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Chapter

7

RF Interconnection and Switching

7.1 Introduction

The components that connect, interface, transfer, and filter RF energy within a given system—or between systems—are critical elements in the operation of vacuum tube devices. Such hardware, usually passive, determines to a large extent the overall performance of the RF generator. To optimize the performance of power vacuum devices, first it is necessary to understand and optimize the components upon which the tube depends.

The mechanical and electrical characteristics of the transmission line, waveguide, and associated hardware that carry power from a power source (usually a transmitter) to the load (usually an antenna) are critical to proper operation of any RF system. Mechanical considerations determine the ability of the components to withstand temperature extremes, lightning, rain, and wind. In other words, they determine the overall reliability of the system.

7.1.1 Skin Effect

The effective resistance offered by a given conductor to radio frequencies is considerably higher than the ohmic resistance measured with direct current. This is because of an action known as the *skin effect*, which causes the currents to be concentrated in certain parts of the conductor and leaves the remainder of the cross section to contribute little or nothing toward carrying the applied current.

When a conductor carries an alternating current, a magnetic field is produced that surrounds the wire. This field continually expands and contracts as the ac wave increases from zero to its maximum positive value and back to zero, then through its negative half-cycle. The changing magnetic lines of force cutting the conductor induce a voltage in the conductor in a direction that tends to retard the normal flow of current in the wire. This effect is more pronounced at the center of the conductor. Thus, current within the conductor tends to flow more easily toward the surface of the wire. The higher the frequency, the greater the tendency for current to flow at the surface. The depth of current flow is a function of frequency, and it is determined from the following equation:

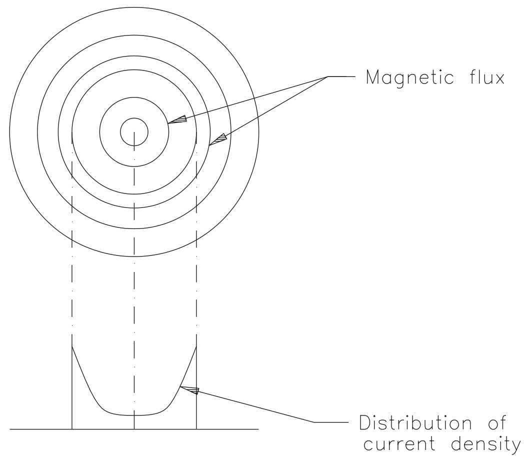


Figure 7.1 Skin effect on an isolated round conductor carrying a moderately high frequency signal.

$$d = \frac{2.6}{\sqrt{\mu f}} \tag{7.1}$$

Where:
d = depth of current in mils
μ = permeability (copper = 1, steel = 300)
f = frequency of signal in mHz

It can be calculated that at a frequency of 100 kHz, current flow penetrates a conductor by 8 mils. At 1 MHz, the skin effect causes current to travel in only the top 2.6 mils in copper, and even less in almost all other conductors. Therefore, the series impedance of conductors at high frequencies is significantly higher than at low frequencies. Figure 7.1 shows the distribution of current in a radial conductor.

When a circuit is operating at high frequencies, the skin effect causes the current to be redistributed over the conductor cross section in such a way as to make most of the current flow where it is encircled by the smallest number of flux lines. This general principle controls the distribution of current regardless of the shape of the conductor involved. With a flat-strip conductor, the current flows primarily along the edges, where it is surrounded by the smallest amount of flux.

7.2 Coaxial Transmission Line

Two types of coaxial transmission line are in common use today: *rigid line* and corrugated (*semiflexible*) line. Rigid coaxial cable is constructed of heavy-wall copper tubes with Teflon or ceramic spacers. (Teflon is a registered trademark of DuPont.)

Rigid line provides electrical performance approaching an ideal transmission line, including:

- High power-handling capability
- Low loss
- Low VSWR (*voltage standing wave ratio*)

Rigid transmission line is, however, expensive to purchase and install.

The primary alternative to rigid coax is semiflexible transmission line made of corrugated outer and inner conductor tubes with a spiral polyethylene (or Teflon) insulator. The internal construction of a semiflexible line is shown in Figure 7.2. Semiflexible line has four primary benefits:

- It is manufactured in a continuous length, rather than the 20-ft sections typically used for rigid line.
- Because of the corrugated construction, the line may be shaped as required for routing from the transmitter to the antenna.
- The corrugated construction permits differential expansion of the outer and inner conductors.
- Each size of line has a minimum bending radius. For most installations, the flexible nature of corrugated line permits the use of a single piece of cable from the transmitter to the antenna, with no elbows or other transition elements. This speeds installation and provides for a more reliable system.

7.2.1 Electrical Parameters

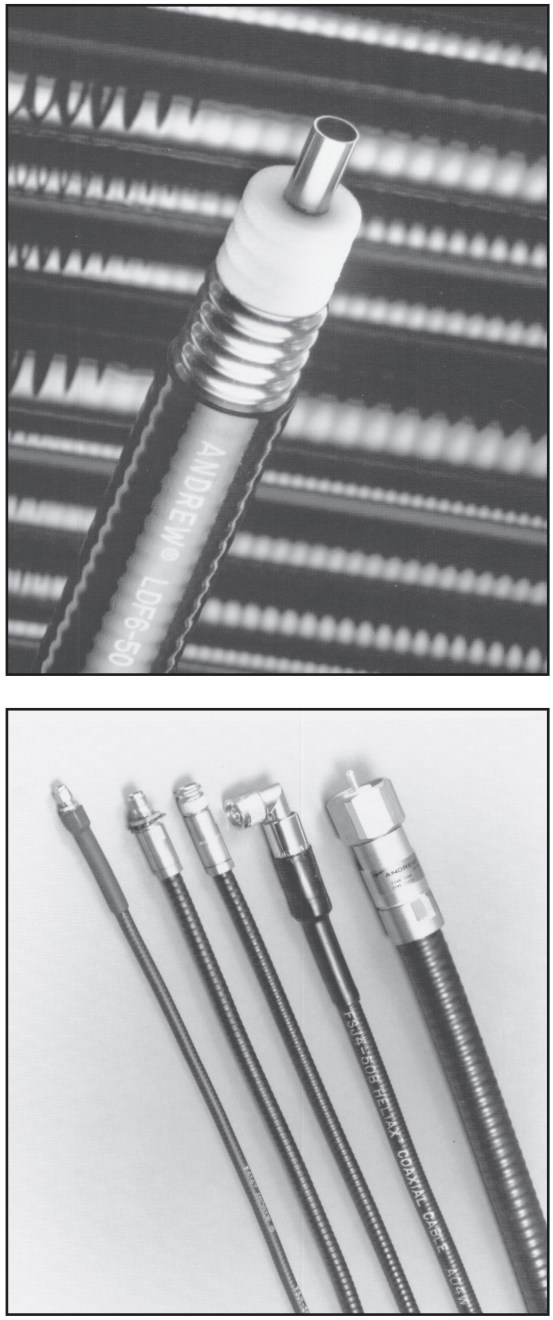
A signal traveling in free space is unimpeded; it has a free-space velocity equal to the speed of light. In a transmission line, capacitance and inductance slow the signal as it propagates along the line. The degree to which the signal is slowed is represented as a percentage of the free-space velocity. This quantity is called the *relative velocity of propagation* and is described by the equation:

$$V_p = \frac{1}{\sqrt{L \times C}} \tag{7.2}$$

Where:
L = inductance in henrys per foot
C = capacitance in farads per foot

and

$$V_r = \frac{V_p}{c} \times 100\% \tag{7.3}$$



(*a*)

(*b*)

Figure 7.2 Semiflexible coaxial cable: (*a*) a section of cable showing the basic construction, (*b*) cable with various terminations. (*Courtesy of Andrew*.)

Where:
V_p = velocity of propagation
$c = 9.842 \times 10^8$ feet per second (free-space velocity)
V_r = velocity of propagation as a percentage of free-space velocity

Transverse Electromagnetic Mode

The principal mode of propagation in a coaxial line is the *transverse electromagnetic mode* (TEM). This mode will not propagate in a waveguide, and that is why coaxial lines can propagate a broad band of frequencies efficiently. The cutoff frequency for a coaxial transmission line is determined by the line dimensions. Above cutoff, modes other than TEM can exist and the transmission properties are no longer defined. The cutoff frequency is equivalent to:

$$F_c = \frac{7.50 \times V_r}{D_i + D_o} \tag{7.4}$$

Where:
F_c = cutoff frequency in gigahertz
V_r = velocity (percent)
D_i = inner diameter of outer conductor in inches
D_o = outer diameter of inner conductor in inches

At dc, current in a conductor flows with uniform density over the cross section of the conductor. At high frequencies, the current is displaced to the conductor surface. The effective cross section of the conductor decreases, and the conductor resistance increases because of the skin effect.

Center conductors are made from copper-clad aluminum or high-purity copper and can be solid, hollow tubular, or corrugated tubular. Solid center conductors are found on semiflexible cable with 1/2 in or smaller diameter. Tubular conductors are found in 7/8 in or larger-diameter cables. Although the tubular center conductor is used primarily to maintain flexibility, it also can be used to pressurize an antenna through the feeder.

Dielectric

Coaxial lines use two types of dielectric construction to isolate the inner conductor from the outer conductor. The first is an air dielectric, with the inner conductor supported by a dielectric spacer and the remaining volume filled with air or nitrogen gas. The spacer, which may be constructed of spiral or discrete rings, typically is made of Teflon or polyethylene. Air-dielectric cable offers lower attenuation and higher average power ratings than foam-filled cable but requires pressurization to prevent moisture entry.

Foam-dielectric cables are ideal for use as feeders with antennas that do not require pressurization. The center conductor is surrounded completely by foam-dielectric material, resulting in a high dielectric breakdown level. The dielectric materials are polyethylene-based formulations, which contain antioxidants to reduce dielectric deterioration at high temperatures.

Impedance

The expression *transmission line impedance* applied to a point on a transmission line signifies the vector ratio of line voltage to line current at that particular point. This is the impedance that would be obtained if the transmission line were cut at the point in question, and the impedance looking toward the receiver were measured.

Because the voltage and current distribution on a line are such that the current tends to be small when the voltage is large (and vice versa), as shown in Figure 7.3, the impedance will, in general, be oscillatory in the same manner as the voltage (large when the voltage is high and small when the voltage is low). Thus, in the case of a short-circuited receiver, the impedance will be high at distances from the receiving end that are odd multiples of 1/4 wavelength, and it will be low at distances that are even multiples of 1/4 wavelength.

The extent to which the impedance fluctuates with distance depends on the standing wave ratio (ratio of reflected to incident waves), being less as the reflected wave is proportionally smaller than the incident wave. In the particular case where the load impedance equals the characteristic impedance, the impedance of the transmission line is equal to the characteristic impedance at all points along the line.

The power factor of the impedance of a transmission line varies according to the standing waves present. When the load impedance equals the characteristic impedance, there is no reflected wave, and the power factor of the impedance is equal to the power factor of the characteristic impedance. At radio frequencies, the power factor under these conditions is accordingly resistive. However, when a reflected wave is present, the power factor is unity (resistive) only at the points on the line where the voltage passes through a maximum or a minimum. At other points the power factor will be reactive, alternating from leading to lagging at intervals of 1/4 wavelength. When the line is short-circuited at the receiver, or when it has a resistive load less than the characteristic impedance so that the voltage distribution is of the short-circuit type, the power factor is inductive for lengths corresponding to less than the distance to the first voltage maximum. Thereafter, it alternates between capacitive and inductive at intervals of 1/4 wavelength. Similarly, with an open-circuited receiver or with a resistive load greater than the characteristic impedance so that the voltage distribution is of the open-circuit type, the power factor is capacitive for lengths corresponding to less than the distance to the first voltage minimum. Thereafter, the power factor alternates between capacitive and inductive at intervals of 1/4 wavelength, as in the short-circuited case.

Resonant Characteristics

A transmission line can be used to perform the functions of a resonant circuit. If the line, for example, is short-circuited at the receiver, at frequencies in the vicinity of a frequency at which the line is an odd number of 1/4 wavelengths long, the impedance

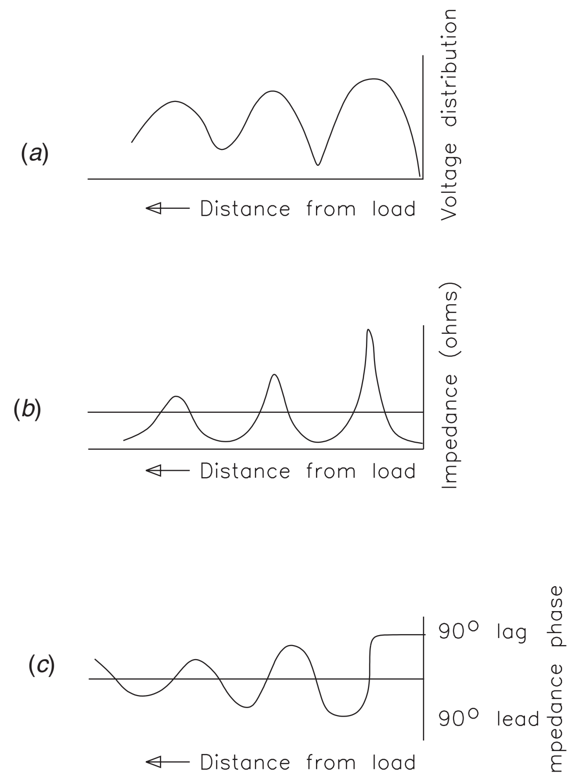


Figure 7.3 Magnitude and power factor of line impedance with increasing distance from the load for the case of a short-circuited receiver and a line with moderate attenuation: (*a*) voltage distribution, (*b*) impedance magnitude, (*c*) impedance phase.

will be high and will vary with frequency in the vicinity of resonance. This characteristic is similar in nature to a conventional parallel resonant circuit. The difference is that with the transmission line, there are a number of resonant frequencies, one for each of the infinite number of frequencies that make the line an odd number of 1/4 wavelengths long. At VHF, the parallel impedance at resonance and the circuit Q obtainable are far higher than can be realized with lumped circuits. Behavior corresponding to that of a series resonant circuit can be obtained from a transmission line that is an odd number of 1/4 wavelengths long and open-circuited at the receiver.

Transmission lines also can be used to provide low-loss inductances or capacitances if the proper combination of length, frequency, and termination is employed. Thus, a line short-circuited at the receiver will offer an inductive reactance when less than 1/4 wavelength, and a capacitive reactance when between 1/4 and 1/2 wavelength. With an open-circuited receiver, the conditions for inductive and capacitive reactances are reversed.

7.2.2 Electrical Considerations

VSWR, attenuation, and power-handling capability are key electrical factors in the application of coaxial cable. High VSWR can cause power loss, voltage breakdown, and thermal degradation of the line. High attenuation means less power delivered to the antenna, higher power consumption at the transmitter, and increased heating of the transmission line itself.

VSWR is a common measure of the quality of a coaxial cable. High VSWR indicates nonuniformities in the cable that can be caused by one or more of the following conditions:

- Variations in the dielectric core diameter
- Variations in the outer conductor
- Poor concentricity of the inner conductor
- Nonhomogeneous or periodic dielectric core

Although each of these conditions may contribute only a small reflection, they can add up to a measurable VSWR at a particular frequency.

Rigid transmission line typically is available in a standard length of 20 ft, and in alternative lengths of 19.5 ft and 19.75 ft. The shorter lines are used to avoid VSWR buildup caused by discontinuities resulting from the physical spacing between line section joints. If the section length selected and the operating frequency have a 1/2-wave correlation, the connector junction discontinuities will add. This effect is known as *flange buildup*. The result can be excessive VSWR. The critical frequency at which a 1/2-wave relationship exists is given by:

$$F_{cr} = \frac{490.4 \times n}{L} \tag{7.5}$$

Where:
F_{cr} = the critical frequency
n = any integer
L = transmission line length in feet

For most applications, the critical frequency for a chosen line length should not fall closer than ±2 MHz of the passband at the operating frequency.

Attenuation is related to the construction of the cable itself and varies with frequency, product dimensions, and dielectric constant. Larger-diameter cable exhibits lower attenuation than smaller-diameter cable of similar construction when operated at the same frequency. It follows, therefore, that larger-diameter cables should be used for long runs.

Air-dielectric coax exhibits less attenuation than comparable-size foam-dielectric cable. The attenuation characteristic of a given cable also is affected by standing waves present on the line resulting from an impedance mismatch. Table 7.1 shows a representative sampling of semiflexible coaxial cable specifications for a variety of line sizes.

Table 7.1 Representative Specifications for Various Types of Flexible Air-Dielectric Coaxial Cable

Cable Size (in.)	Maximum Frequency (MHz)	Velocity (percent)	Peak Power 1 MHz (kW)	Average Power 100 MHz (kW)	Average Power 1 MHz (kW)	Attenuation[1] 100 MHz (dB)	Attenuation[1] 1 Mhz (dB)
1 5/8	2.7	92.1	145	145	14.4	0.020	0.207
3	1.64	93.3	320	320	37	0.013	0.14
4	1.22	92	490	490	56	0.010	0.113
5	0.96	93.1	765	765	73	0.007	0.079

[1] Attenuation specified in dB/100 ft.

7.2.3 Coaxial Cable Ratings

Selection of a type and size of transmission line is determined by a number of parameters, including power-handling capability, attenuation, and phase stability.

Power Rating

Both *peak* and *average* power ratings are required to fully describe the capabilities of a transmission line. In most applications, the peak power rating limits the low frequency or pulse energy, and the average power rating limits high-frequency applications, as shown in Figure 7.4. Peak power ratings usually are stated for the following conditions:

- VSWR = 1.0
- Zero modulation
- One atmosphere of absolute dry air pressure at sea level

The peak power rating of a selected cable must be greater than the following expression, in addition to satisfying the average-power-handling criteria:

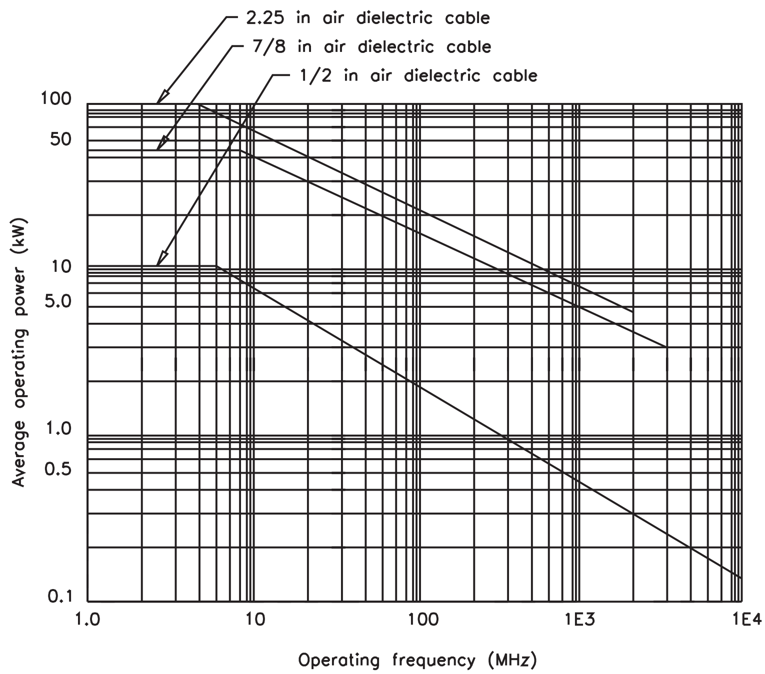
$$E_{pk} > P_t \times (1 + M)^2 \times VSWR \tag{7.6}$$

Where:
P_{pk} = cable peak power rating in kilowatts
P_t = transmitter power in kilowatts
M = amplitude modulation percentage expressed decimally (100 percent = 1.0)
VSWR = voltage standing wave ratio

From this equation, it can be seen that 100 percent amplitude modulation will increase the peak power in the transmission line by a factor of 4. Furthermore, the peak power in the transmission line increases directly with VSWR.

The peak power rating is limited by the voltage breakdown potential between the inner and outer conductors of the line. The breakdown point is independent of frequency.

(*a*)



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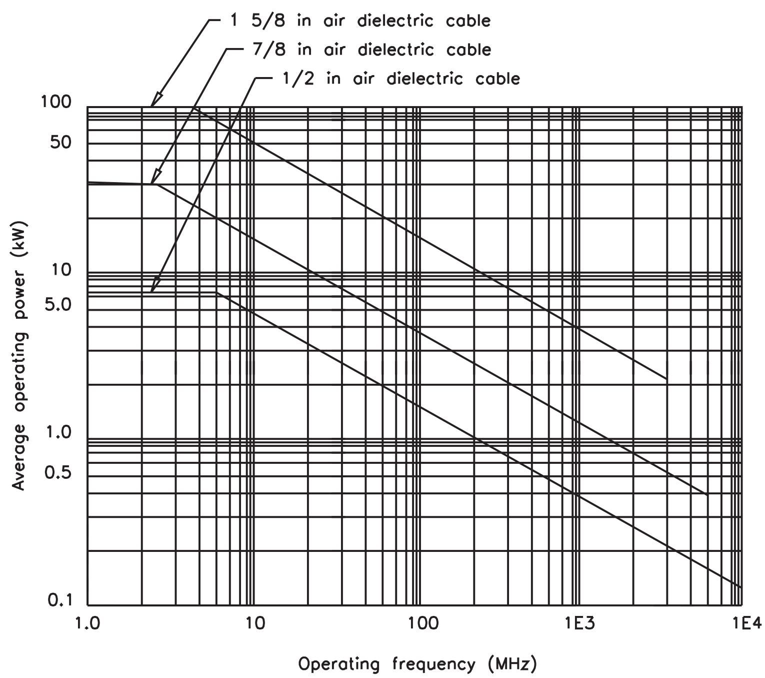


Figure 7.4 Power rating data for a variety of coaxial transmission lines: (*a*) 50 Ω line, (*b*) 75 Ω line.

It varies, however, with the line pressure (for an air-dielectric cable) and the type of pressurizing gas.

The average power rating of a transmission line is limited by the safe long-term operating temperature of the inner conductor and the dielectric. Excessive temperatures on the inner conductor will cause the dielectric material to soften, leading to mechanical instability inside the line.

The primary purpose of pressurization of an air-dielectric cable is to prevent the ingress of moisture. Moisture, if allowed to accumulate in the line, can increase attenuation and reduce the breakdown voltage between the inner and outer conductors. Pressurization with high-density gases can be used to increase both the average power and the peak power ratings of a transmission line. For a given line pressure, the increased power rating is more significant for peak power than for average power. High-density gases used for such applications include Freon 116 and sulfur hexafluoride. Figure 7.5 illustrates the effects of pressurization on cable power rating.

An adequate safety factor is necessary for peak and average power ratings. Most transmission lines are tested at two or more times their rated peak power before shipment to the customer. This safety factor is intended as a provision for transmitter transients, lightning-induced effects, and high-voltage excursions resulting from unforeseen operating conditions.

Connector Effects

Foam-dielectric cables typically have a greater dielectric strength than air-dielectric cables of similar size. For this reason, foam cables would be expected to exhibit higher peak power ratings than air lines. Higher values, however, usually cannot be realized in practice because the connectors commonly used for foam cables have air spaces at the cable/connector interface that limit the allowable RF voltage to "air cable" values.

The peak-power-handling capability of a transmission line is the smaller of the values for the cable and the connectors attached to it. Table 7.2 lists the peak power ratings of several common RF connectors at standard conditions (defined in the previous section).

Attenuation

The attenuation characteristics of a transmission line vary as a function of the operating frequency and the size of the line itself. The relationships are shown in Figure 7.6.

The *efficiency* of a transmission line dictates how much power output by the transmitter actually reaches the antenna. Efficiency is determined by the length of the line and the attenuation per unit length.

The attenuation of a coaxial transmission line is defined by the equation:

$$\alpha = 10 \times \log \left\{ \frac{P_1}{P_2} \right\} \tag{7.7}$$

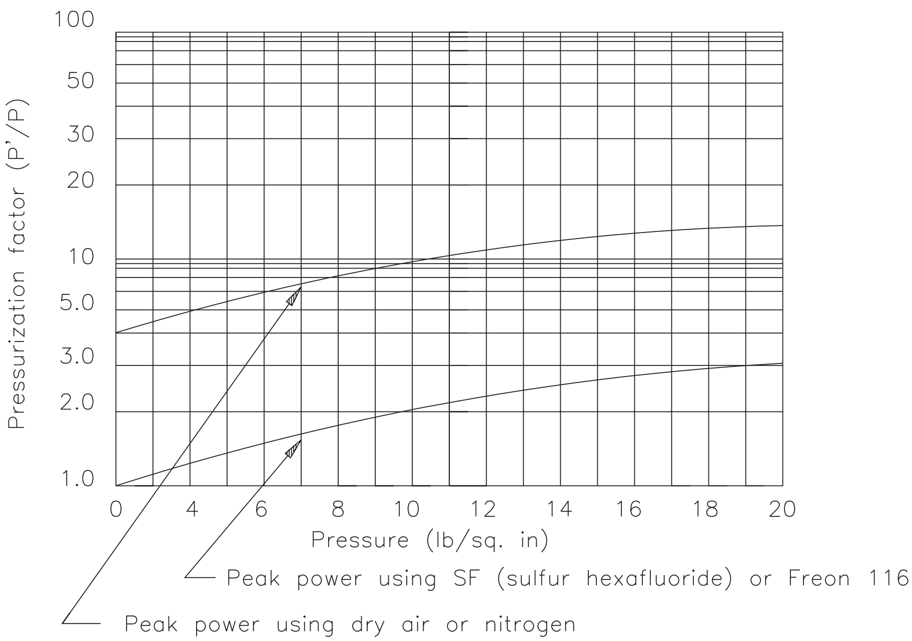


Figure 7.5 Effects of transmission line pressurization on peak power rating. Note that P′ = the rating of the line at the increased pressure, and P = the rating of the line at atmospheric pressure.

Table 7.2 Electrical Characteristics of Common RF Connectors

Connector Type	DC Test Voltage (kW)	Peak Power (kW)
SMA	1.0	1.2
BNC, TNC	1.5	2.8
N, UHF	2.0	4.9
GR	3.0	11
HN, 7/16	4.0	20
LC	5.0	31
7/8 EIA, F Flange	6.0	44
1-5/8 EIA	11.0	150
3-1/8 EIA	19.0	4

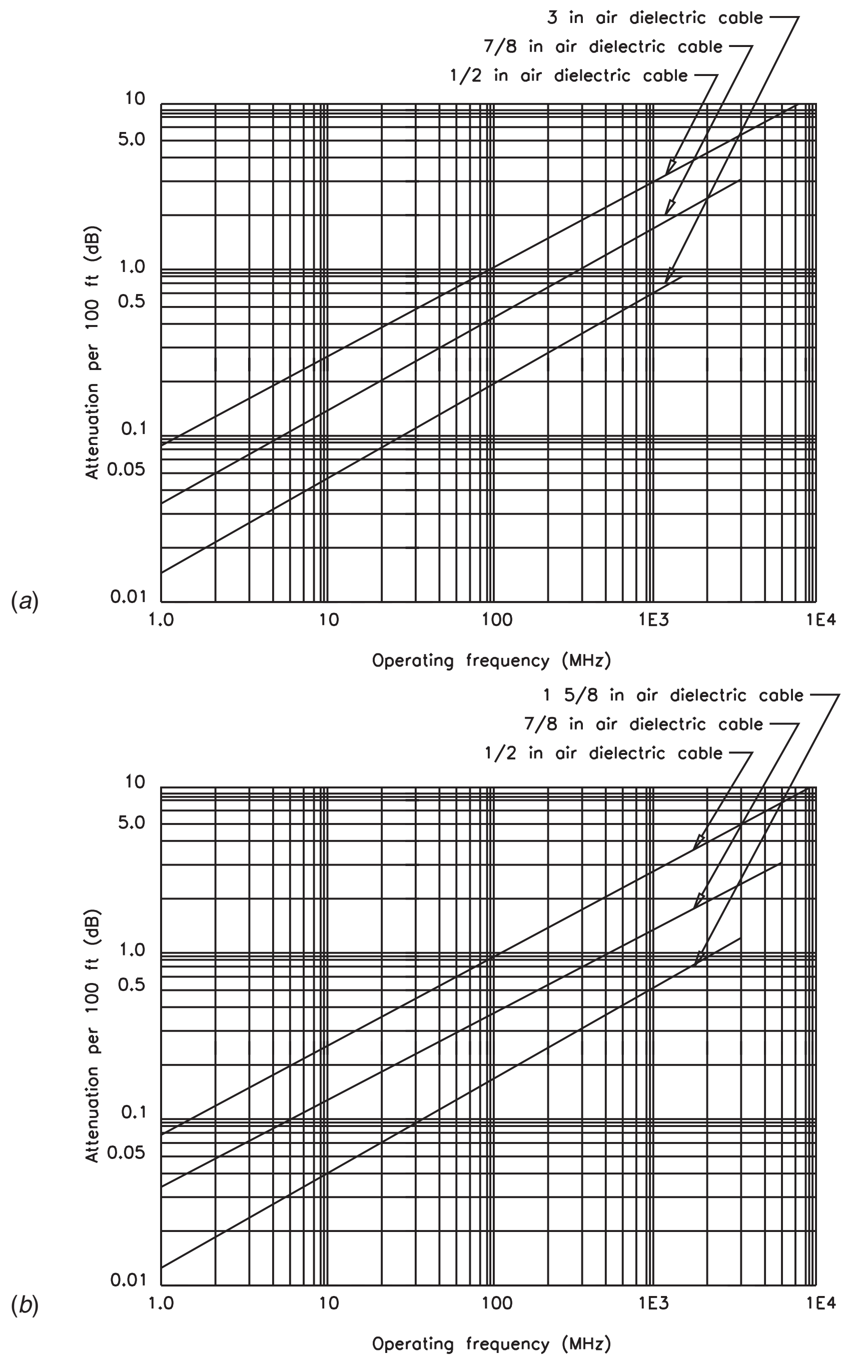


Figure 7.6 Attenuation characteristics for a selection of coaxial cables: (*a*) 50 Ω line, (*b*) 75 Ω line.

Where:
α = attenuation in decibels per 100 meters
P_1 = input power into a 100-meter line terminated with the nominal value of its characteristic impedance
P_2 = power measured at the end of the line

Stated in terms of efficiency (E, percent):

$$E = 100 \times \left\{ \frac{P_o}{P_i} \right\} \tag{7.8}$$

Where:
P_i = power delivered to the input of the transmission line
P_o = power delivered to the antenna

The relationship between efficiency and loss in decibels (*insertion loss*) is illustrated in Figure 7.7.

Manufacturer-supplied attenuation curves typically are guaranteed to within approximately ±5 percent. The values given usually are rated at 24ºC (75ºF) ambient temperature. Attenuation increases slightly with higher temperature or applied power. The effects of ambient temperature on attenuation are illustrated in Figure 7.8.

Loss in connectors is negligible, except for small (SMA and TNC) connectors at frequencies of several gigahertz and higher. Small connectors used at high frequencies typically add 0.1 dB of loss per connector.

When a transmission line is attached to a load, such as an antenna, the VSWR of the load increases the total transmission loss of the system. This effect is small under conditions of low VSWR. Figure 7.9 illustrates the interdependence of these two elements.

Phase Stability

A coaxial cable expands as the temperature of the cable increases, causing the electrical length of the line to increase as well. This factor results in phase changes that are a function of operating temperature. The phase relationship can be described by the equation:

$$\theta = 3.66 \times 10^{-7} \times P \times L \times T \times F \tag{7.9}$$

Where:
θ = phase change in degrees
P = phase-temperature coefficient of the cable
L = length of coax in feet
T = temperature range (minimum-to-maximum operating temperature)
F = frequency in Mhz

Phase changes that are a function of temperature are important in systems utilizing multiple transmission lines, such as a directional array fed from a single phasing

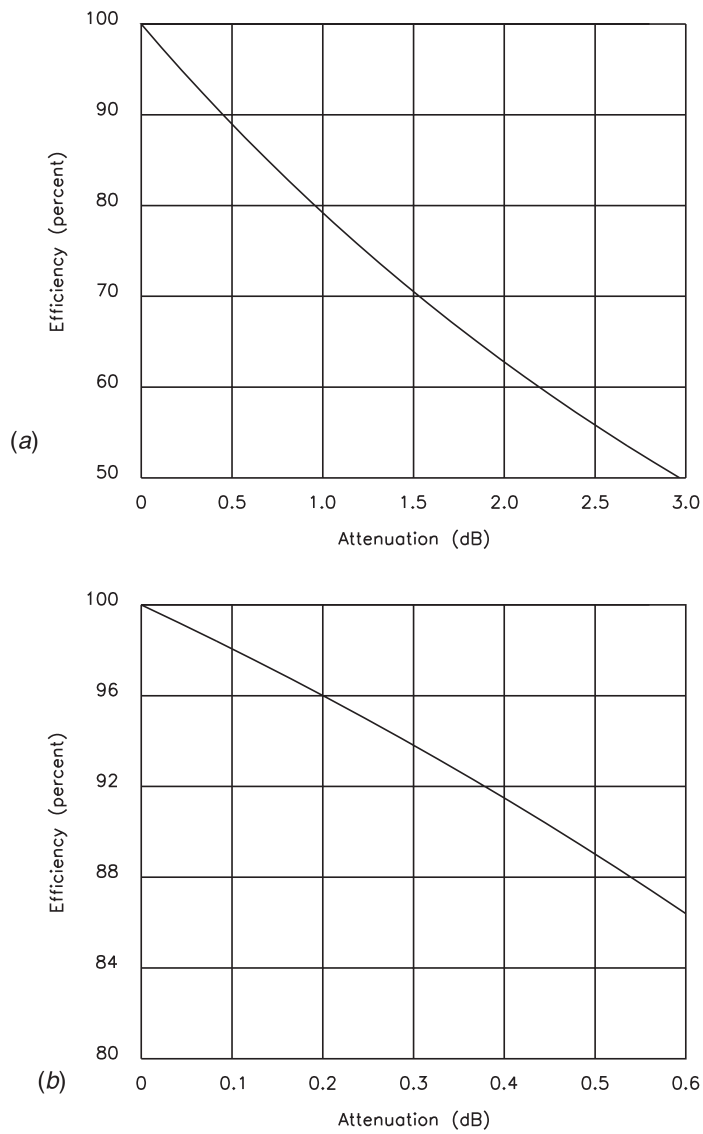


Figure 7.7 Conversion chart showing the relationship between decibel loss and efficiency of a transmission line: (*a*) high-loss line, (*b*) low-loss line.

source. To maintain proper operating parameters, the phase changes of the cables must be minimized. Specially designed coaxial cables that offer low-phase-temperature characteristics are available. Two types of coax are commonly used for this purpose:

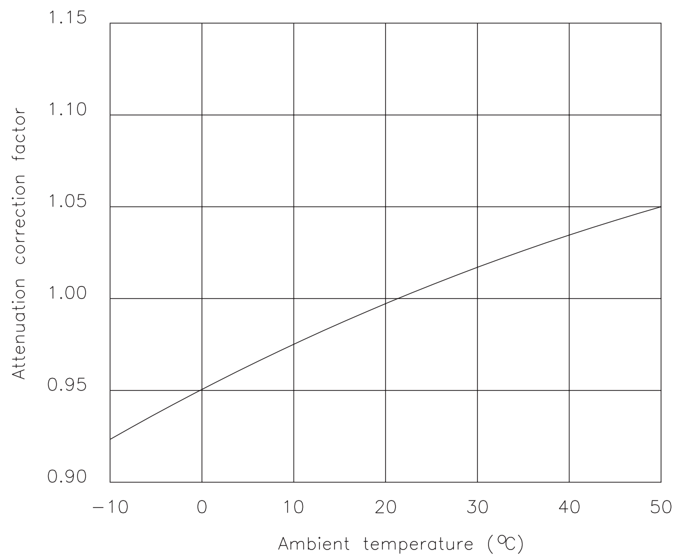


Figure 7.8 The variation of coaxial cable attenuation as a function of ambient temperature.

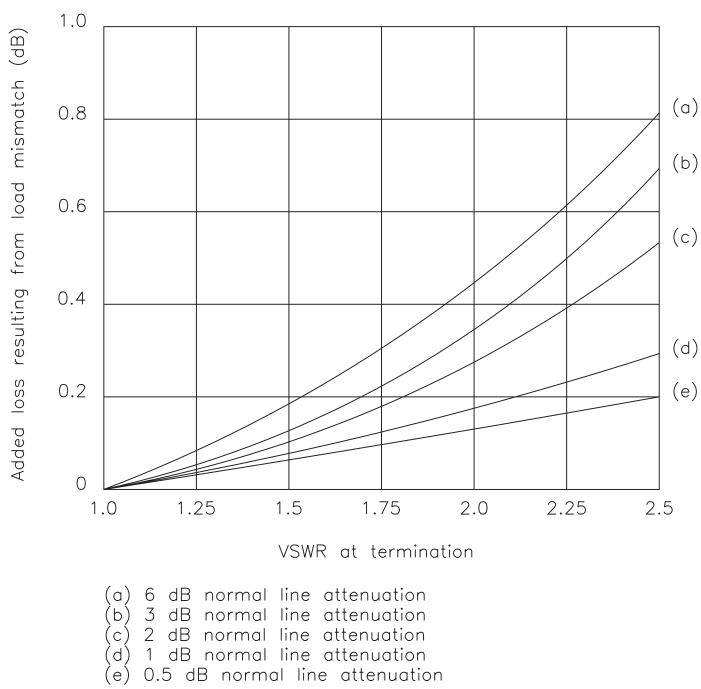


Figure 7.9 The effect of load VSWR on transmission line loss.

- *Phase-stabilized* cables, which have undergone extensive temperature cycling until such time as they exhibit their minimum phase-temperature coefficient
- *Phase-compensated* cables, in which changes in the electrical length have been minimized through adjustment of the mechanical properties of the dielectric and inner/outer conductors

7.2.4 Mechanical Parameters

Corrugated copper cables are designed to withstand bending with no change in properties. Low-density foam- and air-dielectric cables generally have a minimum bending radius of 10 times the cable diameter. Super flexible versions provide a smaller allowable bending radius.

Rigid transmission line will not tolerate bending. Instead, transition elements (elbows) of various sizes are used. Individual sections of rigid line are secured by multiple bolts around the circumference of a coupling flange.

When a large cable must be used to meet attenuation requirements, short lengths of a smaller cable (jumpers or *pigtails*) may be used on either end for ease of installation in low-power systems. The tradeoff is slightly higher attenuation and some additional cost.

The *tensile strength* of a cable is defined as the axial load that may be applied to the line with no more than 0.2 percent permanent deformation after the load is released. When divided by the weight per foot of cable, this gives an indication of the maximum length of cable that is self-supporting and therefore can be installed readily on a tower with a single hoisting grip. This consideration usually applies only to long runs of corrugated line; rigid line is installed one section at a time.

The *crush strength* of a cable is defined as the maximum force per linear inch that may be applied by a flat plate without causing more than a 5 percent deformation of the cable diameter. Crush strength is a good indicator of the ruggedness of a cable and its ability to withstand rough handling during installation.

Cable jacketing affords mechanical protection during installation and service. Semiflexible cables typically are supplied with a jacket consisting of low-density polyethylene blended with 3 percent carbon black for protection from the sun's ultraviolet rays, which can degrade plastics over time. This approach has proved to be effective, yielding a life expectancy of more than 20 years. Rigid transmission line has no covering over the outer conductor.

For indoor applications, where fire-retardant properties are required, cables can be supplied with a fire-retardant jacket, usually listed by Underwriters Laboratories. Note that under the provisions of the National Electrical Code, outside plant cables such as standard black polyethylene-jacketed coaxial line may be run as far as 50 ft inside a building with no additional protection. The line also may be placed in conduit for longer runs.

Low-density foam cables are designed to prevent water from traveling along their length, should it enter through damage to the connector or the cable sheath. This is accomplished by mechanically locking the outer conductor to the foam dielectric by annular corrugations. Annular or ring corrugations, unlike helical or screw-thread-type corrugations, provide a water block at each corrugation. Closed-cell polyethylene dielectric foam is bonded to the inner conductor, completing the moisture seal.

A coaxial cable line is only as good as the connectors used to tie it together. The connector interface must provide a weatherproof bond with the cable to prevent water from penetrating the connection. This is ensured by using O-ring seals. The cable-connector interface also must provide a good electrical bond that does not introduce a mismatch and increase VSWR. Good electrical contact between the connector and the cable ensures that proper RF shielding is maintained.

7.3 Waveguide

As the operating frequency of a system reaches into the UHF band, waveguide-based transmission line systems become practical. From the mechanical standpoint, waveguide is simplicity itself. There is no inner conductor; RF energy is *launched* into the structure and propagates to the load. Several types of waveguide are available, including rectangular, square, circular, and elliptical. Waveguide offers several advantages over coax. First, unlike coax, waveguide can carry more power as the operating frequency increases. Second, efficiency is significantly better with waveguide at higher frequencies.

Rectangular waveguide commonly is used in high-power transmission systems. Circular waveguide also may be used, especially for applications requiring a cylindrical member, such as a rotating joint for an antenna feed. The physical dimensions of the guide are selected to provide for propagation in the *dominant* (lowest-order) mode.

Waveguide is not without its drawbacks, however. Rectangular or square guide constitutes a large windload surface, which places significant structural demands on a tower. Because of the physical configuration of rectangular and square guide, pressurization is limited, depending on the type of waveguide used (0.5 psi is typical). Excessive pressure can deform the guide shape and result in increased VSWR. Wind also may cause deformation and ensuing VSWR problems. These considerations have led to the development of circular and elliptical waveguide.

7.3.1 Propagation Modes

Propagation modes for waveguide fall into two broad categories:

- *Transverse-electric* (TE) waves
- *Transverse-magnetic* (TM) waves

With TE waves, the electric vector (*E vector*) is perpendicular to the direction of propagation. With TM waves, the magnetic vector (*H vector*) is perpendicular to the direction of propagation. These propagation modes take on integers (from 0 or 1 to infinity) that define field configurations. Only a limited number of these modes can be propagated, depending on the dimensions of the guide and the operating frequency.

Energy cannot propagate in waveguide unless the operating frequency is above the *cutoff frequency*. The cutoff frequency for rectangular guide is:

$$F_c = \frac{C}{2 \times a} \tag{7.10}$$

Where:
F_c = waveguide cutoff frequency
$c = 1.179 \times 10^{10}$ inches per second (the velocity of light)
a = the wide dimension of the guide

The cutoff frequency for circular waveguide is defined by the following equation:

$$F_c = \frac{c}{3.41 \times a'} \tag{7.11}$$

Where:
a' = the radius of the guide

There are four common propagation modes in waveguide:

- $TE_{0,1}$, the principal mode in rectangular waveguide.
- $TE_{1,0}$, also used in rectangular waveguide.
- $TE_{1,1}$, the principal mode in circular waveguide. $TE_{1,1}$ develops a complex propagation pattern with electric vectors curving inside the guide. This mode exhibits the lowest cutoff frequency of all modes, which allows a smaller guide diameter for a specified operating frequency.
- $TM_{0,1}$, which has a slightly higher cutoff frequency than $TE_{1,1}$ for the same size guide. Developed as a result of discontinuities in the waveguide, such as flanges and transitions, $TM_{0,1}$ energy is not coupled out by either dominant or cross-polar transitions. The parasitic energy must be filtered out, or the waveguide diameter chosen carefully to reduce the unwanted mode.

The field configuration for the dominant mode in rectangular waveguide is illustrated in Figure 7.10. Note that the electric field is vertical, with intensity maximum at the center of the guide and dropping off sinusoidally to zero intensity at the edges. The magnetic field is in the form of loops that lie in planes that are at right angles to the electric field (parallel to the top and bottom of the guide). The magnetic field distribution is the same for all planes perpendicular to the *Y*-axis. In the *X* direction, the intensity of the component of magnetic field that is transverse to the axis of the waveguide (the component in the direction of *X*) is at any point in the waveguide directly proportional to the intensity of the electric field at that point. This entire configuration of fields travels in the direction of the waveguide axis (the *Z* direction in Figure 7.10).

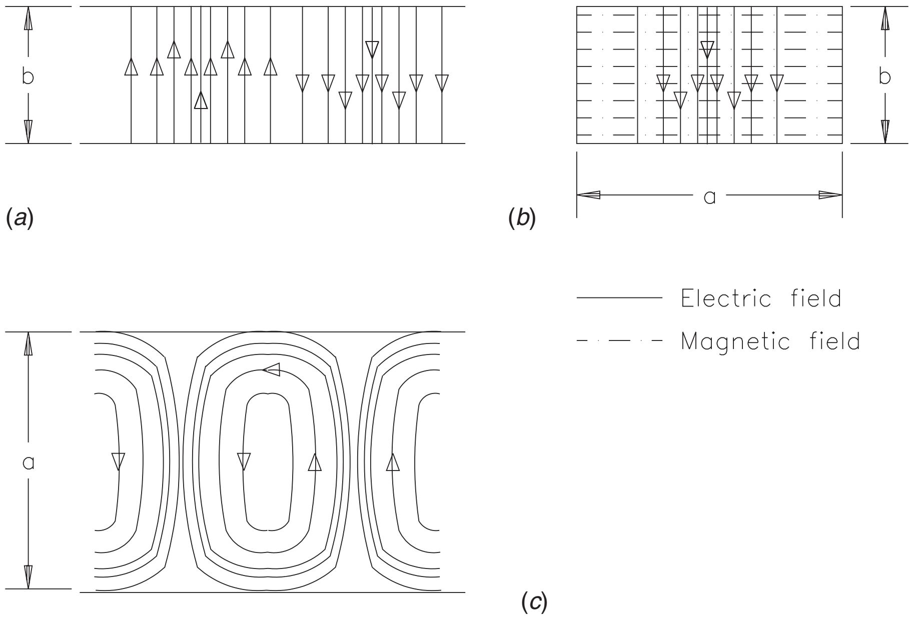


Figure 7.10 Field configuration of the dominant or $TE_{1,0}$ mode in a rectangular waveguide: (*a*) side view, (*b*) end view, (*c*) top view.

The field configuration for the $TE_{1,1}$ mode in circular waveguide is illustrated in Figure 7.11. The $TE_{1,1}$ mode has the longest cutoff wavelength and is, accordingly, the dominant mode. The next higher mode is the $TM_{0,1}$, followed by $TE_{2,1}$.

Dual-Polarity Waveguide

Waveguide will support dual-polarity transmission within a single run of line. A combining element (*dual-polarized transition*) is used at the beginning of the run, and a splitter (polarized transition) is used at the end of the line. Square waveguide has found numerous applications in such systems. Theoretically, the $TE_{1,0}$ and $TE_{0,1}$ modes are capable of propagation without cross coupling, at the same frequency, in lossless waveguide of square cross section. In practice, surface irregularities, manufacturing tolerances, and wall losses give rise to $TE_{1,0}$- and $TE_{0,1}$-mode cross conversion. Because this conversion occurs continuously along the waveguide, long guide runs usually are avoided in dual-polarity systems.

Efficiency

Waveguide losses result from the following:

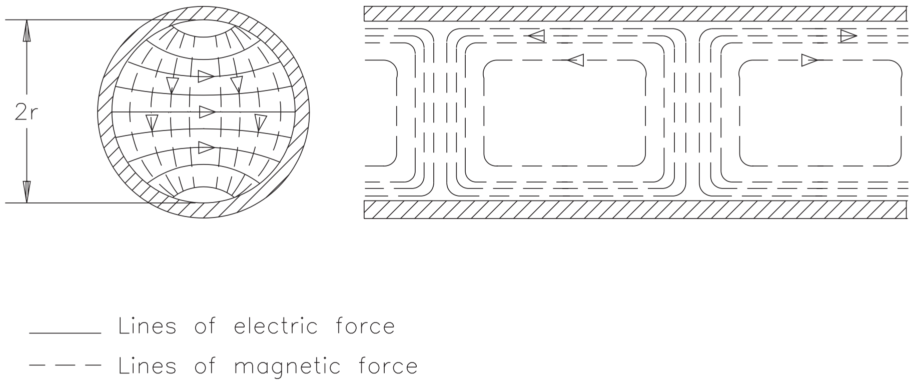


Figure 7.11 Field configuration of the dominant mode in circular waveguide.

- Power dissipation in the waveguide walls and the dielectric material filling the enclosed space
- Leakage through the walls and transition connections of the guide
- Localized power absorption and heating at the connection points

The operating power of waveguide may be increased through pressurization. Sulfur hexafluoride commonly is used as the pressurizing gas.

7.3.2 Ridged Waveguide

Rectangular waveguide may be ridged to provide a lower cutoff frequency, thereby permitting use over a wider frequency band. As illustrated in Figure 7.12, one- and two-ridged guides are used. Increased bandwidth comes at the expense of increased attenuation, relative to an equivalent section of rectangular guide.

7.3.3 Circular Waveguide

Circular waveguide offers several mechanical benefits over rectangular or square guide. The windload of circular guide is two-thirds that of an equivalent run of rectangular waveguide. It also presents lower and more uniform windloading than rectangular waveguide, reducing tower structural requirements.

The same physical properties of circular waveguide that give it good power handling and low attenuation also result in electrical complexities. Circular waveguide has two potentially unwanted modes of propagation: the cross-polarized $TE_{1,1}$ and $TM_{0,1}$ modes.

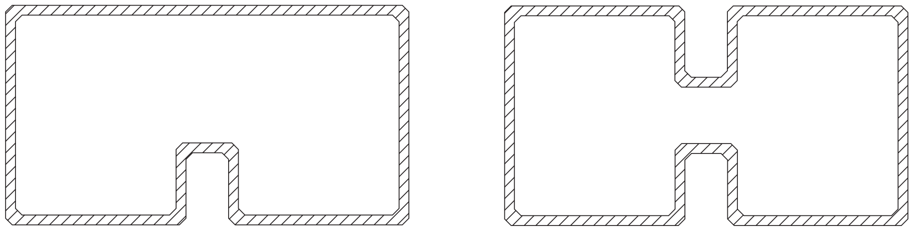


Figure 7.12 Ridged waveguide: (*a*) single-ridged, (*b*) double-ridged.

Circular waveguide, by definition, has no short or long dimension and, consequently, no method to prevent the development of cross-polar or *orthogonal* energy. Cross-polar energy is formed by small ellipticities in the waveguide. If the cross-polar energy is not trapped out, the parasitic energy can recombine with the dominant-mode energy.

Parasitic Energy

Hollow circular waveguide works as a high-Q resonant cavity for some energy and as a transmission medium for the rest. The parasitic energy present in the cavity formed by the guide will appear as increased VSWR if not disposed of. The polarization in the guide meanders and rotates as it propagates from the source to the load. The end pieces of the guide, typically circular-to-rectangular transitions, are polarization-sensitive. See Figure 7.13(*a*). If the polarization of the incidental energy is not matched to the transition, energy will be reflected.

Several factors can cause this undesirable polarization. One cause is out-of-round guides that result from nonstandard manufacturing tolerances. In Figure 7.13, the solid lines depict the situation at launching: perfectly circular guide with perpendicular polarization. However, certain ellipticities cause polarization rotation into unwanted states, while others have no effect. A 0.2 percent change in diameter can produce a –40 dB cross-polarization component per wavelength. This is roughly 0.03 in for 18 in of guide length.

Other sources of cross polarization include twisted and bent guides, out-of-roundness, offset flanges, and transitions. Various methods are used to dispose of this energy trapped in the cavity, including absorbing loads placed at the ground and/or antenna level.

7.3.4 Doubly Truncated Waveguide

The design of *doubly truncated waveguide* (DTW) is intended to overcome the problems that can result from parasitic energy in a circular waveguide. As shown in Figure

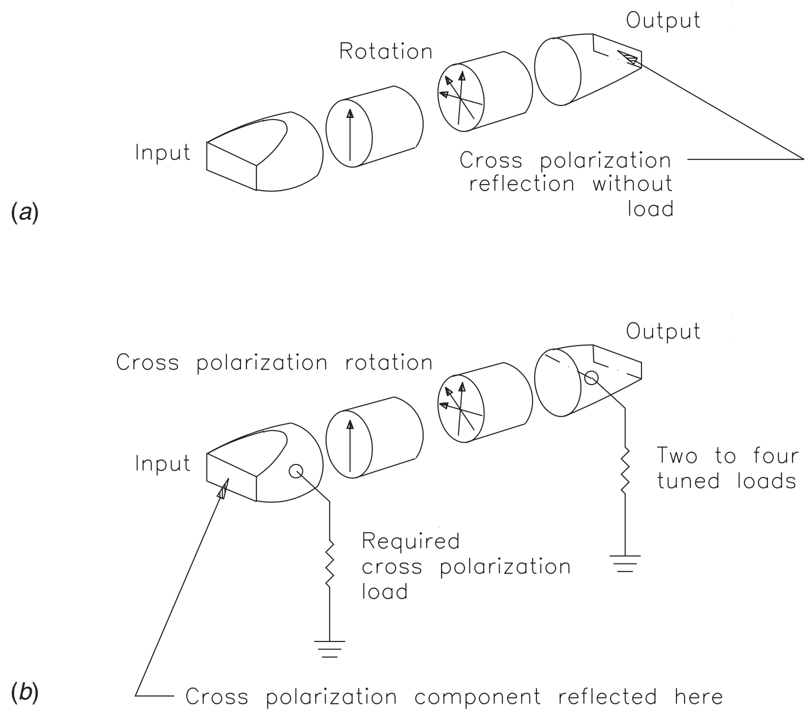


Figure 7.13 The effects of parasitic energy in circular waveguide: (*a*) trapped cross-polarization energy, (*b*) delayed transmission of the trapped energy.

7.14, DTW consists of an almost elliptical guide inside a circular shell. This guide does not support cross polarization; tuners and absorbing loads are not required. The low windload of hollow circular guide is maintained, except for the flange area.

Each length of waveguide is actually two separate pieces: a doubly truncated center section and a circular outer skin, joined at the flanges on each end. A large hole in the broadwall serves to pressurize the circular outer skin. Equal pressure inside the DTW and inside the circular skin ensures that the guide will not "breathe" or buckle as a result of rapid temperature changes.

DTW exhibits about 3 percent higher windloading than an equivalent run of circular waveguide (because of the transition section at the flange joints), and 32 percent lower loading than comparable rectangular waveguide.

7.3.5 Impedance Matching

The efficient flow of power from one type of transmission medium to another requires matching of the field patterns across the boundary to launch the wave into the second medium with a minimum of reflections. Coaxial line typically is matched into rectangular waveguide by extending the center conductor of the coax through the broadwall

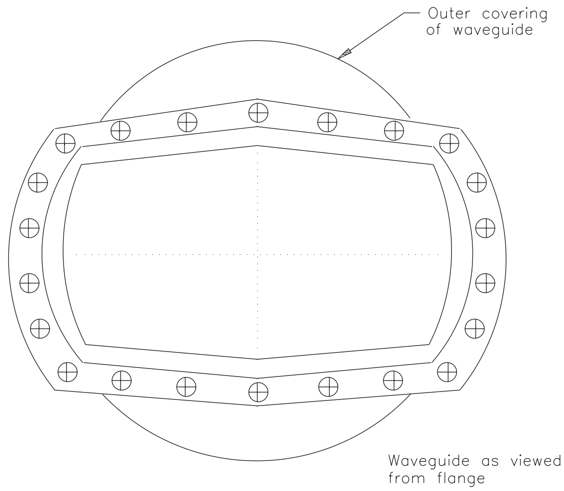


Figure 7.14 Physical construction of doubly truncated waveguide.

of the guide, parallel to the electric field lines across the guide. Alternatively, the center conductor can be formed into a loop and oriented to couple the magnetic field to the guide mode.

Standing waves are generally to be avoided in waveguide for the same reasons that they are to be avoided in transmission lines. Accordingly, it is usually necessary to provide impedance-matching systems in waveguides to eliminate standing waves. One approach involves the introduction of a compensating reflection in the vicinity of a load that neutralizes the standing waves that would exist in the system because of an imperfect match. A probe or tuning screw commonly is used to accomplish this, as illustrated in Figure 7.15. The tuning screw projects into the waveguide in a direction parallel to the electric field. This is equivalent to shunting a capacitive load across the guide. The susceptance of the load increases with extension into the guide up to 1/4 wavelength. When the probe is exactly 1/4 wavelength long, it becomes resonant and causes the guide to behave as though there were an open circuit at the point of the resonant probe. Probes longer than 1/4 wavelength but shorter than 3/4 wavelength introduce inductive loading. The extent to which such a probe projects into the waveguide determines the compensating reflection, and the position of the probe with respect to the standing wave pattern to be eliminated determines the phasing of the reflected wave.

Dielectric slugs produce an effect similar to that of a probe. The magnitude of the effect depends upon the following considerations:

- The dielectric constant of the slug
- Thickness of the slug in an axial direction
- Whether the slug extends entirely across the waveguide

The phase of the reflected wave is controlled by varying the axial position of the slug.

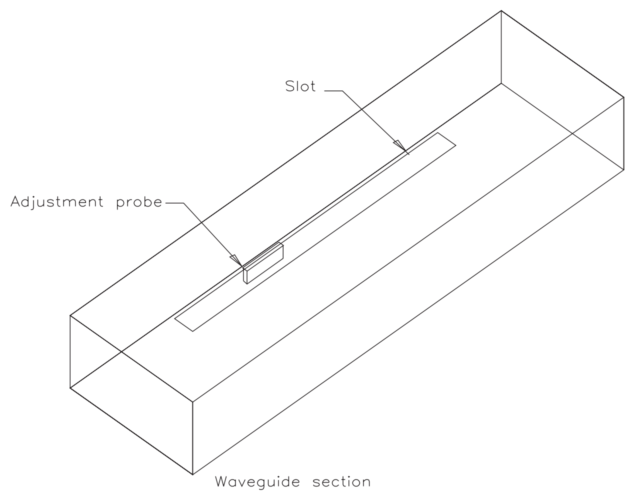


Figure 7.15 A probe configured to introduce a reflection in a waveguide that is adjustable in magnitude and phase.

There are several alternatives to the probe and slug for introducing controllable irregularity for impedance matching, including a metallic barrier or *window* placed at right angles to the axis of the guide, as illustrated in Figure 7.16. Three configurations are shown:

- The arrangement illustrated in Figure 7.16*a* produces an effect equivalent to shunting the waveguide with an inductive reactance.
- The arrangement shown in Figure 7.16*b* produces the effect of a shunt capacitive susceptance.
- The arrangement shown in Figure 7.16*c* produces an inductive shunt susceptance.

The waveguide equivalent of the coaxial cable tuning stub is a *tee* section, illustrated in Figure 7.17. The magnitude of the compensating effect is controlled by the position of the short-circuiting plug in the branch. The phase of the compensating reflected wave produced by the branch is determined by the position of the branch in the guide.

Waveguide Filters

A section of waveguide beyond cutoff constitutes a simple high-pass reflective filter. Loading elements in the form of posts or stubs may be employed to supply the reactances required for conventional lumped-constant filter designs.

Absorption filters prevent the reflection of unwanted energy by incorporating lossy material in secondary guides that are coupled through *leaky walls* (small sections of

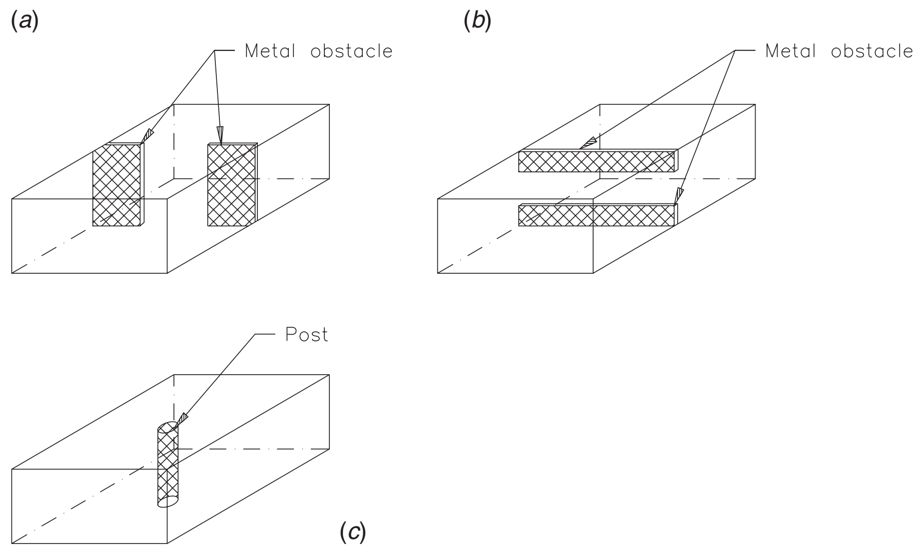


Figure 7.16 Waveguide obstructions used to introduce compensating reflections: (*a*) inductive window, (*b*) capacitive window, (*c*) post (inductive) element.

guide beyond cutoff in the passband). Such filters typically are used to suppress harmonic energy.

7.3.6 Installation Considerations

Waveguide system installation is both easier and more difficult than traditional transmission line installation. There is no inner conductor to align, but alignment pins must be set and more bolts are required per flange. Transition hardware to accommodate loads and coax-to-waveguide interfacing also is required.

Flange reflections can add up in phase at certain frequencies, resulting in high VSWR. The length of the guide must be chosen so that flange reflection buildup does not occur within the operating bandwidth.

Flexible sections of waveguide are used to join rigid sections or components that cannot be aligned otherwise. Flexible sections also permit controlled physical movement resulting from thermal expansion of the line. Such hardware is available in a variety of forms. Corrugated guide commonly is produced by shaping thin-wall seamless rectangular tubing. Flexible waveguide can accommodate only a limited amount of mechanical movement. Depending on the type of link, the manufacturer may specify a maximum number of bends.

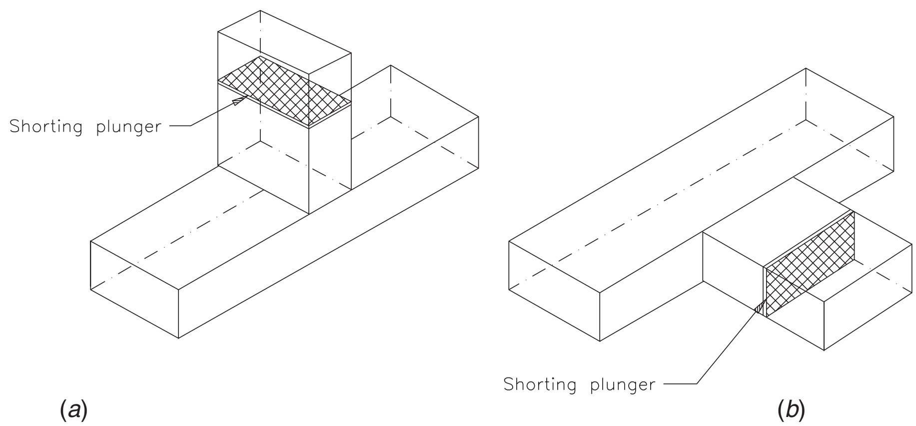


Figure 7.17 Waveguide stub elements used to introduce compensating reflections: (*a*) series *tee* element, (*b*) shunt *tee* element.

Tuning

Circular waveguide must be tuned. This requires a two-step procedure. First, the cross-polar $TE_{1,1}$ component is reduced, primarily through *axial ratio compensators* or *mode optimizers*. These devices counteract the net system ellipticity and indirectly minimize cross-polar energy. The cross-polar filters also may be rotated to achieve maximum isolation between the dominant and cross-polar modes. Cross-polar energy manifests itself as a net signal rotation at the end of the waveguide run. A perfect system would have a net rotation of zero.

In the second step, tuning slugs at both the top and bottom of the waveguide run are adjusted to reduce the overall system VSWR. Tuning waveguide can be a complicated and time-consuming procedure. Once set, however, tuning normally does not drift and must be repeated only if major component changes are made.

Waveguide Hardware

Increased use of waveguide has led to the development of waveguide-based hardware for all elements from the output of the RF generator to the load. Waveguide-based filters, elbows, directional couplers, switches, combiners, and diplexers are currently available. The RF performance of a waveguide component is usually better than the same item in coax. This is especially true in the case of diplexers and filters. Waveguide-based hardware provides lower attenuation and greater power-handling capability for a given physical size.

7.3.7 Cavity Resonators

Any space completely enclosed by conducting walls can contain oscillating electromagnetic fields. Such a cavity possesses certain frequencies at which it will resonate when excited by electrical oscillations. These *cavity resonators* find extensive use as resonant circuits at VHF and above. (The use of cavities in power grid tube amplifiers is discussed in Chapter 4.) Advantages of cavity resonators over conventional *LC* circuits include:

- Simplicity in design.
- Relatively large physical size compared with alternative methods of obtaining resonance. This attribute is important in high-power, high-frequency applications.
- High *Q*.
- The capability to configure the cavity to develop an extremely high shunt impedance.

Cavity resonators commonly are used at wavelengths on the order of 10 cm or less.

The simplest cavity resonator is a section of waveguide shorted at each end with a length *l* equal to:

$$l = \frac{\lambda_g}{2} \tag{7.12}$$

Where:
λ_g = the guide wavelength

This configuration results in a resonance similar to that of a 1/2-wavelength transmission line short-circuited at the receiving end.

A sphere or any other enclosed surface (irrespective of how irregular the outline) also may be used to form a cavity resonator.

Any given cavity is resonant at a number of frequencies, corresponding to the different possible field conditions that can exist within the space. The resonance having the longest wavelength (lowest frequency) is termed the *dominant* or *fundamental resonance*. The resonant wavelength is proportional to the size of the resonator. If all dimensions are doubled, the wavelength corresponding to resonance will likewise be doubled. The resonant frequency of a cavity can be changed by incorporating one or more of the following mechanisms:

- Altering the mechanical dimensions of the cavity. Small changes can be achieved by flexing walls, but large changes require some form of sliding member.
- Coupling reactance into the resonator through a coupling loop.
- Introducing a movable copper paddle into the cavity. A paddle placed inside the resonator will affect the normal distribution of flux and tend to raise the resonant frequency by an amount determined by the orientation of the paddle.

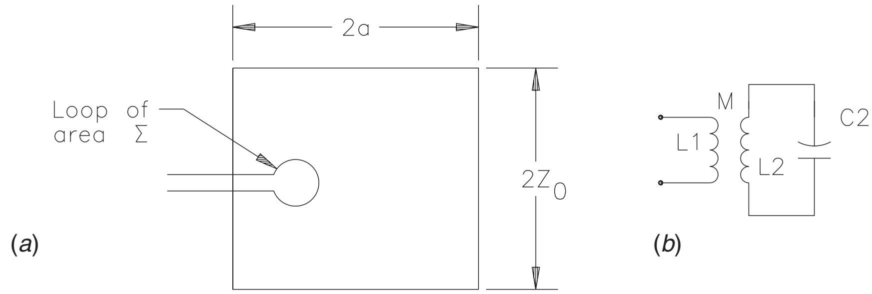


Figure 7.18 Cavity resonator coupling: (*a*) coupling loop, (*b*) equivalent circuit.

The Q of a cavity resonator has the same significance as for a conventional resonant circuit. Q can be defined for a cavity by the following relationship:

$$Q = 2\pi \frac{E_s}{E_l} \tag{7.13}$$

Where:
E_s = energy stored
E_l = energy lost per cycle

The energy stored is proportional to the square of the magnetic flux density integrated throughout the volume of the resonator. The energy lost in the walls is proportional to the square of the magnetic flux density integrated over the surface of the cavity. To obtain high Q, the resonator should have a large ratio of volume to surface area, because it is the volume that stores energy and the surface area that dissipates energy.

Coupling can be obtained from a resonator by means of a coupling loop or coupling electrode. Magnetic coupling is accomplished through the use of a loop oriented so as to enclose magnetic flux lines existing in the desired mode of operation. This technique is illustrated in Figure 7.18. A current passed through the loop will excite oscillations of this mode. Conversely, oscillations existing in the resonator will induce a voltage in the coupling loop. The magnitude of the coupling can be controlled by rotating the loop; the coupling reduces to zero when the plane of the loop is parallel to the magnetic flux.

Coupling of a resonator also may be accomplished through the use of a probe or opening in one wall of the cavity.

7.4 RF Combiner and Diplexer Systems

The basic purpose of an RF combiner is to add two or more signals to produce an output signal that is a composite of the inputs. The combiner performs this signal addition while providing isolation between inputs. Combiners perform other functions as well, and can be found in a wide variety of RF equipment utilizing power vacuum tubes. Combiners are valuable devices because they permit multiple amplifiers to drive a single load. The isolation provided by the combiner permits tuning adjustments to be made on one amplifier—including turning it on or off—without significantly affecting the operation of the other amplifier. In a typical application, two amplifiers drive the hybrid and provide two output signals:

- A combined output representing the sum of the two input signals, typically directed toward the antenna.
- A difference output representing the difference in amplitude and phase between the two input signals. The difference output typically is directed toward a dummy (reject) load.

For systems in which more than two amplifiers must be combined, two or more combiners are cascaded.

Diplexers are similar in nature to combiners but permit the summing of output signals from two or more amplifiers operating at different frequencies. This allows, for example, the outputs of several transmitters operating on different frequencies to utilize a single broadband antenna.

7.4.1 Passive Filters

A *filter* is a multiport-network designed specifically to respond differently to signals of different frequency [1]. This definition excludes *networks*, which incidentally behave as filters, sometimes to the detriment of their main purpose. Passive filters are constructed exclusively with passive elements (i.e., resistors, inductors, and capacitors). Filters are generally categorized by the following general parameters:

- Type
- Alignment (or class)
- Order

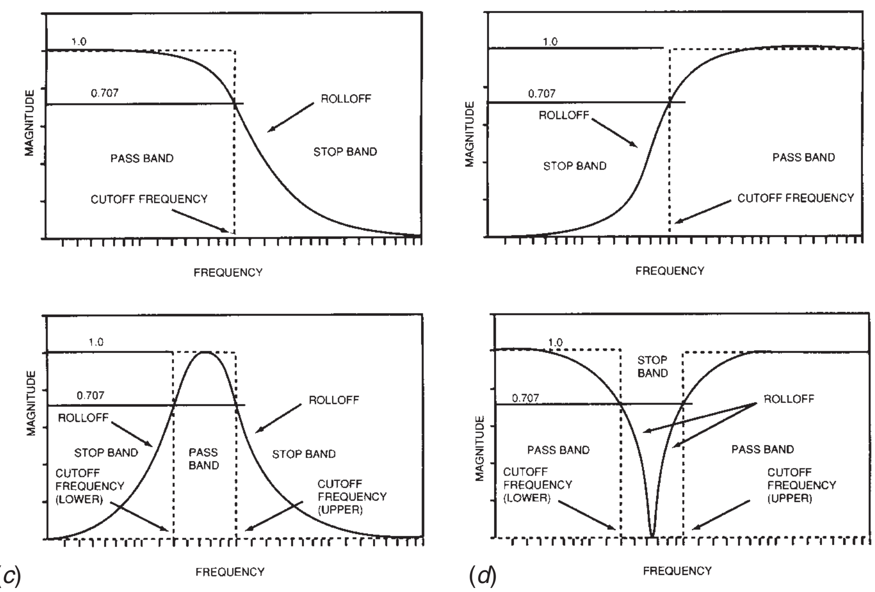
Filter Type

Filters are categorized by type, according to the magnitude of the frequency response, as one of the following [1]:

- *Low-pass* (LP)
- *High-pass* (HP)
- *Band-pass* (BP)
- *Band-stop* (BS).

The terms *band-reject* or *notch* are also used as descriptive of the BS filter. The term *all-pass* is sometimes applied to a filter whose purpose is to alter the phase angle without affecting the magnitude of the frequency response. Ideal and practical interpretations of the types of filters and the associated terminology are illustrated in Figure 7.19.

(*a*) (*b*)



(*c*) (*d*)

Figure 7.19 Filter characteristics by type: (*a*) low-pass, (*b*) high-pass, (*c*) bandpass, (*d*) bandstop. (*From* [1]. *Used with permission*.)

In general, the voltage gain of a filter in the *stop band* (or *attenuation band*) is less than $\sqrt{2}/2\,(\approx\ 0.707)$ times the maximum voltage gain in the pass band. In logarithmic terms, the gain in the stop band is at least 3.01 dB less than the maximum gain in the pass band. The cutoff (*break* or *corner*) frequency separates the pass band from the stop band. In BP and BS filters, there are two cutoff frequencies, sometimes referred to as the *lower* and *upper* cutoff frequencies. Another expression for the cutoff frequency is *half-power frequency*, because the power delivered to a resistive load at cutoff frequency is one-half the maximum power delivered to the same load in the pass band. For BP and BS filters, the center frequency is the frequency of maximum or minimum response magnitude, respectively, and bandwidth is the difference between the upper and lower cutoff frequencies. *Rolloff* is the transition from pass band to stop band and is specified in gain unit per frequency unit (e.g., gain unit/Hz, dB/decade, dB/octave, etc.)

Filter Alignment

The *alignment* (or class) of a filter refers to the shape of the frequency response [1]. Fundamentally, filter alignment is determined by the coefficients of the filter network

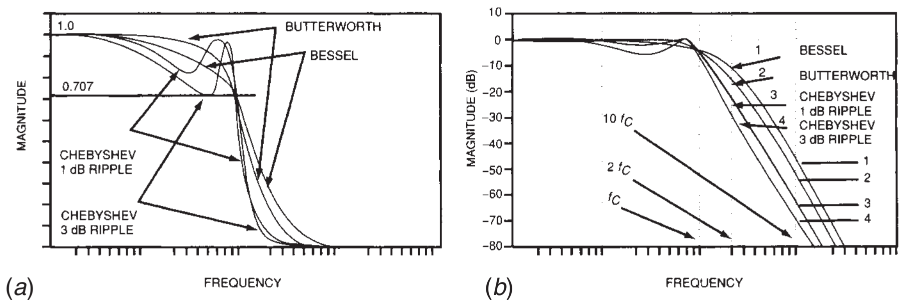


Figure 7.20 Filter characteristics by alignment, third-order, all-pole filters: (*a*) magnitude, (*b*) magnitude in decibels. (*From* [1]. *Used with permission.*)

transfer function, so there are an indefinite number of filter alignments, some of which may not be realizable. The more common alignments are:

- Butterworth
- Chebyshev
- Bessel
- Inverse Chebyshev
- Elliptic (or Cauer)

Each filter alignment has a frequency response with a characteristic shape, which provides some particular advantage. (See Figure 7.20.) Filters with Butterworth, Chebyshev, or Bessel alignment are called *all-pole filters* because their low-pass transfer functions have no zeros. Table 7.3 summarizes the characteristics of the standard filter alignments.

Filter Order

The *order* of a filter is equal to the number of poles in the filter network transfer function [1]. For a lossless *LC* filter with resistive (nonreactive) termination, the number of reactive elements (inductors or capacitors) required to realize a LP or HP filter is equal to the order of the filter. Twice the number of reactive elements are required to realize a BP or a BS filter of the same order. In general, the order of a filter determines the slope of the rolloff—the higher the order, the steeper the rolloff. At frequencies greater than approximately one octave above cutoff (i.e., $f > 2 f_c$), the rolloff for all-pole filters is $20n$ dB/decade (or approximately $6n$ dB/octave), where n is the

Table 7.3 Summary of Standard Filter Alignments (*After* [1].)

Alignment	Pass Band Description	Stop Band Description	Comments
Butterworth	Monotonic	Monotonic	All-pole; maximally flat
Chebyshev	Rippled	Monotonic	All-pole
Bessel	Monotonic	Monotonic	All-pole; constant phase shift
Inverse Chebyshev	Monotonic	Rippled	
Elliptic (or Cauer)	Rippled	Rippled	

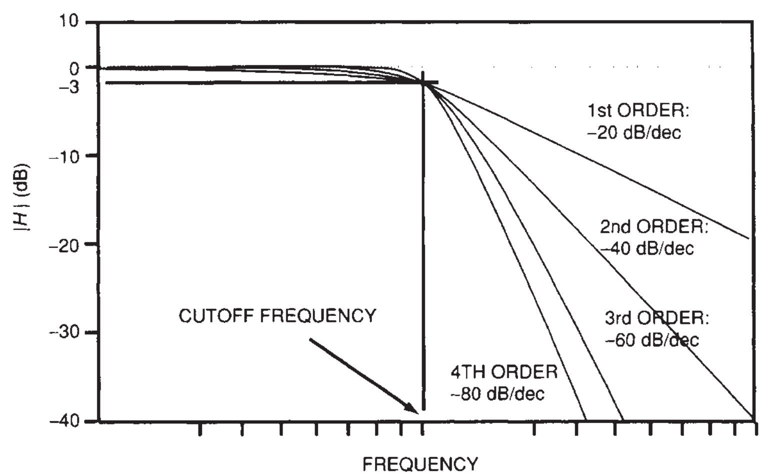


Figure 7.21 The effects of filter order on rolloff (Butterworth alignment). (*From* [1]. *Used with permission.*)

order of the filter (Figure 7.21). In the vicinity of f_c, both filter alignment and filter order determine rolloff.

7.4.2 Four-Port Hybrid Combiner

A hybrid combiner (coupler) is a reciprocal four-port device that can be used for either splitting or combining RF energy over a wide range of frequencies. An exploded view of a typical 3 dB 90° hybrid is illustrated in Figure 7.22. The device consists of two identical parallel transmission lines coupled over a distance of approximately one-quarter wavelength and enclosed within a single outer conductor. Ports at the

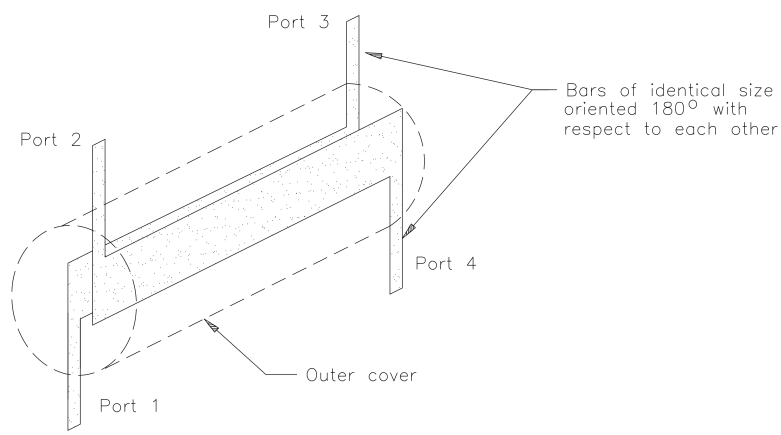


Figure 7.22 Physical model of a 90° hybrid combiner.

same end of the coupler are in phase, and ports at the opposite end of the coupler are in quadrature (90° phase shift) with respect to each other.

The phase shift between the two inputs or outputs is always 90° and is virtually independent of frequency. If the coupler is being used to combine two signals into one output, these two signals must be fed to the hybrid in phase quadrature. When the coupler is used as a power splitter, the division is equal (half-power between the two ports). The hybrid presents a constant impedance to match each source.

Operation of the combiner can best be understood through observation of the device in a practical application. Figure 7.23 shows a four-port hybrid combiner used to add the outputs of two transmitters to feed a single load. The combiner accepts one RF source and splits it equally into two parts. One part arrives at output port *C* with 0° phase (no phase delay; it is the *reference phase*). The other part is delayed by 90° at port *D*. A second RF source connected to input port *B*, but with a phase delay of 90°, also will split in two, but the signal arriving at port *C* now will be in phase with source 1, and the signal arriving at port *D* will cancel, as shown in the figure.

Output port *C*, the summing point of the hybrid, is connected to the load. Output port *D* is connected to a resistive load to absorb any residual power resulting from slight differences in amplitude and/or phase between the two input sources. If one of the RF inputs fails, half of the remaining transmitter output will be absorbed by the resistive load at port *D*.

The four-port hybrid works only when the two signals being mixed are identical in frequency and amplitude, and when their relative phase is 90°.

Operation of the hybrid can best be described by a *scattering matrix* in which vectors are used to show how the device operates. Such a matrix is shown in Table 7.4. In a 3 dB hybrid, two signals are fed to the inputs. An input signal at port 1 with 0° phase will arrive in phase at port 3, and at port 4 with a 90° lag (–90°) referenced to port 1. If the signal at port 2 already contains a 90° lag (–90° referenced to port 1), both input signals will combine in phase at port 4. The signal from port 2 also experiences another 90°

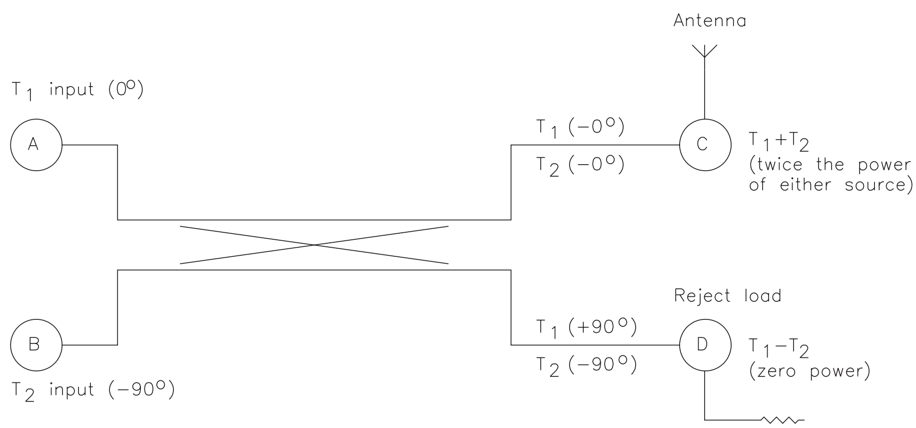


Figure 7.23 Operating principles of a hybrid combiner. This circuit is used to add two identical signals at inputs A and B.

change in the hybrid as it reaches port 3. Therefore, the signals from ports 1 and 2 cancel each other at port 3.

If the signal arriving at port 2 leads by 90° (mode 1 in the table), the combined power from ports 1 and 2 appears at port 4. If the two input signals are matched in phase (mode 4), the output ports (3 and 4) contain one-half of the power from each of the inputs.

If one of the inputs is removed, which would occur in a transmitter failure, only one hybrid input receives power (mode 5). Each output port then would receive one-half the input power of the remaining transmitter, as shown.

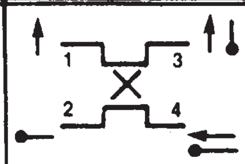
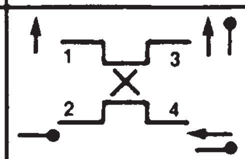
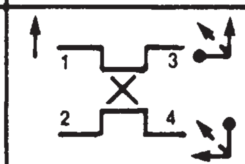
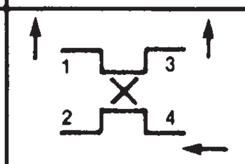
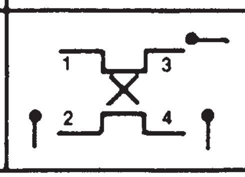
The input ports present a predictable load to each amplifier with a VSWR that is lower than the VSWR at the output port of the combiner. This characteristic results from the action of the difference port, typically connected to a dummy load. Reflected power coming into the output port will be directed to the reject load, and only a portion will be fed back to the amplifiers. Figure 7.24 illustrates the effect of output port VSWR on input port VSWR, and on the isolation between ports.

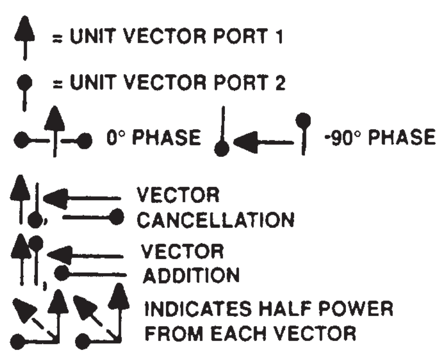
As noted previously, if the two inputs from the separate amplifiers are not equal in amplitude and not exactly in phase quadrature, some power will be dissipated in the difference port reject load. Figure 7.25 plots the effect of power imbalance, and Figure 7.26 plots the effects of phase imbalance. The power lost in the reject load can be reduced to a negligible value by trimming the amplitude and/or phase of one (or both) amplifiers.

7.4.3 Non-Constant-Impedance Diplexer

Diplexers are used to combine amplifiers operating on different frequencies (and at different power levels) into a single output. Such systems typically are utilized to sum different transmitter outputs to feed a single broadband antenna.

Table 7.4 Single 90° Hybrid System Operating Modes

MODE	INPUT 1	INPUT 2	SCHEMATIC	OUTPUT 3	OUTPUT 4
1	$P_1 \angle 0°$	$P_2 \angle -90°$		0	$P_1 + P_2$
2	$P_1 \angle 0°$	$P_2 \angle 90°$		$P_1 + P_2$	0
3	$P_1 \angle 0°$	$P_2 \angle 0°$		$P_{1/2} + P_{2/2}$	$P_{1/2} + P_{2/2}$
4	$P_1 \angle 0°$	$P_2 = 0$		$P_{1/2}$	$P_{1/2}$
5	$P_1 = 0$	$P_2 \angle 0°$		$P_{2/2}$	$P_{2/2}$



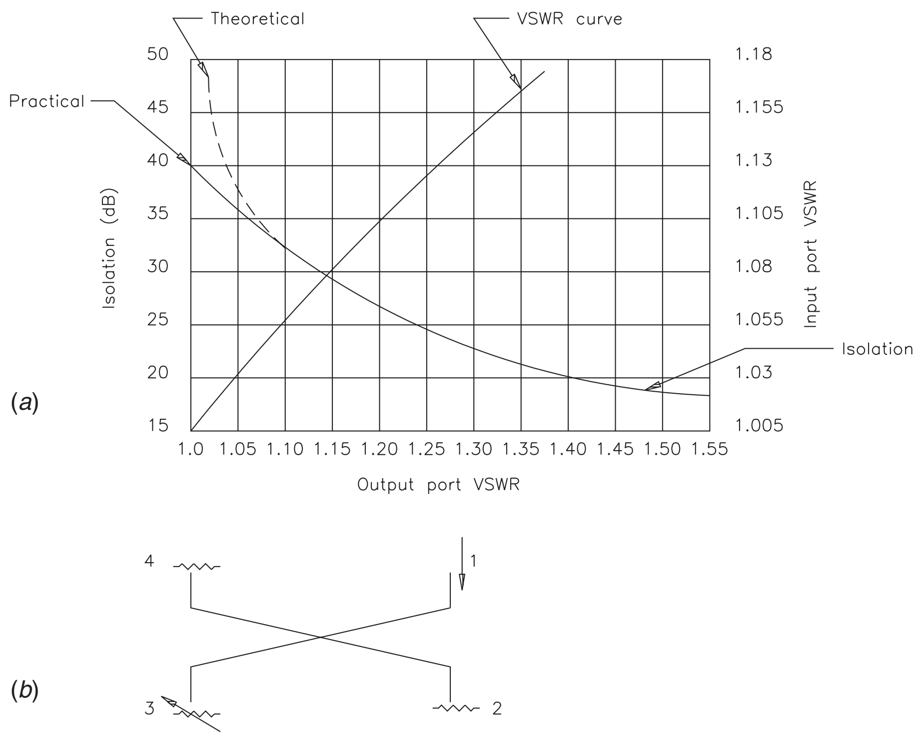


Figure 7.24 The effects of load VSWR on input VSWR and isolation: (*a*) respective curves, (*b*) coupler schematic.

The *branch diplexer* is the typical configuration for a diplexer that does not exhibit constant-impedance inputs. As shown in Figure 7.27, the branch diplexer consists of two banks of filters each feeding into a coaxial tee. The electrical length between each filter output and the centerline of the tee is frequency-sensitive, but this fact is more of a tuning nuisance than a genuine user concern.

For this type of diplexer, all of the electrical parameters are a function of the filter characteristics. The VSWR, insertion loss, group delay, and rejection/isolation will be the same for the overall system as they are for the individual banks of cavities. The major limitation of this type of combiner is the degree of isolation that can be obtained for closely spaced channels.

7.4.4 Constant-Impedance Diplexer

The constant-impedance diplexer employs 3 dB hybrids and filters with a terminating load on the isolated port. The filters in this type of combiner can be either notch-type or bandpass-type. The performance characteristics are noticeably different for each design.

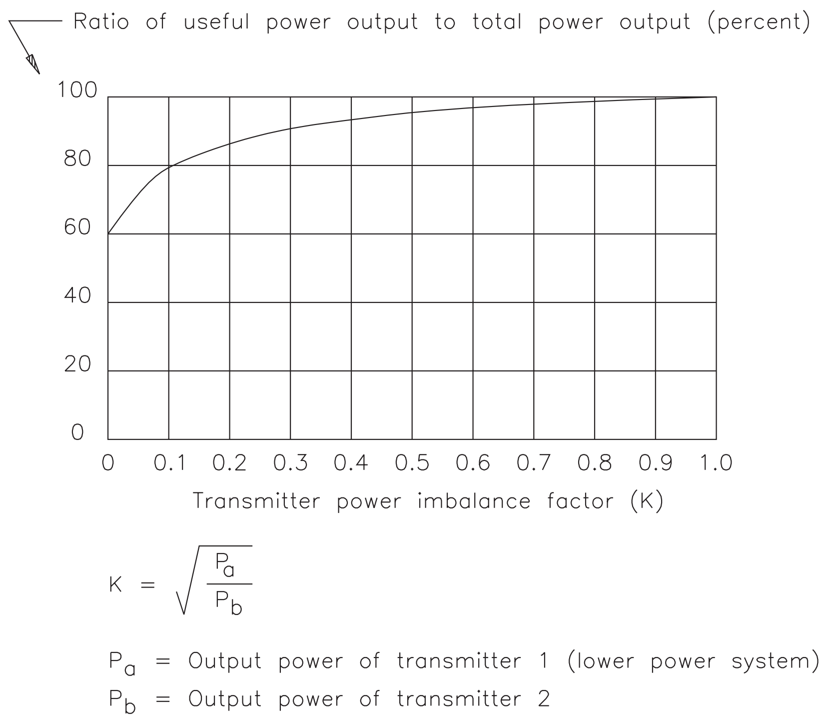


Figure 7.25 The effects of power imbalance at the inputs of a hybrid coupler.

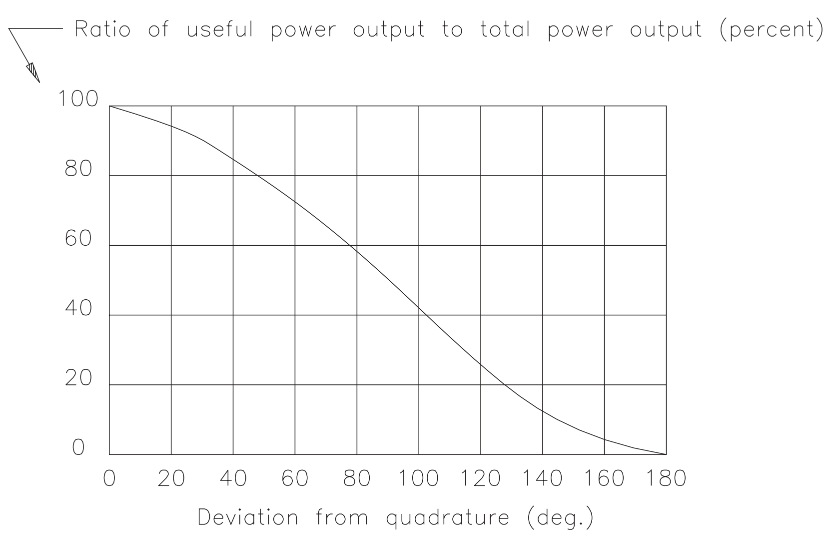


Figure 7.26 Phase sensitivity of a hybrid coupler.

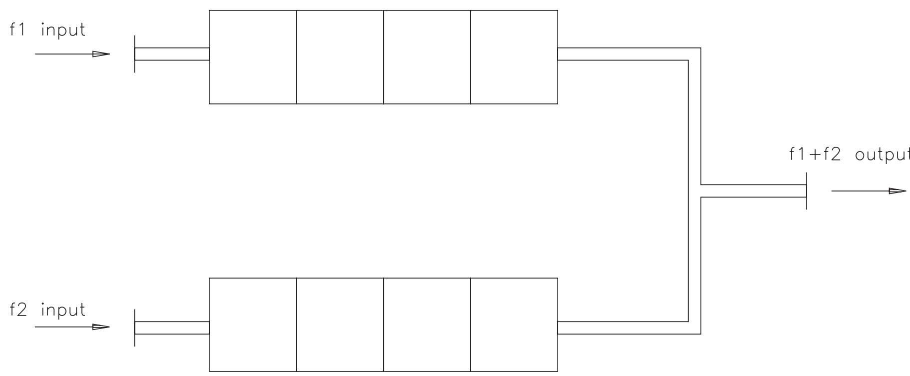


Figure 7.27 Non-constant-impedance branch diplexer. In this configuration, two banks of filters feed into a coaxial tee.

Band-Stop Diplexer

The band-stop (notch) constant-impedance diplexer is configured as shown in Figure 7.28. For this design, the notch filters must have a high Q response to keep insertion loss low in the passband skirts. The high Q characteristic results in a sharp notch. Depending on the bandwidth required of the diplexer, two or more cavities may be located in each leg of the diplexer. They are typically stagger tuned, one high and one low for the two-cavity case. With this dual-cavity reject response in each leg of the band-stop diplexer system, the following analysis explains the key performance specifications.

If frequency f_1 is fed into the top left port of Figure 7.28, it will be split equally into the upper and lower legs of the diplexer. Both of these signals will reach the filters in their respective leg and be rejected/reflected back toward the input hybrid, recombine, and emerge through the lower left port, also known as the *wideband output*.

The VSWR looking into the f_1 input is near 1:1 at all frequencies in the band. Within the bandwidth of the reject skirts, the observed VSWR is equal to the termination of the wideband output. Outside of the passband, the signals will pass by the cavities, enter the rightmost hybrid, recombine, and emerge into the dummy load. Consequently, the out-of-band VSWR is, in fact, the VSWR of the load.

The insertion loss from the f_1 input to the wideband output is low, typically on the order of 0.1 dB at carrier. This insertion loss depends on perfect reflection from the cavities. As the rejection diminishes on the skirts of the filters, the insertion loss from f_1 to the wideband output increases.

The limitation in reject bandwidth of the cavities causes the insertion loss to rise at the edges of the passband. The isolation from f_1 to the wideband input consists of a combination of the reject value of the cavities plus the isolation of the rightmost hybrid.

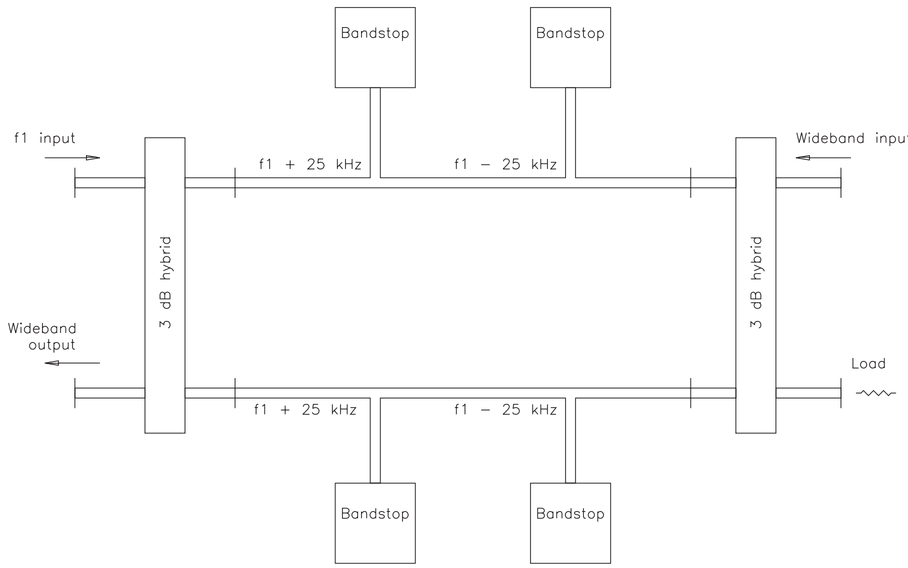


Figure 7.28 Band-stop (notch) constant-impedance diplexer module. This design incorporates two 3 dB hybrids and filters, with a terminating load on the isolated port.

A signal entering at f_1 splits and proceeds in equal halves rightward through both the upper and lower legs of the diplexer. It is rejected by the filters at carrier and, to a lesser extent, on both sides of the carrier. Any residual signal that gets by the cavities reaches the rightmost hybrid. There it recombines and emerges from the *load port*. The hybrid provides a specified isolation from the load port to the wideband input port. This hybrid isolation must be added to the filter rejection to obtain the total isolation from the f_1 input to the wideband input.

If a signal is fed into the wideband input of the combiner shown in Figure 7.28, the energy will split equally and proceed leftward along the upper and lower legs of the diplexer. Normally, f_1 is not fed into the wideband input. If it were, f_1 would be rejected by the notch filters and recombine into the load. All other signals sufficiently removed from f_1 will pass by the cavities with minimal insertion loss and recombine into the wideband output.

The VSWR looking into the wideband input is equal to the VSWR at the output for frequencies other than f_1. If f_1 were fed into the wideband input, the VSWR would be equal to the VSWR of the load.

Isolation from the wideband input to the f_1 input is simply the isolation available in the leftmost 3 dB hybrid. This isolation is usually inadequate for high-power applications. To increase the isolation from the wideband input to the f_1 input, it is necessary to use additional cavities between the f_1 transmitter and the f_1 input that will reject the frequencies fed into the wideband input. Unfortunately, adding these cavities to the input

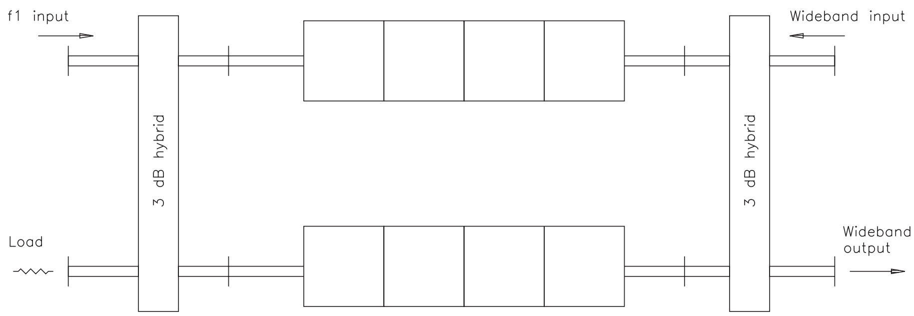


Figure 7.29 Bandpass constant-impedance diplexer.

line also cancels the constant-impedance input. Although the notch diplexer as shown in the figure is truly a constant-impedance type of diplexer, the constant impedance is presented to the two inputs by virtue of using the hybrids at the respective inputs.

The hybrids essentially cause the diplexer to act as an absorptive type of filter to all out-of-band signals. The out-of-band signals generated by the transmitter are absorbed by the load rather than reflected to the transmitter.

If a filter is added to the input to supplement isolation, the filter will reflect some out-of-band signals back at the transmitter. The transmitter then will be seeing the passband impedance of this supplemental filter rather than the constant impedance of the notch diplexer. This is a serious deficiency for applications that require a true constant-impedance input.

Bandpass Constant-Impedance Diplexer

The bandpass constant-impedance diplexer is shown in Figure 7.29. This system takes all of the best features of diplexers and combines them into one unit. It also provides a constant-impedance input that need not be supplemented with input cavities that rob the diplexer of its constant-impedance input.

The bandpass filters exhibit good bandwidth, providing near 1:1 VSWR across the operating bandpass. Insertion is low at carrier (0.28 dB is typical), rising slightly at bandpass extremes. Diplexer rejection, when supplemented by isolation of the hybrids, provides ample transmitter-to-transmitter isolation. Group delay is typically exceptional, providing performance specifications similar to those of a branch-style bandpass system. This configuration has the additional capability of providing high port-to-port isolation between closely spaced operating channels, as well as a true constant-impedance input.

The hybrids shown in Figure 7.29 work in a manner identical to those described for a band-stop diplexer. However, the bandpass filters cause the system to exhibit performance specifications that exceed the band-stop system in every way. Consider a signal entering at the f_1 input.

Within the passbands of the filters, which are tuned to f_1, the VSWR will be near 1:1 at carrier, rising slightly at the bandpass extremes. Because of the characteristic of the leftmost hybrid, the VSWR is, in fact, a measure of the similarity of response of the top and bottom bands of filters. The insertion loss looking from the f_1 input to the wideband output will be similar to the insertion loss of the top and bottom filters individually. A value of approximately 0.28 dB at carrier is typical.

Both the insertion loss and group delay can be determined by the design bandwidth of the filters. Increasing bandwidth causes the insertion loss and group delay deviation to decrease. Unfortunately, as the bandwidth increases with a given number of cavities, the isolation suffers for closely spaced channels because the reject skirt of the filter decreases with increasing bandwidth.

Isolation of f_1 to the wideband input is determined as follows: A signal enters at the f_1 input, splits equally into the upper and lower banks of filters, passes with minimal loss through the filters, and recombines into the wideband output of the rightmost hybrid. Both the load and the wideband input ports are isolated by their respective hybrids to some given value below the f_1 input level. Isolation of the f_1 input to the wideband input is supplemented by the reject skirt of the next module.

A signal fed into the wideband input could be any frequency removed from f_1 by some minimal amount. As the wideband signal enters, it will be split into equal halves by the hybrid, then proceed to the left until the two components reach the reject skirts of the filters. The filters will shunt all frequencies removed from f_1 by some minimal amount. If the shunt energy is in phase for the given frequency when the signal is reflected back to the right hybrid, it will recombine into the wideband output. The VSWR under these conditions will be equal to the termination at the wideband output. If the reject skirts of the filter are sufficient, the insertion loss from wideband input to wideband output will be minimal (0.03 dB is typical).

The isolation from the wideband input to f_1 can be determined as follows: A signal enters at the wideband input, splits equally into upper and lower filters, and is rejected by the filters. Any residual signal that passes through the filters in spite of the rejection still will be in the proper phase to recombine into the load, producing additional isolation to the f_1 input port. Thus, the isolation of the wideband input to the f_1 input is the sum of the rejection from the filters and from the left hybrid.

Isolation of f_1 to f_2

Extending the use of the diplexer module into a multiplexer application supplements the deficient isolation described previously (narrowband input performance), while maintaining the constant-impedance input. In a multiplexer system, the wideband input of one module is connected to the wideband output of the next module, as illustrated in Figure 7.30.

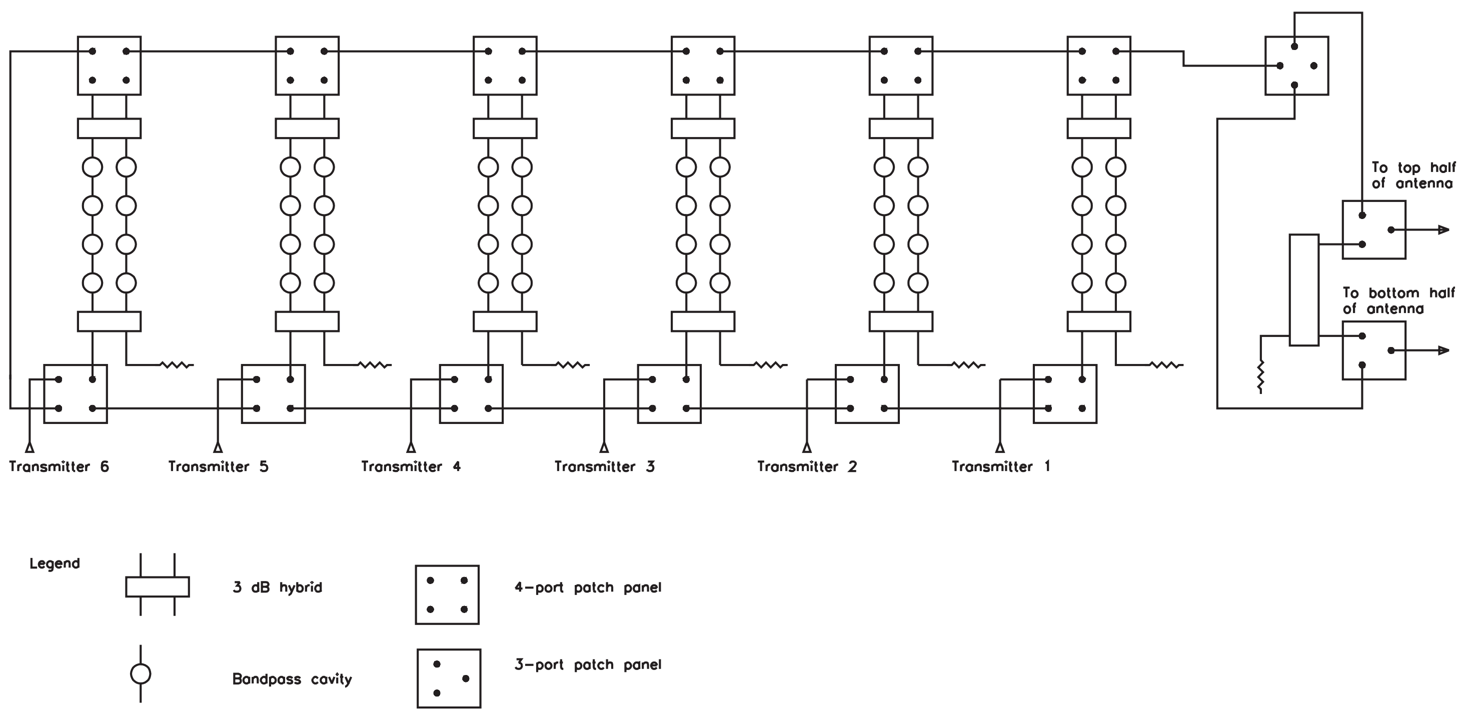


Figure 7.30 Schematic diagram of a six-module bandpass multiplexer. This configuration accommodates a split antenna design and incorporates patch panels for bypass purposes.

It has already been stated that the isolation from the f_1 input to the wideband input is deficient, but additional isolation is provided by the isolation of the wideband output to the f_2 input of the next module. Consider that f_1 has already experienced 30 dB isolation to the wideband input of the same module. When this signal continues to the next module through the wideband output of module 2, it will be split into equal halves and proceed to the left of module 2 until it reaches the reject skirts of the filters in module 2. Assume that these filters are tuned to f_2 and reject f_1 by at least 25 dB. The combined total isolation of f_1 to f_2 is the sum of the 30 dB of the right hybrid in module 1, plus the 25 dB of the reject skirts of module 2, for a total of 55 dB.

Intermodulation Products

The isolation just described is equal in magnitude to that for a band-stop module, but it provides further protection against the generation of intermodulation products. The most troublesome intermodulation (*intermod*) products usually occur when an incoming (secondary) signal mixes with the second harmonic of a primary transmitter. (The properties of intermodulation are discussed in more detail in Chapter 10.)

When the primary transmitter is operating on frequency A, the intermod will occur at the frequency $(2 \times A) - B$. This formula invariably places the intermod from the primary transmitter symmetrically about the operating frequency. By an interesting coincidence, the bandpass filters in the bandpass module also provide symmetrical reject response on both sides of the primary operating frequency.

Assume again that the incoming signal is attenuated by 30 dB in the respective hybrid and by 25 dB in the filter, for a total of 55 dB. If an intermod still is generated in spite of this isolation, it will emerge on the other skirt of the filter attenuated by 25 dB. In the bandpass system, the incoming signal is attenuated by 55 dB and the resulting outgoing spur by 25 dB, for a total of 80 dB suppression.

Interestingly, the entire 80 dB of attenuation is supplied by the diplexer regardless of the *turnaround loss* of the transmitter. The tendency toward wideband final stage amplifiers in transmitters requires constant-impedance inputs. The transmitters also require increased isolation because they offer limited turnaround loss.

Group Delay

Group delay in the bandpass multiplexer module is equal to the sum of the narrowband input group delay and the wideband input group delay of all modules between the input and the load (antenna). (Group delay is discussed in detail in Chapter 10.) The narrowband input group delay is a U-shaped response, with minimum at center and rising to a maximum on both sides at the frequency where the reject rises to 3 dB. Group delay then decreases rapidly at first, then more slowly.

If the bandwidth of the pass band is made such that the group delay is ±25 ns over ±150 kHz (in an example system operating near 100 MHz), the 3 dB points will be at ±400 kHz and the out-of-band group delay will fall rapidly at ±800 kHz and possibly ±1.0 MHz. If there are no frequencies 800 kHz or 1.0 MHz removed upstream in other modules, this poses no problem.

If modules upstream are tuned to 800 kHz or 1.0 MHz on either side, then the group delay (when viewed at the upstream module) will consist of its own narrowband input group delay plus the rapidly falling group delay of the wideband input of the closely spaced downstream module. Under these circumstances, if good group delay is desired, it is possible to utilize a *group delay compensation* module.

A group delay compensation module consists of a hybrid and two cavities used as notch cavities. It typically provides a group delay response that is inverted, compared with a narrowband input group delay. Because group delay is additive, the inverted response subtracts from the standard response, effectively reducing the group delay deviation.

It should be noted that the improvement in group delay is obtained at a cost of insertion loss. In large systems (8 to 10 modules), the insertion loss can be high because of the cumulative total of all wideband losses. Under these conditions, it may be more prudent to accept higher group delay and retain minimal insertion loss.

7.4.5 Microwave Combiners

Hybrid combiners typically are used in microwave amplifiers to add the output energy of individual power modules to provide the necessary output from an RF generator. Quadrature hybrids effect a VSWR-canceling phenomenon that results in well-matched power amplifier inputs and outputs that can be broadbanded with proper selection of hybrid tees. Several hybrid configurations are possible, including the following:

- *Split-tee*
- *Branch-line*
- *Magic-tee*
- *Backward-wave*

Key design parameters include coupling bandwidth, isolation, and ease of fabrication. The equal-amplitude quadrature-phase reverse-coupled TEM 1/4-wave hybrid is particularly attractive because of its bandwidth and amenability to various physical implementations. Such a device is illustrated in Figure 7.31.

7.4.6 Hot Switching Combiners

Switching RF is nothing new. Typically, the process involves coaxial switches, coupled with the necessary logic to ensure that the “switch” takes place with no RF energy on the contacts. This process usually takes the system off-line for a few seconds while the switch is completed. Through the use of hybrid combiners, however, it is possible to redirect RF signals without turning the carrier off. This process is referred to as *hot switching*. Figure 7.32 illustrates two of the most common switching functions (SPST and DPDT) available from hot switchers.

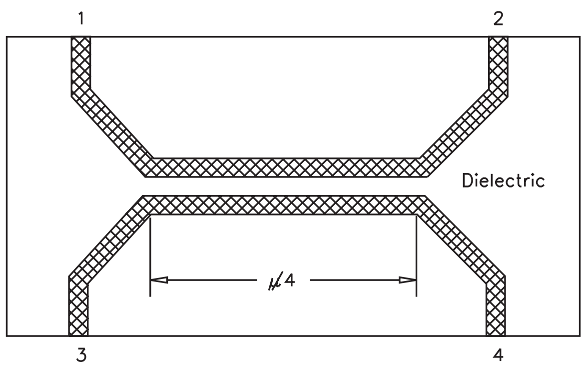


Figure 7.31 Reverse-coupled 1/4-wave hybrid combiner.

The unique phase-related properties of an RF hybrid make it possible to use the device as a switch. The input signals to the hybrid in Figure 7.33*a* are equally powered but differ in phase by 90°. This phase difference results in the combined signals being routed to the output terminal at port 4. If the relative phase between the two input signals is changed by 180°, the summed output then appears on port 3, as shown in Figure 7.33*b*. The 3 dB hybrid combiner, thus, functions as a switch.

This configuration permits the switching of two RF generators to either of two loads. Remember, however, that the switch takes place when the phase difference between the two inputs is 90°. To perform the switch in a useful way requires adding a high-power phase shifter to one input leg of the hybrid. The addition of the phase shifter permits the full power to be combined and switched to either output. This configuration of hybrid and phase shifter, however, will not permit switching a main or standby generator to a main or auxiliary load (DPDT function). To accomplish this additional switch, a second hybrid and phase shifter must be added, as shown in Figure 7.34. This configuration then can perform the following switching functions:

- RF source 1 routed to output *B*
- RF source 2 routed to output *A*
- RF source 1 routed to output *A*
- RF source 2 routed to output *B*

The key element in developing such a switch is a high-power phase shifter that does not exhibit reflection characteristics. In this application, the phase shifter allows the line between the hybrids to be electrically lengthened or shortened. The ability to adjust the relative phase between the two input signals to the second hybrid provides the needed control to switch the input signal between the two output ports.

If a continuous analog phase shifter is used, the transfer switch shown in Figure 7.34 also can act as a hot switchless combiner where RF generators 1 and 2 can be combined

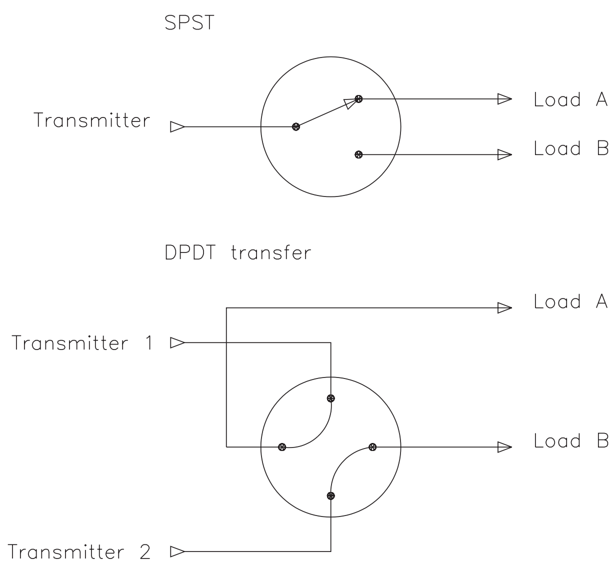


Figure 7.32 Common RF switching configurations.

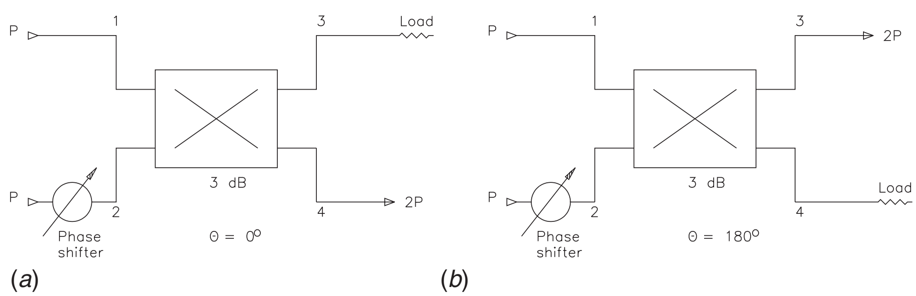


Figure 7.33 Hybrid switching configurations: (*a*) phase set so that the combined energy is delivered to port 4, (*b*) phase set so that the combined energy is delivered to port 3.

and fed to either output *A* or *B*. The switching or combining functions are accomplished by changing the physical position of the phase shifter.

Note that it does not matter whether the phase shifter is in one or both legs of the system. It is the phase difference ($\theta 1 - \theta 2$) between the two input legs of the second hybrid that is important. With 2-phase shifters, dual drives are required. However, the phase shifter needs only two positions. In a 1-phase shifter design, only a single drive is required, but the phase shifter must have four fixed operating positions.

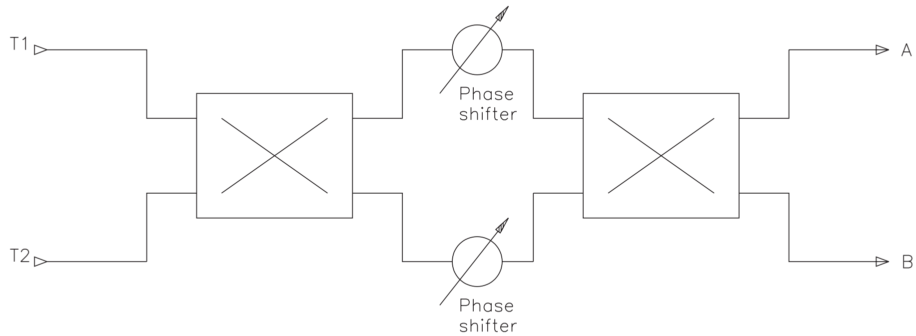


Figure 7.34 Additional switching and combining functions enabled by adding a second hybrid and another phase shifter to a hot switching combiner.

Phase Relationships

To better understand the dual-hybrid switching and combining process, it is necessary to examine the primary switching combinations. Table 7.5 lists the various combinations of inputs, relative phase, and output configurations that are possible with the single-phase shifter design.

Using vector analysis, note that when two input signals arrive in phase (mode 1) at ports 1 and 2 with the phase shifter set to 0°, the circuit acts like a crossover network with the power from input port 1 routed to output port 4. Power from input port 2 is routed to output port 3. If the phase shifter is set to 180°, the routing changes, with port 1 being routed to port 3 and port 2 being routed to port 4.

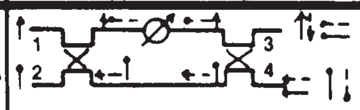
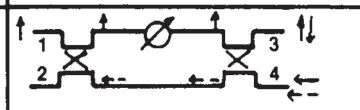
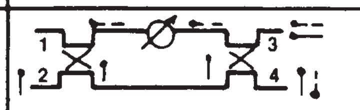
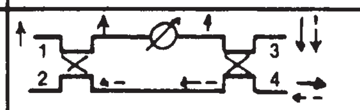
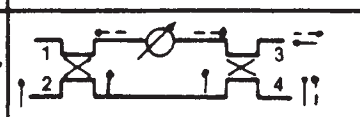
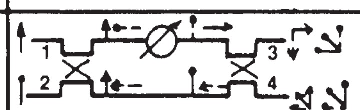
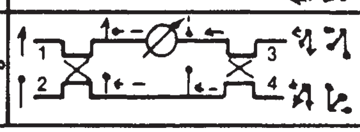
Mode 2 represents the case where one of the dual-input RF generators has failed. The output signal from the first hybrid arrives at the input to the second hybrid with a 90° phase difference. Because the second hybrid introduces a 90° phase shift, the vectors add at port 4 and cancel at port 3. This effectively switches the working transmitter connected to port 1 to output port 4, the load.

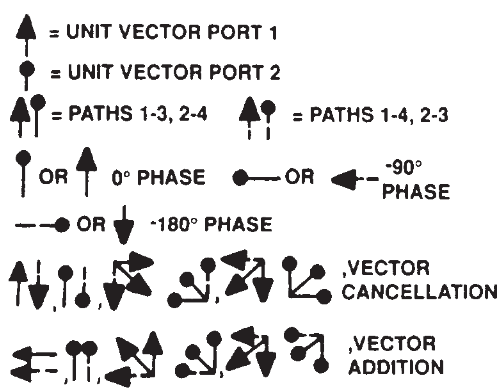
By introducing a 180° phase shift between the hybrids, as shown in modes 4 and 5, it is possible to reverse the circuit. This allows the outputs to be on the same side of the circuit as the inputs. This configuration might be useful if generator 1 fails, and all power from generator 2 is directed to a diplexer connected to output 4.

Normal operating configurations are shown in modes 6 and 7. When both generators are running, it is possible to have the combined power routed to either output port. The switching is accomplished by introducing a ±90º phase shift between the hybrids.

As shown in the table, it is possible to operate in all the listed modes through the use of a single-phase shifter. The phase shifter must provide four different phase positions. A similar analysis would show that a 2-phase shifter design, with two positions for each shifter, is capable of providing the same operational modes.

Table 7.5 Operating Modes of the Dual 90° Hybrid/Single-Phase Shifter Combiner System

MODE	INPUT 1	INPUT 2	Ø	INPUT VECTOR / SCHEMATIC / OUTPUT VECTOR	OUTPUT 3	OUTPUT 4
1	$P_1 / 0°$ $P_1 / 0°$	$P_2 / 0°$ $P_2 / 0°$	0° 180°		P_2 P_1	P_1 P_2
2	$P_1 / 0°$	$P_2 = 0$	0°		0	P_1
3	$P_1 = 0$	$P_2 / 0°$	0°		P_2	0
4	$P_1 / 0°$	$P_2 = 0$	180°		P_1	0
5	$P_1 = 0$	$P_2 / 0°$	180°		0	P_2
6	$P_1 / 0°$	$P_2 / 0°$	90°		0	$P_1 + P_2$
	$P_1 / 0°$	$P_2 / 0°$	– 90°		$P_1 + P_2$	0



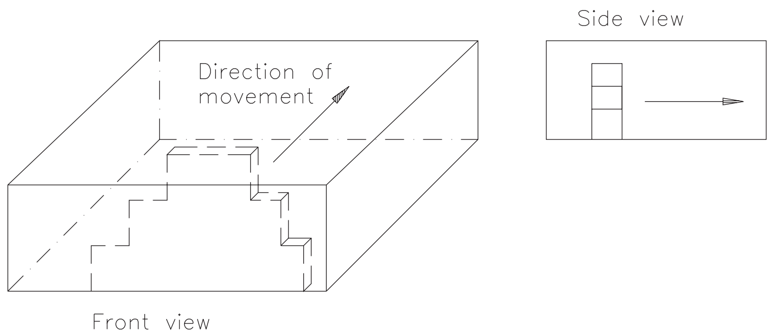


Figure 7.35 The dielectric vane switcher, which consists of a long dielectric sheet mounted within a section of rectangular waveguide.

The key to making hybrid switches work in the real world lies in the phase shifter. The dual 90° hybrid combiner just discussed requires a phase shifter capable of introducing a fixed phase offset of –90°, 0°, +90°, and +180°. This can be accomplished easily at low power levels through the use of a sliding short-circuit (trombone-type) line stretcher. However, when high-frequency and high-power signals are being used, the sliding short circuit is not an appropriate design choice. In a typical case, the phase shifter must be able to handle 100 kW or more at UHF. Under these conditions, sliding short-circuit designs are often unreliable. Therefore, three other methods have been developed:

- *Variable-dielectric vane*
- *Dielectric post*
- *Variable-phase hybrid*

Variable-Dielectric Vane

The variable-dielectric vane consists of a long dielectric sheet mounted in a section of rectangular waveguide, as illustrated in Figure 7.35. The dielectric sheet is long enough to introduce a 270° phase shift when located in the center of the waveguide. As the dielectric sheet is moved toward the wall, into the lower field, the phase shift decreases. A single-sided phase shifter easily can provide the needed four positions. A 2-stage 1/4-wave transformer is used on each end of the sheet to maintain a proper match for any position over the desired operating band. The performance of a typical switchless combiner, using the dielectric vane, is given in Table 7.6.

Dielectric Posts

Dielectric posts, shown in Figure 7.36, operate on the same principle as the dielectric vane. The dielectric posts are positioned 1/4-wavelength apart from each other to cancel any mismatch, and to maintain minimal VSWR.

Table 7.6 Typical Performance of a Dielectric Vane Phase Shifter

Type	Input	Phase Change (deg)	VSWR	Input Attenuation		Output Attenuation	
				1 (dB)	2 (dB)	3 (dB)	4 (dB)
Single input	T1	180	1.06	-	39	0.1	39
	T1	0	1.05	-	39	39	0.1
	T2	180	1.05	39	-	39	0.1
	T2	0	1.06	39	-	0.1	39
Dual input	T1+T2	270	1.06	-	-	0.1	36
	T1+T2	90	1.06	-	-	36	0.1

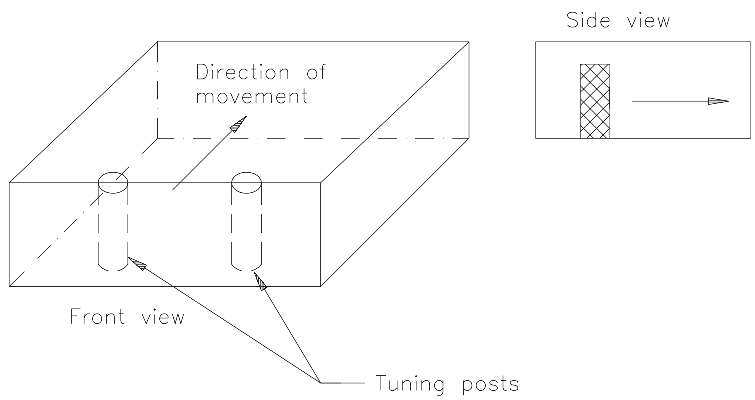


Figure 7.36 Dielectric post waveguide phase shifter.

Variable-Phase Hybrid

The variable-phase hybrid, shown in Figure 7.37, relies on a 90° hybrid, similar to those used in a combiner. With a unit vector incident on port 1, the power is split by the 90° hybrid. The signal at ports 3 and 4 is reflected by the short circuit. These reflected signals are out of phase at port 1 and in phase at port 2. The relative phase of the hybrid can be changed by moving the short circuit.

The variable-phase hybrid is linear with respect to position. Noncontacting choke-type short-circuits, with high front-to-back ratios, are typically used in the device. The performance available from a typical high-power variable-phase switchless combiner is given in Table 7.7.

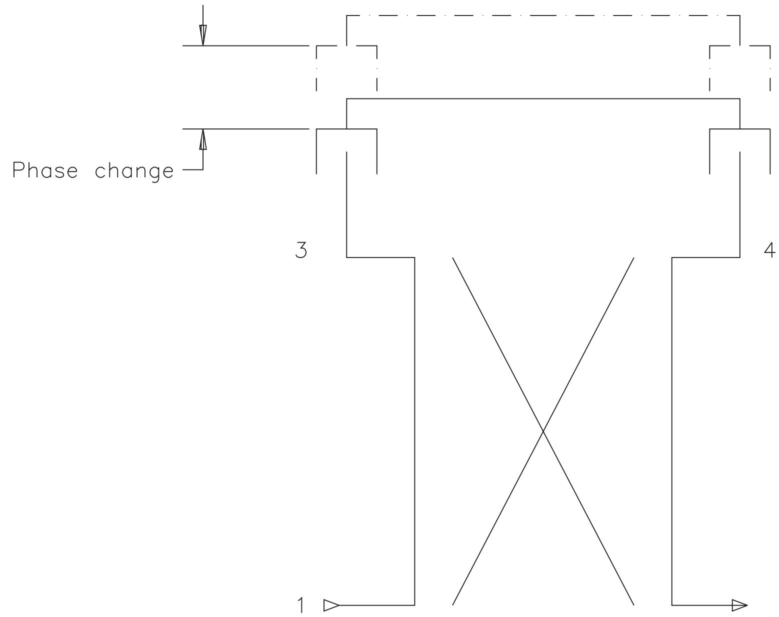


Figure 7.37 Variable-phase hybrid phase shifter.

7.4.7 Phased-Array Antenna Systems

The phased-array antenna, commonly used in radar applications, is a variation on the hot switching combiner systems discussed in the previous section. The hot switching combiner is used to control the flow of RF energy from one or more generators to two or more loads, while the phased-array antenna is used to modify—in a controlled fashion—the radiating characteristics of an antenna.

Phased-array antennas are *steered* by tilting the phase front independently in two orthogonal directions called the *array coordinates*. Scanning in either array coordinate causes the beam to move along a cone whose center is at the center of the array. As the beam is steered away from the array normal, the projected aperture in the beam's direction varies, causing the beamwidth to vary proportionately.

Arrays can be classified as either active or passive. Active arrays contain duplexers and amplifiers behind every element or group of elements of the array. Passive arrays are driven from a single feed point. Active arrays are capable of high-power operation. Both passive and active arrays must divide the signal from a single transmission line among all the elements of the system. This can be accomplished through one of the following methods:

- *Optical feed*: a single feed, usually a monopulse horn, is used to illuminate the array with a spherical phase front, illustrated in Figure 7.38. Power collected by the rear elements of the array is transmitted through the phase shifters to produce a planar front and steer the array. The energy then may be radiated from the other

Table 7.7 Typical Performance of a Variable-Phase Hybrid Phase Shifter

Type	Input	Phase Change (deg)	VSWR	Input Attenuation		Output Attenuation	
				1 (dB)	2 (dB)	3 (dB)	4 (dB)
Single input	T1	180	1.06	-	36	0.1	52
	T1	0	1.04	-	36	50	0.1
	T2	180	1.06	36	-	52	0.1
	T2	0	1.07	36	-	0.1	50
Dual input	T1+T2	270	1.06	-	-	0.1	36
	T1+T2	90	1.06	-	-	36	0.1

side of the array or reflected and reradiated through the collecting elements. In the latter case, the array acts as a *steerable reflector*.

- *Corporate feed*: a system utilizing a series-feed network (Figure 7.39) or parallel-feed network (Figure 7.40). Both designs use transmission-line components to divide the signal among the elements. Phase shifters can be located at the elements or within the dividing network. Both the series- and parallel-feed systems have several variations, as shown in the figures.
- *Multiple-beam network*: a system capable of forming simultaneous beams with a given array. The *Butler matrix*, shown in Figure 7.41, is one such technique. It connects the *N* elements of a linear array to *N* feed points corresponding to *N* beam outputs. The phase shifter is one of the most critical components of the system. It produces controllable phase shift over the operating band of the array. Digital and analog phase shifters have been developed using both ferrites and *pin* diodes.

Frequency scan is another type of multiple-beam network, but one that does not require phase shifters, dividers, or beam-steering computers. Element signals are coupled from points along a transmission line. The electrical path length between elements is longer than the physical separation, so a small frequency change will cause a phase change between elements that is large enough to steer the beam. This technique can be applied only to one array coordinate. If a two-dimensional array is required, phase shifters normally are used to scan the other coordinate.

Phase-Shift Devices

The design of a phase shifter must meet two primary criteria:

- Low transmission loss
- High power-handling capability

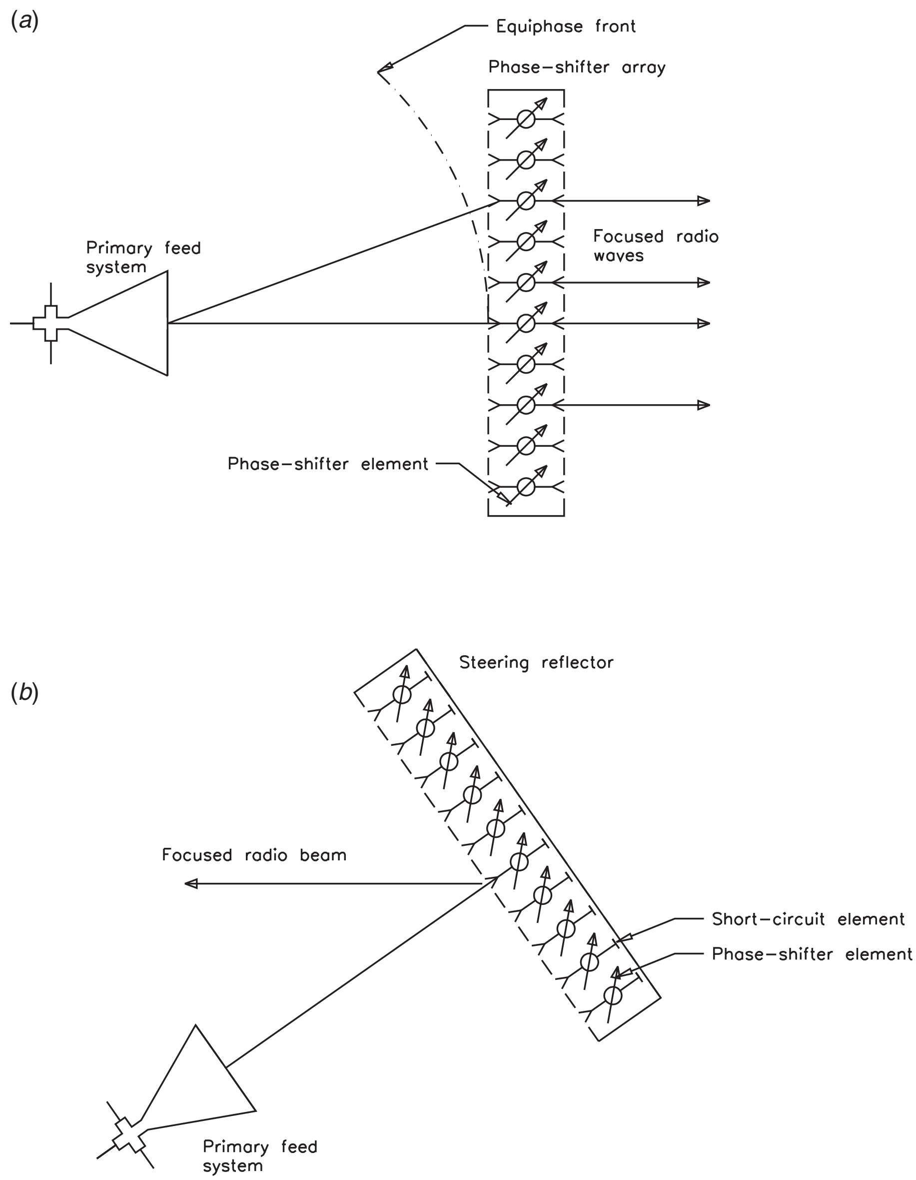


Figure 7.38 Optical antenna feed systems: (*a*) lens, (*b*) reflector.

The *Reggia-Spencer* phase shifter meets both requirements. The device, illustrated in Figure 7.42, consists of a ferrite rod mounted inside a waveguide that delays the RF signal passing through the waveguide, permitting the array to be steered. The amount of phase shift can be controlled by the current in the solenoid, because of the effect a

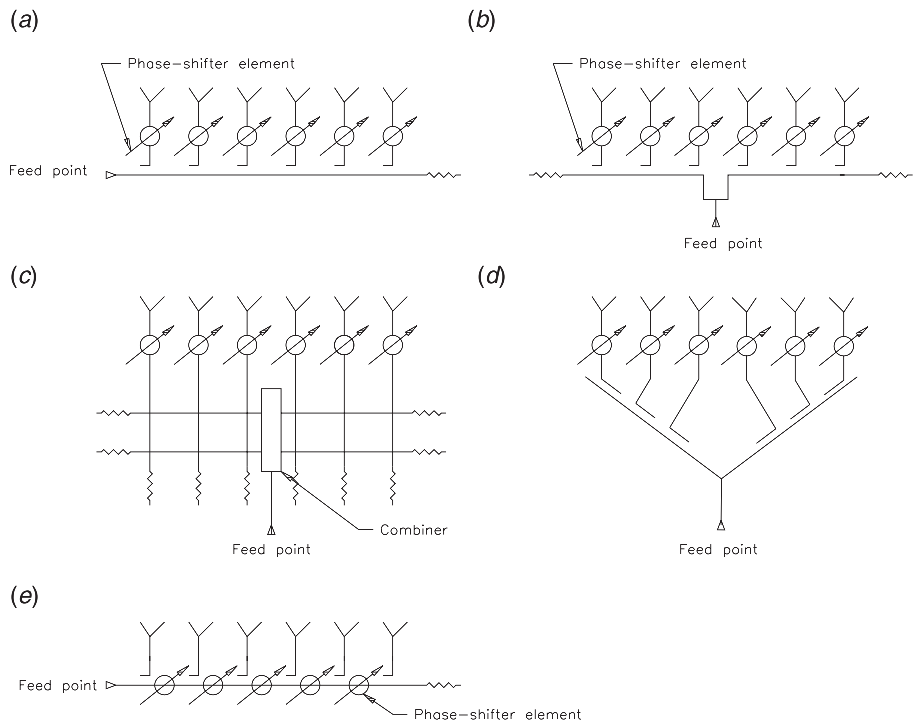


Figure 7.39 Series-feed networks: (*a*) end feed, (*b*) center feed, (*c*) separate optimization, (*d*) equal path length, (*e*) series phase shifters.

magnetic field has on the permeability of the ferrite. This design is a *reciprocal phase shifter*, meaning that the device exhibits the same phase shift for signals passing in either direction (forward or reverse). Nonreciprocal phase shifters also are available, where phase-shift polarity reverses with the direction of propagation.

Phase shifters also may be developed using pin diodes in transmission line networks. One configuration, shown in Figure 7.43, uses diodes as switches to change the signal path length of the network. A second type uses pin diodes as switches to connect reactive loads across a transmission line. When equal loads are connected with 1/4-wave separation, a pure phase shift results.

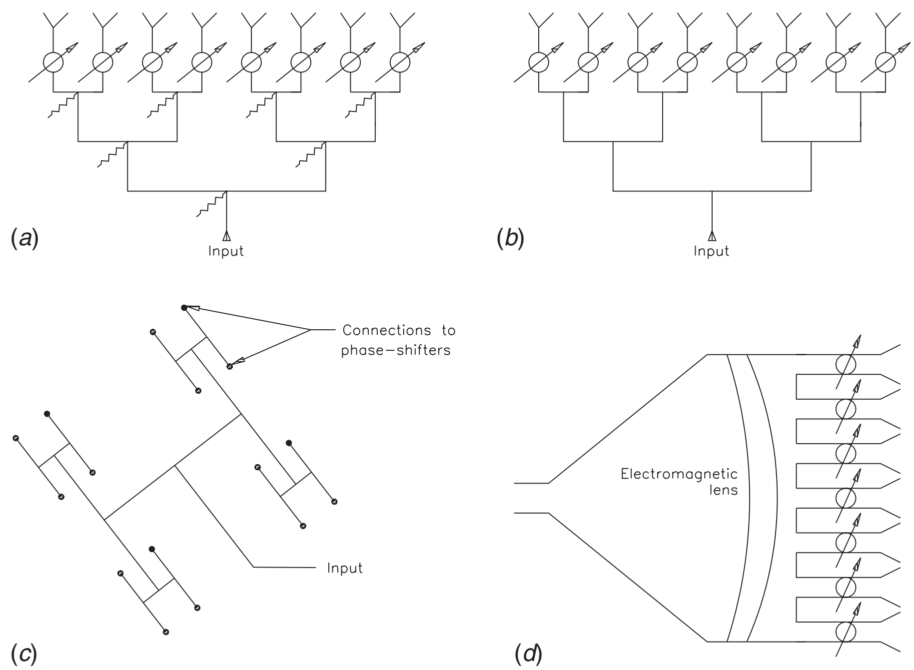


Figure 7.40 Types of parallel-feed networks: (*a*) matched corporate feed, (*b*) reactive corporate feed, (*c*) reactive stripline, (*d*) multiple reactive divider.

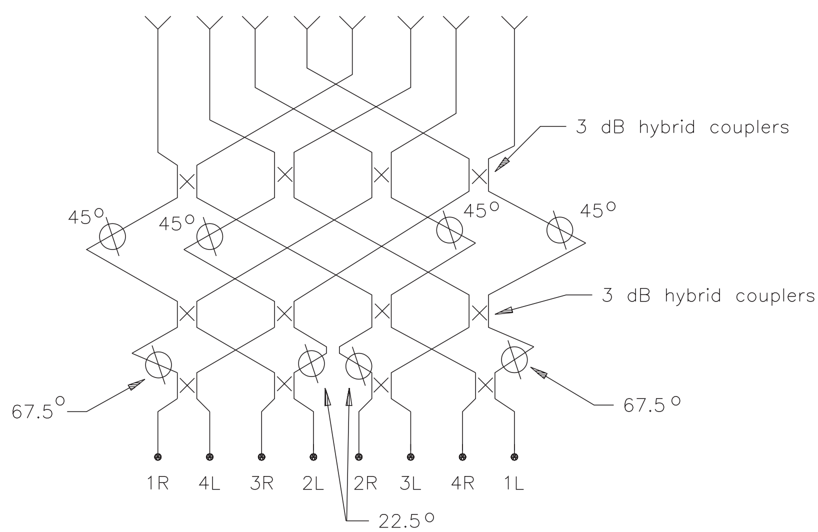


Figure 7.41 The Butler beam-forming network.

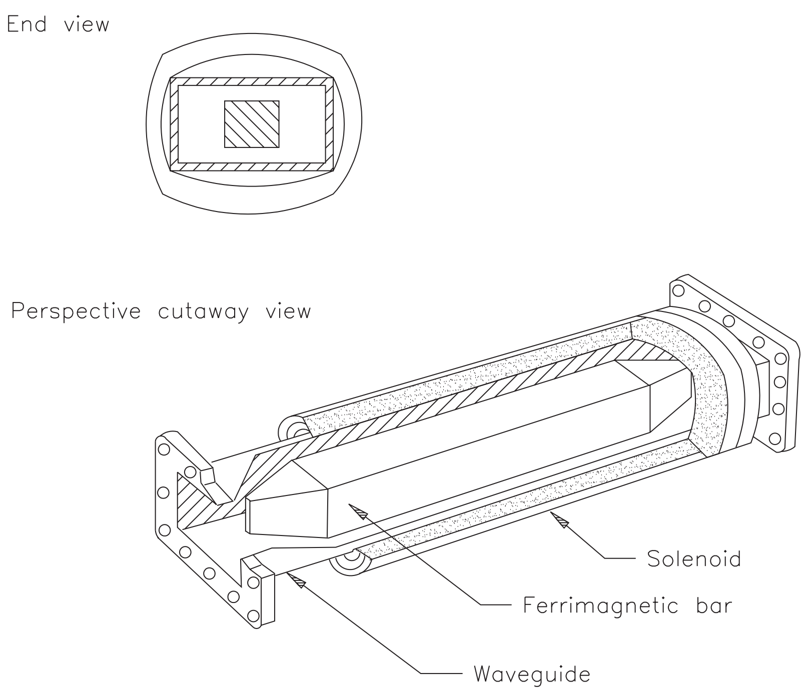


Figure 7.42 Basic concept of a Reggia-Spencer phase shifter.

Radar System Duplexer

The duplexer is an essential component of any radar system. The switching elements used in a duplexer include gas tubes, ferrite circulators, and pin diodes. Gas tubes are the simplest. A typical gas-filled *TR tube* is shown in Figure 7.44. Low-power RF signals pass through the tube with little attenuation. Higher-power signals, however, cause the gas to ionize and present a short circuit to the RF energy.

Figure 7.45 illustrates a *balanced duplexer* using hybrid junctions and TR tubes. When the transmitter is on, the TR tubes fire and reflect the RF power to the antenna port of the input hybrid. During the receive portion of the radar function, signals picked up by the antenna are passed through the TR tubes and on to the receiver port of the output hybrid.

Newer radar systems often use a ferrite circulator as the duplexer. A TR tube is required in the receiver line to protect input circuits from transmitter power reflected by the antenna because of an imperfect match. A four-port circulator generally is used with a load between the transmitter and receiver ports so that power reflected by the TR tube is properly terminated.

Pin diode switches also have been used in duplexers to perform the protective switching function of TR tubes. Pin diodes are more easily applied in coaxial circuitry,

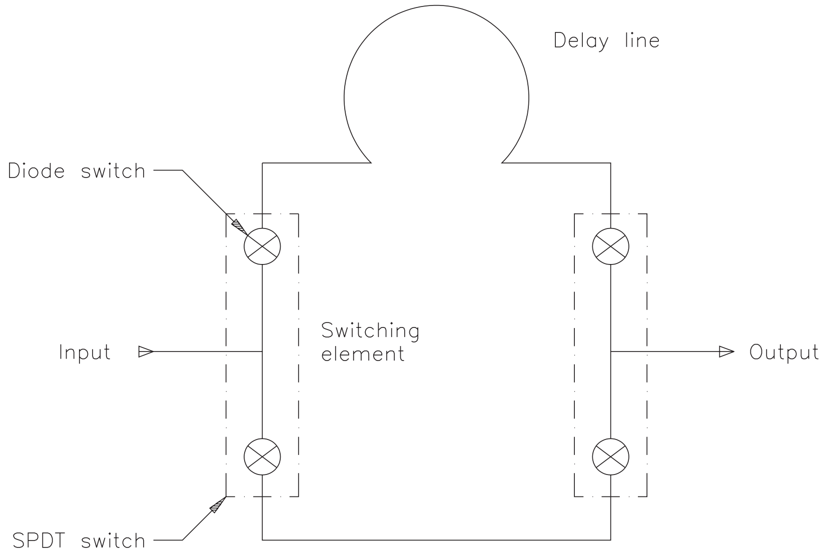


Figure 7.43 Switched-line phase shifter using *pin* diodes.

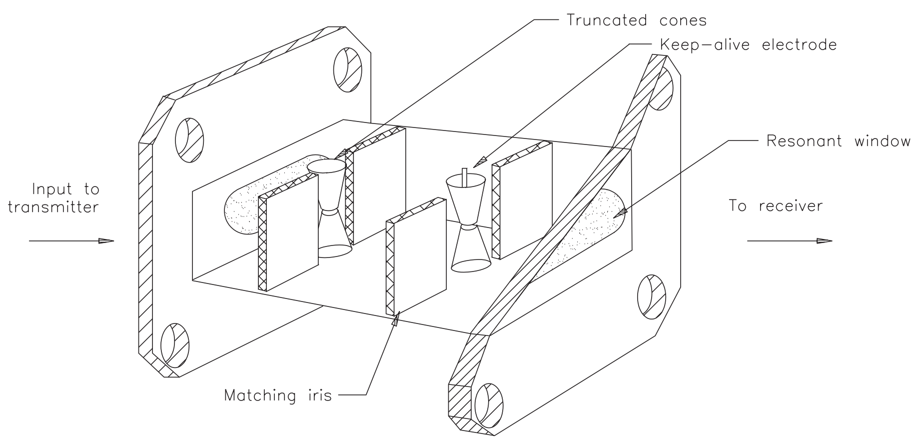


Figure 7.44 Typical construction of a TR tube.

and at lower microwave frequencies. Multiple diodes are used when a single diode cannot withstand the expected power.

Microwave filters sometimes are used in the transmit path of a radar system to suppress spurious radiation, or in the receive signal path to suppress spurious interference.

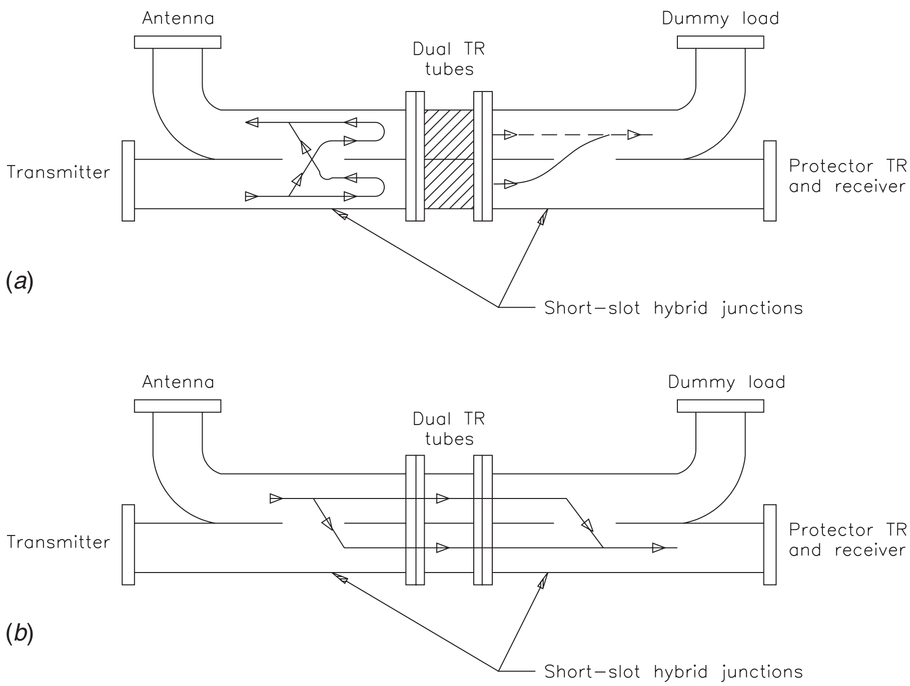


Figure 7.45 Balanced duplexer circuit using dual TR tubes and two short-slot hybrid junctions: (*a*) transmit mode, (*b*) receive mode.

Harmonic filters commonly are used in the transmission chain to absorb harmonic energy output by the system, preventing it from being radiated or reflected back from the antenna. Figure 7.46 shows a filter in which harmonic energy is coupled out through holes in the walls of the waveguide to matched loads.

Narrowband filters in the receive path, often called *preselectors*, are built using mechanically tuned cavity resonators or electrically tuned *TIG resonators*. Preselectors can provide up to 80 dB suppression of signals from other radar transmitters in the same RF band, but at a different operating frequency.

7.5 High-Power Isolators

The high-power ferrite isolator offers the ability to stabilize impedance, isolate the RF generator from load discontinuities, eliminate reflections from the load, and absorb harmonic and intermodulation products. The isolator also can be used to switch between an antenna or load under full power, or to combine two or more generators into a common load.

Isolators commonly are used in microwave transmitters at low power to protect the output stage from reflections. Until recently, however, the insertion loss of the ferrite made use of isolators impractical at high-power levels (25 kW and above). Ferrite isola-

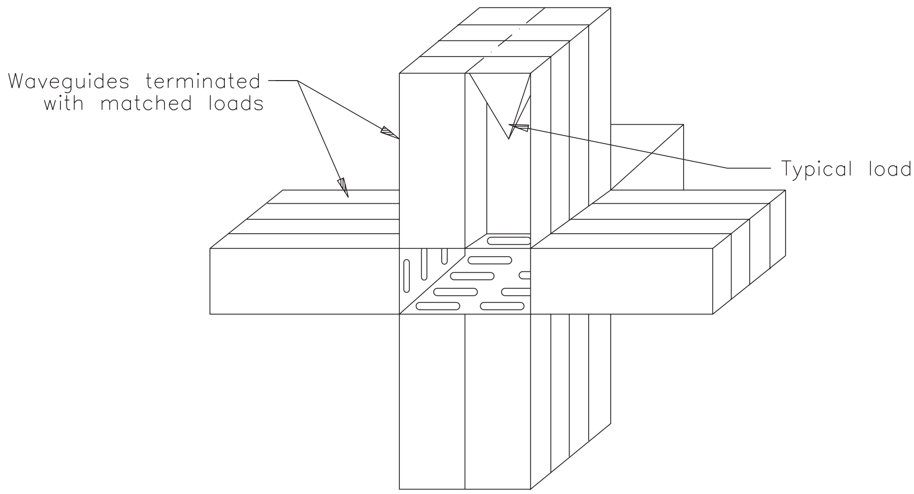


Figure 7.46 Construction of a dissipative waveguide filter.

tors are now available that can handle 500 kW or more of forward power with less than 0.1 dB of forward power loss.

7.5.1 Theory of Operation

High-power isolators are three-port versions of a family of devices known as *circulators*. The circulator derives its name from the fact that a signal applied to one of the input ports can travel in only one direction, as shown in Figure 7.47. The input port is isolated from the output port. A signal entering port 1 appears only at port 2; it does not appear at port 3 unless reflected from port 2. An important benefit of this one-way power transfer is that the input VSWR at port 1 is dependent only on the VSWR of the load placed at port 3. In most applications, this load is a resistive (dummy) load that presents a perfect load to the RF generator.

The unidirectional property of the isolator results from magnetization of a ferrite alloy inside the device. Through correct polarization of the magnetic field of the ferrite, RF energy will travel through the element in only one direction (port 1 to 2, port 2 to 3, and port 3 to 1). Reversing the polarity of the magnetic field makes it possible for RF flow in the opposite direction. Recent developments in ferrite technology have resulted in high isolation with low insertion loss.

In the basic design, the ferrite is placed in the center of a Y-junction of three transmission lines, either waveguide or coax. Sections of the material are bonded together to form a thin cylinder perpendicular to the electric field. Even though the insertion loss is low, the resulting power dissipated in the cylinder can be as high as 2 percent of the for-

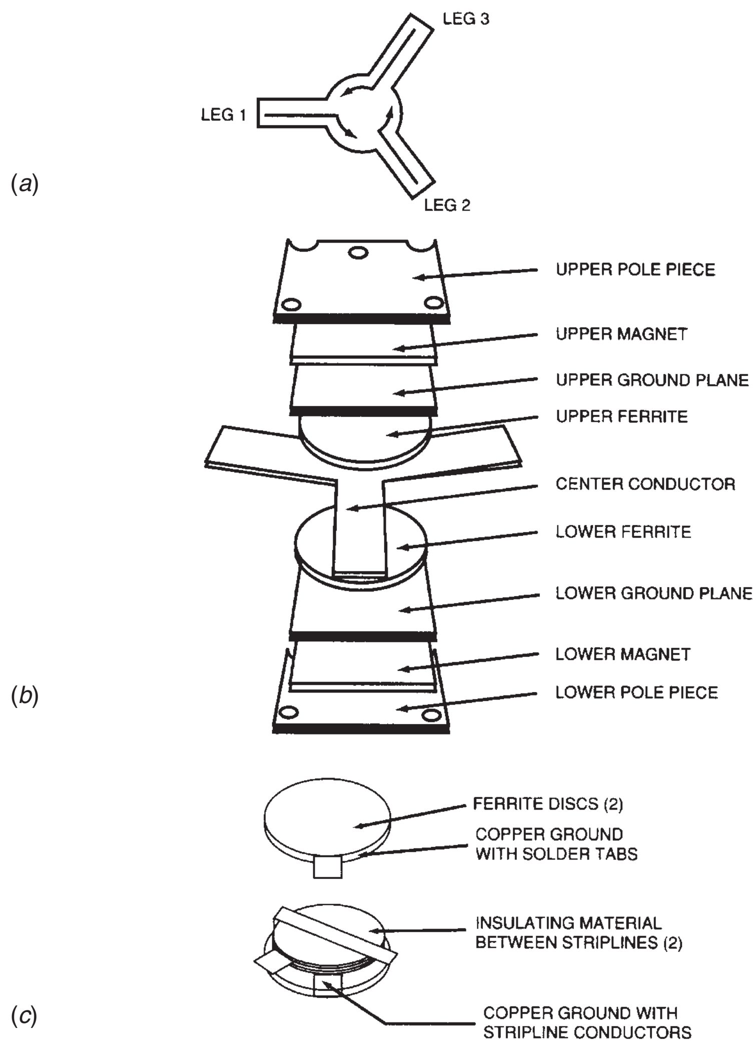


Figure 7.47 Basic characteristics of a circulator: (*a*) operational schematic, (*b*) distributed constant circulator, (*c*) lump constant circulator. (*From* [2]. *Used with permission.*)

ward power. Special provisions must be made for heat removal. It is efficient heat-removal capability that makes high-power operation possible.

The insertion loss of the ferrite must be kept low so that minimal heat is dissipated. Values of ferrite loss on the order of 0.05 dB have been produced. This equates to an efficiency of 98.9 percent. Additional losses from the transmission line and matching structure contribute slightly to loss. The overall loss is typically less than 0.1 dB, or 98 percent efficiency. The ferrite element in a high-power system is usually water-cooled in a closed-loop path that uses an external radiator.

The two basic circulator implementations are shown in Figures 7.47*a* and 7.47*b*. These designs consist of Y-shaped conductors sandwiched between magnetized ferrite discs [2]. The final shape, dimensions, and type of material varies according to frequency of operation, power handling requirements, and the method of coupling. The *distributed constant circulator* is the older design; it is a broad-band device, not quite as efficient in terms of insertion loss and leg-to-leg isolation, and considerably more expensive to produce. It is useful, however, in applications where broad-band isolation is required. More commonly today is the *lump constant circulator*, a less expensive and more efficient, but narrow-band, design.

At least one filter is always installed directly after an isolator, because the ferrite material of the isolator generates harmonic signals. If an ordinary band-pass or band-reject filter is not to be used, a harmonic filter will be needed.

7.5.2 Applications

The high-power isolator permits an RF generator to operate with high performance and reliability despite a load that is less than optimum. The problems presented by ice formations on a transmitting antenna provide a convenient example. Ice buildup will detune an antenna, resulting in reflections back to the transmitter and high VSWR. If the VSWR is severe enough, transmitter power will have to be reduced to keep the system on the air. An isolator, however, permits continued operation with no degradation in signal quality. Power output is affected only to the extent of the reflected energy, which is dissipated in the resistive load.

A high-power isolator also can be used to provide a stable impedance for devices that are sensitive to load variations, such as klystrons. This allows the device to be tuned for optimum performance regardless of the stability of the RF components located after the isolator. Figure 7.48 shows the output of a wideband (6 MHz) klystron operating into a resistive load, and into an antenna system. The power loss is the result of an impedance difference. The periodicity of the ripple shown in the trace is a function of the distance of the reflections from the source.

Hot Switch

The circulator can be made to perform a switching function if a short circuit is placed at the output port. Under this condition, all input power will be reflected back into the third port. The use of a high-power stub on port 2, therefore, permits redirecting the output of an RF generator to port 3.

At odd 1/4-wave positions, the stub appears as a high impedance and has no effect on the output port. At even 1/4-wave positions, the stub appears as a short circuit. Switching between the antenna and a test load, for example, can be accomplished by moving the shorting element 1/4 wavelength.

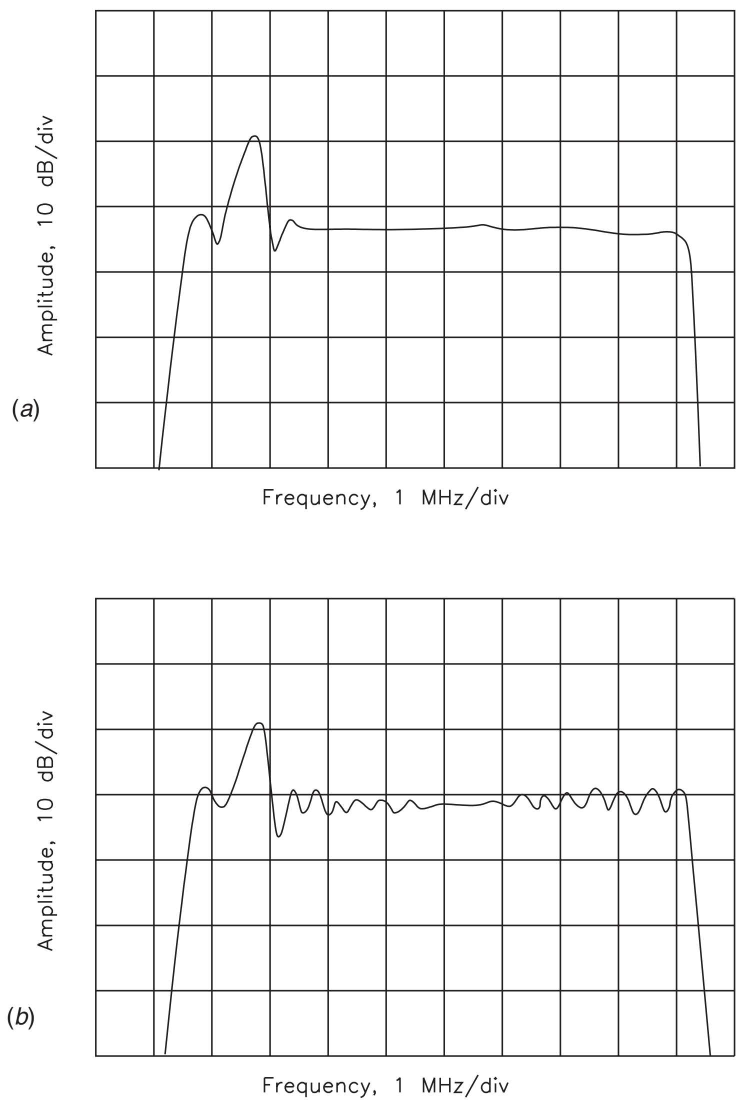


Figure 7.48 Output of a klystron operating into different loads through a high-power isolator: (*a*) resistive load, (*b*) an antenna system.

Diplexer

An isolator may be configured to combine the aural and visual outputs of a TV transmitter into a single output for the antenna. The approach is shown in Figure 7.49. A single notch cavity at the aural frequency is placed on the visual transmitter output

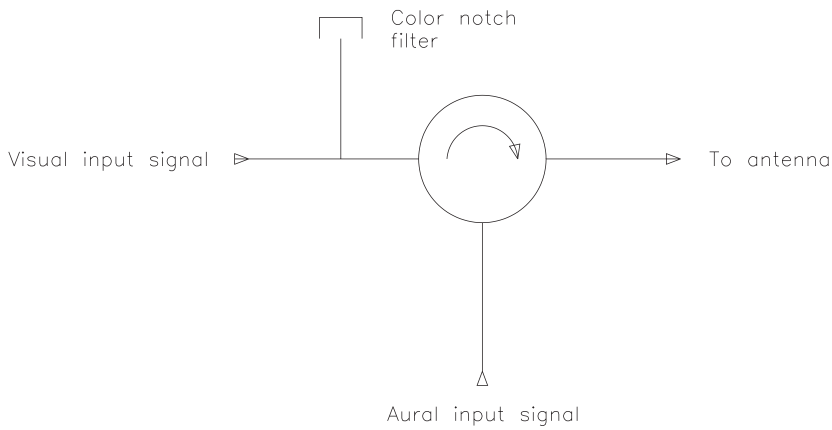


Figure 7.49 Use of a circulator as a diplexer in TV applications.

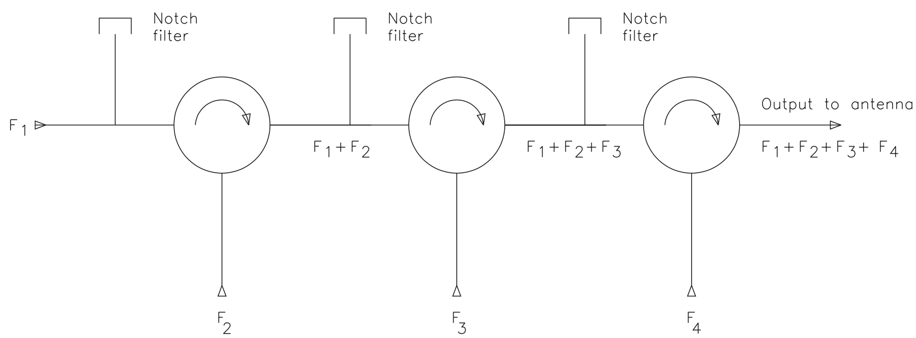


Figure 7.50 Using multiple circulators to form a multiplexer.

(circulator input), and the aural signal is added (as shown). The aural signal will be routed to the antenna in the same manner as it is reflected (because of the hybrid action) in a conventional diplexer.

Multiplexer

A multiplexer can be formed by cascading multiple circulators, as illustrated in Figure 7.50. Filters must be added, as shown. The primary drawback of this approach is the increased power dissipation that occurs in circulators nearest the antenna.

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