INTRODUCTION TO MICROWAVE MEASUREMENTS: DETECTION OF RF POWER AND STANDING WAVES
Lab #2

The purpose of this lab is to gain a basic understanding of typical microwave measurement techniques, as well as improved insight into the behavior of microwave systems based on transmission lines. Topics to be examined include detection of microwave power and the measurement of impedance, reflection coefficient, and voltage standing wave ratio (VSWR). The propagation velocity, loss, and characteristic impedance of various transmission lines will also be measured.

The experiments in this lab fall into three main sections. In the first, we investigate the behavior and performance of "crystal" detectors (Schottky diodes) for detecting microwave power. The second part of the lab utilizes detectors of this type, along with the slotted line and standing wave meter, to measure the impedance, VSWR, propagation velocity, and loss of microwave components, including transmission lines and terminations of various types.

Detection of Microwave Power:

A fundamental requirement for all microwave measurement techniques is a means for detecting microwave power. Most microwave detection schemes fall into one of two categories, so-called "square-law" detectors and heterodyne systems. Heterodyne detection is referred to as linear detection since the output of the detector in this type of system is directly proportional to the electric field (and thus the RF voltage) applied to the detector. Square-law detectors, on the other hand, produce detector outputs that are proportional to the square of the applied RF voltage, and thus the detector output is proportional to the microwave power. Common square-law detectors in use for low-level RF power include crystal detectors, bolometers, and thermistors. Properly calibrated, thermistors are useful for absolute measurement of power; however, in many applications of practical interest, it is sufficient to measure the ratio of the power between two or more different signals. This sort of ratioed measurement can be performed either on unmodulated signals or on modulated signals. The latter technique can provide significantly better measurement sensitivity, as it circumvents the difficulties associated with designing high-gain, low-noise, dc-coupled, stable amplifiers. The range of power levels that can be detected in this way is limited at the low end by detector and amplifier noise, and at high power levels by deviations from square-law behavior due to detector non-linearity. In this experiment, the response of a crystal detector (in this case, a GaAs diode) will be checked by using it to detect RF energy from a microwave source.

a.) Being careful to use correct technique for mating the connectors (be careful not to let the center pin rotate in the receptacle), connect the detector to the ESG-3000 signal source. Set the signal frequency to 1 GHz without modulation. Connect the output of the detector to the HP3455 DMM, in DC voltage mode. For microwave powers from -50 dBm to 10 dBm, tabulate the detector output voltage vs. microwave power, and plot the results on a semilog scale (linear x axis = microwave power in dBm, logarithmic y axis = |detector voltage|).

b.) From this plot, determine the square-law region (the range of input powers for which the detector exhibits square-law operation) and the responsivity (in V/W) of the detector. This information will be needed for later experiments.

c.) Replace the GaAs detector with a General Radio slotted line terminated with a 50 Ω termination. Adjust the stub tuner for maximum signal (tuning at an input power level of -10 dBm usually produces good results), and again measure the DC detector voltage as a function of input power (in dBm). Note that the slotted line contains a crystal detector (Ge-W) similar in function to the HP GaAs detector. Plot the results of this measurement on the same graph as the HP GaAs detector results from part (a) above. Determine the square-law range and responsivity of the slotted-line detector.

Impedance Measurements Using a Slotted Line

Slotted-line measurement techniques are simple and accurate methods for measuring impedance, VSWR, and reflection coefficients at microwave frequencies. The measurement apparatus consists of an air-dielectric transmission line with a small probe that can be slid along the length of the transmission line to pick off a small amount of the signal (electric field), thus giving an indication of the RF voltage as a function of position along the line.

In order to achieve the best possible sensitivity and noise immunity with slotted-line measurements, the microwave source should be modulated (AM), and a narrow-band AC measurement made. This is most conveniently done using an HP 415 SWR meter, which is essentially a calibrated log-scale ac voltmeter. The modulation frequency for the source should match the center frequency of the particular SWR meter being used – typical values are in the vicinity of 1 kHz, and the exact value is labeled on the front of each meter. The SWR meter should be connected to the detector of the slotted line (replacing the DMM), and at each measurement frequency the stub tuner should be adjusted to maximize the received signal. A detailed example outlining the measurement procedure for finding impedances using the slotted line is provided in Example 2.4 (page 71-73) in Pozar, which is included at the end of this lab handout. Please read through this example problem and be familiar with it before performing the slotted-line portion of the lab.

a.) Measure a short (type N female) at 1 GHz. Find more than one minimum along the line; calculate the distance between minima ($\lambda/2$); since the dielectric in the slotted line is air, the propagation velocity of the electromagnetic wave is the speed of light; use this fact to calculate the frequency and compare to the set frequency of the microwave source.

b.) Measure an open at 1 GHz. Again find more than one minimum and calculate the frequency obtained. Compare the locations of the minima obtained for the open and short-terminated lines.

c.) Measure the VSWR and the impedance of the 100 Ω GR termination at 1 GHz. How does the measured VSWR compare to what one would expect for a 100 Ω termination on a 50 Ω transmission line?

d.) Measure the impedance of the microstrip stub circuit (terminate the second port with a 50 Ω type N termination) at frequencies of 500 MHz, 1000 MHz, 1500 MHz, and 2500 MHz. Note that in order to do this you will need to find the location of a null for a short

at each frequency, as well as adjust the stub tuner for best response at each frequency. Plot the results on a smith chart and note which of the microstrip stub circuits you are measuring; in a future lab you will compare the measurements made with the slotted line to those obtained using other techniques.

Transmission Line Measurements Using a Slotted Line
The slotted line can also be used to quickly characterize the performance of other transmission lines. For example, the propagation velocity of non-air dielectric lines is less than the velocity of propagation in free space. One way to find the relative velocity of propagation in a cable is to compare the number of half-wavelengths at a given frequency in the cable to the number of half-wavelengths in the same cable when the velocity of free space is assumed. The procedure for measuring the propagation velocity for RG-8A/U is outlined below.

a.) Attach the 15 foot RG-8A/U cable to the slotted line, and terminate the free end of the cable with a short termination. At a frequency of 1 GHz, find and record the location of the closest minimum to the cable.

b.) Slowly increase the frequency of the source until a null re-appears at the same position on the slotted line (you will need to adjust the tuning stub for best sensitivity as you change the frequency). From knowing that at $f_1 = 1$ GHz there were N half-wavelengths on the line and at f_2 there are N+1 half-wavelengths over the same distance, one can determine the value of N. This is the total number of half-wavelengths in the slotted line and coaxial cable at 1 GHz. To determine the number of half-wavelengths in the cable alone, it is necessary to determine the number of half-wavelengths in the slotted-line section between the carriage and the end of the slotted line. To do this, re-tune the slotted line at 1 GHz terminate it with a type-N short. Find and record the location of the first null on the slotted line, and determine the distance between the nulls for the short-terminated slotted line and the cable-terminated slotted line. The relative velocity of propagation in the cable is then given by the following formula:

$$v_r = \frac{(15\,ft)(12in\,/\,ft)(2.54cm\,/\,in)}{(N-1)\left(\frac{3.0x10^{10}\,cm\,/\,s}{2\cdot1GHz}\right) \pm L}$$

If the null with the slotted line shorted is nearer to the end of the line than the null with the cable shorted, use the negative sign; otherwise use the positive sign in the expression above.

c.) The loss in a cable can also be assessed using the slotted line. At frequencies of 500 MHz, 1000 MHz, 1500 MHz, and 2500 MHz, measure the impedance of the 15 foot RG-8A/U cable terminated with a short. Note that this requires recording the VSWR and null positions at each frequency, as well as the position of the nulls obtained with a short-terminated line previously. From the data collected, compute the "return loss" of the cable (RL(dB) = -20 log(|Γ|)). Since the short (ideally) has a reflection coefficient of -1, it reflects all of the incident power back towards the source. Thus the loss/length of cable at each frequency is the return loss/(2 15'). Plot the cable loss as a function of frequency in units of dB/100 ft.

straightforward to use (2.43) with $\ell = 0$ to find the load impedance from Γ:

$$Z_L = Z_0 \frac{1 + \Gamma}{1 - \Gamma}. \tag{2.60}$$

The use of the Smith chart in solving this problem is best illustrated by an example.

EXAMPLE 2.4   IMPEDANCE MEASUREMENT WITH A SLOTTED LINE

The following two-step procedure has been carried out with a 50 Ω coaxial slotted line to determine an unknown load impedance:

1. A short circuit is placed at the load plane, resulting in a standing wave on the line with infinite SWR, and sharply defined voltage minima, as shown in Figure 2.14a. On the arbitrarily positioned scale on the slotted line, voltage minima are recorded at

$$z = 0.2 \text{ cm}, 2.2 \text{ cm}, 4.2 \text{ cm}.$$

2. The short circuit is removed, and replaced with the unknown load. The standing wave ratio is measured as SWR = 1.5, and voltage minima, which are not as sharply defined as those in step 1, are recorded at

$$z = 0.72 \text{ cm}, 2.72 \text{ cm}, 4.72 \text{ cm},$$

as shown in Figure 2.14b. Find the load impedance.

Solution
Knowing that voltage minima repeat every $\lambda/2$, we have from the data of step 1 above that $\lambda = 4.0$ cm. In addition, because the reflection coefficient and input impedance also repeat every $\lambda/2$, we can consider the load terminals to be effectively located at any of the voltage minima locations listed in step 1. Thus, if we say the load is at 4.2 cm, then the data from step 2 shows that the next voltage minimum away from the load occurs at 2.72 cm, giving $\ell_{min} = 4.2 - 2.72 = 1.48$ cm $= 0.37\lambda$.
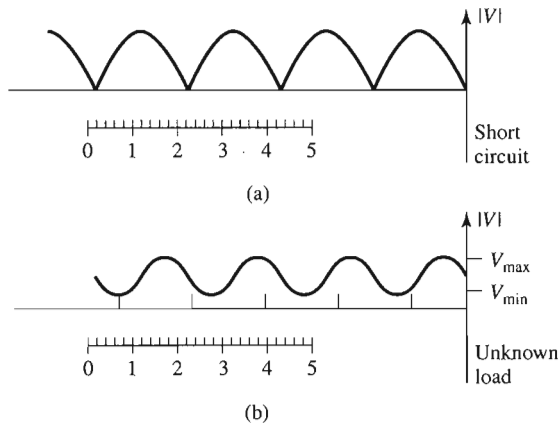
FIGURE 2.14   Voltage standing wave patterns for Example 2.4. (a) Standing wave for short-circuit load. (b) Standing wave for unknown load.
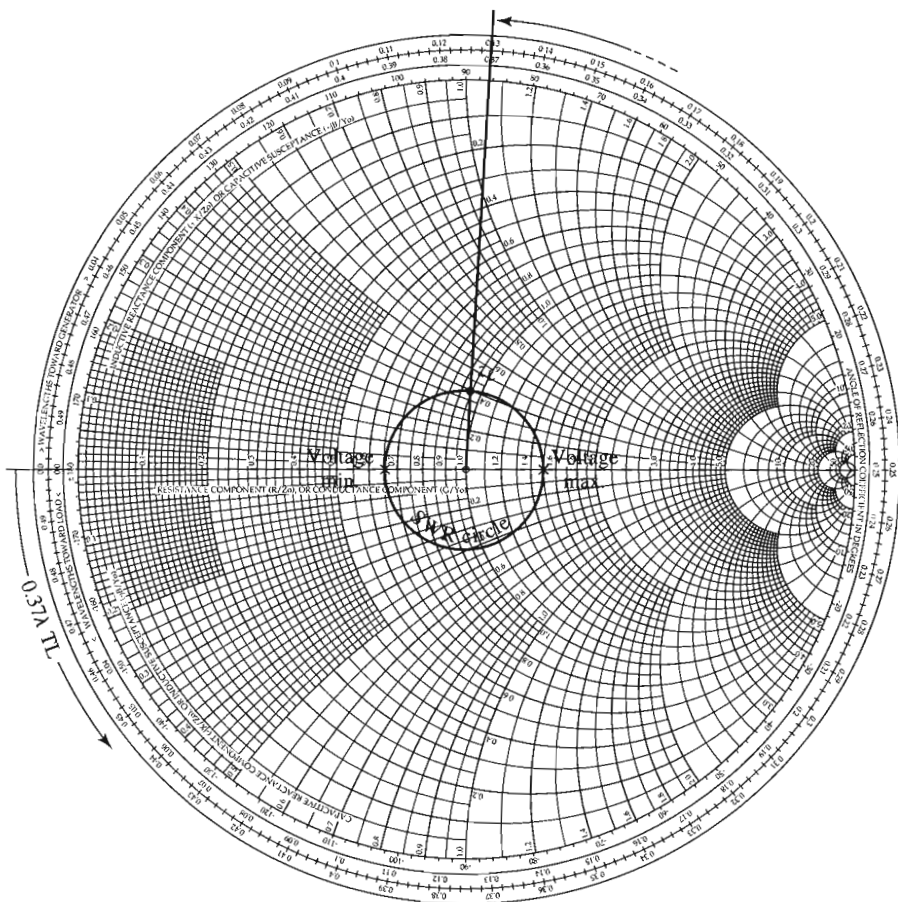
FIGURE 2.15   Smith chart for Example 2.4.

Applying (2.58)–(2.60) to this data gives

$$|\Gamma| = \frac{1.5 - 1}{1.5 + 1} = 0.2,$$

$$\theta = \pi + \frac{4\pi}{4.0}(1.48) = 86.4°,$$

so
$$\Gamma = 0.2e^{j86.4°} = 0.0126 + j0.1996.$$

The load impedance is then

$$Z_L = 50\left(\frac{1 + \Gamma}{1 - \Gamma}\right) = 47.3 + j19.7 \ \Omega.$$

For the Smith chart version of the solution, we begin by drawing the SWR circle for SWR = 1.5, as shown in Figure 2.15; the unknown normalized load impedance must lie on this circle. The reference that we have is that the load is 0.37λ away from the first voltage minimum. On the Smith chart, the position of a voltage minimum corresponds to the minimum impedance point (minimum voltage, maximum current), which is the horizontal axis (zero reactance) to the left of the origin. Thus, we begin at the voltage minimum point and move 0.37λ toward the load

(counterclockwise), to the normalized load impedance point, $z_L = 0.95 + j0.4$, as shown in Figure 2.15. The actual load impedance is then $Z_L = 47.5 + j20 \ \Omega$, in close agreement with the above result using the equations.

Note that, in principle, voltage maxima locations could be used as well as voltage minima positions, but voltage minima are more sharply defined than voltage maxima, and so usually result in greater accuracy.   ∎

2.5   THE QUARTER-WAVE TRANSFORMER

The quarter-wave transformer is a useful and practical circuit for impedance matching and also provides a simple transmission line circuit that further illustrates the properties of standing waves on a mismatched line. Although we will study the design and performance of quarter-wave matching transformers more extensively in Chapter 5, the main purpose here is the application of the previously developed transmission line theory to a basic transmission line circuit. We will first approach the problem from the impedance viewpoint, and then show how this result can also be interpreted in terms of an infinite set of multiple reflections on the matching section.

The Impedance Viewpoint

Figure 2.16 shows a circuit employing a quarter-wave transformer. The load resistance R_L, and the feedline characteristic impedance Z_0, are both real and assumed to be given. These two components are connected with a lossless piece of transmission line of (unknown) characteristic impedance Z_1 and length $\lambda/4$. It is desired to match the load to the Z_0 line, by using the $\lambda/4$ piece of line, and so make $\Gamma = 0$ looking into the $\lambda/4$ matching section. From (2.44) the input impedance Z_{in} can be found as

$$Z_{in} = Z_1 \frac{R_L + jZ_1 \tan \beta\ell}{Z_1 + jR_L \tan \beta\ell}. \tag{2.61}$$

To evaluate this for $\beta\ell = (2\pi/\lambda)(\lambda/4) = \pi/2$, we can divide the numerator and denominator by $\tan \beta\ell$ and take the limit as $\beta\ell \to \pi/2$ to get

$$Z_{in} = \frac{Z_1^2}{R_L}. \tag{2.62}$$

In order for $\Gamma = 0$, we must have $Z_{in} = Z_0$, which yields the characteristic impedance Z_1 as
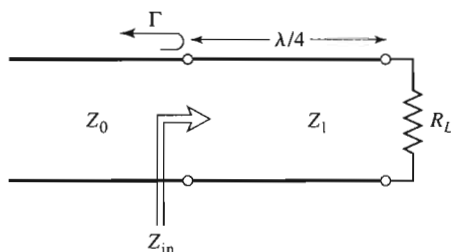
$$Z_1 = \sqrt{Z_0 R_L}, \tag{2.63}$$

FIGURE 2.16   The quarter-wave matching transformer.