# SECTION 7

# UHF AND MICROWAVE COMPONENTS

The three chapters in this section cover passive microwave components, vacuum electronic devices, and microwave semiconductor devices.

A broad range of passive devices (Chap. 7.1) enable the transmission, sampling, filtering, and impedance matching of UHF and microwave power. Rigid rectangular waveguide is widely used for transmission. Flexible coaxial cable is employed for short runs, in cases where its higher attenuation can be tolerated. In lower power applications microstrip technology provides advantages in miniaturization and conformal configurations.

Chapter 7.2 is devoted principally to vacuum electronic devices, both traditional and advanced. Modern materials and design methods have advanced the RF vacuum device orders of magnitude in performance over that of the earlier glass envelope vacuum tube. While magnetrons and klystrons were used in World War II receivers, postwar funding of particle accelerator development resulted in klystrons having a $10^3$ increase in peak power. Later developments designed to overcome bandwidth limitations of klystrons include the clustered cavity technique. The klystrode, an advanced version of the inductive output tube (IOT), offers reduced length and weight as a result of prebunching the beam. The traveling-wave tube (TWT) is a linear-beam device that amplifies microwave signals to high power levels over broad bandwidths. It finds diverse applications in communications, radar, and electronic countermeasure systems, and aerospace and avionics where low weight and volume, and minimizing power consumption are required. A further step to miniaturization of RF systems is the microwave power module (MPM), which exploits both vacuum electronics and solid-state technology. In the MPM, a high-gain monolithic integrated circuit drives a miniaturized helix TWT—both powered by a high-density electronic power conditioner. Gyrotrons (or cyclotron resonance masers) are a class of microwave generators that promise applications in the millimeter-wave region. Gyrotron oscillators have been developed that can produce 1 MW for 1 s at frequencies greater than 100 GHz.

Advances in semiconductor materials, particularly compound semiconductors, have enhanced the performance of microwave and millimeter wave devices (Chap. 7.3). Monolithic microwave integrated circuits (MMICs) enable small, lower cost RF and microwave components like amplifiers, oscillators, and switches for use in high speed/high frequency applications like cell phones and optical fiber communication systems. They are used, for example, in microwave power modules (MPMs). D.C.

**In This Section:**

# CHAPTER 7.1
# PASSIVE MICROWAVE COMPONENTS

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## INTRODUCTION

While the physical concepts and mathematical theory underlying electromagnetic-wave propagation in confined structures were developed at the end of the nineteenth century, the practical utilization of wavelengths shorter than 1 m began during World War II. A wide variety of devices is now available for the transmission, sampling, filtering, and impedance matching of UHF and microwave power. Because of the short wavelengths at these frequencies (10 cm at 3 GHz, 1 cm at 30 GHz), most of these components use distributed elements and obtain specific reactances by judicious use of short-circuited lengths of transmission line. However, the trend toward microcircuitry has led recently to the introduction of lumped elements in some low-power, low-$Q$ applications. In addition, the use of strip and microstrip transmission lines marks a return to open-wire media, with the attendant radiation loss kept low by close spacing and the presence of the dielectric filler.

Except for special applications (such as rotating joints for antenna feeds where a cylindrical member is essential), the use of rectangular waveguide, dimensioned to transmit in the dominant (lowest-order) mode, is standard for high-power transmission. Coaxial cable is used for short-distance runs where the advantage of its flexibility outweighs its higher attenuation. Ridged waveguide is useful for designing matching sections and for providing very-wide-bandwidth single-mode transmission. Oversized cylindrical guide operated in the circular electric mode is finding use in millimeter-wave-carrier telephony, where its extremely low attenuation justifies the special precautions which must be taken to avoid mode conversions.

***Reciprocal and Nonreciprocal Components.*** Of the components which have become standard in this field, perhaps the most unusual are ferrite devices. When biased with the proper magnetic field, ferrites act as nonreciprocal elements with respect to microwave transmission in an appropriate frequency band. This behavior allows isolation of a signal source from reflections and the separation of incident from reflected power along the same transmission line (by means of a ferrite device called a *circulator*). It is also possible to vary the phase of a transmitted wave by adjusting the magnetizing field on the ferrite. Such phase shifters are capable of handling high powers with low loss.

All other types of microwave components, such as hybrid junctions and directional couplers, are reciprocal in their action. The latter are employed for power division and for signal sampling. A wide variety of transmission components such as variable attenuators, matched-load terminations, and slotted lines are used in microwave-measurements and design.

High-$Q$ resonators are formed from completely enclosed short-circuited sections of waveguides with slit or loop coupling. Extremely low loss dielectric cylinders can also be used as high-$Q$ resonators and have the advantages of smaller size and easy coupling to microstrip or other transmission lines. Resonators using lengths of strip, microstrip or coaxial lines are frequently convenient, but have lower $Q$'s. Lumped elements, such as varactor diodes or YIG spheres, provide electrically tunable resonators for microwave oscillators and receivers.

## Transmission-Line Relationships

The basic transmission-line equations are derived for distributed parameters $R$ (series resistance), $G$ (shunt conductance), $L$ (series inductance), and $C$ (shunt capacitance), all defined per unit length of line. Some useful relations are shown in Table 7.1.1. For zero losses ($R$, $G = 0$) one obtains the ideal line expression shown on the right. Table 7.1.2 gives some equations that are useful for relating the measured voltage standing wave ratio (VSWR) to wave transmission and reflection.

## Matching

A microwave circuit typically consists of several devices connected together or connected through sections of transmission line. Mismatches between the devices decrease transmitted power, increase reflected energy and make the performance of the circuit more frequency dependent. Match between devices may be achieved by using capacitive or inductive elements suitably positioned along the transmission line. Another approach is to

**TABLE 7.1.1** Summary of Transmission-Line Equations

| Quantity | General line expression | Ideal line expression |
|---|---|---|
| Propagation constant | $\gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)}$ | $\gamma = j\omega\sqrt{LC}$ |
| Phase constant $\beta$ | Im $\gamma$ | $\beta = \omega\sqrt{LC} = 2\pi/\lambda$ |
| Attenuation constant $\alpha$ | Re $\gamma$ | 0 |
| Impedance, characteristic | $Z_0 = \sqrt{\dfrac{R + j\omega L}{G + j\omega C}}$ | $Z_0 = \sqrt{\dfrac{L}{C}}$ |
| Input | $Z_{-1} = Z_0 \dfrac{Z_r + Z_0 \tanh \gamma l}{Z_0 + Z_r \tanh \gamma l}$ | $Z_{-1} = Z_0 \dfrac{Z_r + jZ_0 \tan \beta l}{Z_0 + jZ_r \tan \beta l}$ |
| Of short-circuited line, $Z_t = 0$ | $Z_{oc} = Z_0 \tanh \gamma l$ | $Z_{oc} = jZ_0 \tan \beta l$ |
| Of open-circuited line, $Z_r = \infty$ | $Z_{oc} = Z_o \coth \gamma l$ | $Z_{oc} = -jZ_0 \cot \beta l$ |
| Of line an odd number of quarter wavelengths long | $Z = Z_0 \dfrac{Z_t + Z_0 \coth \alpha l}{Z_0 + Z_r \coth \alpha l}$ | $Z = \dfrac{Z_0^2}{Z_r}$ |
| Of line an integral number of half wavelengths long | $Z = Z_0 \dfrac{Z_r + Z_0 \tanh \alpha l}{Z_0 + Z_r \tanh \alpha l}$ | $Z = Z_r$ |
| Voltage along line | $V_{-1} = V_i(1 + \Gamma_0 e^{-2\gamma l})$ | $V_{-1} = V_t(1 + \Gamma_0 e^{-2\beta l})$ |
| Current along line | $I_{-1} = I_i(1 - \Gamma_0 e^{-2\gamma l})$ | $I_{-1} = I_l(1 - \Gamma_0 e^{-2\beta l})$ |
| Voltage reflection coefficient | $\Gamma = \dfrac{Z_r - Z_0}{Z_r + Z_0}$ | $\Gamma = \dfrac{Z_r - Z_0}{Z_r + Z_0}$ |

$l$ = length of transmission line.

**TABLE 7.1.2** Some Miscellaneous Relations in Low-Loss Transmission Lines

| Equation | Explanation |
|---|---|
| $r = \dfrac{1+\|\Gamma\|}{1-\|\Gamma\|}$ | $r$ = VSWR |
| $\|\Gamma\| = \dfrac{r-1}{r+1}$ | $\|\Gamma\|$ = magnitude of reflection coefficient |
| $\Gamma = \dfrac{R-Z_0}{R+Z_0}$ | $\Gamma$ = reflection coefficient (real) at a point in a line where impedance is real ($R$) |
| $r = \dfrac{R}{Z_0}$ | $R > Z_0$ (at voltage maximum) |
| $r = \dfrac{Z_0}{R}$ | $V < Z_0$ (at voltage minimum) |
| $\dfrac{P_r}{P_i} = \|\Gamma\|^2 = \left(\dfrac{r-1}{r+1}\right)^2$ | $P_r$ = reflected power<br>$P_i$ = incident power |
| $\dfrac{P_t}{P_i} = 1-\|\Gamma\|^2 = \dfrac{4r}{(r+1)^2}$ | $P_t$ = transmitted power |
| $\dfrac{\alpha_r}{\alpha_m} = \dfrac{1+\Gamma^2}{1-\Gamma^2} = \dfrac{r^2+1}{2r}$ | $\alpha_m$ = attenuation constant when $r = 1$, matched line<br>$\alpha_r$ = attenuation constant allowing for increased ohmic loss caused by standing waves |
| $r_{max} = r_1 r_2$ | $r_{max}$ = maximum VSWR when $r_1$ and $r_2$ combine in worst phase |
| $r_{min} = \dfrac{r_2}{r_1} r_2 > r_1$ | $r_{min}$ = minimum VSWR when $r_1$ and $r_2$ are in best phase |
| $\|\Gamma\| = \dfrac{\|X\|}{\sqrt{X^2+4}}$ | Relations for a normalized reactance $X$ in series with resistance $Z_0$ |
| $\|X\| = \dfrac{r-1}{\sqrt{r}}$ | |
| $\|\Gamma\| = \dfrac{\|B\|}{\sqrt{B^2+4}}$ | Relations for a normalized susceptance $B$ in shunt with admittance $Y_0$ |
| $\|B\| = \dfrac{r-1}{\sqrt{r}}$ | |

use transformers in the transmission line, such transformers having an appropriate length and impedance. Yet another approach is to taper from one impedance to another.

***Smith Chart.*** The matching elements or transformers can be calculated graphically using a Smith chart which plots complex impedance. The standing wave pattern in a transmission line repeats itself every half-wavelength. The angular position of the radius of a circle represents the phase along a transmission line. Rotating the radius of a circle a complete revolution corresponds to a phase change of $\frac{1}{2}$ wavelength. Circles of constant resistance are tangent to the circumference of the Smith chart at a single point. These two features are illustrated in Fig. 7.1.1*a*. The center of the Smith chart has been normalized to a value of 1. To change to a 50-ohm transmission line, all of the resistance values would be multiplied by 50. A constant voltage standing wave

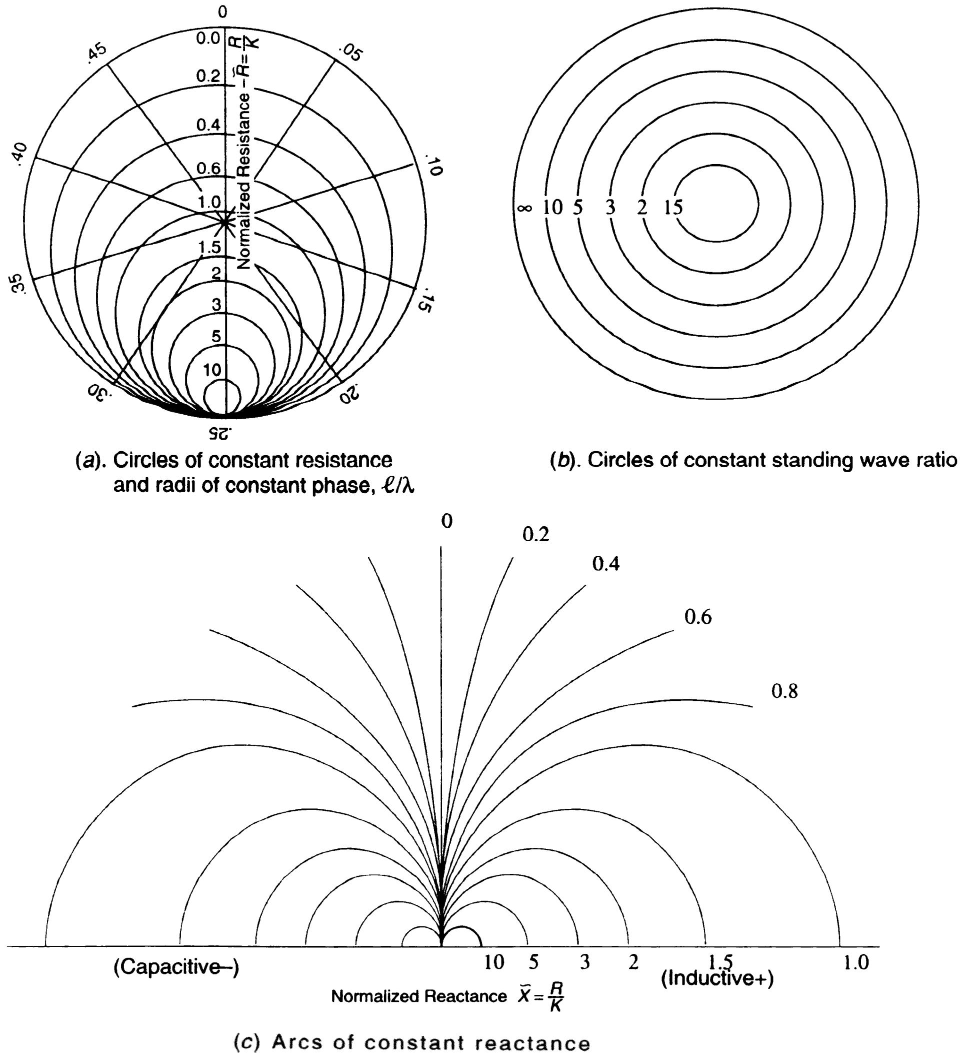



**FIGURE 7.1.1** Elements of a Smith chart.

ratio is represented by a circle concentric with the center of the Smith chart. This is illustrated in Fig. 7.1.1*b*. Circles are not shown on the chart because their value can be determined by measuring the distance from the center of the chart to the point of interest. For the normalized Smith chart, the resistance value along the zero reactance line is numerically equal to the VSWR, varying between 1 and infinity. Normalized reactance is a series of arcs as shown in Fig. 7.1.1*c*. The actual values of reactance can be obtained by multiplying by the normalizing constant. Figures 7.1.1*a* and 7.1.1*c* are superimposed to obtain the Smith chart, Fig. 7.1.2. The impedance of $P_1$ is 0.6 Ω. The impedance of $P_2$ is $0.5 - j1.0$. The impedance of $P_3$ is $0.5 + j0.5$. $P_3$ can be matched by adding a capacitive reactance $-j1.0$ at a distance of 0.72 λ toward the generator, as shown.

The Smith chart may also be used to represent admittance $Y$, where $Y = 1/Z$. For a normalized admittance Smith chart ($Y_0 = 1$ at the center) capacitive reactance ($-jX/Z_0$) is replaced by inductive susceptance ($-jB/Y_0$); inductive reactance ($+jX/Z_0$) is replaced by capacitive susceptance ($+jB/Y_0$).

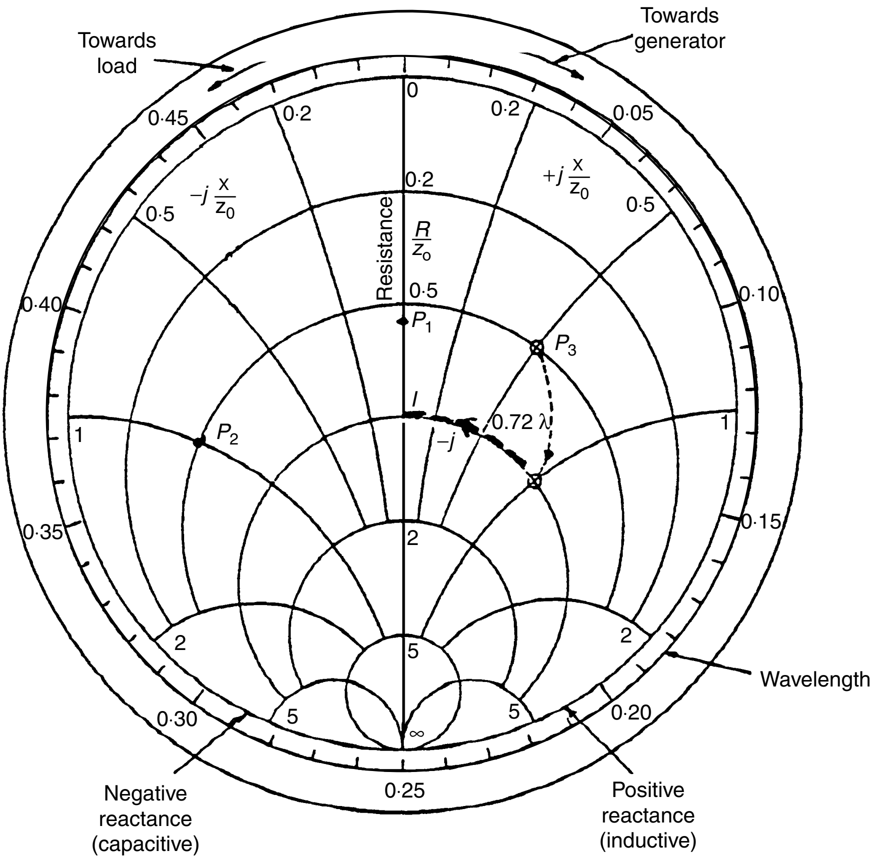



**FIGURE 7.1.2** Smith chart representation of complex impedance. Impedance of $P_1$ is 0.6 ohm, $P_2$ is $0.5 - j1.0$, $P_3$ is $0.5 + j0.5$. $P_3$ is matched with a capacitive impedance $-j1.0$ with a phase of 0.72 $\lambda$ toward the generator.

***Quarter Wavelength Transformers.*** One or more quarter wavelength sections of transmission line may be used to match from one impedance to another. Table 7.1.3 gives the VSWR which will result when one, two, or three transformer sections are used. $R$ is the ratio of the two impedances to be matched. $\omega_q$ is the bandwidth over which match is to be achieved. If a single quarter wavelength section is used, the impedance of the transformer is

$$Z \text{ (impendance of matching transformer)} = \sqrt{Z_L \times Z_H}$$

where $Z_L$ and $Z_H$ are the characteristic impedance of the lower and higher impedance transmission lines, respectively. When two quarter wavelength sections are used and when the impedance of the lower impedance transmission line, $Z_L$, is normalized to unity, the impedance of the first transformer is the same as $Z_1$ in Table 7.1.4*a*. The impedance of the second quarter wavelength section, $Z_2$, is $R/Z_1$. For three matching transformers, the transformer impedances are

$$Z_1 = Z_1 \qquad Z_2 = \sqrt{R} \qquad Z_3 = R/Z_1$$

**TABLE 7.1.3** Maximum VSWR Using Tchebyscheff Quarter Wavelength Transformers

| (*a*) One transformer | | | | | (*b*) Two transformers | | | | | (*c*) Three transformers | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| Impedance ratio, *R* | Bandwidth, $\omega_q$ 0.2 | 0.4 | 0.6 | 0.8 | Impedance ratio, *R* | Bandwidth, $\omega_q$ 0.2 | 0.4 | 0.6 | 0.8 | Impedance ratio, *R* | Bandwidth, $\omega_q$ 0.2 | 0.4 | 0.6 | 0.8 |
| 1.25 | 1.03 | 1.07 | 1.11 | 1.14 | 1.25 | 1.00 | 1.01 | 1.03 | 1.05 | 1.25 | 1.00 | 1.00 | 1.01 | 1.02 |
| 1.50 | 1.06 | 1.13 | 1.20 | 1.27 | 1.50 | 1.01 | 1.02 | 1.05 | 1.09 | 1.50 | 1.00 | 1.00 | 1.01 | 1.03 |
| 1.75 | 1.09 | 1.19 | 1.30 | 1.39 | 1.75 | 1.01 | 1.03 | 1.07 | 1.13 | 1.75 | 1.00 | 1.00 | 1.02 | 1.04 |
| 2.00 | 1.12 | 1.24 | 1.38 | 1.51 | 2.00 | 1.01 | 1.04 | 1.08 | 1.16 | 2.00 | 1.00 | 1.01 | 1.02 | 1.05 |
| 2.50 | 1.16 | 1.34 | 1.53 | 1.73 | 2.50 | 1.01 | 1.05 | 1.12 | 1.22 | 2.50 | 1.00 | 1.01 | 1.03 | 1.07 |
| 3.00 | 1.20 | 1.43 | 1.68 | 1.95 | 3.00 | 1.01 | 1.06 | 1.14 | 1.27 | 3.00 | 1.00 | 1.01 | 1.03 | 1.08 |
| 4.00 | 1.26 | 1.58 | 1.95 | 2.35 | 4.00 | 1.02 | 1.08 | 1.19 | 1.37 | 4.00 | 1.00 | 1.01 | 1.04 | 1.11 |
| 5.00 | 1.32 | 1.73 | 2.21 | 2.74 | 5.00 | 1.02 | 1.09 | 1.23 | 1.45 | 5.00 | 1.00 | 1.01 | 1.05 | 1.13 |
| 6.00 | 1.37 | 1.86 | 2.45 | 3.12 | 6.00 | 1.03 | 1.11 | 1.26 | 1.53 | 6.00 | 1.00 | 1.02 | 1.06 | 1.15 |
| 8.00 | 1.47 | 2.11 | 2.92 | 3.86 | 8.00 | 1.03 | 1.13 | 1.33 | 1.67 | 8.00 | 1.00 | 1.02 | 1.07 | 1.18 |

***Source:*** IRE Trans. PGMTT-8, pp. 478–482 (September 1960). Leo Young, "Tables for Cascaded Homogeneous Quarter-Wave Transformers."

***Matching with a Taper.*** Tapered transitions may be used to match from one transmission line to another. The VSWR is effected by the shape of the taper, i.e., linear or some other shape, and the length of the taper. In general a better match can be achieved using quarter wavelength sections when compared to a taper of the same length. The voltage reflection coefficient Γ is plotted in Fig. 7.1.3 for an impedance ratio of 2:1 for various taper lengths.

## Rectangular Waveguide

The electric field pattern of a dominant-mode ($TE_{10}$) rectangular waveguide is a half sinusoid across the transverse guide dimension with its maximum at the center of the broad wall. Microwave energy cannot be transmitted through a waveguide in frequencies below the cutoff frequency, which has a corresponding cutoff wavelength $\lambda_c$. When the inside broad wall width is *a*, the cutoff wavelength $\lambda_c = 2a$. A different mode (electric field configuration) can be transmitted through the waveguide when the frequency is greater than $c/2b$, where *c* is the speed of light and *b* is the narrow dimension of the waveguide. The normal waveguide operating

**TABLE 7.1.4** $Z_1$ for Tchebyscheff Quarter-Wave Transformers

| (*a*) Two-section transformer | | | | | | (*b*) Three-section transformer | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|
| Impedance ratio, *R* | Bandwidth, $\omega_q$ 0.0 | 0.2 | 0.4 | 0.6 | 0.8 | Impedance ratio, *R* | Bandwidth, $\omega_q$ 0.0 | 0.2 | 0.4 | 0.6 | 0.8 |
| 1.00 | 1.0000 | 1.0000 | 1.0000 | 1.0000 | 1.0000 | 1.00 | 1.0000 | 1.0000 | 1.0000 | 1.0000 | 1.0000 |
| 1.25 | 1.0573 | 1.0581 | 1.0603 | 1.0641 | 1.0697 | 1.25 | 1.0282 | 1.0288 | 1.0305 | 1.0335 | 1.0383 |
| 1.50 | 1.1066 | 1.1080 | 1.1123 | 1.1197 | 1.1305 | 1.50 | 1.0520 | 1.0530 | 1.0561 | 1.0618 | 1.0709 |
| 1.75 | 1.1501 | 1.1521 | 1.1583 | 1.1690 | 1.1846 | 1.75 | 1.0725 | 1.0739 | 1.0783 | 1.0864 | 1.0993 |
| 2.00 | 1.1892 | 1.1918 | 1.1997 | 1.2136 | 1.2338 | 2.00 | 1.0906 | 1.0924 | 1.0980 | 1.1083 | 1.1246 |
| 2.50 | 1.2574 | 1.2611 | 1.2724 | 1.2921 | 1.3211 | 2.50 | 1.1217 | 1.1242 | 1.1319 | 1.1460 | 1.1686 |
| 3.00 | 1.3160 | 1.3207 | 1.3352 | 1.3604 | 1.3976 | 3.00 | 1.1479 | 1.1509 | 1.1605 | 1.1779 | 1.2062 |
| 4.00 | 1.4142 | 1.4208 | 1.4410 | 1.4764 | 1.5289 | 4.00 | 1.1907 | 1.1947 | 1.2074 | 1.2308 | 1.2689 |
| 5.00 | 1.4953 | 1.5036 | 1.5292 | 1.5740 | 1.6408 | 5.00 | 1.2252 | 1.2301 | 1.2455 | 1.2741 | 1.3207 |
| 6.00 | 1.5650 | 1.5750 | 1.6056 | 1.6593 | 1.7397 | 6.00 | 1.2543 | 1.2600 | 1.2779 | 1.3110 | 1.3655 |
| 8.00 | 1.6817 | 1.6947 | 1.7347 | 1.8052 | 1.9110 | 8.00 | 1.3021 | 1.3091 | 1.3312 | 1.3725 | 1.4409 |

***Source:*** IRE Trans. PGMTT-8, pp. 478–482 (September 1960). Leo Young, "Tables for Cascaded Homogenous Quarter-Wave Transformers."

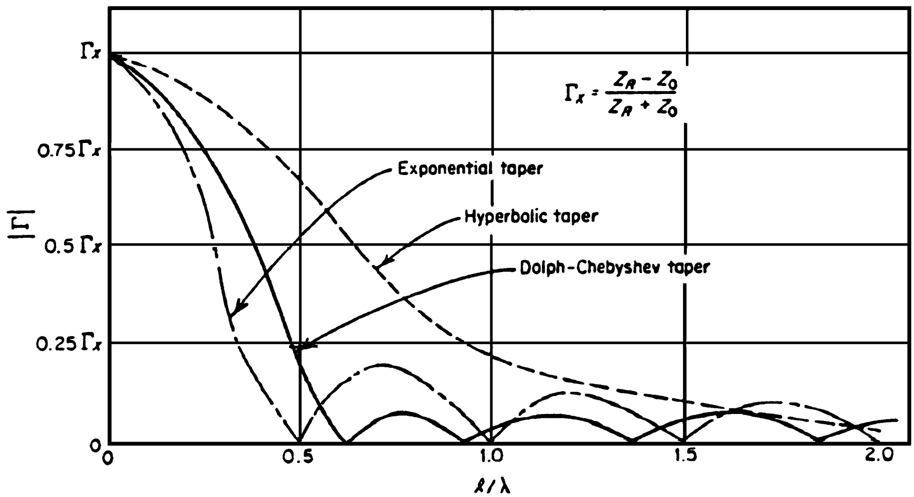



**FIGURE 7.1.3** Use of tapers for impedance matching.

frequency range is significantly higher than the cutoff frequency and significantly lower than the frequency where the second mode can be transmitted.

When higher power must be transmitted than is possible because of breakdown in air with these restrictions on dimensions, pressurization of the air or a sulfur hexafluoride gas fill is generally used.

Unlike the case for the transverse electromagnetic (TEM) mode of a coaxial or parallel-wire line, the guide wavelength $\lambda_g$ in rectangular waveguide departs from the free-space wavelength $\lambda_0$ and varies with frequency, producing phase distortion in a wide-band signal. The relationship is

$$\lambda_g = \lambda_0 / \sqrt{1 - (\lambda_0 / \lambda_c)^2}$$

## Ridge Waveguide

To obtain broader bandwidth in a single mode, as well as increased flexibility in the choice of impedance, ridge waveguide is generally employed. The variation of cutoff wavelength with dimensions for single and double ridges (always inserted on the broad wall of a rectangular guide to give capacitive loading) is given in Fig. 7.1.4. A large increase in bandwidth is made possible by this type of guide.

## Circular-Mode Transmission

The circular electric ($TE_{01}$) mode is currently employed in long-distance broadband communications at millimeter wavelengths because of its low-loss transmission property. The electric field pattern and the wall currents form concentric rings in this unusual mode. Conversion to lower-order modes occurs if the cross section is even slightly elliptical or if the guide axis is curved. Various forms of mode suppression are employed; a common technique consists in fabricating the cylindrical wall from a tightly wound helix with loss between the turns to damp out modes with axial components of current. Attenuation as low as a few decibels per mile has been achieved.

## Coaxial Line

The attenuation and power handling capabilities of rigid coaxial lines of various outer diameters are shown in Fig. 7.1.5. The frequency scale on this figure extends only to 3 GHz because such high-power transmitting coaxial cable is generally not used at higher frequencies.

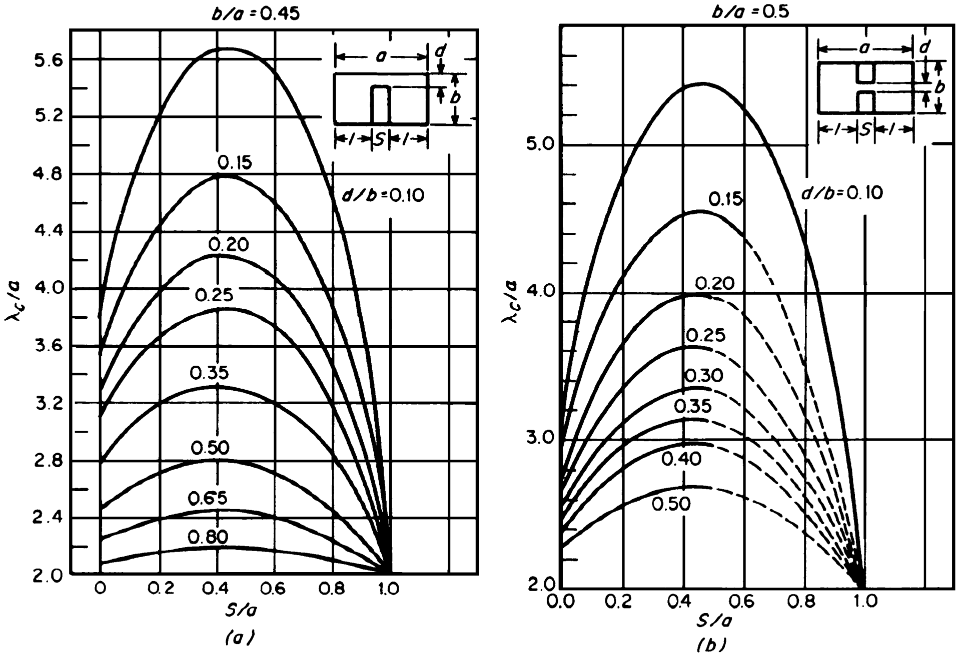



**FIGURE 7.1.4** Cutoff wavelength of ridged waveguide: (*a*) single ridge; (*b*) double ridge.

Care must be taken to avoid operation of a coaxial line at wavelengths where it becomes possible for additional modes to propagate. This occurs when the mean circumference between the inner and outer cylinders forming the transmission path equals a full wavelength; a stable standing wave is then possible in a circumferential direction. The higher attenuation and reduced power-carrying capability which accompany the small dimensions necessary to avoid higher-order modes generally lead to the choice of waveguide as a transmission medium at frequencies above 3 GHz for medium and high-power applications.

## Strip Transmission Line

Strip transmission lines consist of a flat conductor between two ground planes. The characteristics are similar to coaxial line; however, they are easier to change by varying the width of the circuit, the proximity of adjacent lines or by the addition of stubs or other circuit elements. Impedance is controlled by the width of the strip; coupling is controlled by the spacing between lines. The circuit is flat and is easily fabricated. The material between the ground planes and the strip can be air or a low-loss dielectric material such as those enumerated in Table 7.1.5. Components such as directional couples, filters, switches, and isolators are frequently made in stripline. Connectors are usually provided to join one component to another. Characteristic impedance $Z_0$ of a dielectrically loaded strip line ($\epsilon$ = dielectric constant) is given in Fig. 7.1.6. Figure 7.1.7 illustrates typical attenuation curves. A stripline with 0.125-in. ground plane spacing and teflon fiber glass dielectric will usually handle 25 kW peak power and 500 W average power in C-band. The dielectric material and the ground plane spacing must be chosen so that the line propagates only the TEM mode.

## Microstrip Transmission Line

A microstrip transmission line has a conductive circuit on top of a dielectric substrate with a ground plane below the substrate. The advantages of microstrip include low cost production in large volume; the ability

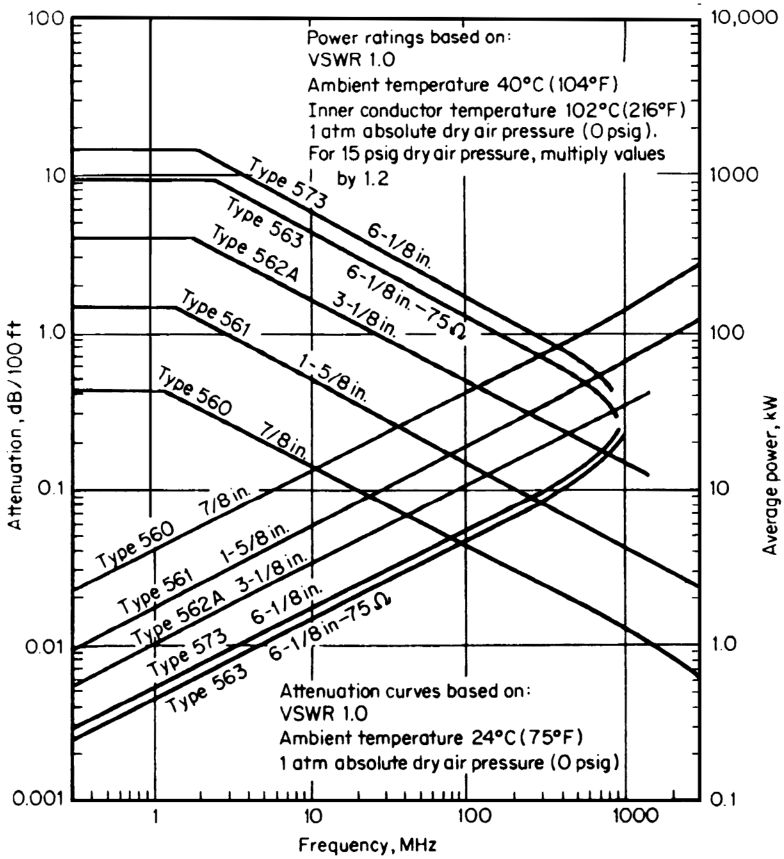



**FIGURE 7.1.5** Power rating and attenuation for rigid coaxial line.

to mount components (such as resistors and dielectric cavities) on top of the circuit; and to adjust the circuit which is open so it can be easily worked on. No connectors are required between circuit elements, thereby reducing size. The circuit itself has somewhat higher loss than stripline, but the elimination of connectors and the reduction in size frequently results in surprisingly low loss. Disadvantages include relatively low power capability (satisfactory performance of 100 W average at 2 GHz), tight tolerances, launching difficulty above 15 GHz, and leakage across the open circuit. Absorptive material and grounding screws may be required. Typical substrate materials are shown in Table 7.1.5. Figure 7.1.8 shows characteristic impedance of microstrip. In microstrip, part of the field is above the circuit in air and part below the circuit in dielectric. The resulting effective dielectric constant is less than the dielectric constant of the substrate as shown in Fig. 7.1.9.

## Finline

A finline consists of rectangular waveguide with a metallic fin attached along the centerline of the top wall of a waveguide and another fin along the centerline of the bottom wall (Fig. 7.1.10). The fin must have electrical continuity with the waveguide walls. A dielectric substrate is located adjacent to the fin and extends from the bottom to the top wall of the waveguide. This structure has low loss and can operate up to 100 GHz.

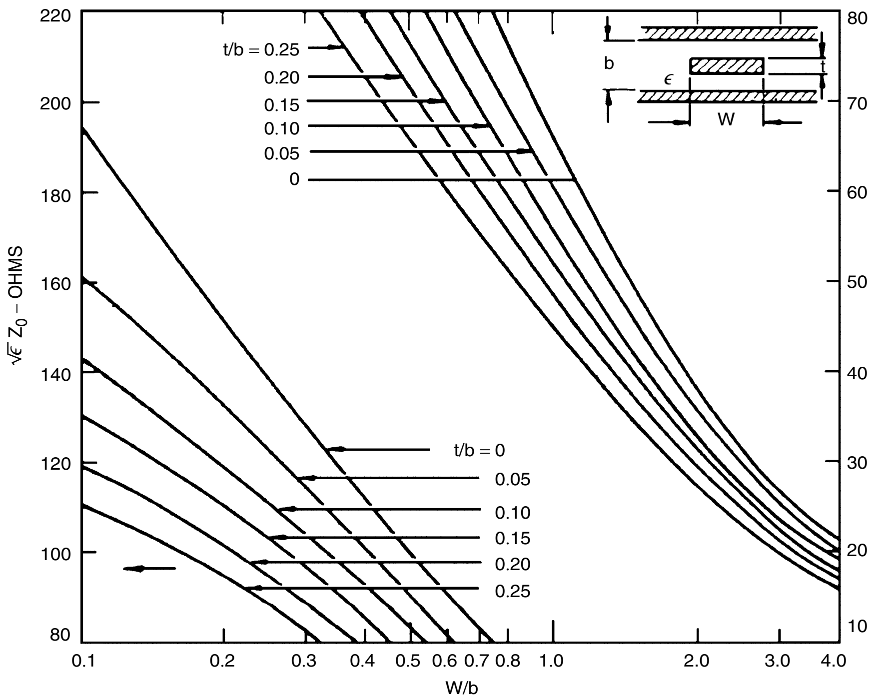



**FIGURE 7.1.6** General curves for characteristic impedance of dielectrically loaded stripline (dielectric constant = $\epsilon$).

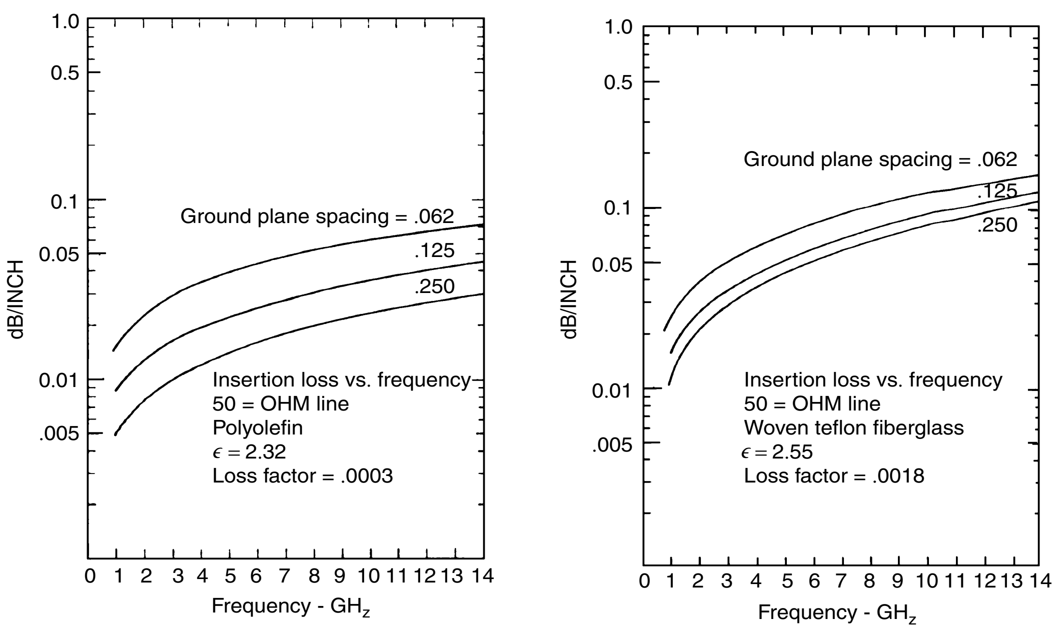



**FIGURE 7.1.7** Insertion loss vs. frequency for polyolefin and fiberglass, 50-ohm line.

**TABLE 7.1.5** Dielectric Materials

| Material | Dielectric constant | Loss factor | Useful temp range, °C | Flexibility | Coeff. of thermal expansion × $10^6$, °C | Surface finish, $\mu$ in* |
|---|---|---|---|---|---|---|
| Woven TFG | 2.55 | 0.0015 | –60 to +200 | Good | 18.5 | N/A |
| Microfiber TFG, | | | | | | |
| Duroid 5870 | 2.33 | 0.0005 | –60 to +200 | Good | 5 | N/A |
| Duroid 5880 | 2.2 | 0.0006 | –60 to +200 | Good | 32 | N/A |
| Polystyrene | 2.53 | 0.0003 | –60 to +100 | Very poor | 7 | N/A |
| Reinforced | 2.62 | 0.002 | –60 to +100 | Poor | 5.7 | N/A |
| Polyphenelene oxide (PPO) | 2.55 | 0.0016 | –60 to +200 | Fair | 29 | N/A |
| Polyolefin | 2.32 | 0.00015 | –60 to +100 | Excellent | 4.4 | N/A |
| Quartz Teflon | 2.47 | 0.0006 | –60 to +200 | Good | 18.5 | N/A |
| Polymide, Micaply 5032 | 4.8 | 0.01 | –60 to +250 | Good | 9 | N/A |
| Epsilam-10 | 10.0 | 0.002 | –60 to +150 | Good | 11–23 | N/A |
| 99.5% alumina | 9.9 | 0.00008 | Up to 500 | Very poor | 7.5 | <3 |
| Quartz (fused silica) | 3.78 | 0.0001 | Up to 500 | Very poor | 0.55 | <1 |
| Sapphire | 9.4 | 0.00008 | Up to 500 | Very poor | 7.7 | <1 |
| | 11.6 | | | | 8.3 | |
| 99.5% BeO | 6.6 | 0.0004 | Up to 500 | Very poor | 7.8 | 3–7 |
| Boron nitride | 4.4 | 0.0003 | Up to 500 | Poor | 1–2 | N/A |

*N/A = Not available.

## Transitions Between Transmission Lines

***Waveguide to Coaxial Line.*** The maximum E field is in the center of the waveguide at a distance $\lambda_g/4$ from a short at the end of the guide. An extension of the center conductor, located at the point of maximum E field, acts like an antenna to couple energy from the coaxial line into the waveguide, Fig. 7.1.11. Typically the diameter of

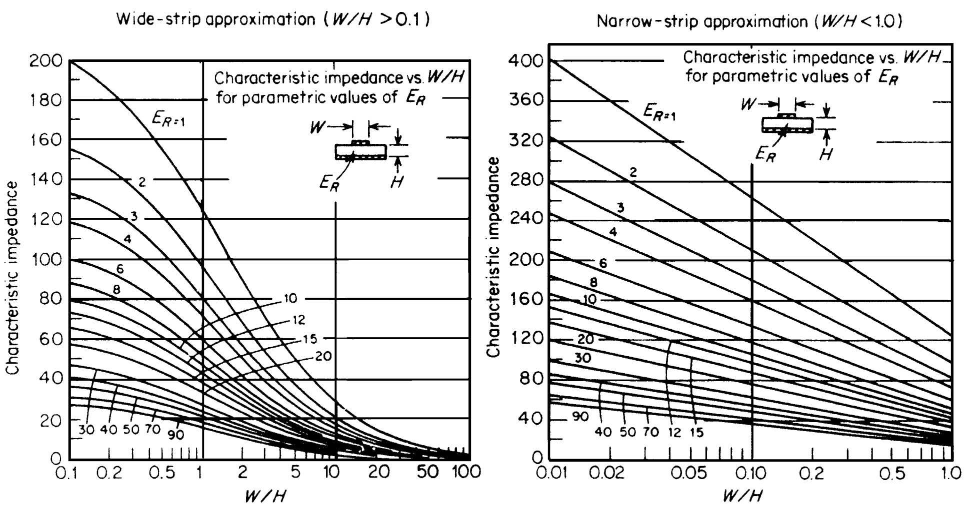



**FIGURE 7.1.8** Characteristic impedance of microstrip line: wide-strip approximation (*left*); narrow-strip approximation (*right*).

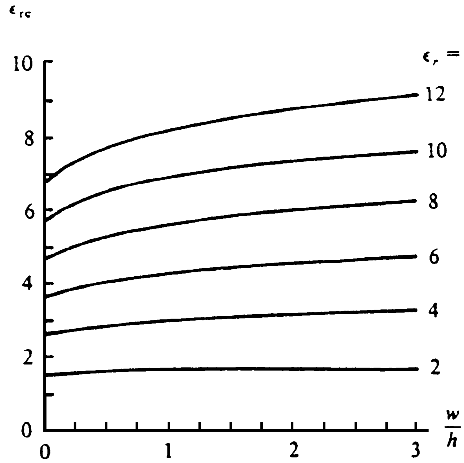



**FIGURE 7.1.9** Plot of the effective dielectric constant vs. $w/h$ for microstrip transmission lines, with the relative dielectric constant as a parameter.

the probe is enlarged and enclosed in a dielectric cylinder to increase bandwidth. Coupling from coaxial line to waveguide may also be achieved by using a loop, as shown in Fig. 7.1.12 which couples to the magnetic fields.

***Waveguide to Stripline or Microstrip.*** The center conductor of a stripline may be extended into a waveguide forming a probe. Increasing the width of the center conductor at the end of the probe may improve bandwidth. Similarly the conductor and substrate of a microstrip circuit, but not the ground plane, may be extended directly into the guide, Fig. 7.1.13.

***Coaxial Line to Microstrip.*** The center conductor of a coaxial line is pressed against or soldered to the strip conductor of microstrip. The outer conductor of the coaxial line is grounded to the microstrip groundplane. At frequencies above 15 GHz, the microstrip substrate thickness may be as little as 0.010 in. which usually requires decreasing the diameter of the coaxial line as illustrated in Fig. 7.1.14. The center conductor must not come so close to the groundplane that it introduces a capacitive discontinuity. It may be necessary to short out unwanted modes with grounded lines or screws near the transition. The exact mechanical configuration and capacitive loading become critical.

## Directional Couplers

***Waveguide Directional Couplers.*** Waveguide directional couplers are employed primarily to sample power for measurements. Two waveguides may be located side by side, one above the other, either parallel to each other or crossing each other. Holes are drilled in the common wall to permit coupling power between guides. In Fig. 7.1.15 the coupling = 10 log $P_1/P_4$, while the directivity = 10 log $P_4/P'_4$ ($P_1$ = power in at port 1, $P_4$ = power out at port 4 with power $P_1$ in a port 1, $P'_4$ is the power at port 4 with power $P_1$ in at port 2). Coupling values range from 3 dB to more than 30 dB with directivity in excess of 40 dB.

***Stripline and Microstrip.*** A stripline or microstrip coupler consists of a main transmission line in close proximity to a secondary line as illustrated in Fig. 7.1.16. The definitions of coupling and directivity are the

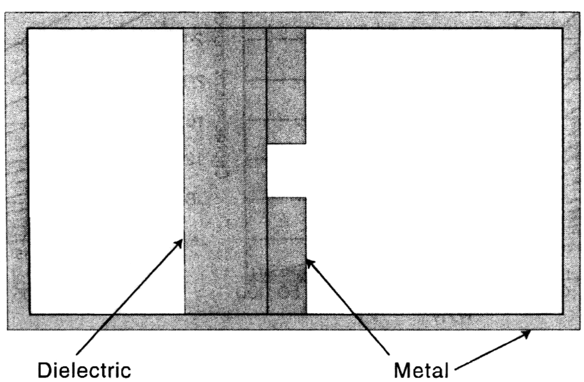



**FIGURE 7.1.10** Finline configuration.

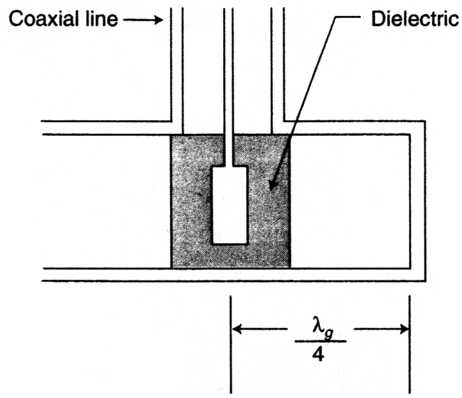



**FIGURE 7.1.11** Electric field coupled waveguide to coax adapter.

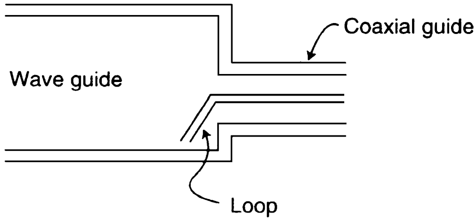



**FIGURE 7.1.12** Magnetic field coupled waveguide to coax adapter.

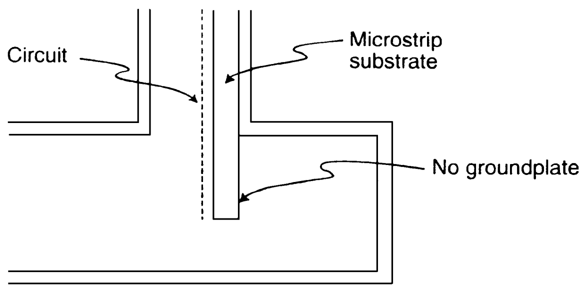



**FIGURE 7.1.13** Waveguide to microstrip adapter using circuit probe.

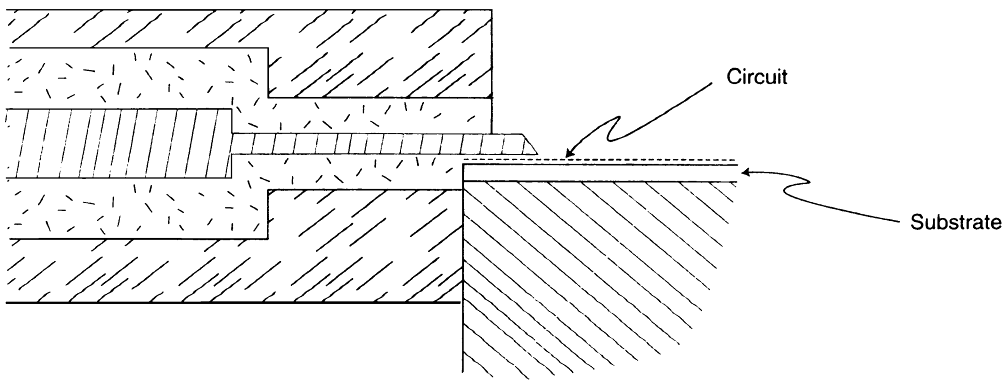



**FIGURE 7.1.14** Coaxial line to microstrip adapter using constant impedance transformer in coaxial line.

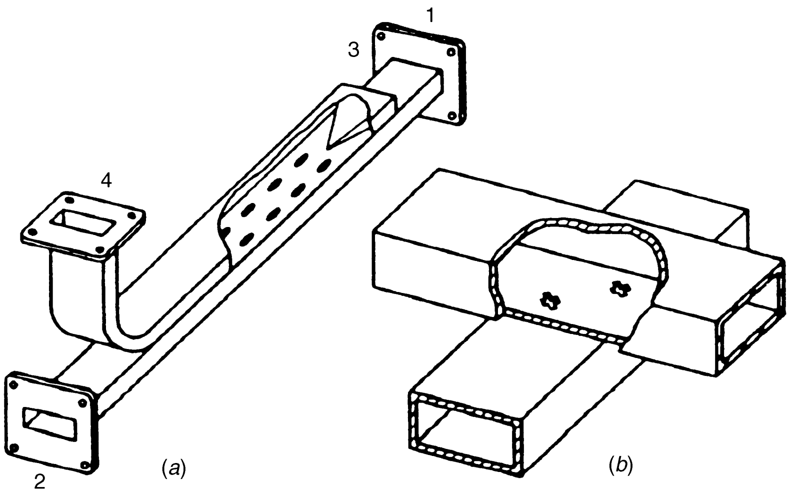



**FIGURE 7.1.15** Directional couplers in waveguide: (*a*) with array of openings, (*b*) crossguide coupler.

same as for waveguide, but the coupled part is adjacent to the input rather than at the far end. Adding quarter wavelength coupling sections on either side of the center section increases bandwidth and reduces ripple. The added quarter-wave sections are less tightly coupled than the center section and are equally disposed about it. For microstrip, the velocity of propagation is different for even and odd modes. The addition of localized capacitances along the line, Fig. 7.1.17, improves the directivity.

***Even and Odd Modes.*** When two lines are in close proximity and the phase of the energy is the same, there is even-mode symmetry of fields as illustrated in Fig. 7.1.18*a*. When the fields are 180° out of phase there is odd mode symmetry, Fig. 7.1.18*b*. The impedance of the two modes are different. In addition, in microstrip, one mode has more field in air than the other, resulting in different propagation velocities.

***In-Line Power Dividers.*** A Wilkinson power divider consists of a single line separated into two quarter wavelength sections as shown in Fig. 7.1.19. The divider can have a VSWR of 1.25:1 or less over an octave. Power division is frequency independent.

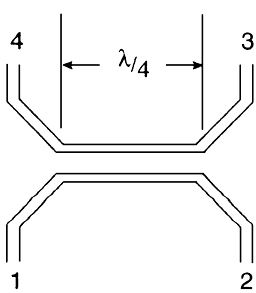



**FIGURE 7.1.16** Stripline directional coupler.

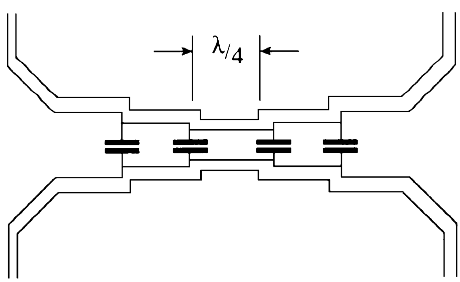



**FIGURE 7.1.17** Microstrip coupler with capacitors to improve the directivity.

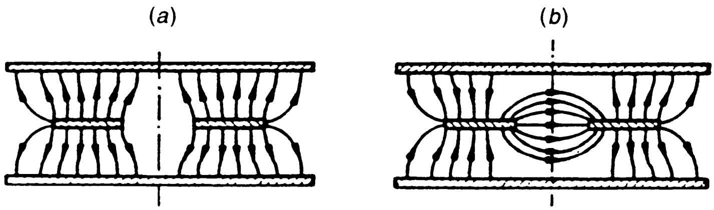



**FIGURE 7.1.18** Even and odd modes on a homogeneous coupled line.

## Resonant Circuits

Resonant circuits are generally formed from a short-circuited length of transmission line. Many modes can exist within the cavity created by the shorted line, as illustrated in Fig. 7.1.20 for a right circular cylinder. The size of the cavity and the quality of resonance, $Q$, is determined by the mode selected. A resonator with a $Q$ of 100 has poor selectivity, while a resonator with a $Q$ of 5000 will have good selectivity. The $Q$ is defined as

$$Q = \frac{\text{Total energy stored in electric and magnetic fields}}{\text{Energy dissipated per cycle}}$$

The operating frequency can be changed by mechanically changing the cavity geometry, or by introducing a metal or dielectric post.

Dielectric disks and cylinders can be used as cavities. They are most easily coupled to microstrip, but can be used with coaxial line or waveguide, or other types of transmission lines. Materials are available from 0.7 to 30 GHz, with typical dielectric constants of 36, which reduces the dimensions by a factor of 6 when compared to air. $Q$ (1/tan $\delta$) is typically 30,000/frequency in GHz, where tan $\delta$ is frequently used as a measure of loss in materials. Tuning of 5 percent or more is achieved by moving metal or dielectric tuning stubs near the dielectric cylinder.

Mechanically tuned resonant cavities are generally used for frequency determination. For a transmission wavemeter, such a cavity is coupled in series into the transmission path, while for an absorptive indication it is coupled in shunt. A dominant-mode ($TE_{111}$) cylindrical resonator is most widely used for this purpose, but for highest selectivity a circular electric-mode ($TE_{011}$) resonator is employed. Dielectric resonators are frequently used to tune or stabilize oscillators.

## Filters

Frequency filters are used to separate the components of a composite waveform for signal-processing purposes or to suppress RF interference (RFI), which results from the spurious output of transmitters. The latter problem

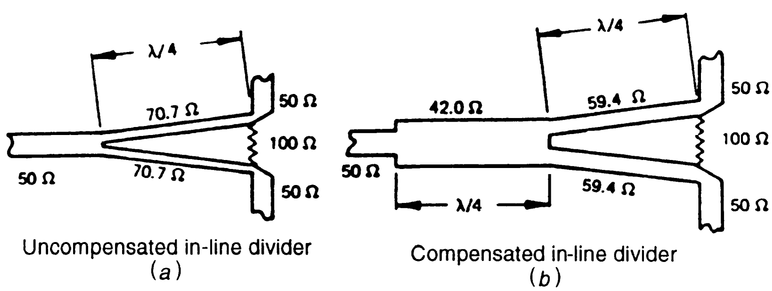



**FIGURE 7.1.19** Compensated and uncompensated in-line three-port power dividers.

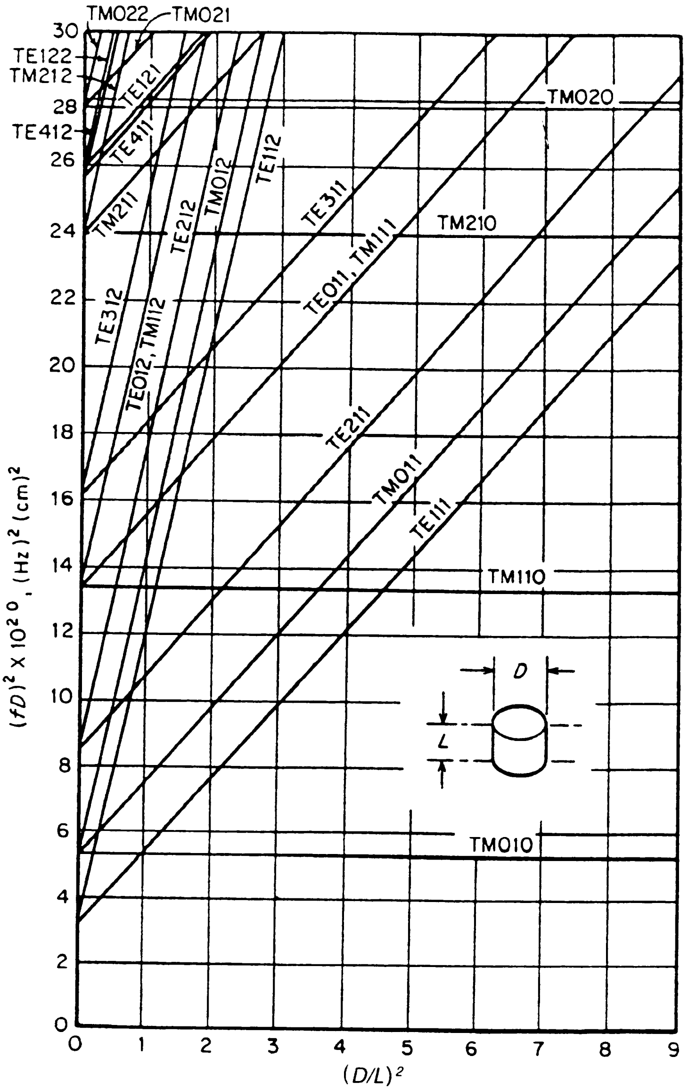
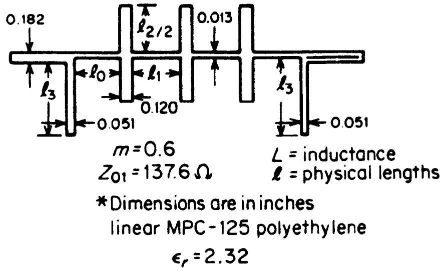



**FIGURE 7.1.20** Mode chart for right circular cylinder.

has only recently become serious at microwave frequencies, as this area of the spectrum has become congested. High-power capability is required for such filters, leading generally to the use of waveguide structures. A section of waveguide beyond cutoff constitutes a simple high-pass reflective filter. Loading elements in the form of posts, irises, or stubs are employed to supply the reactances required for conventional lumped-constant-filter design.

The desired skirt steepness and stopband attenuation determine the number of sections, as at lower frequencies. A disk-loaded coaxial line is generally used as a low-pass high-power filter. Insertion loss of reflective filters is typically 0.1 to 0.2 dB, with stopband attenuation of the order of 50 dB. Absorption filters avoid the reflection of unwanted energy by incorporating lossy material in secondary guides which are coupled through leaky walls (typically small sections of guide beyond cutoff in the passband). These filters are effective primarily against harmonics.




| Frequency $f_c$ | Dimensions $\ell_0$ $0.084\lambda_{\epsilon_r}$ | $\ell_1$ $0.100\lambda_{\epsilon_r}$ | $\ell_{2/2}$ $0.144\lambda_{\epsilon_r}$ | $\ell_3$ $0.200\lambda_{\epsilon_r}$ |
|---|---|---|---|---|
| 1 GHz | 0.6653 | 0.7920 | 1.093 | 1.548 |
| 2 GHz | 0.3326 | 0.3960 | 0.533 | 0.756 |
| 3 GHz | 0.2220 | 0.2640 | 0.333 | 0.493 |

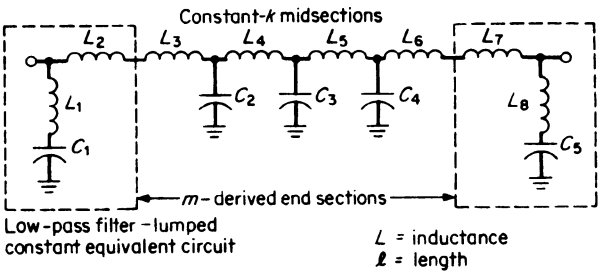



**FIGURE 7.1.21** Strip-line filter design.

For low-power applications, strip-line filters are widely used because of their compact sizes and low cost. Typical dimensions for a low-pass filter of this type are shown in Fig. 7.1.21.

## Other Components

Among the components useful for measurement purposes are wavemeters, attenuators, and matched terminations. In waveguide, variable attenuators take the form of thin absorptive material introduced tangential to the electric field typically through a slot in the broad wall of rectangular guide to produce minimum reflection. Load terminations are tapered attenuators designed to produce at least 40 dB of return loss, while maintaining a good match (maximum VSWR less than 1.2) through the specified band. For high power, water cooling is provided either around a loaded ceramic or dielectric absorber or by introducing a tube of water directly into the guide to act as the absorber. Calorimetric determination of power is possible with such water loads. A wide range of grounded resistive elements are used as terminations.

## Ferrite Components

***Isolators.*** Isolators transmit microwave power in one direction with little attenuation, while power transmitted in the opposite direction is absorbed. Attenuation in the transmit direction may be less than 0.5 dB, while isolation in the opposite direction may exceed 20 dB. Isolators are relatively well matched in both directions, a VWSR less than 1.20:1 is typical. Such an isolator will reduce a mismatch from 2:1 to less than 1.3:1. When used between stages of an amplifier they may eliminate the need for tuning between stages and may improve gain flatness. Isolators are available in waveguide, in stripline with or without coaxial connectors and in microstrip. At 1.0 GHz a differential phase shift type waveguide isolator can handle 4000 kW peak power and

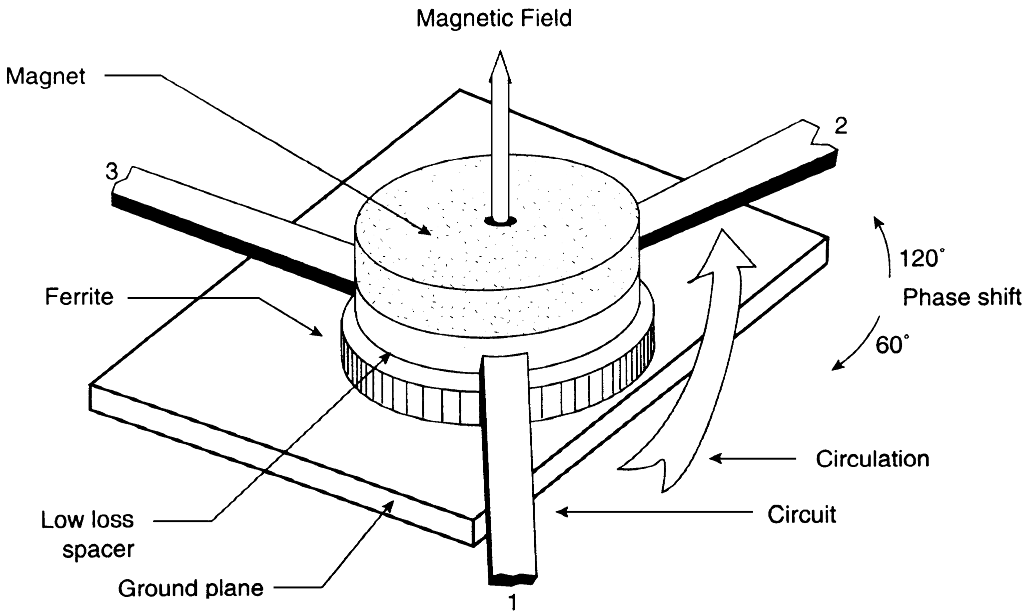



**FIGURE 7.1.22** Configuration of microstrip junction circulator.

over 250 kW average power. At higher frequencies, such as Ku-band, a microstrip isolator may handle only a few watts and weigh only a few grams.

Isolators and circulators are based on the interaction of a circularly polarized magnetic component of microwave energy with the magnetic field of a ferromagnetic molecule, usually iron, which has been aligned by the application of a steady dc magnetic field. The ferromagnetic molecules are held in a nonconductive crystalline lattice through which microwave energy can pass with little loss. The interaction between the microwave magnetic field and the magnetic field of the molecule is a vector product interaction which tends to cause the molecule to process like a gyroscope, and is frequently referred to as gyromagnetic interaction. The direction of the applied dc magnetic field determines the direction for transmission and attenuation of the isolator. Reversing the direction of the applied field reverses the direction of the isolator.

***Circulators.*** The most common type of circulator uses a single junction and can be built in waveguide, stripline or microstrip. Figure 7.1.22 illustrates a single junction microstrip circulator. The stack-up of parts consists of a groundplane (frequently metalized on the ferrite), a ferrite disk, a conductive metal circuit with arms at 120° relative to each other, a spacer to keep the microwave fields out of the magnet, and a magnet to supply the dc magnetic field. Due to the nonreciprocity of the magnetically biased ferrite, the phase shift between ports in the circulation direction is 120° while the phase shift in the opposite direction is 60°. Energy transmitted from port one to port two is shifted 120°, while energy from 1 to 3 is shifted 60° and energy from 2 to 3 is shifted 60°. Energy going either direction is in phase at port 2 and adds together. Energy at port 3 is out of phase and cancels so no energy is transmitted to this port. Due to symmetry, this applies to all ports. Typical characteristics of junction circulators is shown in Table 7.1.6. Placing a termination at port 3 converts the circulator to an isolator.

***Differential Phase Shift Circulators.*** High-power circulators are generally built in waveguide as shown diagrammatically in Fig. 7.1.23. If the phases are added up as shown, it will be seen that power in at 1 goes to 2, power in at 2 goes to 3, and so on. It should be noted that in a folded hybrid Tee, power in at 1 has the same phase in path A and path B. Power in at the vertical arm of the Tee is 180° out of phase for the two paths. A similar phase shift occurs for power traveling in the opposite direction. There is a 90° phase shift across the short slot hybrid as shown. Placing termination at ports 3 and 4 converts the circulator into an isolator.

**TABLE 7.1.6** Characteristics of Typical Junction Circulators

| Circulator type | Center frequency, GHz | Bandwidth, % | Isolation, dB | Insertion loss, dB | Power capacity Avg, W | Power capacity Peak kW |
|---|---|---|---|---|---|---|
| Waveguide | 12.4–18 | . . . . | 20 | 0.3 | 20 | 1 |
| | 18–26.5 | . . . . | 17 | 0.5 | 15 | 0.5 |
| Strip transmission | 4–8 | . . . . | 20 | 0.4 | 35 | |
| line | 12.4–18 | . . . . | 18 | 0.5 | 25 | 0.25 |
| Switching | 2.9 | 8.9 | 26 | 0.35 | | 15 |
| | 35 | 5 | 15 | 0.5 | | 15 |
| Lumped constant | 1.2 | 30 | 20 | 0.6 | | |
| | 0.4–0.5 | 30 | 20 | 0.4 | | |

***Source:*** *Microwave J.*, November 1978.

Circulators may be used to simultaneously connect a transmitter and receiver to a single antenna as shown in Fig. 7.1.24. Transmitted power in at 1 goes to the antenna at 2. Received energy entering the antenna at 2 travels to the receiver at 3. Transmitter energy reflected from the antenna also goes to the receiver. The reflected power must be low enough it does not damage the receiver or a fast-acting switch must be provided to protect the receiver. If the transmitter and the receiver are operating at different frequencies, protection may be provided by a filter.

***Phase Shifters.*** The simplest reciprocal phase shifter uses a thin ferrite rode centered within a rectangular waveguide. An axially directed magnetic field provides a full 180° phase variation at 9 GHz with a change of a few hundred gauss. A more cost-effective approach is the dual-mode reciprocal phase shifter, shown in Fig. 7.1.25, which uses Faraday rotation. Toroidal nonreciprocal phase shifters, in which the ferrite takes the form of a rectangular closed loop in waveguide, switched by means of an axial current-carrying conductor, are also popular. Digital phase shifting is generally obtained by latching ferrites, since the remanent magnetization states of the hysteresis loop require relatively small bias fields.

*Limiters.* The nonlinear behavior of ferrites at high power levels is used in ferrite limiters. Such limiters have replaced TR gas-discharge tubes in radar. Peak powers of tens of kilowatts at X band can be handled, but with considerable spike-energy leakage, which requires a follow-on solid-state (*pin*) diode stage. Typical insertion loss is

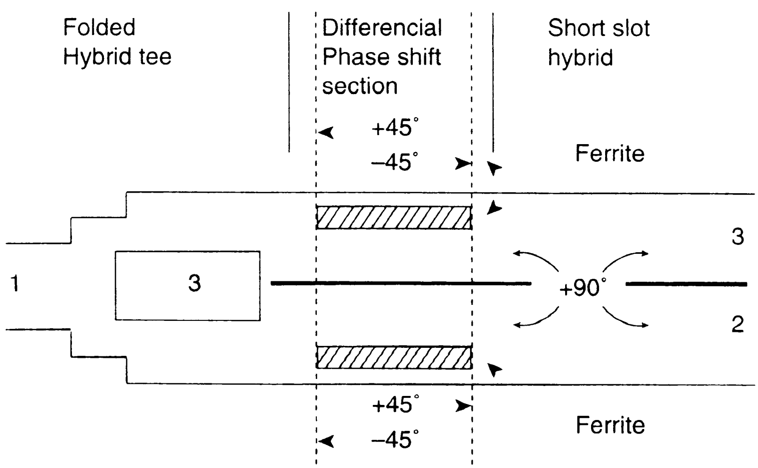



**FIGURE 7.1.23** Diagram of differential phase shift waveguide circulator.

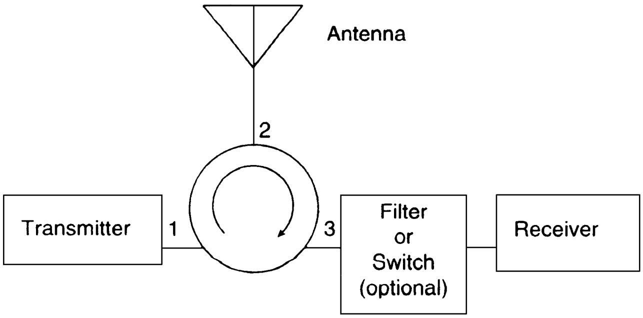



**FIGURE 7.1.24** A transmitter and receiver connected to same antenna.

about 0.5 dB, with 30 dB of flat limiting and about 6 dB of spike limiting for the ferrite alone. The diode stage increases the insertion loss to about 1 dB but reduces all leakage an additional 30 dB. The recovery time is very short, typically 100 ns, and is determined primarily by the diode section.

## Acoustic-Wave Devices

Acoustic waves owe their usefulness to their relatively low velocity, typically $10^{-5}$ times the electromagnetic velocity, permitting relatively long electric-signal delay times to be obtained in a physically small space. Both bulk-mode propagation and surface waves have been used, the latter gaining in popularity because of the relative ease of access to intermediate points along the propagation path for structure shaping and taping.

***Transducers.*** The transducers designed for the two types of acoustic modes are physically quite different, but both contain electrodes spaced either a quarter or half an acoustic wavelength in a piezoelectric material. Microwave bulk-wave transducers consist of multiple films of ZnO or CdS, each approximately $\lambda/4$ thick, separated by similar films of gold or other nonpiezoelectric material deposited on the bulk medium to provide electric-to-acoustic wave coupling and impedance transformation.

Surface acoustic waves (SAW) are generally excited by interdigital transducers. An interdigital transducer, illustrated in Fig. 7.1.26, consists of two sets of interleaved metal electrodes, called *fingers*, deposited on the piezoelectric substrate. To generate a wave an RF potential is applied between the adjacent sets of fingers, which are spaced by a distance equal to one-half wavelength at the transducer design frequency. A typical 100-MHz transducer on $LiNbO_3$ has aluminum fingers 0.2 $\mu$m thick by 9 $\mu$m wide with 9-$\mu$m gaps.

The wave excited by the RF potential between a pair of fingers travels at the surface-wave velocity. By the time the wave arrives midway between the next pair of fingers, the RF excitation potential has reversed sign, and the wave excited by the second pair of fingers will be in phase with the wave from the first pair. Thus the

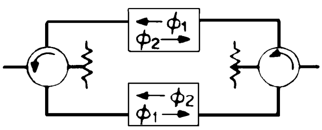



**FIGURE 7.1.25** Dual-mode reciprocal phase shifter.

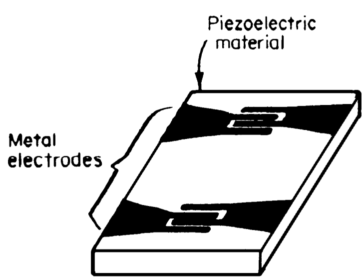



**FIGURE 7.1.26** Piezoelectric surface-wave microwave device with interdigital electrodes.

PASSIVE MICROWAVE COMPONENTS **7.23**

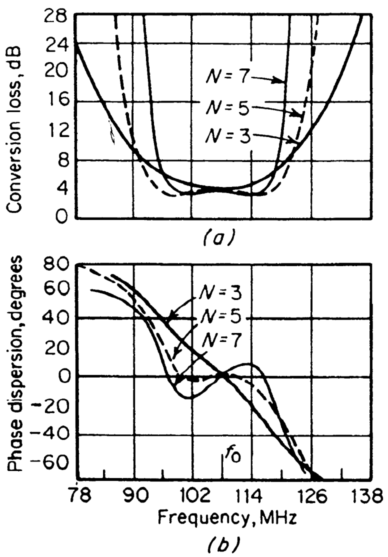



**FIGURE 7.1.27** Coupling between electric port and one acoustic port for transducers with three, five, and seven interdigital periods: (*a*) theoretical conversion loss; (*b*) phase dispersion.

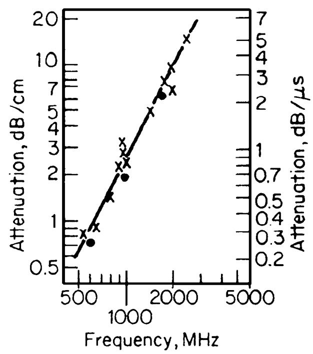



**FIGURE 7.1.28** Surface-wave attenuation of Y-cut lithium niobate.

excitation due to the second pair is added to the excitation from the first, and so on. The mechanism is reciprocal, and hence the transducer that excites a wave will also detect it.

The transducer has a fractional bandwidth of $1/N$, where $N$ is the number of finger parts. Electrically, the transducer is represented by a capacitance shunted by a radiation resistance which depends on the choice of finger length.

A surface-wave transducer is a three-port device, i.e., with one electric and two elastic ports. Figure 7.1.27 shows the conversion loss as a function of frequency from a 50-Ω source to one of the two acoustic outputs for three different transducers on a lithium niobate surface. These calculated curves show that for this particular case the use of five finger pairs provides the widest bandwidth and smallest conversion loss. The attenuation of the wave, once it has been launched, is given in Fig. 7.1.28 for lithium niobate.

The lowest operating frequency of acoustic surface-wave devices is limited by the allowable size to typically 10 MHz. At present the upper operating frequency is limited by fabrication difficulties to about 1 GHz with possible operation to 3 GHz.

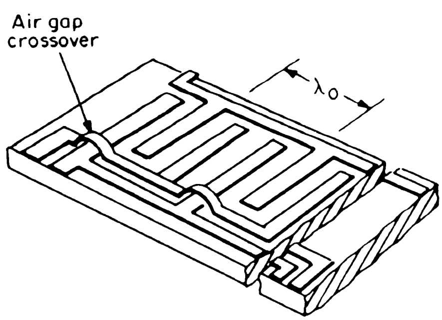



**FIGURE 7.1.29** Three-phase unidirectional transducer. Each electrode is 120° out of phase, so that the acoustic waves add in one direction and cancel in the other.

***Other Acoustic-Wave Devices.*** Bandpass filtering is the principal commercial application of SAW technology. Such devices are replacing *LC* filters in television receiver i.f. circuits. Minimum stop-band rejection of the order of 60 dB with in-band response flat to ±0.1 dB and phase deviation from linearity of only a few degrees are typical of these filters. The major drawback of SAW devices, their high insertion loss (of the order of 15 dB), can be reduced to less than 3 dB by using a three-phase unidirectional transducer structure, as shown in Fig. 7.1.29. Other acoustic-wave devices include: (a) dispersive filters used primarily for pulse compression; (b) chirp transformers that can be used to obtain a Fourier transform of an input signal; and (c) compact, low-cost SAW resonators, which are useful below 1 GHz.

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