Chapter 13: Microwave Communication Systems

Chapter 13 Objectives

At the conclusion of this chapter, the reader will be able to:

- Describe the differences between microwave and lower-frequency communications techniques.
- Describe the different modes of propagation within a waveguide.
- Explain the operation of a magnetron and other microwave oscillators.
- Explain the operation of traveling wave tube (TWT) and klystron amplifiers.
- Expain the operation of a parametric amplifier.
- Describe the operation of radar systems using time-of-arrival and Doppler shift methods.

Microwave communications systems are those that operate above 2 GHz (2000 MHz), approximately. Microwave communications applications include satellites, terrestrial (earth-based) relay links, radar, plus some consumer applications such as cordless telephones, the global positioning system (GPS), IEEE 802.11 wireless LANs, and many others. At these high radio frequencies, conventional circuit construction techniques using *lumped* (discrete) L and C components are not very effective for reasons we'll soon discuss. The circuit elements that you have used in your studies of electronic fundamentals (resistors, capacitors, inductors, and so on) are referred to as *lumped* components, because the "works" of each part are contained or "lumped" within a single package. You can easily identify the individual parts in a circuit containing these elements. Each part performs one function.

Microwave circuits often appear to be little more than a collection of tubing and other "plumbing parts" to the novice technician. Each component in a microwave system has a function similar to the *groups* of lumped components used in lower-frequency approaches. There's nothing mysterious or magic about microwaves, once you learn a few new principles.

13-1 Microwave Construction Techniques

Limitations of Conventional Components

There are undesired or *parasitic* elements in all conventional electronic components. At low frequencies, the effects of these additional elements can largely be ignored; this is not true at UHF and SHF! For example, the 100 Ω resistor shown in Figure 13-1 actually consists of several equivalent components. The inductors L_1 and L_2 represent the *lead inductance* of the component. These values are dependent upon the length of the resistor's leads, and are generally a few nH or less. In addition, the resistor appears shunted by a *package capacitance* of approximately 1 pF (this value often ranges up to 5 pF for certain types of resistors).

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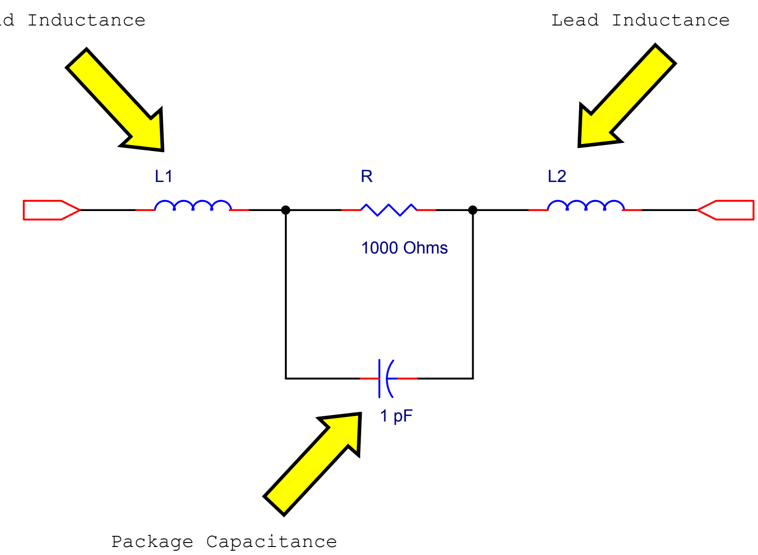

Figure 13-1: Equivalent Circuit of a 100 Ω Resistor

You might be inclined to think that these additional values are too small to have any real effect upon the operation of the resistor; of course, this is correct at *low* frequencies. At even the "low" microwave frequency of 2 GHz, this is no longer true!

Example 13-1

Referring to Figure 13-1, if L_1 and L_2 are 1 nH, and the operating frequency is 2 GHz, calculate the reactance of each component, and calculate the total impedance of the resistor.

Solution
The inductive reactance X_L for L_1 and L_2 can be calculated as:

$$X_{L1} = X_{L2} = 2\pi fL = 2\pi(2GHz)(1nH) = \underline{\underline{12.57\Omega}}$$

Therefore, the lead inductances add about 25 Ω to the circuit impedance!
The capacitive reactance X_C shunting the resistance can be calculated as:

$$X_C = \frac{1}{2\pi fC} = \frac{1}{2\pi(2GHz)1pF} = \underline{\underline{79.5\Omega}}$$

The capacitance therefore reduces the impedance of the resistor, because it is in parallel with it.

The *total impedance* is the phasor total of all the reactances and resistances in the circuit. To calculate this, we have to use appropriate rectangular and polar notation for the individual reactances in the circuit. We get:

$$Z_{total} = jX_{L1} + jX_{L2} + 100\Omega \,||\, (-jX_C) =$$
$$j25.14 + (100\Omega) \,||\, (-j79.5\Omega) = \underline{\underline{38.7 - j23.57\Omega}} = \underline{\underline{45.34\Omega\angle -31.3^\circ}}$$

This is very strange; the resistor hardly looks like a resistor at all! In fact, because there are both L and C components, there will also be at least one frequency where the component looks like a resonant circuit -- it may look very close to a *short* at the resonant frequency.

Example 13-1 demonstrates that there are real problems with using conventional components at microwave frequencies. Conventional capacitors and inductors also have hidden resistances, capacitances, and inductances that prevent their use at microwave frequencies.

So how can we build microwave communications circuits at all? There are three important elements we must consider. First, we must abandon many of the components used at lower frequencies and adopt devices that are built for UHF and SHF application. In general, the use of *surface mount* components greatly reduces the amount of parasitic lead inductance by eliminating the wire leads. Figure 13-2 shows several surface-mount RF semiconductors on a demonstration board.

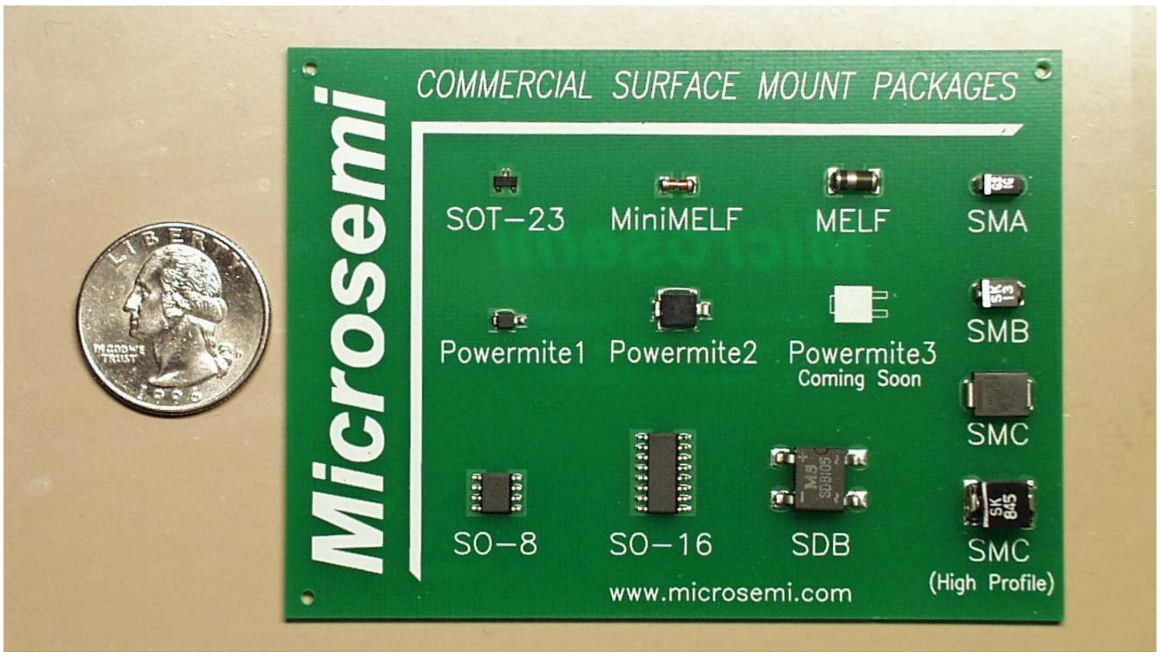

Figure 13-2: Surface Mount RF Semiconductors

Second, we must begin thinking in terms of *distributed* circuit elements, rather than *lumped* elements. *Distributed* circuit elements are contained throughout the physical circuit, rather than being contained within individual locations within it (the *lumped* approach). Finally, because of the high frequencies, wavelengths are very short. Even an inch of wire can be an effective antenna at microwave frequencies. Unless the entire circuit is treated as a *transmission line*, it is likely that unintended radiation and feedback will occur.

Example 13-2

What is the length, in inches, of a 1/4-wave antenna at a frequency of 10.24 GHz?

Solution
The wavelength equation from Chapter 1 can be used:

$$\lambda = \frac{v}{f} = \frac{3\times10^{8}\, m/s}{10.24 GHz} = \underline{0.029m}$$

Only 1/4 of this wavelength is required, so we get:

$$\lambda / 4 = 0.029m / 4 = 0.00732m = \underline{\underline{0.288"}}$$

This is a *very* short wavelength. In conventional circuits, it is not hard to imagine a lead length of a quarter-inch. In a 10.24 GHz circuit, a stray quarter-inch lead becomes a very effective antenna, disrupting the circuit's operation. At microwave frequencies, construction methods are very critical. At these frequencies, the dimensions of all wiring must be accurately calculated and fabricated, or circuits won't work properly.

Distributed Circuit Techniques

By looking at electrical circuits in a new way, we can develop microwave *equivalents* to circuits that we used at lower frequencies. The microwave versions of the circuits depend on the overall layout of the conductors and can be thought of as *distributed* circuits.

Suppose that we wish to build an LC resonant circuit to act as a bandpass filter at a frequency of 10.24 GHz. We *could* try building the circuit from a discrete capacitor-inductor combination (and we can still think in those terms, of course). The circuit might look a little like Figure 13-3(a).

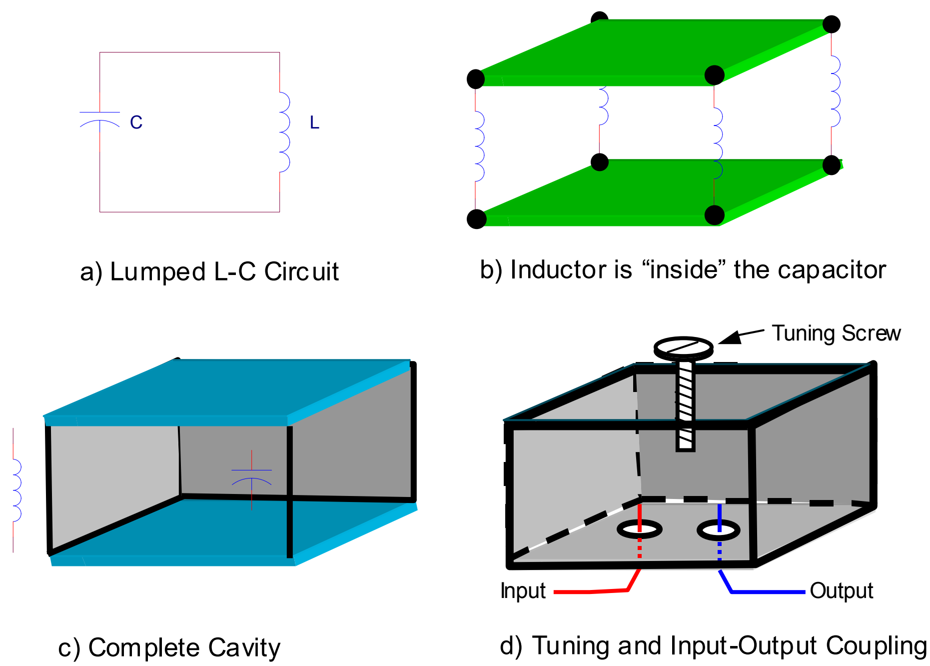

Figure 13-3: Development of a Cavity Resonator

Now we need to reduce the leads to near zero length to get this circuit work at microwave frequencies. Imagine that we can somehow put the inductor right inside the capacitor, in between its plates. In order to reduce the effect of the inductor's own lead lengths, we will *distribute* it evenly in four parts around the capacitor's plates. This is Figure 13-3(b). Figure 13-3(c) is the completed filter, which is implemented as a *cavity resonator*. Where did the inductor go? It is *distributed* within the walls of the cavity. The longer the cavity is built, the more inductance it will have, and the lower its resonant frequency will become.

The *capacitance* is also distributed in the cavity, between the top and bottom lids. With this arrangement, how can we get energy in and out of the filter, since the top and bottom are closed? There are two common methods. First, we can poke coupling wires or *stubs* into the cavity through one end. The coupling stubs (red and blue wires in the figure) act as transmitting and receiving antennas, and the cavity then filters the signal that passes in between them. Second, the cavity can be attached to a printed circuit board assembly, and the conductors on the board can be etched in a manner so that they pass underneath the filter cavity.

Finally, we may need to fine-tune the cavity resonator filter. Tuning can be accomplished by the simple addition of a screw and locknut on top of the unit. As the screw is turned, it will move in or out of the open cavity, which will slightly change the capacitance of the filter, which in turn changes the filter's resonant frequency. The cavity can also be tuned by the addition of a sliding piston, which acts to vary the length of the cavity, and thus changes its resonant frequency.

Microwave Circuits and Transmission Lines

Since the wavelength of microwave frequencies is very short, even a conductor measuring a mere fraction of an inch must be accounted for as a *transmission line* in order for the circuit to work as intended. From Chapter 11, we know that transmission lines can perform two primary circuit functions. First, they can carry RF energy between two points in a circuit with minimum radiation and loss. Second, they can *match* or *transform* one impedance into another in order to obtain maximum power transfer between two stages. Let's take a look at how this can be done.

Figure 13-4 shows a low-noise RF preamplifier for the 10 GHz band employing a special microwave JFET called a GaSFET (gallium-arsenide field effect transistor). The GaSFET offers far superior performance at microwave frequencies than a conventional bipolar or field-effect transistor.

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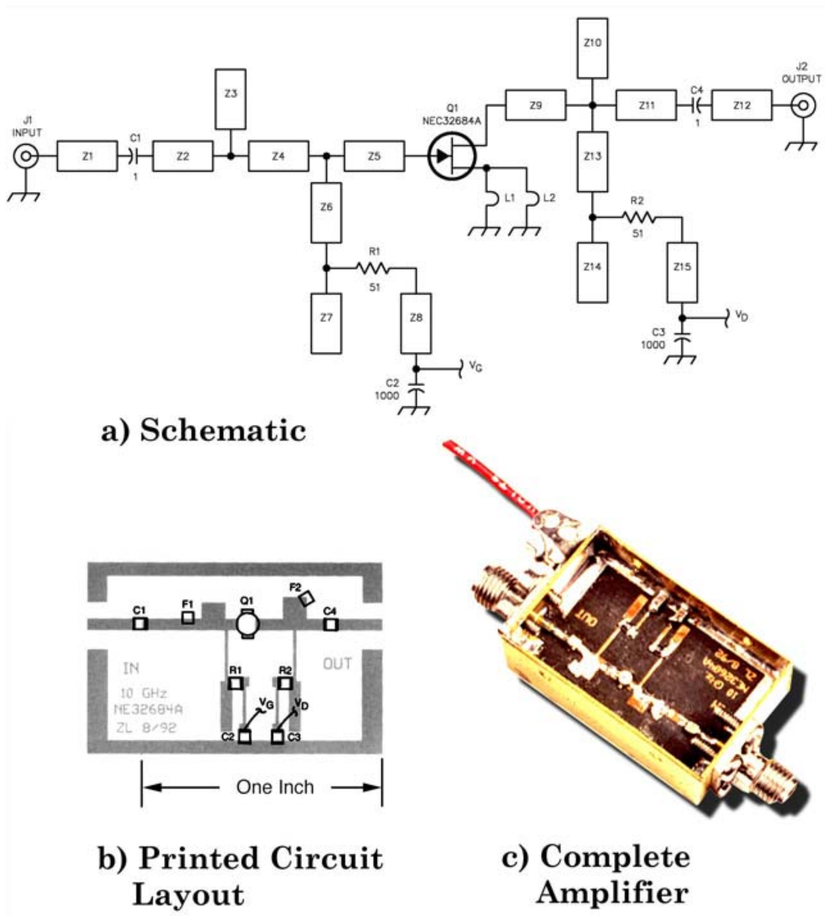

Figure 13-4: A GaSFET Preamplifier for the 10 GHz Band
(a) Schematic (b) Printed Circuit Layout (c) Complete amplifier
(Source: National Association for Amateur Radio. Reprinted with permission.)

The schematic of Figure 13-4(a) looks very strange. You can recognize the transistor and few passive components (such as R_1, C_2, and so on), but the rest of the circuit consists of rectangles labelled "Z_1," "Z_2," and so on. What are these?

The rectangles in this circuit are *transmission line sections*. These transmission line sections are used to impedance match the input and output of the amplifier to 50 ohms, and also provide bandpass filtering. They are fabricated right on the printed circuit board. Transmission lines formed on a PC board are often called *microstrip* lines. Their physical dimensions must be very accurate, and the PC board material must have a known (and stable) dielectric constant. Figure 13-4(b) shows the PC board layout; look carefully at the board, and you can identify each of the "Z" microstrip line elements.

Figure 13-4(c) shows a completed amplifier. Note the extensive use of surface mount components! Should you ever have to repair a unit such as this, remember that the circuit board traces are transmission lines. If they are damaged in any way (such as from overheating during soldering), their characteristic impedances will be altered, and the circuit will no longer function as designed!

Remember that these transmission line sections also depend upon the dielectric constant of the PC board material. Overheating a PC board trace can burn or melt the PC board insulating material, which changes its dielectric constant. Even a small amount of flux left behind after soldering can throw the circuit off frequency. Proper construction and repair techniques therefore include using the minimum amount of heat for soldering, and thorough cleaning (defluxing) of circuits after repairs have been completed.

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Section Checkpoint

13-1 What is the lowest approximate frequency where microwave techniques are used?
13-2 Explain the concept of *lumped* components; give an example.
13-3 How can a resistor also behave as an inductor? What part of the package provides the inductance?
13-4 What three principles must be considered when constructing microwave circuitry?
13-5 What are *distributed* circuit elements?
13-6 Explain how the resonator of Figure 13-2 acts as a bandpass filter.
13-7 What are two functions carried out by transmission lines?
13-8 Describe how the transmission line sections in the amplifier of Figure 13-3 are fabricated.
13-9 What can happen if a microstrip section is overheated during soldering?
13-10 After servicing a microwave PC board assembly, solder flux must be removed. Why?

13-2 Microwave Transmission Lines

At microwave frequencies the conventional transmission lines discussed in Chapter 11 become very inefficient. There are two reasons for this. First, the *wavelength* of microwave signals is very short; and second, most of the dielectric (insulator) materials in transmission lines become lossy at SHF.

The *open* transmission lines (such as ladder line or ribbon cable) will become *antennas* at microwave frequencies because the distance between the two conductors in the line approaches a sizable fraction (more than 1/8th) of a wavelength. The fields set up by the transmission line conductors can no longer cancel, resulting in *radiation loss*. This is very undesirable!

Even coaxial cable performs miserably at microwave frequencies. In a coaxial cable, the dielectric is often made of a plastic such as polyethylene or polystyrene, in solid or foam form. The waves moving down the coax must pass through the dielectric, since they are confined within the shield. As the frequency is increased, the dielectric begins to absorb more of the RF energy and convert it to heat. In addition, the *skin effect* (where most of the RF energy flows on the outer surface of conductors) also increases at SHF, increasing the effective RF resistance of the transmission line conductors.

There is only one transmission line that can efficiently conduct microwave energy over any distance at all, and that is *waveguide*. For short runs (a few inches or less) on printed circuit boards, *microstrip* and *stripline* lines can be used. Figure 13-5 shows many of the types of waveguide that are available.

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Figure 13-5: Several Types of Waveguide
Clockwise, from Upper Left: Straight; H-Bend; Twist; Series Tee

Waveguide Operation

Waveguide is a hollow metal tube that conducts RF energy between two points in a circuit. It can be build in both rectangular and round configurations. Figure 13-6 shows a section of rectangular waveguide.

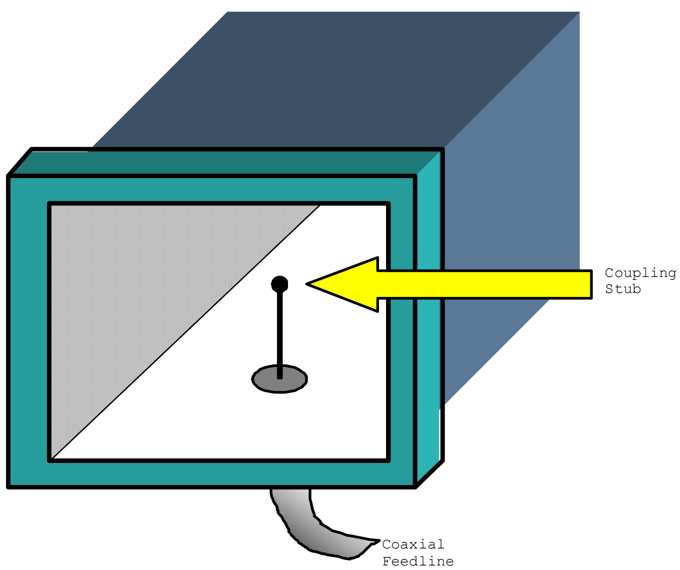

Figure 13-6: A Rectangular Waveguide Section

In Figure 13-6, the RF energy is *coupled* (passed) into the waveguide by means of a quarter-wave *stub* antenna. This is often referred to as *capacitive* coupling. Once the RF energy is inside the waveguide, it travels along inside, *reflecting* off the inside walls as it

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proceeds. Efficient reflection of RF energy requires that the inside surface of the waveguide have a low electrical resistance. Thus, the inside of waveguide is often plated with excellent (and expensive) conductors, such as silver or gold. The plating need not be very thick, because RF currents will only flow along the "skin" of the wall.

Waveguide Cutoff Frequency

You might wonder why waveguide isn't used for *all* RF transmission line needs, since it's much more efficient than conventional types of line. There are two reasons for this. First, waveguide is very expensive when compared to other transmission lines. Second, there is a *minimum* frequency that can be passed by any particular type of waveguide. This frequency is referred to as the *cutoff frequency*.

The cutoff frequency is determined by the longest cross dimension of the waveguide. When the wavelength of the energy being introduced into the waveguide becomes too long, the waveguide "shorts" out the electric and magnetic fields of the wave. This occurs when one-half (or more) of the signal wavelength occupies the waveguide's cross-section. Therefore, to determine the cutoff frequency of a section of waveguide, simply measure the longest cross-dimension, and then find the frequency where that length is one-half of a wavelength. The following equation summarizes this relationship:

$$(13\text{-}1) \quad f_{co} = \frac{v}{2d}$$

Where v is the velocity of radio energy (3×10^8 m/s), and d is the longest cross dimension, in meters. Figure 13-7 shows the measurement of the dimension d in a section of rectangular waveguide.

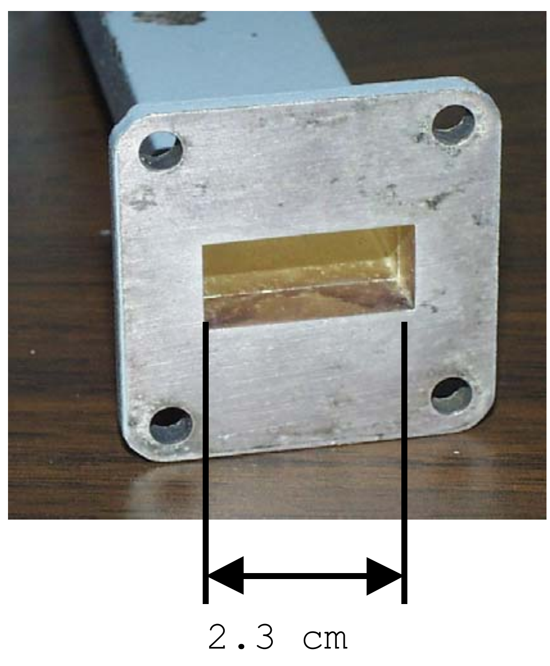

Figure 13-7: Measurement of the Cutoff Frequency in Rectangular Waveguide

Example 13-3

The long dimension of the waveguide measures 2.3 cm in Figure 13-7. Calculate the cutoff frequency f_{co} for the unit.

Solution
Equation 13-1 directly solves this problem. We must not forget to express the value of 2.3 cm as 0.023 m (100 cm = 1 m):

$$f_{co} = \frac{3 \times 10^8\, m/s}{2(.023m)} = \underline{\underline{6.52\ GHz}}$$

Therefore, frequencies below 6.52 GHz cannot pass through this section of waveguide. In effect, the waveguide acts as a *high-pass filter*. The manufacturer will typically rate waveguide in terms of recommended frequency range; these ranges are usually 10% to 20% higher than the cutoff frequency, for the simplest possible propagation mode. We will discuss propagation modes shortly.

Waveguide is not practical for frequencies below the microwave region. The next example demonstrates this point.

Example 13-4

How long must the long dimension of a rectangular waveguide be in order to support a lower cutoff frequency of 30 MHz?

Solution
Equation 13-1 can be manipulated to solve this problem. We must solve for the dimension *d*:

$$f_{co} = \frac{v}{2d}$$

$$d = \frac{v}{2f_{co}} = \frac{3 \times 10^8\, m/s}{2(30\ MHz)} = \underline{\underline{5\ meters}}$$

This is an *enormous* physical size! The waveguide would have to be at least 5 meters (16.4 feet) wide inside to support the propagation of 30 MHz signals. This would obviously not be very practical!

Propagation Modes in Waveguide

There are many ways in which RF energy can move through a waveguide. Each of these is called a *mode* of propagation. There are two primary modes, called *transverse electric* and *transverse magnetic*, abbreviated by the initials TE and TM, respectively. Recall that a radio wave is composed of electric and magnetic fields. These fields are at 90-degree angles with respect to each other. In TE propagation, there is no component of the *electric* (E) field aligned with the direction of propagation; the electric field is perpendicular to the line of movement. In TM propagation, the magnetic (M) field is perpendicular to the line of propagation.

A shorthand notation is used to describe TM and TE modes within a waveguide. Two numbers follow the TE or TM designator. The first indicates the number of half-wave patterns in the *long* cross dimension of the waveguide, and the second indicates the number of half-wave patterns in the *short* cross dimension of the waveguide.

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Dominant Propagation Mode

Figure 13-8 shows TE_{10} propagation, which is often called the *dominant* propagation mode for waveguide. This is the most common, and simplest, mode of energy travel.

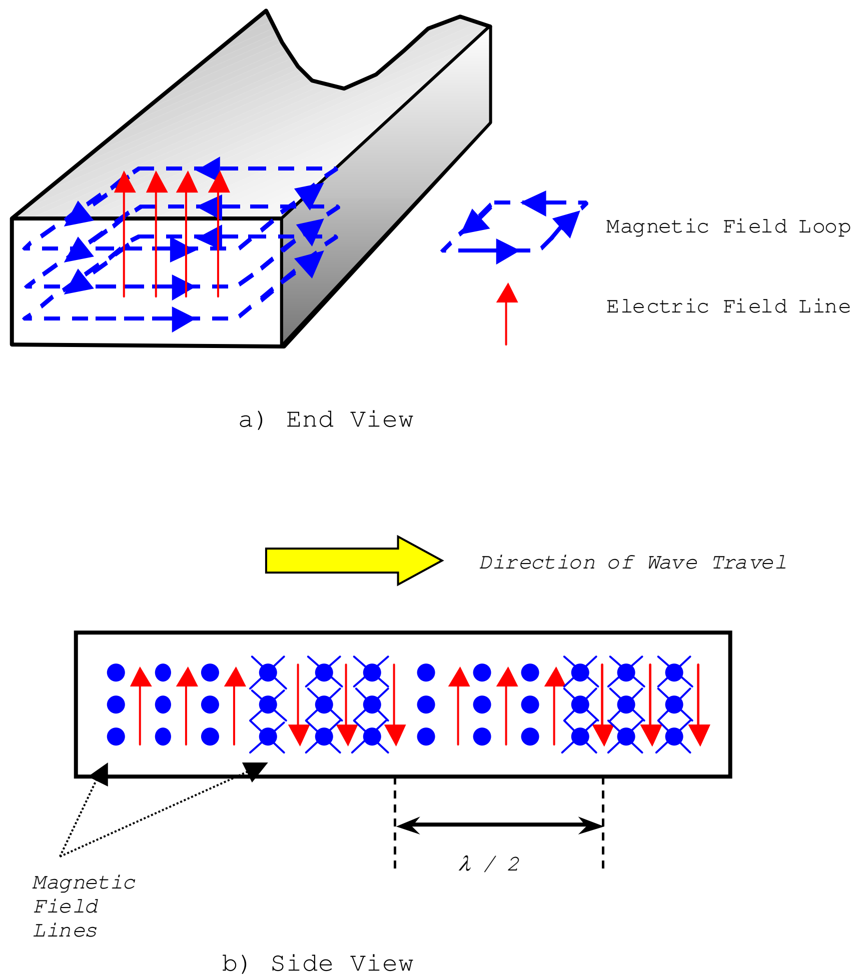

Figure 13-8: TE_{10} Propagation

In the TE_{10} mode, there is one half-wave pattern of the electric (E) field across the long cross dimension of the waveguide, and *no* half-wave E-pattern across the short dimension. The lines of the E-field are perpendicular to the direction of wave movement. The magnetic (H) field lines are perpendicular to the E-field lines. In practical work with waveguides, TE_{10} is the most commonly used mode. The TE_{10} mode also provides the lowest possible *cut-off frequency*, as predicted by Equation 13-1.

Higher Modes

Higher modes of propagation are obtained when the operating frequency is increased sufficiently to allow one or more wavelength to fit within the long cross dimension of the waveguide. The use of higher modes is generally restricted to special applications; it is much simpler to work in the dominant TE_{10} mode. Figure 13-9 is a comparison of the various propagation modes.

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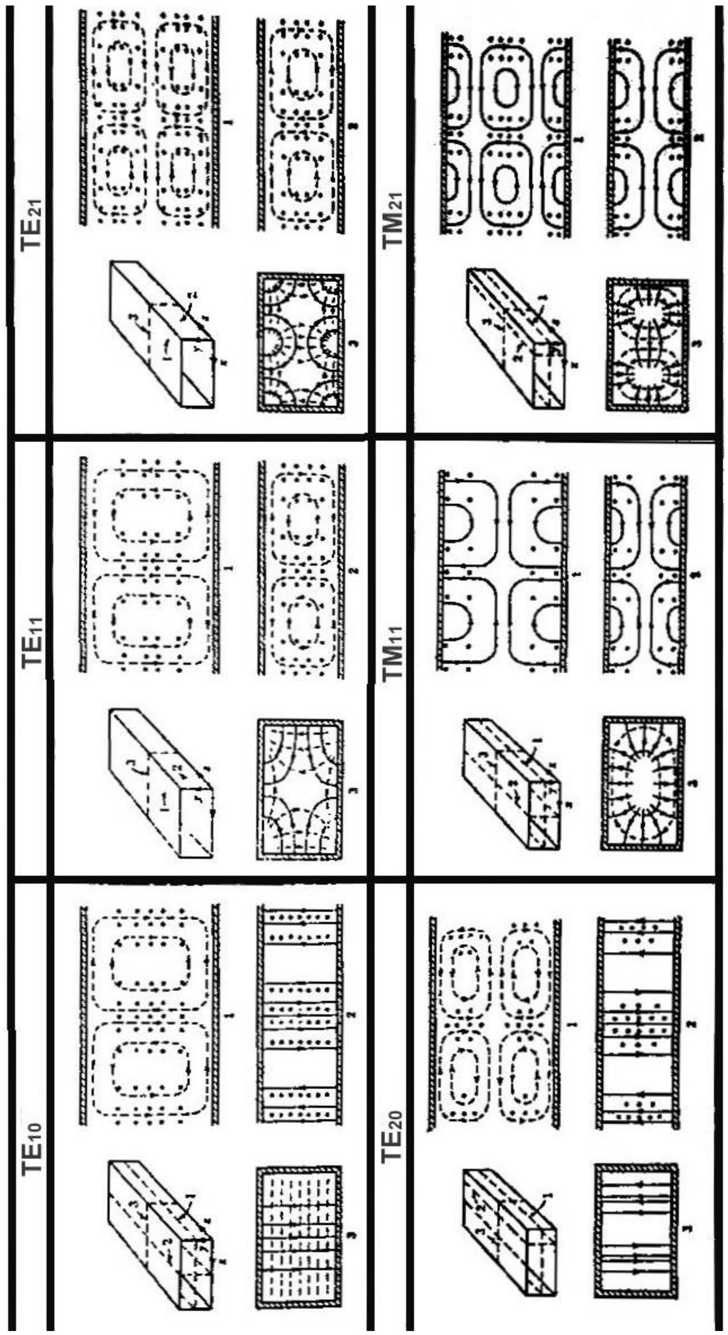

Figure 13-9: Field Configurations in Various Propagation Modes (Source: S. Ramo, Fields and Waves in Modern Radio. Reprinted with permission of John Wiley & Sons, Inc.)

In the TE_{20} mode, there are now two half-wave patterns across the long dimension of the waveguide, but still no patterns across the short-dimension. By increasing the size of the short dimension, it is possible to get TE_{11}, where one half-wave field pattern now appears across *both* dimensions of the waveguide. Compare the pictures of TE_{10} and TE_{11} propagation; the latter is *much* more complicated in terms of where the electric and magnetic field lines will lie. This fact makes coupling (getting energy in and out) much more critical for higher modes of propagation, and is one reason why the dominant or fundamental mode, TE_{10}, is so popular.

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To completely describe higher propagation modes requires the use of Maxwell's equations, which describe the conditions under which electromagnetic energy can exist. These equations are well beyond what a technician needs to know.

Special Waveguide Sections

There are several special purpose waveguide sections that are often used. Among these sections are *bends*, *tees*, *attenuators, terminators* and *directional couplers*.

Bends

The path between a microwave antenna and radio is seldom a straight line. *Bends* are sections of waveguide that can accommodate the necessary turns in an installation while preserving the desired mode of propagation. Figure 13-10 shows several of these.

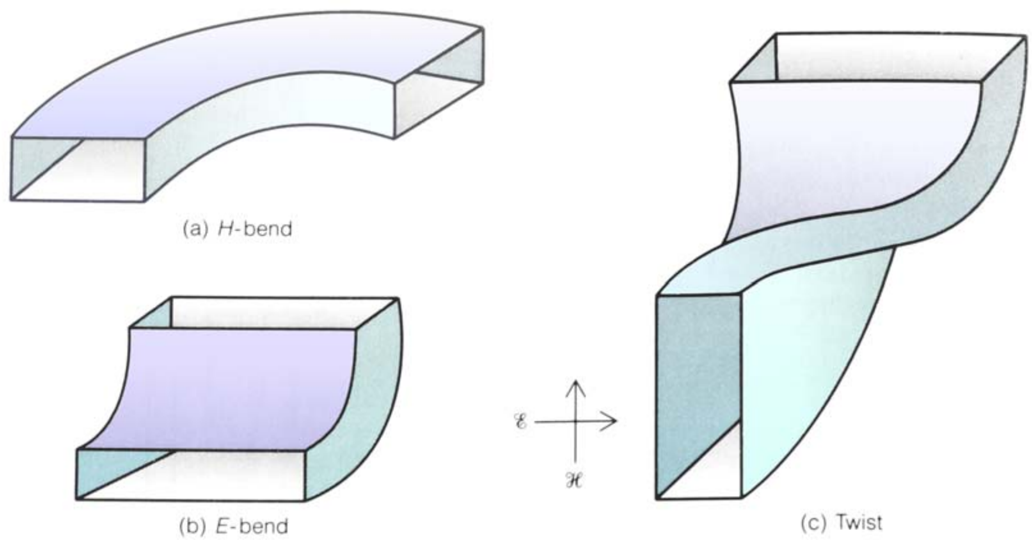

Figure 13-10: Waveguide Bends and Twists

The *H* bend of Figure 13-10(a) is used to turn a 90 degree corner. The propagation is undisturbed, except that the magnetic field is bent 90 degrees as it passed through the section. The polarization (orientation of the electric (E) field) is unaffected. The *E* bend also completes a 90-degree turn, in either an upward or downward direction. The magnetic field is undisturbed, but the *E* field is bent by 90 degrees within. The *twist* of Figure 13-10(c) is used to effect a shift in the polarization of the wave. The electric and magnetic fields maintain the same orientation within the waveguide, but because of the gradual twist, their orientation is different at the opposite ends with respect to the outside world.

The choice of bends and twists is largely determined by mechanical considerations; in other words, an engineer will simply specify a part that fits in a given space! By combining these three basic shapes, nearly any electrical and mechanical requirement can be met.

Tees

Tees are often used to act as power combiners, power splitters, and isolators. A *power combiner* adds the RF power from two sources (which must supply in-phase energy). The *shunt tee* of Figure 13-11(a) can be used as a power combiner.

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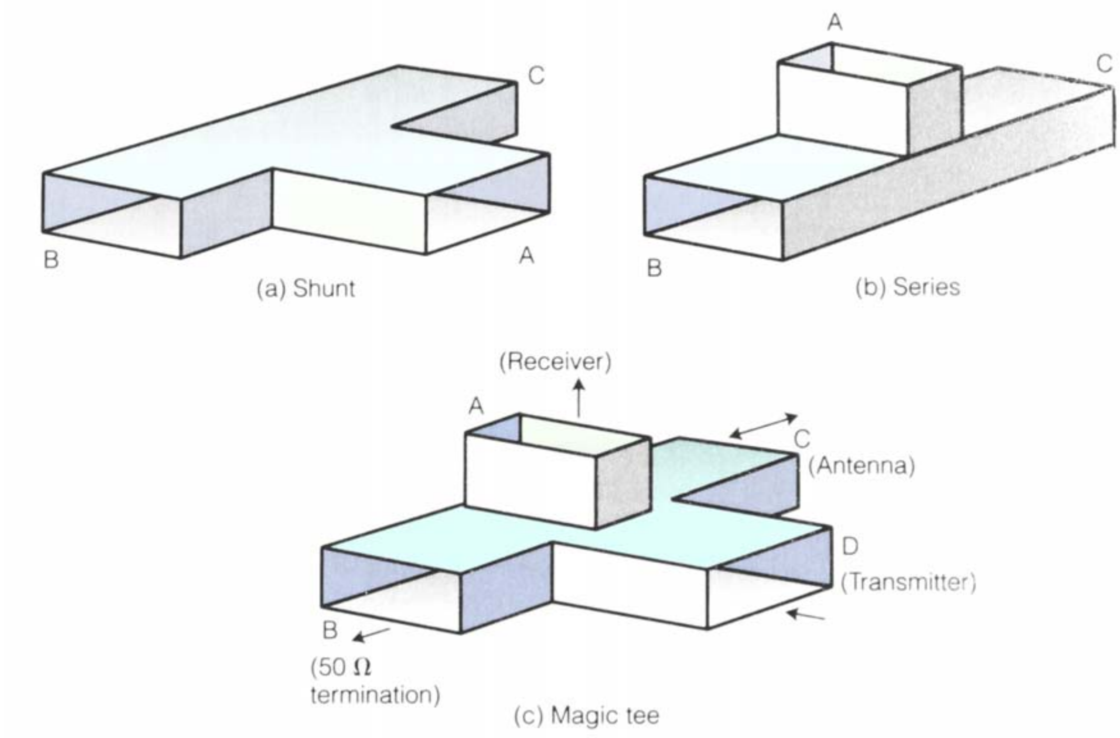

Figure 13-11: Three Tees

If RF energy is applied to port A in Figure 13-11(a), ports B and C will each have an in-phase of the input signal. The power is equally divided among ports B and C. Likewise, if two RF power sources are connected to B and C, the total power output at A will be the sum of the individual input powers. Notice that since there is no disturbance in the electric field orientation, there is no phase shift produced in the shunt tee.

The *series tee* of Figure 13-11(b) operates in a similar manner. If a signal is inserted into port A, it travels downward into the body of the tee. As it arrives at the junction of the body and stub, the electric (E) field must bend around the corner. Electrically, the body appears to be in *series* with the electric field lines in the stub, so each half of the body of the device must have *opposite* polarization. In other words, the series tee acts as a phase-splitter transformer; if a signal is applied at A, signals are produced at B and C that are 180 degrees out of phase. The power of the signals at B and C are still half that of the inserted power, just like that of the shunt tee. Because of the 180 degree phase shift, the series tee is useful for creating transmission line impedance-matching networks at microwave frequencies; a sliding short (piston) is often connected at port C in such an application.

The *magic tee* of Figure 13-11(c) combines the characteristics of both the series and shunt tees. It is often used as a transmit-receive switch in non-critical applications. For example, if a transmitter's output is coupled into port D, the transmitter's energy will be divided equally among ports B and C. Very little of the transmitter's energy will leave port A, because the polarization of stub A does not agree with the fundamental mode (TE_{10}) of the propagation in the body of the device. The sensitive receiver is therefore protected from the high power of the transmitter.

When the transmitter power enters port D, it's equally split amongst ports B and C. Port B must be terminated in a matched load (to be discussed shortly) to prevent reflection, and thus, 3 dB (one half) of the transmitter's power is wasted as heat. Without the matched load at port B, standing waves will form within the body of the tee, some of which will couple excessively into the receiver port (A), possibly damaging the receiver front end.

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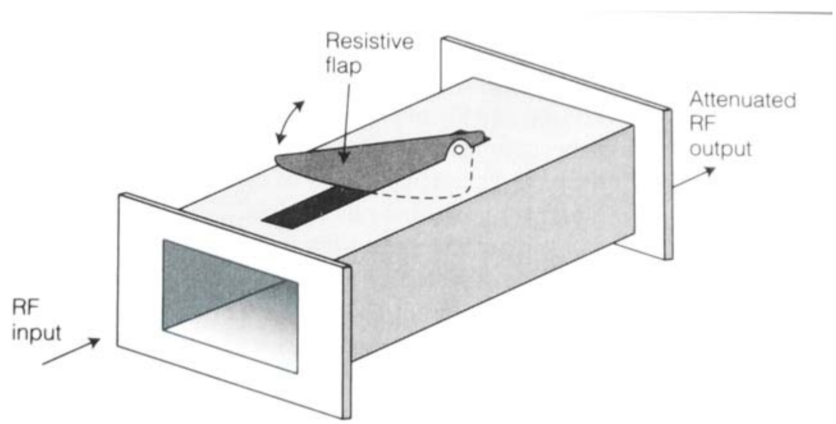
In receive mode, the energy from the antenna enters port C, and it encounters a 90 degree shift (which is harmless) as it bends into port A to make its way to the receiver. The receiver can "see" the antenna's energy because the E-field orientation of ports C and A compatible. Some of the antenna energy also flows into ports B and D, where it is dissipated. Thus, slightly more than 3 dB of attenuation is inserted in the receive signal. Because of the 3 dB transmit and receive losses, this approach to T/R switching isn't the most practical one for high-power or high-sensitivity applications.

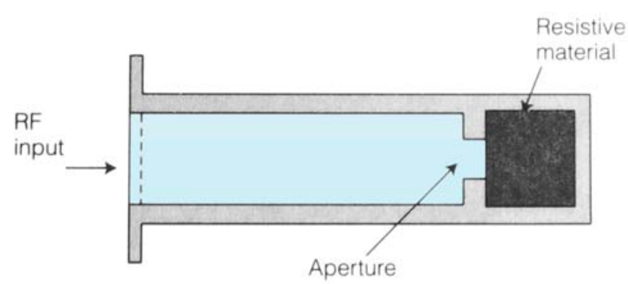
Attenuators and Terminators

An *attenuator* is a device that weakens a signal in a controlled fashion. Often an attenuator is used to precisely control the output power level of an oscillator or other RF power source before combining it in a mixer, bridge, or amplifier. In order to attenuate microwave signals, a resistive material (such as plastic impregnated with carbon) is placed within the waveguide. Passing electric and magnetic field components set up currents within the resistive material, which cause power to be converted to heat (I^2R loss).

A *variable attenuator* can be built by using a flap of resistive material that can be lowered into the waveguide, as shown below in Figure 13-12(a). The amount of attenuation is controlled by how far the flap is inserted into the waveguide; the farther it is inserted, the more power that will be diverted from the output. All commercially made attenuators are rated for a *maximum input power*, and this power level must not be exceeded, or the device will literally go up in smoke (the resistive material in the flap can only handle a limited amount of heat, depending upon its design).



(a) Flap attenuator



(b) Terminator

Figure 13-12: Variable Attenuator and Fixed Termination

A *terminator* is a device connected to the end of a transmission line in order to provide a matched load. Signals enter a terminator and do not leave! (It is also possible to

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provide *mismatched* loads as desired.) A terminator is therefore a dummy antenna for microwave frequencies. There is no place to hook a discrete resistor for terminating a waveguide (as might be done to terminate coaxial line), so a different approach is necessary. The *fixed termination* of Figure 13-12(b) is in fact very similar to an attenuator; signal flows into the device, and moves past the *aperture* (provided to obtain an impedance match), and into the resistive material, where it is converted to heat.

Like an attenuator, a terminator can only handle a specified level of RF power; excessive power will cause overheating, degradation of impedance characteristics (wave reflections will result due to the impedance mismatch), and possibly even fire in severe cases. Unlike conventional electronic components, a technician can't normally eyeball an attenuator or terminator and see this type of damage (unless the device has been hot enough to discolor the metal or cause the paint to bubble) .

Directional Couplers

A *directional coupler* is a device that can be used to measure both the *amount* and *direction* (forward or reverse) of RF energy flow. One type is shown in Figure 13-13.

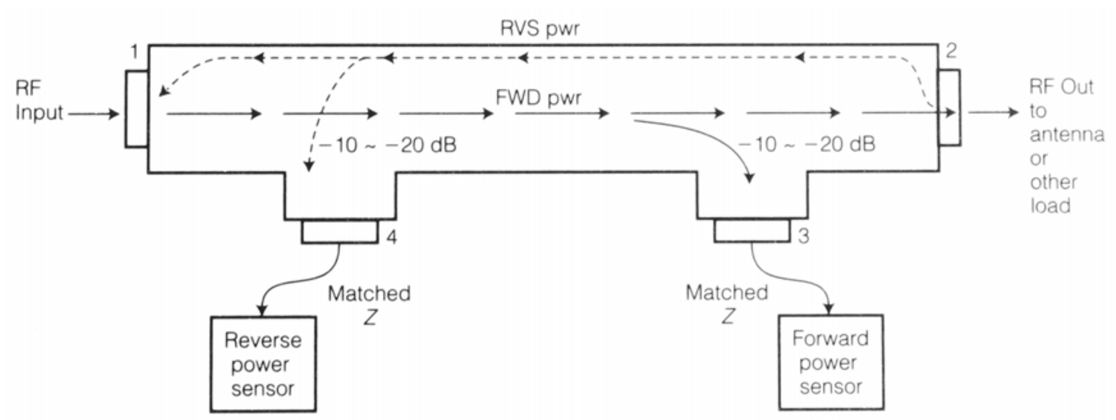

Figure 13-13: A Directional Coupler

The coupler is designed to allow RF energy to move very easily between the *input and output ports*, labeled 1 and 2. For example, if 1 watt of RF is placed into port 1, then very nearly 1 watt will be output at port 2. The *insertion loss* of a directional coupler is the measure of how much RF power is lost between ports 1 and 2, and it's measured in dB.

Ports 3 and 4, which must both be terminated in a matched load, are called the *sampling ports*. The power output at port 3 is related to the *forward* power (flowing from port 1 to 2 by definition). Since most of the RF energy flows from 1 to 2, only a small fraction is available at port 3; the fraction is related by the *forward coupling factor*, which is given in dB. Coupling factors from 10 dB to 20 dB (10:1 power ratio, 100:1 power ratio) are very common. The power output at port 4 is related to the *reflected power* (flowing "backward" from port 2 to 1 by definition.) In a system with a perfectly matched load at port 2, there is no reflection and the sampled output at port 4 is zero. In reality, the *directivity* of the sampling ports isn't perfect, so port 4 will read a small amount of RF energy, even with a matched load.

Directional couplers are commonly used for automatic power control (keeping a transmitter's output power constant by using feedback, which can compensate for gain variations due to component aging), protection from high SWR (the transmitter can automatically turned off if a load fault occurs), or manual measurement of the SWR along a waveguide section. The forward sampling port can also be used for direct measurement of transmitted signal quality using a spectrum analyzer (and inline attenuator, if needed).

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