11 Microwave measurements

11.1 INTRODUCTION

The quantities which are measured at microwave frequencies are essentially the same as those measured at lower frequencies. But, because the wavelengths of the signals are comparable with the dimensions of the equipment, it is not possible to use the same techniques. In this chapter we shall examine how frequency, signal level, impedance, attenuation and other quantities are measured at microwave frequencies.

All measurements contain sources of error and it is important in any particular case to know what these are and to have an estimate of the magnitude. Strictly speaking no measurement is of any value unless an estimate of its accuracy can be given. We shall therefore pay attention to the errors which occur in microwave measuring systems in the discussion which follows.

11.2 MEASUREMENT OF FREQUENCY

One simple way of measuring frequency is to measure the wavelength of a standing wave on an airspaced coaxial line. This wavelength is half the free-space wavelength so the frequency can be calculated. Figure 11.1 shows the arrangement of a slotted coaxial line. The strength of the electric field on the line is sampled by a wire probe which protrudes a short way into the space between the conductors. The signal picked up is passed via a detector diode and a sensitive amplifier to a meter. The probe must draw some current in order for a measurement to be possible. This, therefore, limits the accuracy of the measurement. As the probe is moved along the line maxima and minima of the standing wave are detected. For a perfect standing wave the minima are zeroes and their positions can therefore be determined with considerable accuracy. In practice there is some uncertainty about the position of a minimum beause the signal detected falls below the noise level of the detection system. This error can be reduced by measuring the positions of as many minima as possible. In that way several

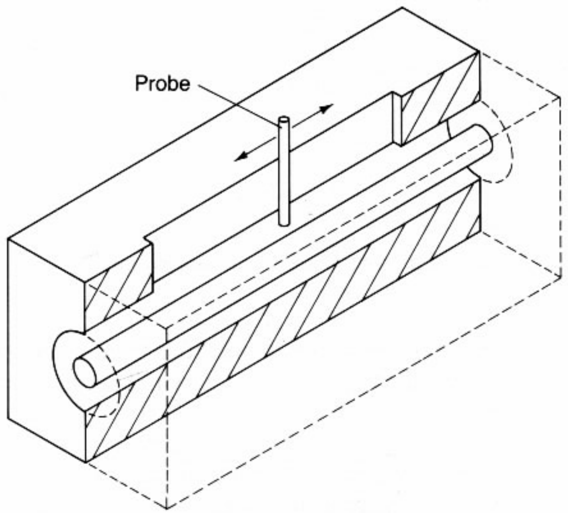

Fig. 11.1 Sectioned view of a slotted section of coaxial line.

different values for the wavelength can be obtained and the average taken to reduce the standard deviation. (Topping, 1962). This approach has the advantage of directness but the accuracy which can be obtained is low (perhaps 0.1% at best) and the measurements are time consuming.

For most microwave laboratory measurements it is much better to have a direct reading of frequency. Originally this was achieved by using a calibrated resonant cavity. By careful design of the cavity the Q factor could be kept high to give a sharp response. The tuning mechanism could also be made to give a direct reading of frequency. Cavity resonance wavemeters, as these devices are called, are still sometimes encountered but they have been superceded by microwave frequency counters. The accuracy of a cavity resonance wavemeter is typically 0.1%.

Microwave frequencies are too high for it to be possible to use the direct counting technique employed at lower frequencies. The way around this is to mix the signal to be measured with that from a crystal controlled local oscillator as shown in Fig. 11.2. If the local oscillator waveform is rich in harmonics then the output from the mixer will be a set of frequencies given by

$$f_i = f_x - nf_1, \tag{11.1}$$

where n is the order of the harmonic and it is assumed that $f_x > nf_i$. The mixer output is fed through a bandpass filter which selects just one frequency out of the set generated. This frequency can be chosen to be low enough for it to be measured with a conventional counter. Since f_1 is known it is possible to compute the source frequency f_x if n can be determined. To

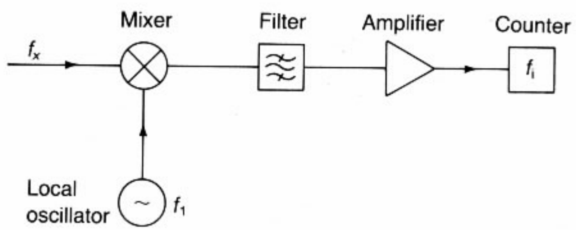

Fig. 11.2 Block diagram of a microwave frequency counter.

do this a second measurement is taken with the local oscillator frequency reduced to f_2 where the offset $(f_1 - f_2)$ is known. The unknown frequency is then given by

$$f_x = nf_1 + f_{i1} \tag{11.2}$$

and

$$f_x = nf_2 + f_{i2} \tag{11.3}$$

where it is assumed that the frequency offset is small enough so that the same harmonic is responsible for the output measured. Eliminating f_x from these two equations gives

$$n = \frac{f_{i2} - f_{i1}}{f_1 - f_2} \tag{11.4}$$

so that n can be computed. The unknown frequency can then be found by substitution back into (11.2).

In practice it is necessary for the method to be a little more complicated to take account of the possibility that one or both of the harmonic frequencies may lie above the unknown. It is also necessary to take steps to ensure that the measurement is accurate even if the incoming signal is frequency modulated.

11.3 MEASUREMENT OF POWER

When a simple detection of a microwave signal is required it is usual to employ a semiconductor diode of the kind shown in Fig. 9.14. At frequencies above 1 GHz it becomes difficult to match the diode satisfactorily because its impedance varies with power level and alternative techniques based on converting the microwave power into heat are used.

At low power levels (a few milliwatts) the detecting element is either a thermistor or a bolometer. A thermistor is manufactured from a mixture of semiconducting oxides and has a negative temperature coefficient of resistance. A bolometer is a thin film resistor deposited on an insulating substrate. Bolometers have response times of less than a millisecond but are very easily damaged by being exposed to too much power.

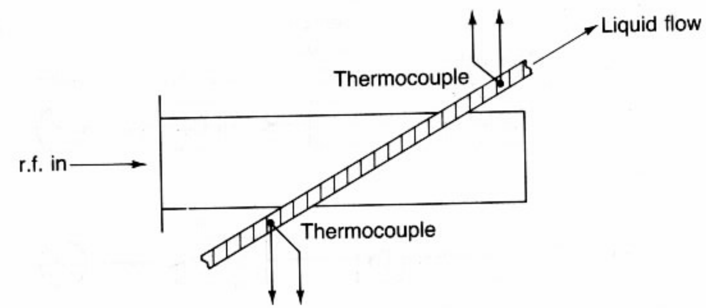

Fig. 11.3 A liquid-flow calorimeter for measuring microwave power.

Thermistors are more rugged but have response times of up to a second. In either case the resistance of the sensing element varies with ambient temperature as well as with the microwave power absorbed. A power meter head therefore normally incorporates two matched thermistors or bolometers which are connected to two arms of a Wheatstone bridge. Only one of the devices is exposed to microwave power. The result is that the balance of the bridge is unaffected by changes in ambient temperature. The bridge is balanced automatically and the output displayed directly in milliwatts on a meter.

At higher power levels (a few watts) the power meter head must be protected from the full power by a calibrated attenuator which is capable of dissipating the full power. An alternative technique is to use a directional coupler to sample the power.

Direct measurement of high power levels is carried out by using a continuous-flow calorimeter as shown in Fig. 11.3. The input power (normally in a waveguide) is absorbed by liquid flowing in a dielectric tube. The tube crosses the guide at an oblique angle to ensure a good match. Very often water flowing in a glass tube is used. The temperature rise in the liquid is measured by a pair of thermocouples. The device is calibrated for a particular flow rate and the flow rate carefully controlled. Alternatively an electric heating element is used as a calibrating heat source.

11.4 MEASUREMENT OF GAIN AND LOSS

In many microwave systems it is necessary to know the gain or loss of each component in order to compute the system performance. These quantities are commonly measured by comparison with standard attenuators. The two possible configurations shown in Fig. 11.4 are r.f. and d.c. substitution. In both cases there is a signal source, a standard attenuator, a detector and some kind of signal level indicator. In r.f. substitution (Fig. 11.4(a)) the attenuator would be a rotary-vane attenuator in a waveguide or a switched

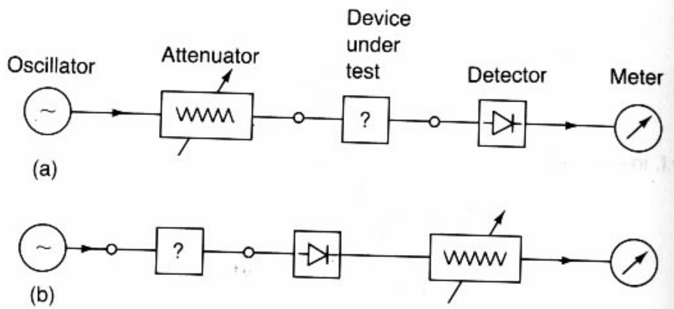

Fig. 11.4 Block diagrams of systems for measuring microwave attenuation: (a) r.f. substitution, and (b) d.c. substitution.

attenuator in a coaxial line. In d.c. substitution (Fig. 11.4(b)) the attenuator could be made in the form of a switched network of precision resistors. The indicator could be a meter or an oscilloscope. The procedure in either case is to set the signal level to a convenient value with the device under test (DUT) in position. It is then removed and the attenuator adjusted to bring the signal back to the same level. This method avoids errors caused by non-linearity in the detector.

In general the gain or loss measured is made up of two components namely that caused by the gain or attenuation inherent in the device and that caused by reflection at mismatches. As the device under test can never be perfectly matched some of the input signal is reflected back towards the source at both the input and the output terminals. Unless the source is very well matched to the connecting transmission lines there will be multiple reflections of the signal producing errors which vary with frequency. A common practice is to put a 10 dB attenuator (a 'pad') between the source and the system to reduce the possibility of multiple reflections as far as possible.

Frequently the measurement is to be made over a band of frequencies. The signal source would then be a sweep oscillator set to sweep repeatedly over the band required and the output could be fed to an $x-y$ plotter to provide a permanent record. A simple r.f. substitution system might use a power meter or a VSWR meter as a detector as shown in Fig. 11.5. Because the output of the oscillator and the sensitivity of the detector vary with frequency it is necessary to produce a set of calibration lines with the attenuator. The performance of the device under test can then be deduced by interpolation between them as shown in Fig. 11.5. Commonly the oscillator is levelled by an external or internal feedback loop to reduce the variation of its output power with frequency.

Better plots of the gain or loss against frequency can be produced if a scalar network analyser system is used. Figure 11.6 shows the general

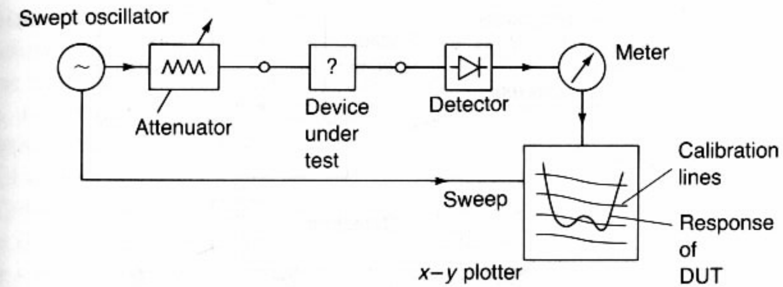

Fig. 11.5 Block diagram of equipment for making swept-frequency insertion-loss measurements.

arrangement. The signal from the sweep oscillator is sampled by high-directivity directional couplers before and after passing through the device under test. The signals in the coupler side-arms are detected and passed to the scalar analyser which is able to display the two signal levels and their ratio in dB against frequency. The output from the scalar analyser can be fed to an $x-y$ plotter to provide a permanent record of the performance of the device under test. The signal-to-noise ratio of the system is enhanced by square-wave modulation of the signal and the use of a tuned amplifier in the scalar analyser.

This arrangement removes errors produced by variations in the output of the oscillator by taking the ratio of the signal levels. It is still liable to errors from a number of sources including the finite directivity of the couplers and any differences in the frequency responses of the couplers and the detectors. Systematic errors which are independent of frequency can be eliminated by removing the device under test and setting the zero level on the analyser. Some systems incorporate a storage normalizer which is able to store the characteristics of the system in the absence of the device under test and correct for them when the result of the measurement is displayed. It is tempting to regard the results produced by such a system as being free from errors though this can never be the case. If the device under test has a high reflection coefficient, for example, then the measurements will be appreciably affected by multiple reflections between it and the source (which can never be a perfect match).

Example

In the system shown in Fig. 11.6 the reflection coefficients of the device under test and the source are 0.3 and 0.05 respectively and the couplers have 20 dB coupling and 40 dB directivity. Investigate the possible errors in the measurement of the insertion loss if all other components can regarded as perfect.

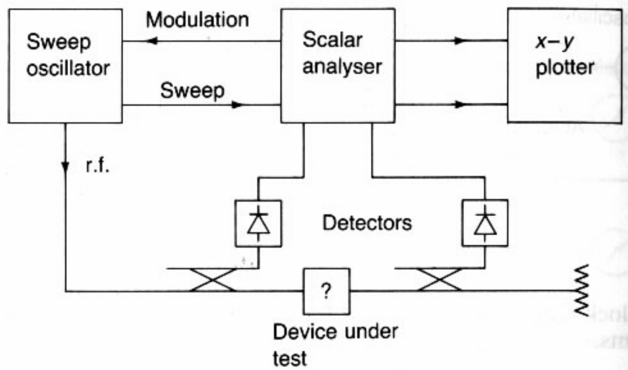

Fig. 11.6 Block diagram of a scalar analyser system for making insertion-loss measurements.

Solution

Multiple reflection of the incident signal between the device under test and the source is illustrated in Fig. 11.7(a). The incident signal level at the device under test is therefore

$$V_1 = \frac{V_0}{1 \pm \varrho_0\varrho_1}, \tag{11.5}$$

where the positive and negative signs represent the extreme cases of the phase of the reflected signal relative to V_1. V_1 can therefore vary by ± 0.015 (0.13 dB) about its nominal level.

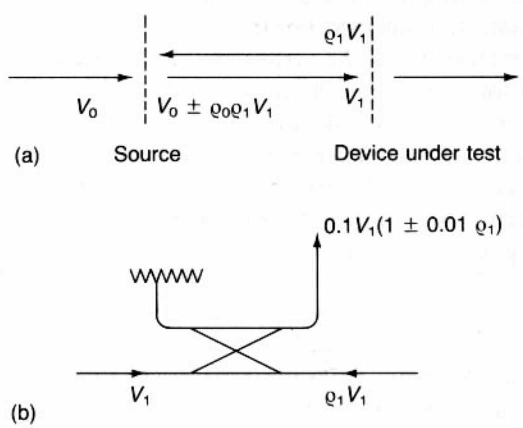

Fig. 11.7 Sources of error in microwave measurements: (a) multiple reflections between the source and the device under test, and (b) finite directivity of directional couplers.

Since the couplers have 40 dB directivity the signal level in the side arm produced by the backward wave is 0.01 of that of a forward wave of the same amplitude. Thus the signal detected is $0.1(1 \pm 0.003)V_1$, as shown in Fig. 11.7(b), giving a possible variation of ± 0.026 dB depending upon the relative phases of the signals. The variation is likely to be below the noise level of the equipment.

Conservation of power at the input port of the device under test requires that the actual signal input is $(1 - 0.09)V_1$. If we suppose that the insertion loss (including any effects of internal reflections) is nominally 20 dB then the signal detected by the second detector is $0.01(1 - 0.09)V_1$. The scalar analyser compares the signal levels at the two detectors. The result will be in error to the extent that the analyser does not form an exact ratio of the signals.

In this example the errors calculated would be likely to lie below the noise level of the system so that they would not be detectable. In other cases this might not be so and it is necessary to be aware of the ways in which the result of a measurement might be in error. For more complicated microwave systems the method of signal flow graphs (Seely, 1972) is used to analyse the errors.

11.5 MEASUREMENT OF RETURN LOSS

A simple modification to the system shown in Fig. 11.6 allows us to measure the return loss of a component directly (see Fig. 11.8). This time the two directional couplers are set to measure the incident and reflected power in the transmission line connected to the input port of the device under test. The arrangement, known as a reflectometer, is widely used for the adjustment of the matches of devices during manufacture. The errors involved in this measuring system can be estimated in the way illustrated in the example above. Notice that errors caused by the directivity of the coupler detecting

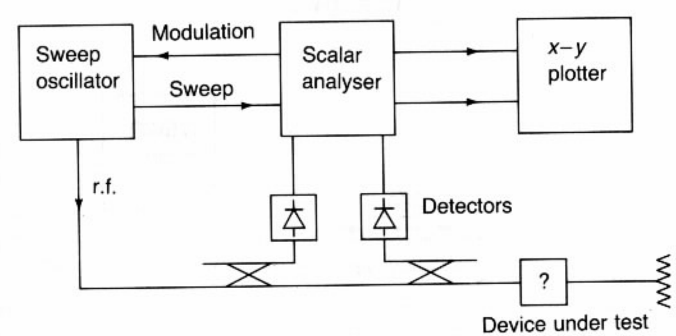

Fig. 11.8 Block diagram of a reflectometer for measuring reflection coefficients.

the reflected power increase as the match of the device under test is improved. Note also that if the transmission loss of the device is caused by internal reflections rather than by dissipation of power then the result will be affected by match of the matched load used to terminate the output line.

11.6 MEASUREMENT OF IMPEDANCE

Although a reflectometer can be a very useful tool for checking and adjusting the match of a device it suffers from the disadvantage that it cannot provide any information about the phase of the reflection. If the requirement is to match a completely unknown impedance then phase information is often necessary.

A basic technique which is still useful though rather tedious is to use a slotted line. These exist for both coaxial line (see Fig. 11.1) and waveguide. The method depends upon the measurement of the positions of the standing-wave mimima and the voltage standing-wave ratio (VSWR) produced by the unknown.

The equipment used for slotted line measurements is illustrated in Fig. 11.9. The signal picked up by the probe on the slotted line is passed from the detector diode to a special instrument known as a VSWR meter. The oscillator is square-wave modulated (normally at 1 kHz.) and the VSWR meter incorporates a tuned amplifier which rejects all signals outside a narrow band centred on the modulation frequency. The purpose of this is to give the best possible signal-to-noise ratio. The VSWR meter also contains both step and infinitely variable attenuators which are used to set the level displayed on the meter and to change from range to range. The meter is calibrated directly in VSWR and usually has scales displaying decibels as well. The way in which measurements are made is described below.

If the reflection coefficient of the unknown is ϱ then the reflected wave amplitude is

$$V_r = \varrho V_i. \tag{11.6}$$

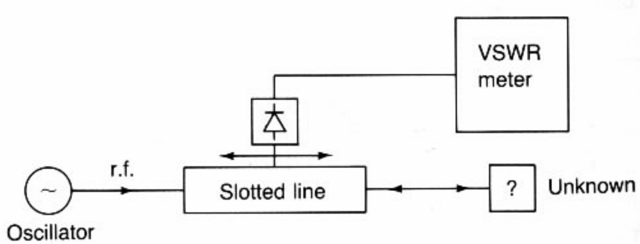

Fig. 11.9 Block diagram of equipment for making slotted line measurements of impedance.

The maximum of the standing wave has amplitude

$$V_{\max} = V_i(1 + |\varrho|) \tag{11.7}$$

and the minimum

$$V_{\min} = V_i(1 - |\varrho|) \tag{11.8}$$

so that the voltage standing-wave ratio is

$$S = \frac{1 + |\varrho|}{1 - |\varrho|}. \tag{11.9}$$

The VSWR is measured by moving the probe to a signal maximum and setting the meter to full-scale deflection (marked '1' on the VSWR scale) using the attenuators. The probe is then moved to a signal minimum and the VSWR read directly from the meter. Greater accuracy can be obtained by taking the average of several measurements.

By rearranging (11.9) we find that the magnitude of the reflection coefficient is given by

$$|\varrho| = \frac{S - 1}{S + 1}. \tag{11.10}$$

The equipment can therefore be used to make measurements of reflection coefficient as an alternative to a reflectometer. It has the disadvantage that the measurements must be made at spot frequencies.

The positions of the minima can be used to determine the magnitude and phase angle of the impedance at a reference plane. At a standing-wave minimum the current is given by

$$I_{\min} = \frac{V_i}{Z_0}(1 + |\varrho|) \tag{11.11}$$

so that the apparent impedance of the unknown at that plane is

$$\begin{aligned} Z'_L &= V_{\min}/I_{\min} \\ &= Z_0\left(\frac{1 - |\varrho|}{1 + |\varrho|}\right) \\ &= Z_0/S. \end{aligned} \tag{11.12}$$

Thus, at a standing wave minimum, the apparent impedance is real.

If the unknown is replaced by a short circuit as shown in Fig. 11.10 a new set of minima can be detected which are spaced at half wavelength intervals from the short circuit. Suppose that there is a minimum at P in the presence of the unknown and one at P′ with the short circuit. Then, because the apparent impedance of the unknown is known at P, it can be calculated at P′ using the usual formula for transformation of impedance on a transmission line (Carter, 1986)

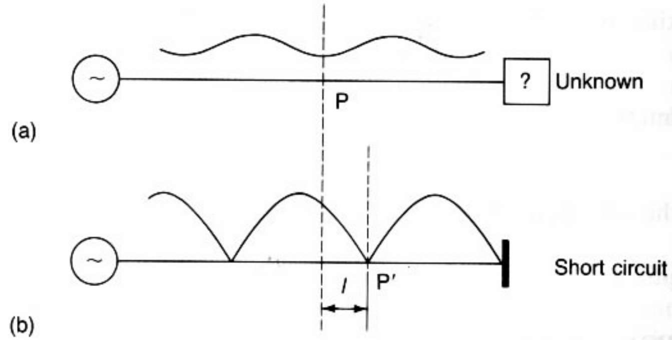

Fig. 11.10 Measurement of impedance using a slotted line: (a) standing wave when the line is terminated by the unknown, and (b) standing-wave pattern with a short-circuit termination.

$$\frac{Z_{P'}}{Z_0} = \frac{Z_P + jZ_0 \tan kl}{jZ_P \tan kl + Z_0}, \tag{11.13}$$

where l is positive if P′ lies closer to the generator than P. The transformation of impedance can be carried out graphically using a Smith chart (see Appendix A).

Although, in theory, the maxima of the standing wave could be used for these measurements the minima are always used because their positions can be found more accurately.

Another way of measuring the phase angle of a reflected signal is to use a phase bridge. The arrangement of the bridge is shown in Fig. 11.11. The reflected signal is sampled by a directional coupler and passed through a calibrated variable phase shifter. The incident signal is sampled likewise

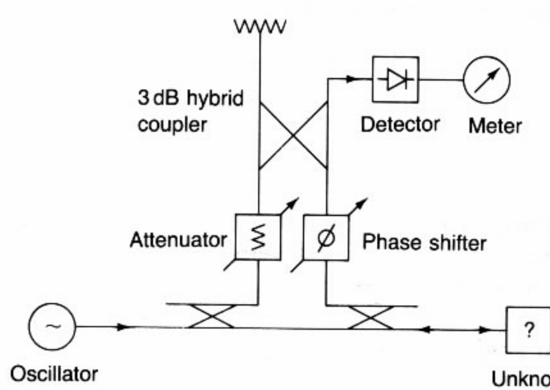

Fig. 11.11 Block diagram of a phase bridge for measuring the phase angle of the reflection coefficient of the unknown.

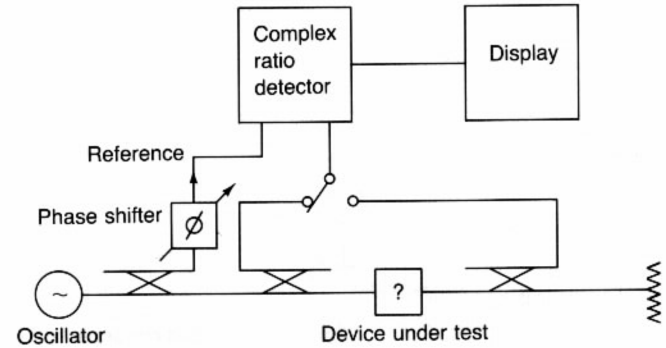

Fig. 11.12 Block diagram of a vector network analyser for measuring complex transmission and reflection coefficients.

and passed through a variable attenuator. These two signals are then combined using a 3 dB hybrid coupler so that the phasor sum is detected. The attenuator is adjusted so that the amplitudes of the signals are the same. Then the phase shifter is adjusted to produce a signal null at the detector. If the unknown is replaced by a short circuit the measurement can be repeated to give a phase reference.

The phase bridge just described is an r.f. substitution method. It has largely been supplanted by an instrument in which the reference and reflected signals are converted to an intermediate frequency at which the phase comparison is made. This system, known as a vector network analyser, is shown in Fig. 11.12. The signals reflected from or transmitted by the unknown are compared with a reference signal by a complex ratio detector. The phase of the reference signal is adjusted by a phase shifter. A modern vector analyser usually contains an accurate signal source and a computer which is able to carry out error correction and calibration as well as displaying the results of the measurements in a variety of forms. For a more detailed discussion of the vector network analyser see Bryant (1988).

11.7 TIME-DOMAIN REFLECTOMETRY

In a complex microwave system it is sometimes easier to measure the mismatch at a port than to say exactly what part of the system is responsible for it. Time-domain reflectometry (TDR) provides a complementary technique to the frequency-domain reflectometry described in Section 11.5. The equipment used is illustrated in Fig. 11.13.

The step generator applies a step function to the system under test and the voltage on the input line is detected with a high-impedance probe. If the system is perfectly matched then the probe voltage is just half the generator voltage. The presence of mismatches within the system causes pulses of

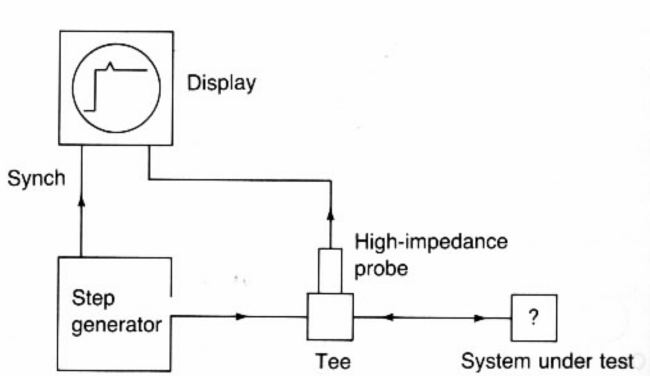

Fig. 11.13 Block diagram of equipment for making measurements by time-domain reflectometry.

returned power which add to, or subtract from, the input pulse depending upon the sign of the reflection (Carter, 1988, p. 113 ff.). The time delay of the returned pulse gives a measure of the position of the mismatch within the system. This technique is used at lower frequencies to find faults in telephone cables. If the rise time of the pulse is very fast (of the order of a picosecond) then it is possible to resolve discontinuities which are a few millimetres apart.

The method just described can only be used with systems which use two conductor lines so that they can carry signals down to d.c. For systems of limited bandwidth such as waveguides it is necessary to use pulsed microwave signals instead (Gardiol, 1984).

The time-domain and frequency-domain responses of a circuit are related to one another (Dunlop and Smith, 1984, Ch. 1). Suppose that the reflection coefficient measured at the circuit input as a function of frequency is $\varrho(\omega)$ and that the input signal is $V_i(t)$. The input signal can be expressed as the superposition of sinusoidal waves whose amplitude varies with frequency as

$$G(\omega) = \frac{1}{2\pi}\int_{-\infty}^{\infty} V_i(t)e^{-j\omega t}\, dt. \qquad (11.14)$$

The amplitude of the reflected signal at any frequency is obtained by multiplying $G(\omega)$ by the reflection coefficient. Finally the system response in the time domain is obtained by Fourier synthesis as

$$f(t) = \int_{-\infty}^{\infty} G(\omega)\varrho(\omega)e^{j\omega t}\, d\omega. \qquad (11.15)$$

The relationship between the time- and frequency-domain descriptions of the reflection can be used to compute the one from the other. Some vector network analysers can compute the TDR response from a swept frequency

measurement using the fast Fourier transform (FFT) method (Dunlop and Smith, 1984, p. 21).

11.8 SPECTRUM ANALYSER MEASUREMENTS

When measurements are to be made on non-linear or active microwave systems the waveforms may not be sinusoidal. This may be because they are amplitude or frequency modulated, or contain harmonics or spurious frequencies caused by parasitic oscillations. In all these cases it is useful to be able to take the actual waveform in the time domain and analyse it into its frequency components. An instrument which performs this function is called a spectrum analyser.

In principle spectrum analysis could be carried out using a tunable narrow-band filter. The output from the filter would then be proportional to the harmonic amplitude at each frequency. In practice it is difficult to make such a filter except for small frequency ranges. An alternative is to mix the signal to be analysed with that of a swept-frequency local oscillator as shown in Fig. 11.14. The result is the production of sum and difference frequencies of which the latter is selected by a narrow-band i.f. amplifier and passed through an envelope detector to the display. There can be problems with this arrangement if the swept frequency has appreciable harmonic content. The output from the mixer would then be

$$f_i = f \pm nf_0 \qquad (11.16)$$

so that the same f_i could be generated by more than one input frequency depending upon the value of n. A possible solution to this problem is to pass the input signal through a swept filter before it enters the mixer. This filter can be quite broad band because it only needs to eliminate those frequencies which are far enough from the centre frequency to produce spurious output. Thus if the analyser is designed to work with the nth harmonic of the swept oscillator the nearest frequencies which can give spurious output are

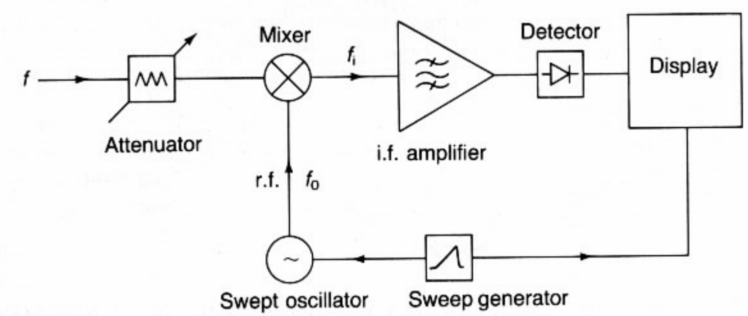

Fig. 11.14 Block diagram of a spectrum analyser.

$$f = f_i + (n + 1)f_0 \tag{11.17}$$

and

$$f = f_i + (n - 1)f_0 \tag{11.18}$$

so that the pre-filter bandwidth must be less than $2f_0$.

11.9 ELECTROMAGNETIC COMPATIBILITY MEASUREMENTS

Electromagnetic compatibility measurements fall into two main classes: the susceptibility of equipment to external electromagnetic fields and the emission of radiation from equipment.

Measurements of immunity to external fields are made by placing the equipment under test (EUT) in a known radiation environment. Field strengths as high as 200 V m^{-1} are required by military test specifications. At low frequencies this can be achieved by placing the EUT between the conductors of a specially constructed section of TEM transmission line. Figure 11.15 shows the arrangement of one such TEM cell. A section of 50 Ω parallel-plate line is connected by tapers to coaxial lines. This arrangement is satisfactory provided that the dimensions of the EUT do not exceed 30% of those of the cell and that the width of the line is less than half a wavelength. The method is used at frequencies up to 500 MHz.

At microwave frequencies the test field is provided by antennas of known gain radiating within anechoic chambers. There is, however, the difficulty that the field is perturbed by the presence of the EUT since the latter

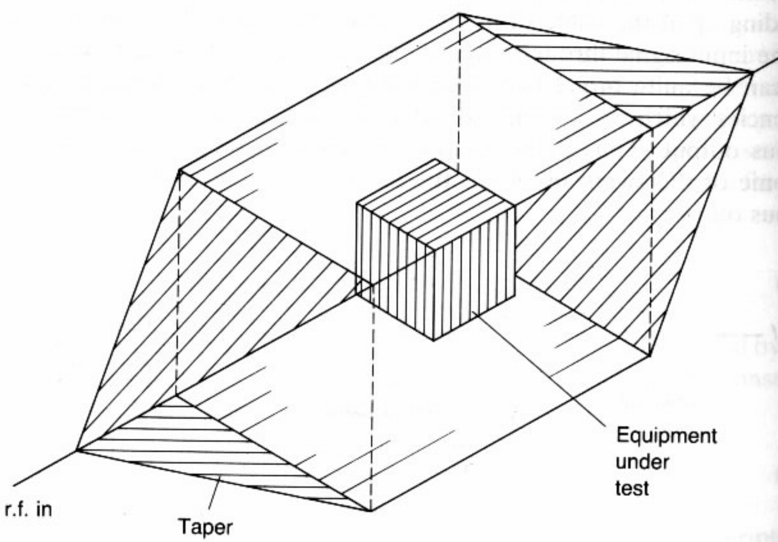

Fig. 11.15 Arrangement of a TEM cell for measuring the electromagnetic susceptibilities of electronic equipment.

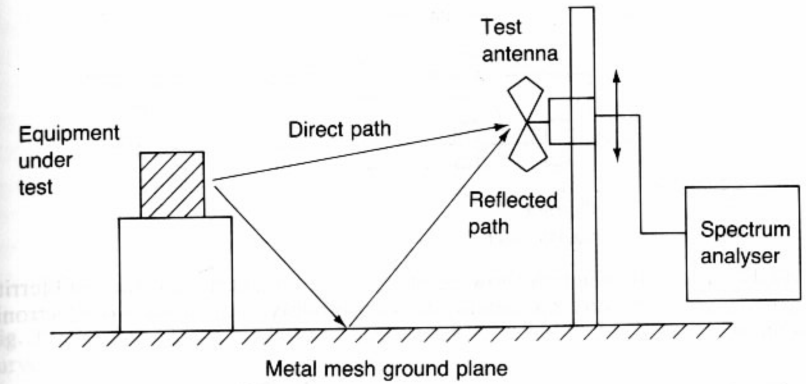

Fig. 11.16 Arrangement of an open field test site for measuring the radiation emitted from electronic equipment.

commonly has a metal case. Thus it is necessary to specify whether the test field is to be set up in the presence or absence of the EUT.

Radiated emission measurements are commonly made on an open field test site as shown in Fig. 11.16. The radiation from the EUT is received by a standard antenna and fed to a spectrum analyser. Because reflection of the radiation from the ground is unavoidable a metal mesh ground plane is used to ensure that the reflection is as close to that predicted by theory as possible. When the test antenna is moved vertically on its mast maxima and minima of the signal are observed corresponding to constructive and destructive interference between the direct and reflected signals.

Open field test sites suffer from the disadvantage that the test antenna also picks up the signals from nearby radio transmitters. To avoid this problem an anechoic chamber can be used, but it is difficult to make one which is satisfactory below 100 MHz. At lower frequencies screened rooms have been used to make measurements of radiated emission. In order to do this satisfactorily it is necessary to characterize the room at all frequencies in the range of interest because of the effects of reflections from the walls. An added problem with low-frequency measurements is that the receiving antenna is inevitably in the induction field rather than the radiation field of the EUT.

In many cases emission or susceptibility of the EUT has more to do with conduction along cables rather than direct radiation. Measurements are then made by using current transformers to couple signals off or on to the cables as shown in Fig. 11.17. The ferrite rings provide matched terminations for the signals to be measured.

Electromagnetic compatibility measurements are becoming much more important because of the introduction in 1992 of an EEC directive that all electrical equipment marketed and put to use within the EEC must satisfy

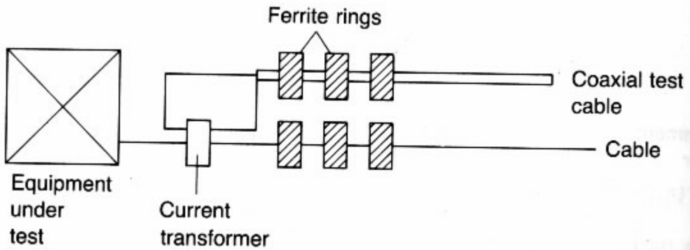

Fig. 11.17 Schematic diagram showing the use of a current transformer and ferrite damping rings to measure the conducted susceptibility and emission of electronic equipment.

regulations for both emission and immunity. Further information is given by Keiser (1983) and Jackson (1989).

11.10 MEASUREMENT OF RESONATORS

A microwave resonator is characterized by its resonant frequency, Q factor and shunt impedance (R/Q). These parameters can be measured using the equipment illustrated in Fig. 11.18. The signal from a microwave sweep oscillator is coupled into the resonator by a probe. The coupling must be weak to avoid loading of the resonator by the measuring system. A similar probe couples the output to a scalar analyser. The frequency of the oscillator is measured by coupling some of the signal into a microwave counter. Because the counter requires an unmodulated input and the scalar analyser requires a modulated input it is necessary to modulate the output signal from the resonator before it is detected.

If the oscillator is swept over a wide band of frequencies a series of peaks appears on the screen of the scalar analyser corresponding to the different modes of the resonator. The oscillator signal must not have appreciable

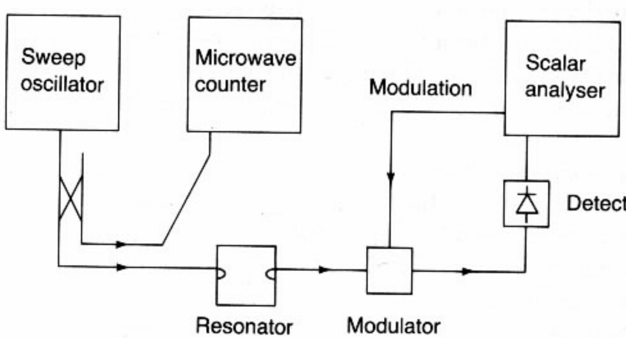

Fig. 11.18 Block diagram of equipment for measuring the properties of microwave resonators.

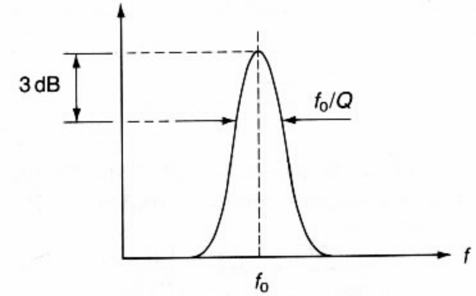

Fig. 11.19 Measuring the Q of a resonator from the width of its frequency-response curve.

harmonic content or spurious responses will appear. The sweep can then be narrowed down to isolate a single resonance (see Fig. 11.19) and the frequency swept manually to locate the peak of the response. This frequency is read directly from the counter to an accuracy of around $0.1/Q$. The vertical scale of the scalar analyser display is calibrated in decibels so it is easy to locate the 3 dB points, measure their frequencies, and calculate the Q factor of the resonator from (7.10).

To measure the shunt impedance of a cavity resonator a perturbation technique is used. This can either involve a perturbation of the cavity boundary by a metal probe or the insertion of a dielectric rod (Waldron, 1967). The latter technique will be described here with reference to the TM_{01} mode of a cylindrical pillbox cavity.

Figure 11.20(a) shows such a cavity with a thin dielectric rod inserted along its axis. For this mode of oscillation the electric field is axial and maximum on the axis. Provided that the perturbing effect of the rod is small we can assume that the fields in the cavity outside the rod are unaffected by its presence. The equivalent circuit of the perturbed cavity is then as shown in Fig. 11.20(b) where L_0 and C_0 are the inductance and capacitance of the unperturbed cavity and C_1 represents the effect of the dielectric rod. C_1 is calculated by treating the rod as a parallel-plate capacitor

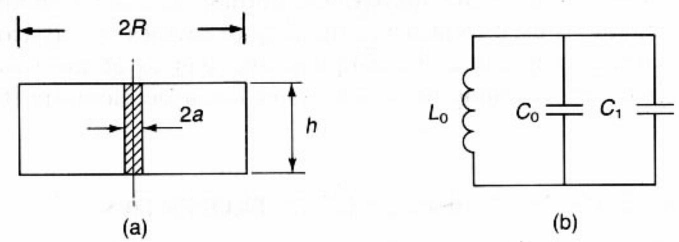

Fig. 11.20 Perturbation of a cylindrical microwave resonant cavity by a dielectric rod: (a) general arrangement, and (b) equivalent circuit.

and subtracting the capacitance of an air-spaced capacitor having the same shape. The result is

$$C_1 = \frac{\varepsilon_0 \pi a^2}{h}(\varepsilon_r - 1), \tag{11.19}$$

where ε_r is the relative permittivity of the rod. If the frequency perturbation is $\Delta\omega$ then, since the frequency is proportional to $1/\sqrt{C}$,

$$\frac{\omega_0 + \Delta\omega}{\omega_0} = \sqrt{\left(\frac{C_0}{C_0 + C_1}\right)}. \tag{11.20}$$

If $C_1 \ll C_0$ the square root can be expanded by the binomial theorem to give

$$\frac{\Delta\omega}{\omega_0} \simeq -\frac{C_1}{2C_0}. \tag{11.21}$$

Substituting for C_1 from (11.19) gives

$$\frac{\Delta\omega}{\omega_0} \simeq -\frac{\varepsilon_0 \pi a^2(\varepsilon_r - 1)}{2hC_0}. \tag{11.22}$$

Now from (7.4)

$$\frac{1}{C_0} = \omega_0\left(\frac{R}{Q}\right) \tag{11.23}$$

so that (11.22) becomes

$$\frac{\Delta\omega}{\omega_0^2} \simeq -\frac{\varepsilon_0 \pi a^2(\varepsilon_r - 1)}{2h}\left(\frac{R}{Q}\right). \tag{11.24}$$

Thus, if the dimensions of the rod and its relative permittivity are known the R/Q of the cavity can be calculated from measurements of ω_0 and $\Delta\omega$. Note that the negative sign in (11.24) means that the frequency is perturbed downwards by the rod.

A more exact analysis takes account of the changes in the cavity fields as a result of the presence of the rod. For a ratio $a/R = 0.06$ the error in R/Q resulting from the use of the approximate formula (11.24) is about 3%.

This technique is used to measure the R/Q of cavities for klystrons and can be adapted to measure the coupling impedances of the slow-wave structures used in travelling-wave tubes and linear accelerators (Ch. 10) (Connolly, 1976).

11.11 MEASUREMENT OF DIELECTRIC PROPERTIES

The dielectric properties of a material can be measured by essentially the same method as that described for measuring R/Q in the previous section.

If the cavity resonator has a simple shape such as a cylindrial pillbox then its R/Q can be calculated from theory (see Section 7.3). Equation (11.24) can then be used to calculate the relative permittivity of the perturbing rod from the frequency shift. The change in the Q of the cavity produced by the rod can be used to calculate the loss tangent of the material.

This method suffers from a number of disadvantages. It requires the sample of material to be of a particular size and shape, it suffers from errors caused by the approximations made in deriving equation (11.24) and it is not good for measuring the loss tangents of low-loss materials because the change in Q is too small to measure accurately. The first two problems can be overcome by using a method in which the sample protrudes into a cavity through a slot and making measurements for different depths of insertion. When the frequency shift is plotted against the insertion depth the slope of the line gives the permittivity free from errors produced by the finite size of the sample and the slot in the cavity wall through which it protrudes.

Another technique for measuring the dielectric properties of materials depends on measuring the properties of a section of coaxial line or waveguide which is filled with the material. The transmission loss and return

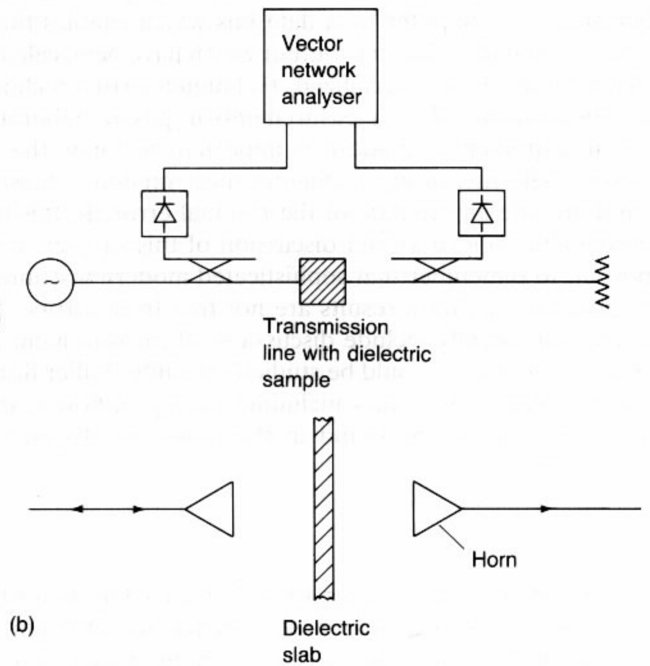

Fig. 11.21 Measurement of the dielectric properties of materials: (a) using a section of transmission line filled with the dielectric, and (b) using a slab of dielectric.

loss can be measured with a vector network analyser (see Fig. 11.21(a)) and the dielectric properties of the sample deduced.

When the sample of the material is in the form of a large slab its properties can be measured by placing the slab between two microwave horns as shown in Fig. 11.21(b). The slab has the effect of changing the path length between the horns and also of reflecting some of the incident power because of the mismatch of wave impedance at its surface. The transmitted and relected signals are then analysed and the dielectric properties calculated as in the previous paragraph.

A useful review of methods of measuring dielectric properties and a table giving figures for a wide range of materials are given by Metaxas and Meredith (1983).

11.12 CONCLUSION

This chapter has set out to show how the devices and theory discussed earlier in the book are employed in a variety of microwave measuring systems. Most of the techniques and instruments likely to be found in an industrial or university microwave laboratory have been described. For the most part they are used routinely as part of the process of product development or manufacture. Very often the instruments are connected to each other and to a computer by a data bus which enables the measurements to be automated and results output which have been calculated from the raw data. Other, more specialized, techniques exist which are used in standards laboratories and for measurements in physics laboratories.

When a measurement is made it is important to know the sources of possible error. The measuring technique must minimize these as far as possible and provide an estimate of the residual error. In this book it has only been possible to give a brief discussion of this subject. It is particularly important to remember that sophisticated modern instruments which appear to give very accurate results are not free from errors. The manufacturers' manuals usually include discussions of measurement errors and their correction and these should be studied carefully. Fuller discussions of microwave measuring techniques including the signal-flow graph method for estimating errors will be found in the books by Bryant (1988) and Laverghetta (1976).

EXERCISES

11.1 When a slotted section of air-spaced coaxial line is terminated by a short circuit the standing-wave minima are 18.6 mm, 16.3 mm, 12.7 mm and 9.6 mm apart at four different frequencies. What are those frequencies?

11.2 If the figures given in Question 11.1 were obtained with a WG16 waveguide slotted line what would the frequencies be?

11.3 Examine the effect of increasing the source reflection coefficient in the example on p. 269 to 0.1.

11.4 In a series of slotted-line measurements the separation of the minima is 36.5 mm. When the unknown loads are replaced by a short circuit the minima move towards it by 18.2 mm, 23.7 mm and 31.8 mm. The corresponding VSWR figures are 1.07, 1.20, 1.43. Find the normalized impedances of the unknown loads.

11.5 A cylindrically symmetrical cavity resonator is 11 mm high and resonates at 3.56 GHz with the electric field directed vertically. When a 3 mm glass rod ($\varepsilon_r = 4.1$) is inserted along the axis the resonant frequency drops by 45 MHz. Calculate the R/Q of the cavity.

11.6 The resonator described in Question 11.5 is perturbed by a different dielectric rod 2.54 mm in diameter. If the frequency drop is 94 MHz what is the relative permittivity of the rod?