled by a number of commercial software packages, which have been developed for the purpose of analysis of arbitrarily shaped 3-D passive structures (103). This approach is usually time-consuming, especially in the case when the performance of these structures has to be optimized. This is why it is desirable to have approximate fast analyses. For this purpose, the problem can be simplified by assuming that the field in an adjacent cylindrical region containing a post or a coaxial probe (including the coaxial aperture) is axially symmetric. By making this assumption, the field in these cylindrical regions can be represented in term of radial TM modes having only vertical component of the electric field and the azimuthal component of the magnetic field. These fields can generally vary in the vertical direction. The fields outside the cylindrical regions containing the probes or posts can be expressed as sums of waveguide modes. Special accelerating routines are applied to achieve their fast convergence (104,105,106). For a given excitation, the field within the combiner can be determined using, for example, a field matching method. Using this approach, a number of elegant solutions to single or multiple probes in rectangular or radial guide have been presented in the microwave literature. The solutions for rectangular waveguides include single and double coaxial probe (104,105), cylindrical posts and equivalent strip (106,107), and single and dual disk-ended probes (108). Please note that the problems of coaxially driven probes can also be solved in an alternative manner by replacing coaxial entries with equivalent gaps in post (108). Excellent insights into the operation of variety of coaxial-to-waveguide transitions with practical solutions have been presented in Ref. 109.

As in rectangular waveguide cavities, electromagnetic problems involving radial cavities with different types of posts and probes can be solved. Field matching solutions for for radial cavities with different types of probes were demonstrated in Refs. 34 and 38.

Radial or coaxial cavities with rectangular waveguide ports connected in their E- or H-plane were modeled in Refs. 110 and 111.

Based on the described analyses, a number of computer programs have been developed. These programs can be run on any PC. They are able to determine equivalent circuit parameters for a rectangular or radial cavity combiner with a moderate ($<$20) number of coaxial or waveguide ports at a single frequency point in a matter of seconds of central processing unit (CPU) time. Due to this feature, these programs are very useful in the analysis and design of many power-combining structures.

The developed analyses of parallel-plate type, rectangular and radial cavity combiners with coaxial probes and cylindrical posts can also be useful in the design of conical and hemispherical cavity combiners. In this case, the design is divided into separate designs of central and peripheral probes. The design of the central probe is similar to the design of a good transition between a coaxial line and an infinite radial guide. The design of peripheral probes is equivalent to the design of probes in a rectangular waveguide with electric bottom and top walls and magnetic sidewalls. This design strategy was presented in Ref. 112.

Analysis of Spatial-Type Power Combining Structures

As in the circuit-type combiner, in order to obtain a predictable design of the spatial-type combiner it is important to know the circuit equivalent parameters of these structures as seen from the terminals of active devices. The knowledge of these parameters is necessary to obtain suitable impedance conditions for proper operation of active devices. As the spatial-type combiner is a form of active array antenna, obtaining the radiation pattern of the array through a suitable analysis is also of great importance. According to the classification presented in the last section, grids, arrays of patches, or open-ended waveguides are of concern. Determining equivalent circuit parameters for such arrays is not a trivial task, due to usually a large number of elements, which require large computing resources. To ease the complexity of the problem, simplified, approximate analysis can be used. In this case, the combiner's symmetry can be explored.

For large arrays of identical radiating elements with perfect symmetry in two planes, the analysis can often be confined to the unit-cell problem. For example, this concept is explored in planar grid combiners, which is explained as follows. By assuming that the grid is infinite and active devices are matched and synchronized in magnitude, frequency, and phase, an identical environment is experienced by each one of these elements. Consequently, the analysis reduces to a single element cell. The unit cell often resembles the post or strip waveguide diode mounting structure that was initially analyzed in the 1970s (107). The difference with the present case is that a perfect magnetic conductor, instead of an electric conductor, forms sidewalls of the equivalent rectangular waveguide. Using this equivalent waveguide approach, grids including straight, bow-tie, and double-vee strips can be analyzed (113,114).

For finite arrays, this approximate analysis can result in a relatively accurate equivalent circuit or the center section of the array. This is because the elements in this section experience an environment similar to that in an infinite array. This is not the case for the end array elements.

The absence of identical environment in a finite array makes the in-phase, unit-cell analysis less accurate with respect to the array's radiation pattern. Due to different neighboring environments, active elements may be out of phase. This results in effective combined power losses. Additionally, progressive phase shift between adjacent elements may result in an end-fire operation of the array. To make the elements radiate in the broadside direction, the elements particularly at the array's end may require some means of tuning.

The above example shows that for a finite size grid array a strict electromagnetic field analysis is required to predict its real performance. This task can be accomplished with present-day software for the analysis of multilayer microstrip patch antennas (97). This statement also applies to spatial combiners using patch antennas in so-called strong-coupled configurations.

For weakly coupled arrays of patches, the analysis of a single element may be sufficient for an ample design of the array. This is because impedance characteristics of the patch are only slightly affected by their placement in an array. However, the prediction of radiation performances of both weakly and strongly coupled patch arrays may still require a strict electromagnetic field analysis. For finite arrays, commercial software packages may be the best answer.

Spatial combiners with open-ended rectangular waveguides can be handled by approximate analyses, which have already been described for circuit-type rectangular waveguide combiners (112). One extension, which is required in the present application, is to include open space conditions at the apertures of these waveguides. An alternative way is to use commercial software packages, which are able to analyze 3-dimensional open-ended waveguide diode mounting structures. The required tasks are within the reach of the present day commercial packages using powerful workstations (115).

APPLICATIONS

The previous sections have dealt with the classification and analysis methods of a variety of power combining/dividing structures. This section discusses the application of circuit-level and spatial-level combiners to build powerful solid-state sources. Much of this discussion is concerned with the circuit-level power combining structures. The discussion concludes with examples of the most impressive circuit- and spatial-level combiners.

Solid-State Power Amplifiers Using Circuit-Level Combiners

As can be gathered from the available literature concerning circuit-level power combining of solid-state devices, the early structures predominantly dealt with two-terminal devices such as the Gunn or IMPATT diodes, whose outputs were combined to produce a more powerful source (2,3). Waveguide cavities were mainly used as combining structures. The reason was that the early stage solid-state power combiners operated as oscillators. In this case, a high Q waveguide cavity was used for the synchronization purposes.

Later developments in the field had seen the emergence of powerful three-terminal solid-state devices, transistors. These devices were not only able to produce a higher output level, but also had a higher dc–RF conversion efficiency, which was an extra advantage over their 2-terminal counterparts. While the Gunn diode could achieve a few percent and IMPATT diode could reach several percent, the transistor was able to achieve over 20% percent dc–RF conversion rate, with 50% its limit.

As a 3-terminal device, the transistor is better suited with planar rather than waveguide technology. Due to its likely broadband operation, the amplifier-mode rather than the oscillator-mode of operation is preferred. Using the planar approach, cascaded amplifiers with tens of watts output power can be built. However, even at this power level, these amplifiers require a good heat sink arrangement. This in turn requires a suitable substrate, which has good thermal properties. Proper mounting of transistors including good quality heat sinks is also necessary.

The cascade configuration has one advantage in that a large amplifying factor can be achieved using even a moderate number of amplifying stages. Its deficiency is that if one of the intermediate transistors fails the entire amplifier may also record the failure. Thus this configuration does not exhibit graceful degradation and therefore is explored in building preamplifier modules. Hence, a new strategy is required in the high-power stages to avoid catastrophic failure. Suitable power combining techniques can be used to achieve this goal.

If one decides to continue with the planar technology to build the next amplifying stage, hybrid combiners are a good choice (39,116). At the early stage, 2-way Wilkinson combiners can be used to build a corporate combining network (39). This usually involves up to $N = 4$ individual modules. The configuration of Fig. 1(b) is employed in this process. A corporate divider splits the input signal into N ways. These in turn are input to N amplifying modules. A corporate combiner, identical to the divider, sums outputs of these amplifiers. As the next-size corporate combiner/divider handles $2^3 = 8$ amplifiers, it is avoided due to its significant size. It has to be mentioned that some designers prefer to use 3 dB quadrature couplers instead of Wilkinson's dividers to build a corporate network (116) at this amplification stage.

Following this amplifying stage, the use of the corporate network faces some problems. This is due to the parasitics of isolating resistors (39) and general power losses that corporate networks incur. Dissipation of a considerable level of power in isolating resistors creates extra problems as well. Due to this situation, a preferable option is to share the failure between a larger number of dissipating units. A waveguide radial divider/combiner and its like (hemispherical and conical combiners) provides a suitable solution (39,116). The advantages of this device are a relatively large operational bandwidth (almost one octave), low insertion losses, good amplitude and phase balance as well as good isolation, which generally increases with the number of ports. Fine examples of radial divider/combiner in the last stages of solid-state transmitters have been demonstrated (33,39,40,41,42). A 110-way radial divider/combiner was used in a 100 kW pulse L-band transmitter (39). A 50-way radial divider/combiner was applied in a 22 kW pulse S-band transmitter (116). A 20-way hemispherical divider/combiner was used in 60 W and 80 W CW Ku-band SSPAs (40,41). Similarly, a 30-way radial divider/combiner was employed in a 26 W CW X-band SSPA (42). The design of a 60 GHz, 16-way power divider/combiner was presented in (33).

Solid-State Sources Using Spatial Level Combiners

The previous section has shown that circuit-level combiners can offer the design of very reliable high-level power solid-state amplifiers at microwave frequencies and the lower end of millimeter-waves. However, the extension of this type of combiners to upper millimeter-wave frequencies faces some difficulties. This is due to the tight manufacturing tolerances, which are required at these frequency bands. For example, miniature coaxial probes are difficult to manufacture in some waveguide-type combiners. Other problems are associated with increased conductive losses. For example, the quality factor for a resonant conductor cavity becomes low, making it less attractive for oscillator synchronization purposes.

The requirement for low conductive losses demands that a new generation of millimeter-wave combiners should better utilize free space as a low-loss combining medium. At this frequency range, transversal dimensions of combining structures, in terms of multiples of wavelength, do not create a problem as their low frequency counterparts. This means that a planar approach to building of spatial combiners becomes a very feasible option. To reduce transversal dimensions, a stack (cascade) configuration (117) can be explored. However, at this stage of research, this solution has not found much success due to the reduced combining efficiency (117). As the combiner becomes an active array with a large number of identical solid-state devices, there is no need to use the multi-level combining approach, as for the circuit-type combiners.

A lot of early work on spatial power combiners has been concerned with oscillators. However, many recent investigations concentrate on amplifiers. The reason is that amplifiers offer a larger operational bandwidth. In addition, the amplifier design seems to be less critical, as compared to the oscillator, knowledge of large-signal parameters of the solid-state device is not necessary. The required small-signal parameters are easier to obtain from measurements or the manufacturer's data, making the design more predictable.

Although a large number of spatial-level combiners with solid-state devices have appeared in the microwave literature (60), they have to be critically viewed. This is because the spatial combining concept is still in its exploration phase. For example, many presented structures are only low-cost prototypes. Often, they are formed by small-size arrays operating at the microwave frequency band. Contrary to some expectations, spatial combiners may suffer from problems similar to those of their circuit-type counterparts. For example, in spatial resonators solid-state devices have to be properly spaced. Otherwise, a sudden drop in the combining efficiency is observed (62). Patch oscillators with a strong coupling mechanism may suffer from the multimoding problem unless special configurations are employed (87) or resistive elements are used (72). In addition, due to their radiative nature, spatial combiners face problems which have not been known for circuit-level combiners. For example, a random phase distribution in weakly or strongly coupled oscillators may be responsible for significant reduction in power combining efficiency, while progressive phase shift can lead to an undesired beam squint (68,70–72) that effectively can be regarded as power loss in the boresight direction.

The most impressive examples of spatial combiners concern structures with the largest number of passive or active elements. In this case, grid structures have a leading role. Examples include 100-element MESFET and HBT amplifiers or oscillator (74), (117), (118), a 464 element *pin* diode switch (119) and a 4800 Schottky diode transmittance controller (120), to name a few. The work to achieve higher amplification or directional active gains with spatially combined solid-state devices is still under way and it may take some time before significant breakthroughs in this area take place.

BIBLIOGRAPHY

1. S. Y. Liao, *Microwave Devices and Circuits,* 2nd Ed. Englewood Cliffs, NJ: Prentice-Hall, 167–168, 1985.
2. K. J. Russell, Microwave power combining techniques, *IEEE Trans. Microw. Theory Tech.,* **MTT-27**: 472–478, May 1979.
3. K. Chang and C. Sun, Millimetre-wave power-combining techniques, *IEEE Trans. Microw. Theory Tech.,* **MTT-31**: 91–107, Feb. 1983.
4. J. G. Josenhans, Diamond as an insulating heat sink for a series combination of IMPATT diodes, *Proc. IEEE,* **56**: 762–763, 1968.
5. C. T. Rucker et al., Symmetry experiments with four-mesa IMPATT-diodes, *IEEE Trans. Microw. Theory Tech.,* **MTT-25**: 75–76, 1977.
6. C. T. Rucker et al., Series-connected GaAs and Si diode chips: some new results, *Electron. Lett.,* **13** (11): 331–332, 1977.
7. C. T. Rucker et al., Multichip IMPATT power combining, a summary with new analytical and experimental results, *IEEE Trans. Microw. Theory Tech.,* **MTT-27**: 951–957, 1979.
8. D. Staiman, M. E. Breese, and W. T. Patton, New technique for combining solid-state sources, *IEEE J. Solid-State Circuits,* **SC-3**: 238–243, 1968.
9. J. C. Wiltse and J. W. Mink, Quasi-optical power combining of solid-state sources, *Microwave J.,* 144–156, 1992.
10. R. S. Harp and H. L. Stover, Power combining of X-Band IMPATT circuit modules, *1973 IEEE-ISSCC Dig. Tech. Papers,* **XVI**: 118–119, 1973.
11. N. Okubo et al., A 6 GHz 80 W GaAs FET amplifier with TM-mode cavity power combiner, *1983 IEEE MTT-S International Microwave Symposium Digest,* 276–278.
12. S. Nogi, M. Kojima, and K. Fukui, Wideband microwave power divider/combiner using conical-radial line, *1990 Asia-Pacific Microwave Conference Proceedings,* 121–124, 1990.
13. R. Aston, Techniques for increasing the bandwidth of a TM_{010}-mode power combiner, *IEEE Trans. Microw. Theory Tech.,* **MTT-27**: 479–482, 1979.
14. J. A. Navarro and K. Chang, *Integrated Active Antennas and Spatial Power Combining,* Wiley Series in Microwave and Optical Engineering, ch. 4, New York: Wiley, 1996.
15. K. Kurokawa and F. M. Magalhaes, An X-Band 10-Watt multiple-IMPATT oscillator, *Proc. IEEE,* 102–103, 1971.
16. K. Kurokawa, The single-cavity multiple-device oscillator, *IEEE Trans. Microw. Theory Tech.,* **MIT-19**: 91–107, 1971.

17. K. Chang and R. L. Ebert, W-Band power combiner design, *IEEE Trans. Microw. Theory Tech.,* **MTT-28**: 295–305, 1980.

18. K. Chang, W. F. Thrower, and G. M. Hayashibara, Millimeter-wave silicon IMPATT sources and combiners for 110–260 GHz range, *IEEE Trans. Microw. Theory Tech.,* **MTT-29**: 1278–1284, 1981.

19. K. Fukui et al., Ladder-type microwave power divider/combiners, *Electronics and Communications in Japan, Part 2,* **74** (8): 25–37, 1991.

20. A. Sanada, K. Fukui, and S. Nogi, A wideband microwave power divider/combiner with multiple-port double-ladder structure, *1992 Asia-Pacific Microwave Conference Proceedings,* 885–888, Adelaide, 1992.

21. A. Sanada, K. Fukui, and S. Nogi, Traveling-wave microwave power divider composed of reflectionless dividing units, *IEEE Trans. Microw. Theory Tech.,* **MTT-43**: 4–20, 1995.

22. M. Dydyk, Efficient power combining, *IEEE Trans. Microw. Theory Tech.,* **MTT-28**: 755–762, 1980.

23. C. A. Drubin et al., A 1 kW peak, 300 W avg IMPATT diode injection locked oscillator, *1982 MTT-S International Microwave Symposium Digest,* 126–128.

24. S. Tanaka et al., Crucial factors in power combining by oversized cylindrical cavity multiple-device structures, *IEEE Trans. Microw. Theory Tech.,* **MTT-37**: 1755–1759, 1989.

25. H. Matsumara and H. Mizuno, Design of microwave power combiner with circular TM_{0m0} mode cavity, *Electronics and Communications in Japan, Part 2,* **70** (9): 1–10, 1987.

26. H. Matsumara, Analysis of a microwave power amplifier using a combiner/divider with circular cavities, *IEEE Trans. Microw. Theory Tech.,* **MTT-38**: 928–933, 1990.

27. K. Fukui and S. Nogi, Mode analytical study of cylindrical cavity power combiners, *IEEE Trans. Microw. Theory Tech.,* **MTT-34** (9): 943–951, Sept. 1986.

28. S. Nogi and K. Fukui, TM0n0- and TMm10- mode oversized cylindrical cavity power combiners, *IEEE Trans. Microw. Theory Tech.,* **MTT-35**: 835–842, 1987.

29. G. W. Swift et al., A radial wave power combiner designed by broadband techniques, *15th European Microwave Conference Proceedings,* 973–978, 1985.

30. G. W. Swift and D. I. Stones, A comprehensive design technique for the radial wave power combiner, *1988 IEEE MTT-S International Microwave Symposium Digest,* 279–281.

31. M. Oz, Wideband double-ridged 8-way radial combiner, *14th European Microwave Conference Proceedings,* 317–322, Sept. 1984.

32. M. Oz et al., High efficiency, single-ridged, 16-way radial power combiner, *21st European Microwave Conference Proceedings,* 1029–1034, 1991.

33. T. Hsu and M. D. Simonoutti, A wideband 60 GHz 16-way power divider/combiner, *1984 IEEE MTT-S International Microwave Symposium Digest,* 175–177.

34. M. E. Bialkowski and V. P. Waris, Electromagnetic model of a planar radial-waveguide divider/combiner incorporating probes, *IEEE Trans. Microw. Theory Tech.,* **41**: 1126–1134, 1993.

35. A. Thompson and J. V. Yelland, A sixty-way S-band radial waveguide, *14th European Microwave Conference Proceedings,* 335–340, 1984.

36. S. Nogi, M. Kojima, and K. Fukui, Wideband microwave power divider/combiner using conical-radial line, *1990 Asia-Pacific Microwave Conference Proceedings,* 121–124, 1990.

37. S. Nogi, F. Okazaki, and K. Fukui, A broadband conical-radial wave power divider/combiner, *1994 Asia-Pacific Microwave Conference Proceedings,* 507–510, 1994.

38. M. E. Bialkowski and V. P. Waris, Analysis of an N-way radial cavity divider with a coaxial central port and waveguide output ports, *IEEE Trans. Microw. Theory Tech.,* **MTT-44**: 2010–2016, 1996.

39. B. J. Sanders, Radial combiner runs circles around hybrids, *Microwaves,* 55–58, Nov. 1980.

40. P. J. Allen and J. B. Ness, A hemispherical radial power-waveguide divider/combiner for high power amplifiers, *IREECON 1987 Proceedings,* Sydney, 102–104.

41. P. Martin, P. J. Allen, and J. B. Ness, A 14 GHz 80 watt GaAs FET amplifier, *21st European Microwave Conference Proceedings,* 1465–1468, 1991.

42. E. Belohoubek et al., 30-way radial power combiner for miniature GaAs FET power amplifiers, *1986 IEEE MTT-S International Microwave Symposium Digest,* 515–518.

43. A. Fathy and D. Kalokitis, Analysis and design of a 30-way radial combiner for Ku-band applications, *RCA Review,* **47**: 487–508, 1986.

44. A. A. M. Saleh, Planar electrically symmetric n-way hybrid power dividers/combiners, *IEEE Trans. Microw. Theory Tech.,* **MTT-28**: 555–563, 1980.

45. M. D. Abouzahra and K. C. Gupta, Multiple-port power divider/combiner circuits using circular microstrip disk configurations, *IEEE Trans. Microw. Theory Tech,* **MTT-35**: 1296–1302, 1987.

46. J. Goel, A K-band GaAs FET amplifier with 8.2-W output power, *IEEE Trans. Microw. Theory Tech.,* **MTT-32**: 317–323, 1984.

47. M. Cohn, B. D. Geller, and J. M. Schellenberg, A 10 Watt broadband FET combiner/amplifier, *1979 IEEE MTT-S International Microwave Symposium Digest,* 292–297.

48. E. J. Wilkinson, An N-way hybrid power divider, *IRE Trans. Microw. Theory Tech.,* **MTT-8**: 116–118, 1960.

49. H. Y. Yee, F. C. Chang, and N. F. Audeh, N-way TEM-mode broadband power dividers, *IEEE Trans. Microw. Theory Tech.,* **MTT-18**: 682–688, 1970.

50. N. Nagai, M. Maekawa, and K. Ono, Ne N-way power dividers, *IEEE Trans. Microw. Theory Tech.,* **MTT-25**: 1008–1012, 1977.

51. Z. Galani and S. J. Temple, A broadband planar N-way combiner/divider, *1977 IEEE MTT-S International Microwave Symposium Digest,* 499–502, June 1977.

52. U. H. Gysel, A new N-way power divider/combiner suitable for high power applications, *1975 IEEE MTT-S International Microwave Symposium Digest,* 116–119, June 1975.

53. L. I. Parad and R. L. Moynihan, Split-tee power divider, *IEEE Trans. Microw. Theory Tech.,* **MTT-13**: 91–95, 1965.

54. M. D. Abouzahra and K. C. Gupta, Multiport power divider-combiner circuits using circular-sector-shaped planar components, *IEEE Trans. Microw. Theory Tech.,* **MTT-36**: 1747–1751, 1988.

55. P. S. Kooi et al., A multi-way broadband sectorial power divider using exponential taper, *Proc. 1995 Asia-Pacific Microwave Conference,* KAIST, Taejon, Korea, 48–51, 9–13 October 1995.

56. Q-C. Sun, K. W. Eccleston, and S. P. Yeo, The design of optimum planar power dividers/combiners for SSPA applications, *Proc. 1995 Asia-Pacific Microwave Conference,* KAIST, Taejon, Korea, 56–59, 9–13 October 1995.

57. K. W. Eccleston, Q-C. Sun, and S. P. Yeo, Tapered microstrip-line power combiners with co-linear input ports, *Microwave and Optical Technology Letters,* August 1997.

58. E. A. Wolff and R. Kaul, *Microwave Engineering and Systems Applications,* ch. 8, New York: Wiley, 1988.

59. S. J. Foti, R. P. Flam, and W. J. Scharpf, Jr, 60-way radial combiner uses no isolators, *Microwaves & RF,* 96–118, July 1984.

60. J. A. Navarro and K. Chang, *Integrated Active Antennas and Spatial Power Combining,* New York: Wiley, 1996.

61. L. Wandinger and V. Nalbandian, Millimeter-Wave Power Combining Using Quasi-Optical Techniques, *IEEE Trans. Microw. Theory Tech.,* **MTT-31**: 189–193, 1983.

62. J. W. Mink, Quasi-optical power combining of solid-state millimeter-wave sources, *IEEE Trans. Microw. Theory Tech.,* **MTT-34**: 273–278, 1986.

63. J. C. Wiltse and J. W. Mink, Quasi-optical power combining of solid-state sources, *Microwave J.,* 144–150, Feb. 1992.

64. S. Young and K. D. Stephan, Stabilisation and power combining of planar microwave oscillators with an open resonator, *1987 IEEE MTT-S International Microwave Symposium Digest,* 185–188.

65. M. Nakayama et al., Quasi-optical resonator for millimeter and submillimeter wave solid-state sources, *1990 IEEE MTT-S International Microwave Symposium Digest,* 1209–1212.

66. K. D. Stephan, Inter-injection-locked oscillators for power combining and phased arrays, *IEEE Trans. Microw. Theory Tech.,* **MTT-34**: 1017–1025, 1986.

67. K. D. Stephan, Analysis of Inter-Injection-Locked Oscillators for Integrated Phased Arrays, *IEEE Trans. Antennas Propag.,* **35**: 771–781, 1987.

68. R. A. York and R. C. Compton, Quasi-optical for power combining using mutually synchronized oscillator arrays, *IEEE Trans. Microw. Theory Tech.,* **MTT-39**: 1000–1009, 1991.

69. R. A. York, Nonlinear analysis of phase relationships in quasi-optical oscillator arrays, *IEEE Trans. Microw. Theory Tech.,* **MTT-41**: 1799–1809, 1993.

70. P. Liao and R. A. York, Phase-shifterless beam-scanning using coupled-oscillators: theory and experiment, *IEEE Trans. Microw. Theory Tech.,* **MTT-41**: 1810–1815, 1993.

71. P. Liao and R. A. York, A six-element beam-scanning array, *IEEE Microw. Guided Wave Lett.,* **4**: 20–22, 1994.

72. S. Nogi, J. Lin, and T. Itoh, Mode analysis and stabilisation of a spatial power combining array with strongly coupled oscillators, *IEEE Trans. Microw. Theory Tech.,* **MTT-41**: 1827–1837, 1993.

73. Z. B. Popovic et al., Bar-grid oscillators, *IEEE Trans. Microw. Theory Tech.,* **MTT-38**: 225–229, 1990.

74. Z. B. Popovic et al., A 100-MESFET planar grid oscillator, *IEEE Trans. Microw. Theory Tech.,* **MTT-39**: 193–200, 1991.

75. K. Chang, K. A. Hummer, and J. L. Klein, Experiments on injection locking of active antenna elements for active phased arrays and spatial power combiners, *IEEE Trans. Microw. Theory Tech.,* **MTT-37**: 1078–1084, 1989.

76. R. Adler, A study of locking phenomena in oscillators, *Proc. of the I.R.E. and Waves and Electrons,* 351–357, June 1946.

77. J. D. Kraus, A backward angle-fire array antenna, *IEEE Trans. Antennas Propag.,* **12**: 48–50, 1964.

78. C. Chen, Transmission of microwaves through perforated flat plates of finite thickness, *IEEE Trans. Microw. Theory Tech.,* **MTT-37**: 1078–1084, 1989.

79. W. C. Brown, Experiments in transportation of energy by microwave beam, *1964 IEEE MTT-S International Microwave Symposium Digest,* 8–17.

80. C. Chen, The history of power transmission by radio waves, *IEEE Trans. Microw. Theory Tech.,* **MTT-32**: 1230–1242, 1984.

81. Z. B. Popovic and D. B. Rutledge, Diode grid oscillators, *Proc. 1988 IEEE Antennas Propag. Symp.,* 442–445.

82. Z. B. Popovic, M. Kim, and D. B. Rutledge, Grid oscillators, *Int. J. Infrared Millimeter Waves,* **9**: 647–654, 1988.

83. H. J. Thomas, D. L. Fudge, and G. Morris, Gunn source integrated with microstrip patch, *Microwaves and RF,* **24** (2): 87–91, Feb. 1985.

84. M. Dydyk, Planar radial resonator oscillator, *1986 IEEE MTT-S International Microwave Symposium Digest,* 167–168.

85. J. Lin and T. Itoh, Two-dimensional quasi-optical power combining arrays using strongly coupled oscillators, *IEEE Trans. Microw. Theory Tech.,* **MTT-42**: 734–741, 1994.

86. R. A. York, R. D. Martinez, and R. C. Compton, Active patch antenna element for array applications, *Electronics Letters,* **26** (7): 494–495, March 1990.

87. A. Balasubramaniyan and A. Mortazawi, Two-dimensional MESFET-based spatial power combiners, *IEEE Microw. Guided Wave Lett.,* **3**: 366–368, 1993.

88. A. W. Robinson and M. E. Bialkowski, Ku-band planar amplifier unit cell, *1995 Asia-Pacific Microwave Conference Proc.,* 724–727, Korea, Oct. 10–13, 1995.

89. S. Kawasaki and T. Itoh, Quasi-optical power planar arrays with FETs and slots, *IEEE Trans. Microw. Theory Tech.,* **MTT-41**: 1838–1844, 1993.

90. R. N. Simmons and R. Q. Lee, Spatial frequency multiplier with active linearly tapered slot antenna array, *1994 IEEE MTT-S International Microwave Symposium Digest,* 1557–1560.

91. J. Bae et al., Millimeter and submillimeter wave quasi-optical oscillator with Gunn diodes, *IEEE Trans. Microw. Theory Tech.,* **MTT-41**: 1851–1854, 1993.

92. M. Sanagi et al., Fabry Perot multiple-device oscillator using an axially symmetric mode, *Proc. 1994 Asia-Pacific Microwave Conference,* 859–862, Tokyo, 29–2.

93. J. A. Navarro, Y. H. Shu, and K. Chang, Broadband electronically tunable planar active radiating elements and spatial power combiners using notch antennas, *IEEE Trans. Microw. Theory Tech.,* **MTT-40**: 323–328, 1992.

94. D. M. Pozar, *Microwave Engineering,* Reading MA: Addison-Wesley, 1990.

95. S. Rosloniec, *Algorithms for Computer-Aided Design of Linear Microwave Circuits,* Norwood, MA: Artech House, 1990.

96. Touchstone, HPEEsof, Westlake Village, CA.

97. Ensemble, Boulder Microwave Technologies, Boulder, CO.

98. H. J. Song and M. E. Bialkowski, A multi-layer microstrip patch antenna sub-array design using CAD, *Microw. J.* **40** (3): 22–34, March 1997.

99. F. Tefiku and E. Yamashita, Improved analysis method for multiport microstrip annular-ring power-dividers, *IEEE Trans. Microw. Theory Tech.,* **MTT-42**: 376–382, 1994.

100. T. Kawai and I. Ohta, Planar-circuit-type 3 dB quadrature hybrids, *1994 IEEE MTT-S International Microwave Symposium Digest,* 205–208.

101. M. E. Bialkowski and S. T. Jellett, A theoretical analysis of a planar elliptically shaped 3 dB quadrature coupler, *Microwave Optical Tech. Lett.,* **8**: 140–144, 1995.

102. T. Okoshi and T. Miyoshi, The planar circuit—an approach to microwave integrated circuitry, *IEEE Trans. Microw. Theory Tech.,* **MTT-20**: 245–252, 1972.

103. High Frequency Structure Simulator, Hewlett Packard.

104. A. G. Williamson, Analysis and modeling of a coaxial-line/rectangular waveguide junction, *Proc. IEE,* Part H, **129**: 262–270, 1982.

105. M. E. Bialkowski, Modelling of a coaxial-waveguide power combining structure *IEEE Trans. Microw. Theory Tech.,* **MTT-34**: 937–942, 1986.

106. A. G. Williamson, Analysis and modeling of a single-post waveguide mounting structure, *Proc. IEE,* Part H, **129**: 271–277, 1982.

107. R. L. Eisenhart and P. J. Khan, Theoretical and experimental analysis of a waveguide mounting structure, *IEEE Trans. Microw. Theory Tech.,* **MTT-19**: 706–719, 1971.

108. M. E. Bialkowski, Analysis of a coaxial-to-waveguide adaptor including a disc-ended probe and a tuning post, *IEEE Trans. Microw. Theory Tech.,* **43**: 344–349, 1995.

109. F. C. de Ronde, Ideal w.g. to coax transitions using a f.b.m. monopole, *1988 IEEE MTT-S International Microwave Symposium Digest,* 591–593.

110. M. E. Bialkowski et al., Simplified mode-matching techniques for the analysis of coaxial-cavity-coupled radial e-plane power dividers, *IEEE Trans. Microw. Theory Tech.,* **MTT-43**: 1875–1880, 1995.

111. M. E. Bialkowski, V. P. Waris, and P. W. Davis, Field matching analysis of a hybrid structure including a radial cavity and h-plane coupled rectangular waveguides, *IEICE Trans. on Electronics, Special Issue on Microwave and Millimeter-wave Technology,* **E78-C** (8): 1146–1150, August 1995, Japan.

112. V. Waris, *Field Matching Analyses of Radial Waveguide Power Combiners and Dividers,* PhD thesis, University of Queensland, June 1997.

113. A. Pance and M. J. Wengler, Microwave modeling of 2-D active grid array antenna arrays, *IEEE Trans. Microw. Theory Tech.,* **MTT-43**: 1875–1880, 1995.

114. S. C. Bundy and Z. B. Popovic, Analysis of planar grid oscillators, *1994 IEEE MTT-S International Microwave Symposium Digest,* 827–830.

115. P. W. Davis and M. E. Bialkowski, Performance of field matching and finite element methods on analyzing coaxial-to-waveguide transitions, *Journal of Electrical and Electronics Engineering, Special Issue on Antennas,* **15** (1): 23–29, March 1995.

116. M. Kumar et al., 22 kW next generation low cost s-band solid state transmitter for surveillance and air traffic control radars, *1995 IEEE MTT-S International Microwave Symposium Digest,* 1601–1604.

117. W. A. Shiroma, B. L. Shaw, and Z. B. Popovic, Three dimensional power combiners, *1994 IEEE MTT-S International Microwave Symposium Digest,* 831–834.

118. M. Kim et al., A 100-Element HBT Grid Amplifier, *IEEE Trans. Microw. Theory Tech.,* **MTT-41**: 1762–1771, 1993.

119. K. D. Stephan, F. H. Spooner, and P. F. Goldsmith, Quasi-optical millimeter-wave hybrid and monolithic PIN diode switches, *IEEE Trans. Microw. Theory Tech.,* **MTT-41**: 1791–1798, 1993.

120. L. B. Sjogren et al., A monolithic diode array millimeter-wave beam transmittance controller, *IEEE Trans. Microw. Theory Tech.,* **MTT-41**: 1782–1790, 1993.

MAREK E. BIALKOWSKI
University of Queensland and
Nanyang Technological
University

POWER BLACKOUT PREVENTION. See POWER SYSTEM SECURITY.

POWER CABLE, SHIELDED. See SHIELDED POWER CABLE.

POWER COMBINERS. See POWER COMBINERS AND DIVIDERS.

POWER CONTROL. See LOAD REGULATION OF POWER PLANTS.

POWER CONVERTERS. See HVDC POWER CONVERTERS; RESONANT POWER CONVERTERS; VARIABLE SPEED DRIVES.

POWER, DC, TRANSMISSIONS. See HVDC POWER TRANSMISSION.