Microwave Theory and Measurements

PRENTICE-HALL INTERNATIONAL, INC.
London • Tokyo • Sydney • Paris
PRENTICE-HALL OF CANADA, LTD.
PRENTICE-HALL DE MEXICO, S. A.

Engineering Staff
of the Microwave Division
HEWLETT-PACKARD COMPANY

Microwave Theory and Measurements

Under the Editorship of
Dr. Irving L. Kosow

PRENTICE-HALL, INC.
Englewood Cliffs, N.J. 1962

 Printed in the United States of America. Library of Congress Catalog Card Number: 62–17424

5 8 1 5 2–C
5 8 1 5 3–C

Preface

This book has been prepared to meet a continuing and increasing demand in the electronic industry and the academic field for basic information on microwave theory and microwave measurements. The contents provide analyses of microwave theory, microwave instrumentation, and microwave measurement techniques. The material is so presented that it is readily understandable to any person with a fundamental engineering background. No calculus is required. Technical institute students, as well as college undergraduate and graduate students, can use this publication as a basic text or as a supplement to other books. Engineers can use the book to learn the fundamentals of microwave theory and techniques.

The book consists of four major sections. Section 1, an introduction, contains a description of the importance and behavior of microwaves. Microwave theory is covered in Section 2. The first part of this section describes basic microwaves, explains the difference between microwave and low-frequency circuits, and discusses voltages and currents in a transmission line. These considerations are extended to waveguide systems in the second part of Section 2. Section 3 presents information on the basic types of microwave measurements. Section 4 contains fifteen experiments designed to acquaint the reader with microwave equipment and with the techniques employed in making microwave measurements. In addition to the four Sections, three Appendices contain a glossary of microwave terms, descriptive literature showing the individual microwave instruments, and a short bibliography of selected reference works.

The experiments in Section 4 are arranged to permit maximum flexibility, depending upon the amount of time available and the experience of the experimentor. Because compatibility of equipment is important in microwave instrumentation, the experiments are based on the use of a specially selected group of microwave instruments currently available from Hewlett-Packard Company.

The data from which this book was compiled were selected from experiments and technical discussions developed by the staff of the hp Microwave Development Laboratory for use in training seminars. In addition to the credit due the individual engineers of the Laboratory, special recognition must be given to John Minck, Microwave Application Engineer, and Carl Anderson, Application Note Editor, for their work in arranging and editing the book.

Engineering Staff
of the Microwave Division
Hewlett-Packard Company

Contents

EXPERIMENT

SECTION ONE

Introduction

Microwaves have a broad range of application in modern technology. In the field of entertainment, for example, television programs are transmitted from coast to coast via a transcontinental microwave network. The same network also carries hundreds of telephone and telegraph circuits. Microwaves are also used for inter-city communications and for local transmission of television between studios and remote transmitters.

Microwaves also form an important element of local and national security programs. Microwave applications in radar, guidance systems, and communications are of major importance in our over-all national defense. At the level of local security, microwave is used extensively for police and fire department communications. There are also many strictly commercial uses of microwaves. These include private communication systems within large organizations such as railroads, pipeline companies, and public utility organizations. Microwave techniques are being utilized to develop improved methods of air navigation and airport control.

Most of the areas of microwave application involve "communication" in the broadest sense of the word. Microwaves are important to communications for reasons that are, in general, related to the manner in which they are generated and transmitted.

1-1 Background Material

Electromagnetic waves. Electromagnetic waves are wave motions produced by electric and magnetic fields whose intensity and orientation vary as a function of time. If we neglect the manner in which these fields are produced, and consider only that their existence produces an electromagnetic wave, we find that the *rate* at which these fields vary with time can range over a tremendous magnitude. They extend from the extremely slow rates resulting from the rotation of a coil in a magnetic field to the tremendously high rates caused by molecular and atomic disturbances.

Included in this wide spectrum are the various rates of variation (frequencies) at which electromagnetic waves produce such tangible and well-known effects as visible light, radiant heat, and, at much lower rates, radio waves.

Except for the manner of production of their fields, all electromagnetic waves obey the same physical laws. For this reason the microwave engineer can borrow extensively from the techniques of both the optical engineer and the communications engineer who works with the more familiar lower radio frequencies.

It can be demonstrated both analytically and experimentally that wherever electromagnetic waves are propagated they inherently carry energy in the direction of propagation. The manner of production of the time-varying fields and the manner of launching the resultant electromagnetic wave into free space is the concern of the microwave engineer. Also, the accuracy and dependability of communication is closely related to the quantity of energy that is propagated, and the matter of efficiency in handling this energy is, therefore, of as much importance to the microwave engineer as it is, for example, to the power engineer.

Optics. Many familiar examples of how light may be focused into beams and reflected or refracted have their counterparts in microwave engineering. In fact, the extensive knowledge of optics has contributed much to advancements in the microwave field. Behavior of optical waves and microwaves becomes increasingly similar as frequency of the microwaves increases. The similarities are particularly important when we consider the transmission of microwaves in free space (between antennas).

Wavelength and frequency. The characteristics of electromagnetic waves and the manner of their physical behavior are functions of the rate at which the electric and magnetic fields vary. In single frequencies, these variations are periodic and sinusoidal, and therefore can be considered in terms of frequency in cycles (complete alternations) per second.

The wave motions produced by electromagnetic fields propagate and carry energy. In microwave communications the medium of propagation is usually the "free space" surrounding the earth. Regardless of the frequency of the wave, the velocity of propagation is constant within any one medium, and in the most usual medium (free space), the propagation velocity is the speed of light.

Wavelength is the measure of the distance a wave travels during one alternation of the producing field. The relationship is written in the form:

$$\text{Wavelength} = \frac{\text{Velocity of propagation}}{\text{Frequency}}$$

Note that for a given medium, wavelength is inversely proportional to frequency. A typical example of long wavelength is that of the field produced by 60 cycle current in power transmission lines. This wavelength is on the order of 3100 miles. On the other hand, the center of the amplitude-modulation broadcasting band involves wavelength on the order of 300 meters. The frequencies generally considered to be microwaves involve wavelengths on the order of 30 cm to a fraction of a centimeter. (The light used for viewing this page involves wavelengths averaging 60 millionths of a centimeter.) The lower limit of the microwave frequency spectrum is not well defined. Some communications companies refer to all frequencies above 890 mc (about 30 cm wavelength) as microwaves. Other engineering authorities say the microwave region begins at about 2000 mc. Waveguide, the most convenient closed microwave transmission system, is commonly used for frequencies above 2.6 gigacycles (gc)*, although some extremely high power systems utilize waveguide for delivering power at frequencies in the range of 500 mc.

History. Much of the analytical work for microwave development is derived from the basic equations formulated by James Clerk Maxwell in 1864. Even at that early date Maxwell was able to predict that if

* 1 gc = 1 kmc = 10^9 cycles.

radio waves could be produced, they would act in the same manner as light waves. It was not until some 25 years later that Heinrich Hertz produced electromagnetic waves having a wavelength of around 60 cm, and demonstrated that the effects anticipated by Maxwell did, in fact, exist.

In 1897 (still before the days of radio communications as known today), Lord Rayleigh, making use of Maxwell's equations, showed that electromagnetic waves may be propagated in hollow tubes or, in fact, in any medium having specified boundaries between two electrically different media. Finally, at the turn of the century, Marconi demonstrated the first practical radio communication. The story of radio in the ensuing 60 years is well known.

Generation. As we mentioned earlier, electromagnetic fields can be generated by widely divergent physical phenomena. It is this aspect in which there are major differences between light and radio waves. At the wavelengths of light, energy is produced by disturbances in the chemical or physical structure of material itself. A natural result is that the frequencies produced are dispersed over a narrow spectrum and are not, in general, coherent. Visible light, then, consists not of single frequencies of constant amplitude, but of dispersion of frequency, amplitude, and phase.

In the early spark-gap experiments, at the time of Hertz, radio waves were incoherent in the same sense. However, thanks to the vacuum tube and other devices, coherent waves (those which are, essentially, of single frequency and constant amplitude) can now be produced easily.

In the early days, the vacuum tube served merely as a method for supplying the necessary electrical energy to maintain constancy of amplitude of the electrical oscillation. However, the physical dimensions of vacuum tubes became excessively large compared with the wavelengths encountered as the communications art worked up toward the microwave range. As a result, it became necessary to incorporate these parameters into the basic structure of the tube. This technique gave rise to such microwave sources as the klystron and the magnetron. Detailed knowledge of the operation of these tubes is not necessary to the consideration of microwave systems. It is, however, important to note that these sources are generally of rather low efficiency, so that maximum efficiency in transmitting the energy is a matter of prime importance.

Transmission of microwaves. Microwaves travel through free space when they are transmitted between antennas. The transmission system extending from the microwave generator to the antenna usually consists of either coaxial lines or hollow metal pipes (rectangular or circular) called *waveguides.* Coaxial lines have higher losses than waveguides. They are usually limited in application to very short distances or to special situations in which the broader bandwidth of a coaxial system is needed.

Design and construction of appropriate transmission systems is now a major field of engineering. Most of the material in this publication is meant to help the reader understand waveguide systems so that he may use them for any specific application in the microwave field.

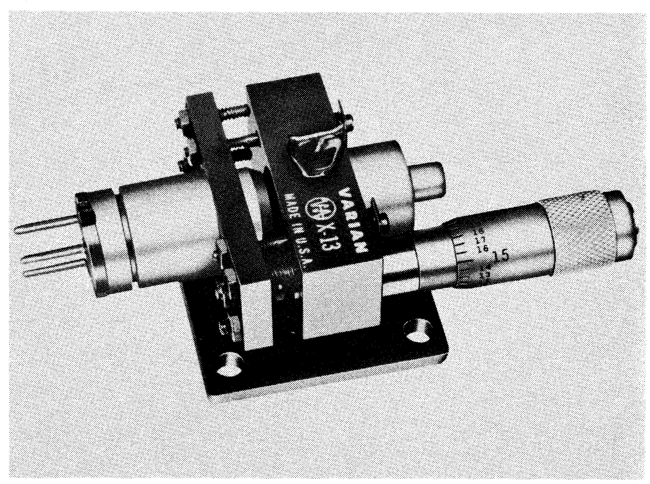


FIGURE 1–1 *The reflex klystron. Reproduced by permission of Varian Associates.*

1-2 Why Use Microwaves?

Bandwidth considerations. There is a direct relationship between bandwidth and the information-carrying capacity of a microwave system. The more information one wishes to communicate, the broader the bandwidth that is required in the transmission system. A single-frequency source of electromagnetic energy can transmit intelligence only with some form of modulation.

Regardless of the method of modulation, its effect is to produce sideband frequencies with respect to the original carrier. The extent to which these sideband frequencies deviate from the carrier is directly proportional to the information transmitted. Consider, for example, a typical amplitude-modulated broadcasting system, operating at 1 mc, which carries a radio program having frequency components of, say, 10,000 cycles. That station requires exclusive use of a range of frequencies approximately 20,000 cycles wide, centered around the carrier frequency. The allotted frequency range for commercial AM broadcasting is only from 0.5 to 1.5 mc. Accordingly, there is a limit to the number of stations which can operate in the same geographical area without interfering with each other.

As another example, consider that a television station requires a bandwidth of about 5 mc to handle all the information necessary to the production of a good television picture. Obviously, 1 mc cannot, in the normal sense, be used as the carrier. However, a center frequency of 100 mc can easily be used. A major reason for the importance of microwaves in communications is that it is possible to place a large number of information channels in a frequency range that occupies only a small percentage of the carrier frequency.

Antenna directivity. If electromagnetic energy can be radiated in such a manner that the outward energy flow from the radiating source is increased in a particular direction instead of being equal in all directions, the result is equivalent to an increase in power. At microwave frequencies such radiation can be made very highly directive. In fact, a good antenna arrangement can effectively produce a beam in much the same way as a spotlight is beamed. Also, as a result of this high directivity of radiation, a certain amount of information security is achieved. This security can be useful in terms of military needs or, probably more important, it provides protection against interference by other transmitting systems in the same area. Finally, the very fact that microwave beams can be made sharp makes it possible to give directional information in applications such as radar.

Waveguides as transmission systems. Both coaxial lines and waveguide are used to transmit microwaves up to 12.4 gc. The choice depends on several factors, such as bandwidth to be transmitted, length of the system, and ease of setting up and changing the system. Coaxial systems have broader bandwidth and are easier to handle; waveguide has lower loss and greater power-handling capacity. In general, coaxial systems predominate below about 2 gc, and waveguide systems are more widely used at higher frequencies.

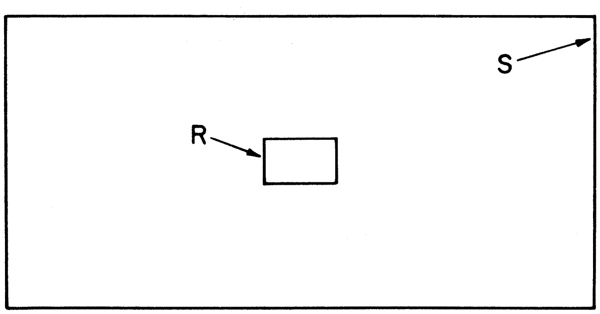


FIGURE 1–2 *Waveguide size is directly related to the transmitted frequency (and wavelength). Shown here are the relative interior sizes of waveguide for "S" band (2.6 to 3.95 gc) and "R" band (26.5 to 40 gc).*

1-3 Some Basic Considerations

Because generation of power at microwave frequencies is not a particularly efficient process, every effort is made within the transmission and radiation systems to obtain high efficiency in order to conserve the power available. To do so, the microwave engineer pays considerable attention to the problem of reducing any electrical defects that waste power in the system.

There are two important causes of power loss in a transmission system Some power is lost because of inherent imperfections in the conductors. This loss, which can be considered to be power dissipation, can be reduced to almost negligible amounts in waveguide by means of modern manufacturing techniques. The most common practice is to plate the interior waveguide surface with a thin coating of high-conductivity material, such as silver.

In many cases, the power lost through reflection of transmitted energy is more significant than the power lost through attenuation or dissipation. Power is reflected whenever the impedances of connected sections of the transmission system are not perfectly matched. A major part of microwave engineering is the measurement and minimization of reflections. Fortunately, the nature of electromagnetic waves

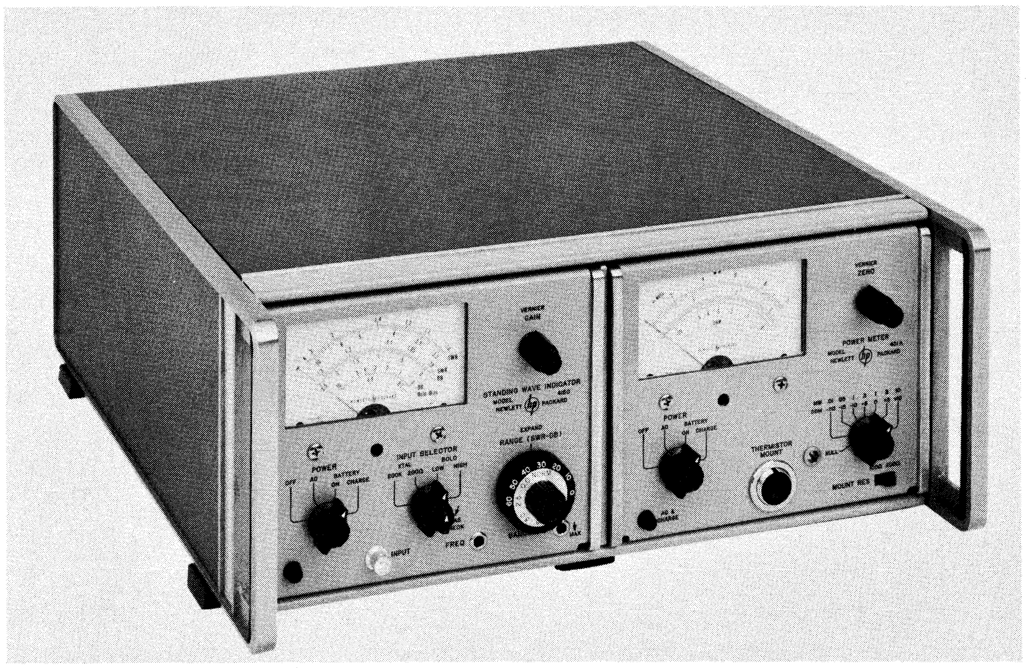


FIGURE 1–3 *Two important microwave measuring instruments. The standing wave indicator (left) used when determining microwave reflections, and the microwave power meter (right).*

is such that reflection can easily be detected by measuring the standing wave pattern in the system. This pattern is essentially an interference effect caused by power flowing in both directions. A comparable physical phenomenon occurs when a wave is introduced by means of a transverse motion on a rope with one end tied to a fixed point. Most of the power in the wave is reflected from the fixed end to form nodes or standing waves along the rope.

Impedance matching. If a transmission system were infinitely long and there were no impedance discontinuities, there would be no reflections. In a waveguide system, any abrupt variation in the transmission path appears as an impedance discontinuity and causes a reflection. Note that abrupt changes in the transmission path, rather than gradual changes, are the cause of discontinuities. If changes in physical dimensions or electrical characteristics must be made, one way of avoiding reflections is to make the changes gradually.

Characteristic impedance and normalized impedance. Every transmission system has a characteristic impedance. If the system were terminated by a load whose impedance is equivalent to the characteristic impedance, there would be no reflections. Further, every condition of reflection is caused by an impedance which can be related to the characteristic impedance. Thus, by determining the nature of the reflection, the nature of the impedance irregularity can also be determined. With this knowledge, corrective measures can be taken to compensate for the irregularity and to obtain increased transmission efficiency.

Most waveguide impedance calculations are made on the basis of what is called "normalized" impedance. This is simply the actual impedance divided by the characteristic impedance. The normalized impedance is always unity (1) for a system in which there are no reflections, since this condition exists only when the system is terminated with its characteristic impedance.

SECTION TWO

Microwave Transmission Theory

2-1 Transmission Lines

Circuit constants. At low frequencies circuit elements are lumped. Therefore, there is no variation of parameters with position. Another way of expressing the same concept is to say that low-frequency circuits are electrically short compared to the wavelength of the signals transmitted.

Microwave frequencies, on the other hand, have much shorter wavelengths (30 cm and less). The result is that circuits have electrical lengths either comparable to or greater than the wavelength of the signals under consideration. Because of this relationship, the various circuit parameters vary with position—a fact which must be taken into account in microwave work.

In general, all the well-known low-frequency circuit concepts apply to microwave circuits. In addition, however, the basic concept of distributed circuits must be considered.

A distributed circuit is simply a transmission path in which the basic elements of the circuit (resistance, inductance, and capacity) are spread evenly over the entire length of the path. The importance of the distributed-circuit concept can be demonstrated by comparing equivalent circuits for low-frequency and microwave work. First, consider a section of ordinary transmission line, as shown in Fig. 2–1. In this illustration, a generator feeds a load through a section of transmission line. We want to find out how this section of line affects the transmission from the generator to the load, first at a representative low frequency, and then at higher, or microwave, frequencies.

Current flowing in the conductors sets up a magnetic field encircling the conductors. Hence, the transmission line has inductance. Furthermore, the conductors making up the

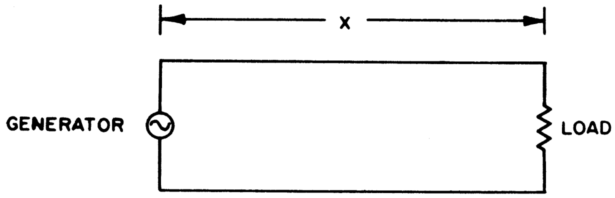


FIGURE 2–1 *Transmission line section.*

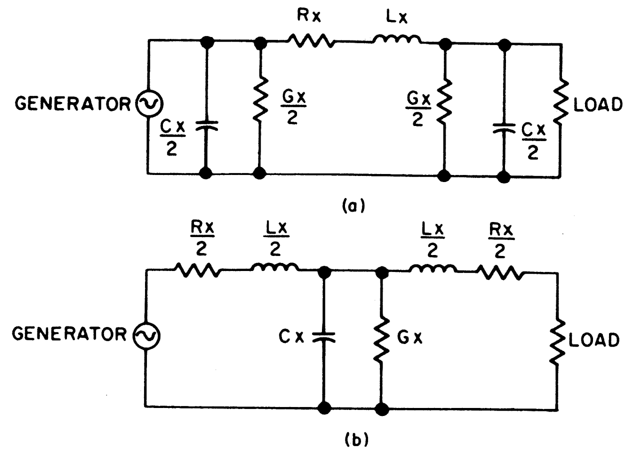


FIGURE 2–2 *Low-frequency equivalent circuit of transmission line.*

transmission line will have resistance associated with them. For a unit length of this line the resistance and inductance can be calculated. Let this resistance be R ohms per unit length and let this inductance be L henries per unit length. Since all other sections of the transmission line are in series for these basic elements, the total line resistance will be Rx ohms, and the total line inductance will be Lx henries.

There also will be voltage between and charges on the conductors, so there will be a capacity between the conductors. Furthermore, in the usual situation, the lines may be embedded in a lossy dielectric material, so that a conductive element must be assumed between the lines to account for this loss. Here again, a unit length can be used to calculate the capacity and conductance. Let the capacity be C farads per unit length and the conductance be G mhos per unit length. Now, for this capacity and conductance all sections of the line can be considered to be in parallel. Hence, the total capacity is Cx farads and the total conductance is Gx mhos.

At very low frequencies the effect of the inductance, capacity, resistance, and conductance of the line can be taken into effect by means of either equivalent circuit shown in Fig. 2–2.

The circuits shown in Fig. 2–2 obviously incorporate the correct amount of series inductance and resistance, as well as the proper amount of shunt capacity and conductance. There is, therefore, a temptation to treat them as exact equivalent circuits. If they were, their effects should be the same under all conditions. Further investigation shows that this simple assumption is not true. At very high frequencies the circuits of Figs. 2–2(a) and 2–2(b) present different impedances to the generator. In Fig. 2–2(a) the impedance presented to the generator at sufficiently high frequencies is $2/j\omega\, Cx$. In Fig. 2–2(b) this impedance is $j\omega\, Lx/2$. Obviously, these are not the same, because the impedance presented to the generator in Fig. 2–2(a) goes to 0 as the frequency becomes very high, and the impedance in Fig. 2–2(b) goes to infinity as the frequency becomes very high.

A simple means of making the impedances more nearly alike is to split up the total inductance, resistance, capacity, and conductance into more sections. In the ultimate case these basic elements could be divided into an infinite number of sections. When they are so divided, the two circuits shown in Fig. 2–2 do, indeed, become one and the same. This ultimate equivalent circuit, shown in Fig. 2–3, is the essence of distributed constant circuits.

Characteristic impedance and propagation constant. Many circuit concepts apply to microwave circuits in essentially the same manner as they do in low-frequency work. Perhaps the most important of these are (1) characteristic impedance and (2) propagation constant.

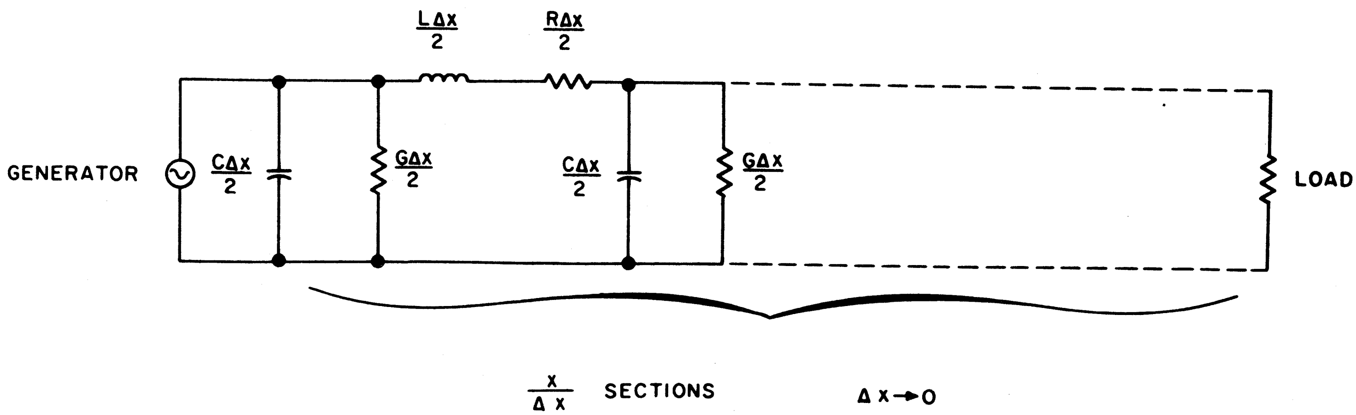


FIGURE 2–3 *Exact equivalent circuit of transmission line.*

1. Characteristic impedance, Z_0, is the input impedance of an infinitely long line. In terms of the inductance, resistance, capacity, and conductance, per unit length of the line, it is given by:

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \text{ ohms} \tag{1}$$

2. Propagation constant, γ, is a measure of the phase shift and attenuation along the line. It is a complex quantity given by:

$$\gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)} \text{ per unit length} \tag{2}$$

There are two units for propagation constant. Attenuation, α, is expressed in nepers per unit length; phase shift, β, in radians per unit length.

As in low-frequency circuits, γ is the exponent of an exponential function which relates the amplitude and phase of a voltage or current at two different points in a system. Consider, for example, a section of an infinitely long transmission line as shown in Fig. 2–4. In this figure let a sinusoidal signal E_1 be imposed at the left-hand terminal. Let E_x be the voltage a distance x down the line.

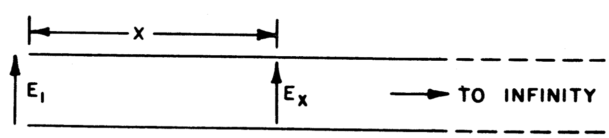


FIGURE 2–4 *Section of infinite transmission line.*

Now, E_x is related to E_1 by the propagation constant in the following manner:

$$\frac{E_x}{E_1} = e^{-\gamma x} = e^{-\alpha x - j\beta x} = e^{-\alpha x} \cdot e^{-j\beta x} \tag{3}$$

The first term in this expression, $e^{-\alpha x}$, is a real quantity and gets smaller as x increases. It tells us how much the signal is attenuated. The second term, $e^{-j\beta x}$, is an imaginary exponential quantity. Since it represents a vector of unit amplitude with a phase angle of βx rad, this quantity tells us how much the phase of the voltage E_x lags the phase of the voltage E_1. These two terms together completely define the phase and amplitude relations existing between voltages E_x and E_1. The propagation constant can be applied similarly to the current.

3. Guide wavelength and velocity of propagation. The lag in phase down the line is characteristic of distributed constant circuits; it comes about because there is a finite velocity at which a signal travels down the transmission line.

The imposed sinusoidal signal applied to the left-hand terminals of the infinitely long line in Fig. 2–4 can be represented as the projection of a rotating vector, as shown in Fig. 2–5. Here, the rotating vector of magnitude E rotating counterclockwise at a speed of ω rad/sec is shown at the left. Imagine that the transmission line is to the right of the figure beginning with the vertical line at $x = 0$, and that the distance x is measured to the right. The signal propagates along this line with a velocity v. In this figure the applied signal is represented by the projection of the rotating vector E on the vertical line at $x = 0$, and it is a sinusoidal wave. The signal impressed upon the transmission line at any particular instant of time propagates to the right with a velocity v. Consequently, at any particular instant the signal appearing at various points along the line is represented in the example shown in Fig. 2–5.

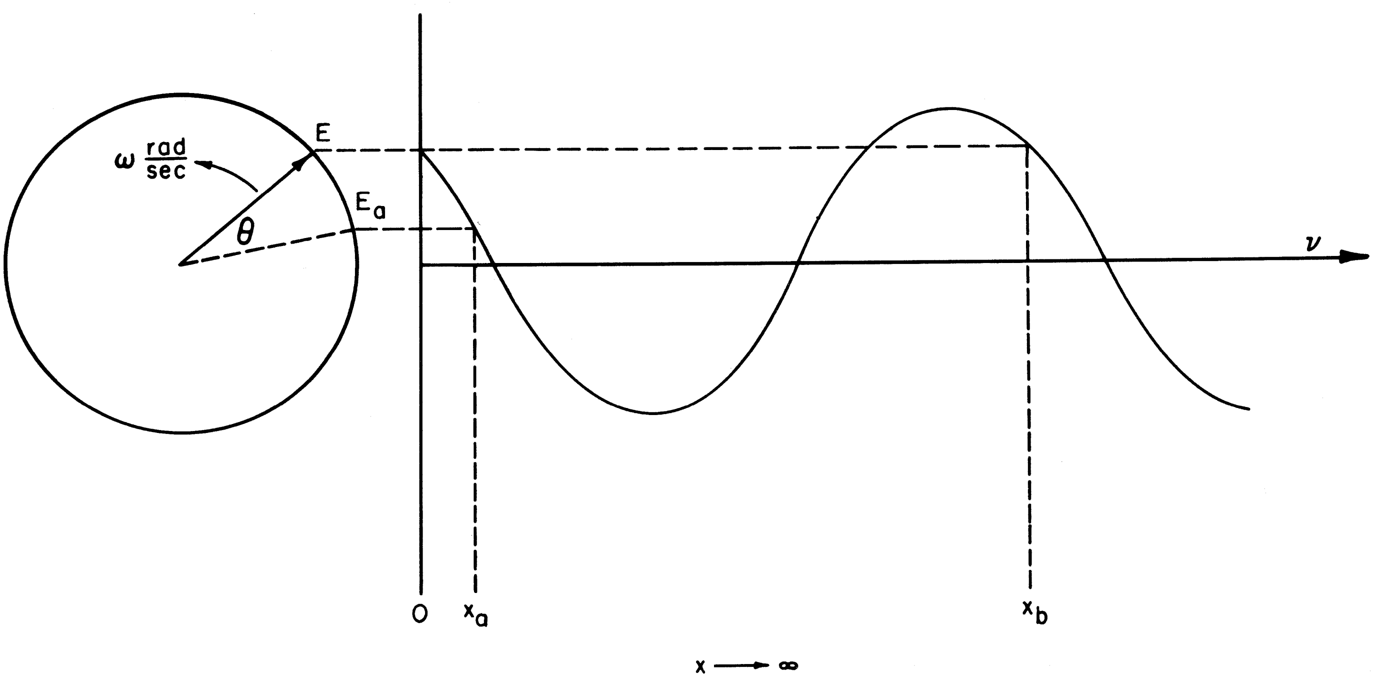


FIGURE 2–5 *Signal progression along infinite transmission line.*

At $x = 0$, the signal is represented by the rotating vector E. At x_a the signal can still be represented by a rotating vector (in this case E_a). This new vector has the same magnitude as the original vector, but it has been delayed by an angle θ. As the reference point is moved down the line (to the right), the phase lag increases until at point b the phase lag is 2π rad with respect to the input signal.

The distance x_b is known as a guide wavelength λ_g, and from the definition of β (phase constant) we have

$$\beta\lambda_g = 2\pi \tag{4}$$

$$\beta = \frac{2\pi}{\lambda_g} \tag{5}$$

Also, the time required for the signal to propagate to the point x_b is the same as the period of the applied signal, so

$$\frac{2\pi}{\omega} = \frac{x_b}{v} = \frac{\lambda_g}{v} \tag{6}$$

or

$$\lambda_g = \frac{2\pi v}{\omega} = \frac{v}{f} \tag{7}$$

and

$$\beta = \frac{2\pi}{\lambda_g} = \frac{\omega}{v} = \frac{2\pi f}{v} \tag{8}$$

Velocity of propagation is the speed at which a signal is transmitted along a transmission line. This velocity of propagation depends upon such factors as the mode of transmission in the line and the values of the elements of its equivalent circuit. For normal lossless two-wire lines, such as coaxial, with air dielectric between the lines, and transmitting in its lowest order mode (the normal mode used for two-wire lines), the velocity of propagation is equal to the velocity of light. If the velocity of light is represented by v_c, then from Eq. 7 we have:

$$\lambda_g = \frac{v_c}{f} = \text{free space wavelength } \lambda \tag{9}$$

In normal coaxial transmission systems the loss is small, so R and G can be neglected in determining β. Hence, from Eq. 2, for an air dielectric coaxial line

$$j\beta = j\omega\sqrt{LC_a} \tag{10}$$

and from Eqs. 10 and 8

$$\beta = \omega\sqrt{LC_a} = \frac{2\pi f}{v_c}$$

or

$$v_c = \frac{1}{\sqrt{LC_a}} \text{ (air dielectric line)} \tag{11}$$

If there were a dielectric in the line having a dielectric constant k, then the capacity would be

$$C = kC_a$$

and

$$v = \frac{1}{\sqrt{LkC_a}} = \frac{v_c}{\sqrt{k}} \tag{12}$$

Also, from Eqs. 7 and 9,

$$\lambda_g = \frac{v}{f} = \frac{v_c}{f\sqrt{k}} = \frac{\lambda}{\sqrt{k}} \tag{13}$$

Infinitely long line. Some of the properties of an infinitely long line are discussed in the previous section. The infinitely long line, though never encountered in practice, may be used to develop an understanding of the basic concepts of microwave circuits. Refer to Fig. 2–4, in which a voltage E_1 was impressed on the sending end of an infinitely long line. At a distance x from the sending end, we have a voltage E_x given by

$$E_x = E_1 e^{-\gamma x} \tag{3A}$$

Also, we know from the definition of characteristic impedance that the impedance looking into the sending end of this infinitely long line is Z_0. Hence,

$$I_1 = \frac{E_1}{Z_0} \tag{14}$$

and

$$I_x = I_1 e^{-\gamma x} \tag{15}$$

and

$$\frac{E_x}{I_x} = \frac{E_1 e^{-\gamma x}}{I_1 e^{-\gamma x}} = Z_0 \tag{16}$$

Voltage and current at any position along an infinitely long line are related to the characteristic impedance and the propagation constant by Eq. 16.

In almost all practical cases the line can be considered lossless, so that $R = G = 0$ and (from Eqs. 1 and 2)

$$Z_0 = \sqrt{\frac{L}{C}} \quad \text{a purely real quantity}$$

and $\gamma = j\omega\sqrt{LC}$ a purely imaginary quantity (no attenuation).

Hence, in the lossless case,

$$E_x = E_1 e^{-j\beta x}$$
$$I_x = I_1 e^{-j\beta x}$$
$$\frac{E_x}{I_x} = Z_0 \quad \text{a pure resistance}$$

and power being transmitted from left to right in the line is

$$P = E_x I_x = E_1 I_1 = \frac{E_1^2}{Z_0} = I_1^2 Z_0 \tag{17}$$

These relations for the lossless case are shown in Fig. 2–6.

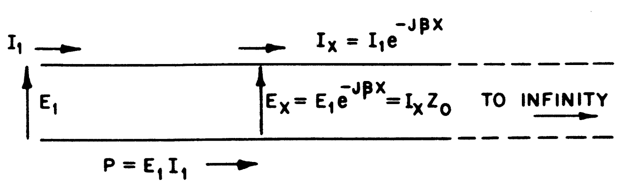


FIGURE 2–6 *Voltage, currents, and power in infinitely long transmission line.*

Matched finite line. With the concept of the infinite line firmly in mind, lines of finite length can be considered. Start with an infinitely long line and split it into two sections, as shown in Fig. 2–7. This action produces a line of finite length on the left terminated by a line infinitely long on the right. Insofar as the finite section on the left is concerned there would be no difference in the voltages and currents existing on it if the infinite section to the right could be replaced by the exact equivalent of its impedance. This substitution can be made. The input impedance of an infinitely long line is Z_0. If the section of infinite line in Fig. 2–7 is replaced by Z_0, none of the conditions existing on the left-hand section of finite length will be affected. If the line is considered to be lossless, the conditions shown in

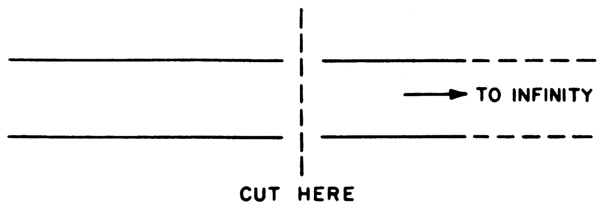


FIGURE 2–7 *Infinitely long line split into two sections.*

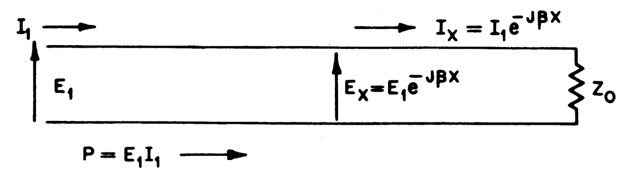


FIGURE 2–8 *Finite matched transmission line.*

Fig. 2–8 are obtained by placing Z_0 in the position of the infinitely long line section of Fig. 2–7. This procedure is known as matched operation.

Finite line with arbitrary load. In the above section we made the transition from an infinitely long line to one of finite length. This change produced a rather special case in which the finite line section is terminated in its characteristic impedance. Of considerably more practical importance is a finite line with an arbitrary load. Note that in the special case of a finite line terminated in its characteristic impedance, the ratio of voltage to current at any point in the line was equal to Z_0, and that all the power the generator delivered into the line was absorbed in the load. Furthermore, in this special case the voltages and currents were being propagated only from the generator to the load. In other words, there was no signal traveling from right to left.

There is only one more condition to be added to permit handling of arbitrary loads. Since the forward wave at all points on the transmission line maintains the ratio between E_i and I_i as Z_0, the wave meeting an impedance not equal to Z_0 can no longer support the same ratio between E_i and I_i. However, the addition of a signal traveling from right to left permits the resulting voltage and current at that point to have the proper ratio, which is the impedance of the load. There are then two current and voltage waves on the line. They are defined as follows:

Incident wave. The voltage or current wave being propagated from the generator towards the load. These waves will be denoted by the subscript *i*.
Reflected wave. The voltage or current wave being propagated from the load towards the generator. These waves will be denoted by the subscript *r*.

The general case of a finite line with arbitrary load is shown in Fig. 2–9.

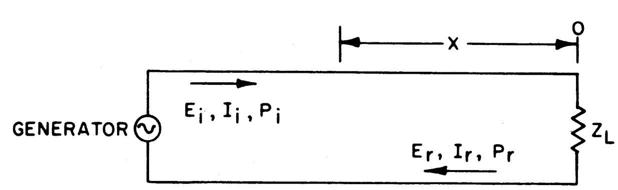


FIGURE 2–9 *Finite line terminated in arbitrary load.*

Since the relation between the reflected and incident waves is completely determined by the load and the line, we will measure distance from right to left with zero distance at the load. Voltages and currents at the load will be denoted by the subscript 0, and voltages and currents at other points will be denoted by the subscript x.

Reflection coefficient. There is a definite relation between the reflected and the incident waves. It is determined by the load and characteristic impedance of the line. Called voltage reflection coefficient ρ_v and current reflection coefficient ρ_i, the relations are defined as follows:

$$\frac{E_{r0}}{E_{i0}} = \frac{Z_L - Z_0}{Z_L + Z_0} = \rho_v \text{ voltage reflection coefficient} \tag{18}$$

$$\frac{I_{r0}}{I_{i0}} = \frac{Z_0 - Z_L}{Z_0 + Z_L} = \rho_i = -\rho_v \text{ current reflection coefficient} \tag{19}$$

Furthermore, at any point in the line the incident and reflected voltages and currents add vectorially to give the total voltage and current. Thus, at the load or position 0 the total or load voltage and current is given by

$$E_L = E_{r0} + E_{i0}$$
$$I_L = I_{r0} + I_{i0}$$

Voltages and currents along a line. If we know the voltages and currents at the load or position zero, we may apply the basic concepts of propagation constant and characteristic impedance to see how the voltage, currents, and impedances vary at other positions in the line. At a distance x from the load toward the generator, we have a voltage E_x composed of the incident and reflected wave given by

$$\begin{aligned} E_x &= E_{rx} + E_{ix} = E_{i0}e^{\gamma x} + E_{r0}e^{-\gamma x} \\ &= E_{i0}e^{\gamma x} + \rho_v E_{i0}e^{-\gamma x} \\ &= E_{i0}e^{\gamma x}(1 + \rho_v e^{-2\gamma x}) \end{aligned} \tag{20}$$

Similarly, I_x is given by

$$\begin{aligned} I_x &= I_{rx} + I_{ix} = I_{i0}e^{\gamma x} + I_{r0}e^{-\gamma x} \\ &= I_{i0}e^{\gamma x} + \rho_i I_{i0}e^{-\gamma x} \\ &= I_{i0}e^{\gamma x}(1 + \rho_i e^{-2\gamma x}) \\ &= I_{i0}e^{\gamma x}(1 - \rho_v e^{-2\gamma x}) \end{aligned} \tag{21}$$

Don't let the sign of the exponent used for the propagation constant confuse you. Just remember that when we move along a line in the direction of propagation of a particular wave, the wave must get smaller (if the line has loss) and suffer a lag in phase. When we move in a direction opposite to the direction of propagation, the wave must get larger and have a leading component of phase.

Standing-wave ratios. Now let us look at the voltage at an arbitrary position x along the line. The value of this voltage, given in Eq. 20, becomes, in the lossless case, ($\alpha = 0$),

$$E_x = E_{i0}e^{j\beta x}(1 + \rho_v e^{-2j\beta x}) \tag{20A}$$

The term $E_{i0}e^{j\beta x}$ is a multiplying term and denotes signal level. The amplitude variation with position is determined by the term within the parentheses, so this term should be examined more closely. The term in parentheses can be represented vectorially, as shown in Fig. 2–10. As position x is varied, the small vector

$\rho_v e^{-j2\beta x}$ will rotate around the end of the unit vector, and the distance from 0 to this circle will represent the magnitude and phase of the resultant vector. The maximum of this resultant vector will be $1 + |\rho_v|$ and the minimum will be $1 - |\rho_v|$.

Hence
$$E_{\max} = E_{i0}e^{j\beta x}(1 + |\rho_v|)$$
$$E_{\min} = E_{i0}e^{j\beta x}(1 + |\rho_v|)$$

The relation between maximum and minimum voltage is called the voltage standing-wave ratio σ, which is

$$\sigma = \frac{E_{\max}}{E_{\min}} = \frac{1 + |\rho_v|}{1 - |\rho_v|} \tag{22}$$

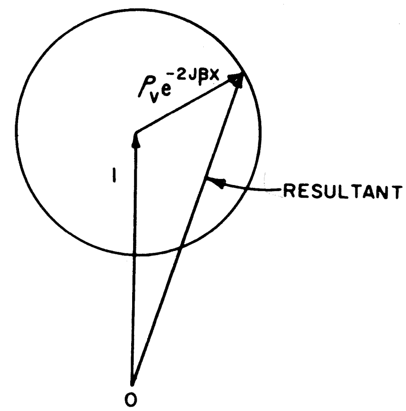


FIGURE 2–10 *Vector relations of voltage in transmission line.*

Also, $2\beta x$ has to change by 180 deg or π rad between a maximum and minimum. The distance between maximum and minimum can be determined from Eq. 8, as follows:

$$2\beta(X_{\max} - X_{\min}) = \pi$$

But
$$\beta = \frac{2\pi}{\lambda_g}$$

Hence
$$\frac{4\pi}{\lambda_g}(X_{\max} - X_{\min}) = \pi$$

or
$$X_{\max} - X_{\min} = \frac{\lambda_g}{4} \tag{23}$$

A similar process can be applied to the current to show that the current swr is the same as the voltage swr. However, since there is a 180-deg phase difference between the current and voltage swr ($\rho_i = -\rho_v$), the current maximum will occur where the voltage is a minimum, and vice versa. Typical current and voltage patterns are shown in Fig. 2–11.

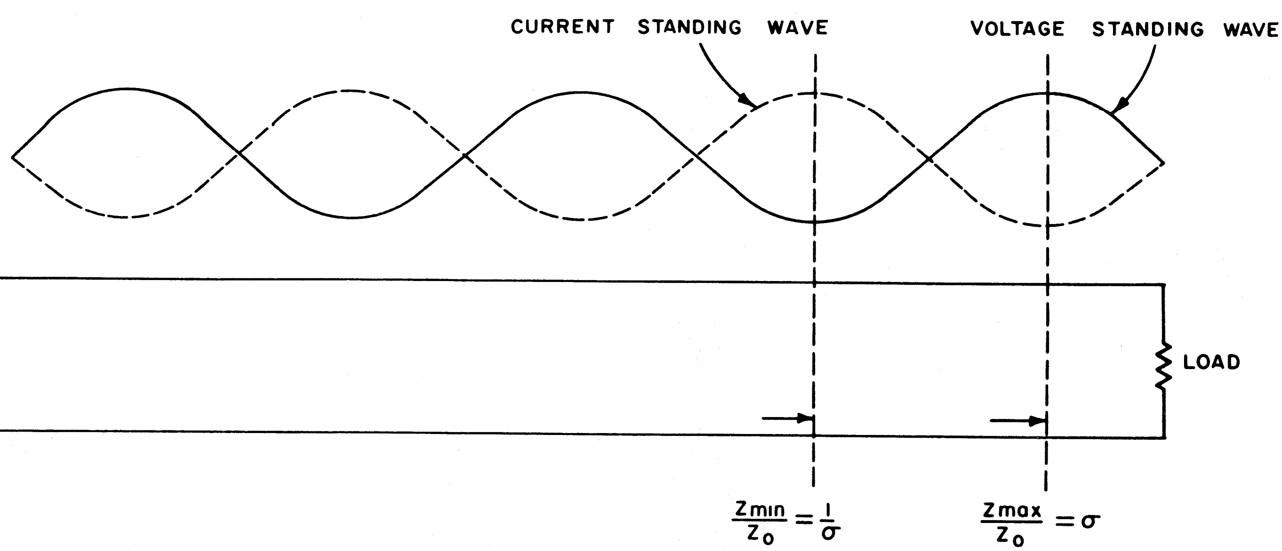


FIGURE 2–11 *Current and voltage standing wave on transmission line.*

Impedance at voltage maximum and minimum. At a voltage maximum we have

$$E_{\max} = E_{i0}e^{j\beta x}(1 + |\rho_v|)$$
$$I_{min} = I_{i0}e^{j\beta x}(1 - |\rho_v|)$$

or the line $Z_{\max}$ is given by

$$Z_{\max} = \frac{E_{\max}}{I_{\min}} = \frac{E_{i0}}{I_{i0}} \frac{1 + |\rho_v|}{1 - |\rho_v|} = Z_0\,\sigma \tag{24}$$

and, similarly, at a voltage minimum

$$E_{\min} = E_{i0}e^{j\beta x}(1 - |\rho_v|)$$
$$I_{\max} = I_{i0}e^{j\beta x}(1 + |\rho_v|) \tag{25}$$
$$Z_{\min} = \frac{Z_0}{\sigma}$$

Note that in the last two special cases, and with a lossless transmission line, the impedances are purely resistive.

Impedance at any point along a line. The only other factor to be determined is the impedance at any point a distance x from the load. Voltage and current at a point x (from Eqs. 20 and 21) are

$$E_x = E_{i0}e^{\gamma x} + \rho_v E_{i0}e^{-\gamma x} \tag{20B}$$
$$I_x = I_{i0}e^{\gamma x} - \rho_v I_{i0}e^{-\gamma x} \tag{21A}$$

Hence,

$$\begin{aligned} Z_x &= \frac{E_{i0}}{I_{i0}} \frac{e^{\gamma x} + \rho_v e^{-\gamma x}}{e^{\gamma x} - \rho_v e^{-\gamma x}} \\ &= Z_0 \frac{e^{\gamma x} + \dfrac{Z_L - Z_0}{Z_L + Z_0}\, e^{-\gamma x}}{e^{\gamma x} - \dfrac{Z_L - Z_0}{Z_L + Z_0}\, e^{-\gamma x}} \\ &= Z_0 \frac{(Z_L + Z_0)\, e^{\gamma x} + (Z_L - Z_0)\, e^{-\gamma x}}{(Z_L + Z_0)\, e^{\gamma x} - (Z_L - Z_0)\, e^{-\gamma x}} \\ &= Z_0 \frac{Z_L \cosh \gamma x + Z_0 \sinh \gamma x}{Z_0 \cosh \gamma x + Z_L \sinh \gamma x} \end{aligned} \tag{26}$$

where

$$\cosh \gamma x = \frac{e^{\gamma x} + e^{-\gamma x}}{2}$$
$$\sinh \gamma x = \frac{e^{\gamma x} - e^{-\gamma x}}{2}$$

For the lossless case, Eq. 26 reduces to

$$Z_x = Z_0 \frac{Z_L \cos \beta x + jZ_0 \sin \beta x}{Z_0 \cos \beta x + jZ_L \sin \beta x} \tag{27}$$

Impedances for short- and open-circuited lines. A lossless transmission line has been assumed in most of the work done so far. This assumption is usually justifiable in practice except in certain specific cases. One of the important exceptions occurs when the line is terminated in an open or short circuit. Since an open circuit is the same as a short circuit a quarter of a wavelength farther down the line, the two conditions can be handled in a similar fashion.

For a short-circuited line, the impedance Z_x is given by

$$\begin{aligned} Z_x &= Z_0 \tanh \gamma x \\ &= Z_0 \tanh (\alpha x + j\beta x) \\ &= Z_0 \frac{\tanh \alpha x + \tanh j\beta x}{1 + \tanh \alpha x \tanh j\beta x} \\ &= Z_0 \frac{\tanh \alpha x + j \tan \beta x}{1 + j \tanh \alpha x \tan \beta x} \end{aligned} \tag{28}$$

The voltage will be zero at a short, and hence will be a maximum when $x = \dfrac{\lambda_g}{4}$, so

$$\tan \beta x = \tan \frac{2\pi}{\lambda_g} \cdot \frac{\lambda_g}{4} = \tan \frac{\pi}{2} = \infty$$

and here

$$Z_{\lambda_g}/4 = \frac{Z_0}{\tanh \alpha x} \tag{29}$$

At a distance $\lambda_g/2$ from the short,

$$\tan \beta x = \tan \frac{2\pi}{\lambda_g} \cdot \frac{\lambda_g}{2} = \tan \pi = 0$$

and $Z_{\lambda_g}/2 = Z_0 \tanh \alpha x$ (30)

Equations 29 and 30 serve as the basis of a very useful way for measuring attenuation of transmission lines. The term "tanh αx" (and thus attenuation per unit length) can be determined by measuring impedance at ¼ wavelength from the short. The measured value is then substituted in Eq. 30. This procedure provides very accurate determination of small attenuation values.

Normalization. The process of normalization refers all transmission-line equations to unity characteristic impedance in order to simplify equations so that they are the same for a line of any characteristic impedance. Various engineering aids, such as Smith charts, are much easier to use if impedances are normalized. True impedance values are obtained by multiplying the normalized answers by the characteristic impedances.

Note that, in the various equations in which impedance has appeared, we could have divided each impedance by Z_0. Had we done so, each impedance would have been expressed as a ratio of that impedance to Z_0. For instance, consider Eq. 18:

$$\rho_v = \frac{Z_L - Z_0}{Z_L + Z_0} = \frac{\dfrac{Z_L}{Z_0} - 1}{\dfrac{Z_L}{Z_0} + 1} \tag{18A}$$

or Eqs. 24 and 25, which can be rewritten:

$$\frac{Z_{\max}}{Z_0} = \sigma \tag{24A}$$

$$\frac{Z_{\min}}{Z_0} = \frac{1}{\sigma} \tag{25A}$$

or Eq. 26, which can also be rewritten:

$$\frac{Z_x}{Z_0} = \frac{\dfrac{Z_L}{Z_0} \cosh \gamma x + \sinh \gamma x}{\cosh \gamma x + \dfrac{Z_L}{Z_0} \sinh \gamma x} \tag{26A}$$

In each case, any particular impedance appears as the ratio of itself to Z_0. These ratios are called normalized impedances and are normally written as lower-case "z." If we use this notation, the preceding equations become

$$\rho_v = \frac{z_L - 1}{z_L + 1} \tag{31}$$

$$z_{\max} = \sigma \tag{32}$$

$$z_{\min} = \frac{1}{\sigma} \tag{33}$$

$$z_x = \frac{z_L \cosh \gamma x + \sinh \gamma x}{\cosh \gamma x + z_L \sinh \gamma x} \tag{34}$$

Characteristic impedance does not have to be known in many microwave measurements. Consequently, the concept of normalized impedance is particularly useful when one is working with waveguide, which does not have the well-defined concept of impedance that coaxial systems have.

Power relations. Impedance relations in a transmission system are primarily important because of their effect on transmitted power. If the reflection coefficient for a particular system is known, the percentage of power transmitted to the load can be determined.

Equation 17 showed that the power being transmitted by the incident wave from the generator towards the load is given by P_i, where

$$P_i = E_i I_i = \frac{E_i^2}{Z_0} \tag{35}$$

Furthermore, the power being transmitted by the reflected wave from the load towards the generator is given by

$$P_r = E_r I_r = \frac{E_r^2}{Z_0} \tag{36}$$

Hence, power transmitted to the load is

$$P_L = P_i - P_r \tag{37}$$

$$= \frac{E_i^2}{Z_0} - \frac{E_r^2}{Z_0}$$

$$\frac{P_L}{P_i} = \frac{E_i^2 - E_r^2}{E_i^2} = \frac{1 - |\rho_v|^2}{1} = \frac{4\sigma}{(1+\sigma)^2} \tag{38}$$

$$\frac{P_r}{P_i} = \frac{E_r^2}{E_i^2} = |\rho_v|^2 = \left(\frac{1-\sigma}{1+\sigma}\right)^2 \tag{39}$$

2-2 Waveguide

The term waveguide denotes a hollow conducting tube used for the transmission of electromagnetic waves. Coaxial systems have two conductors. The basic difference between coaxial lines and waveguides is the mode in which they propagate energy. There is an infinite number of modes by which energy can be propagated down a transmission system. These various modes of propagation in waveguide are all characterized by having a cutoff frequency below which they cannot propagate energy. The higher the order of the mode, the higher is this cutoff frequency. All waveguide transmission systems act as a high-pass filter does.

Modes in transmission systems are described as TE (Transverse Electric Field) or TM (Transverse Magnetic Field). A TE mode has no component of *electrical* field in the direction of propagation, whereas a TM mode has no component of *magnetic* field in the direction of propagation.

A coaxial line or any two-wire system also has the property of being able to propagate in a mode having neither electric nor magnetic field components in the direction of propagation. In this mode, called TEM, the low-frequency cutoff is at 0 frequency. In other words, all frequencies down to d-c can be transmitted by a two-conductor system operating in the TEM mode. The field configurations in a coaxial system are shown in Fig. 2–12. The field configurations in a rectangular guide transmitting in the commonly used TE_{10} mode are shown in Fig. 2–13.

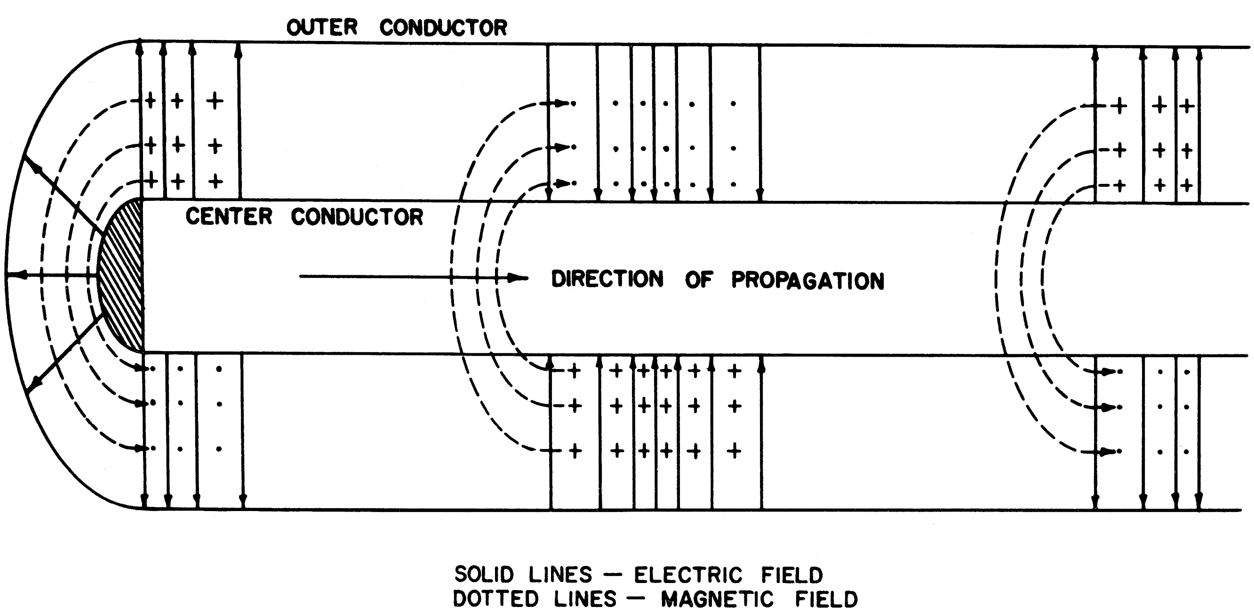


FIGURE 2–12 *Fields in a coaxial line TEM mode.*

In all types of transmission systems, it is desirable to select sizes so that only one mode of propagation is possible. In other words, the physical size of the guide is related to the frequency band under consideration. Because of the possibility of higher-order modes, it has become common to operate waveguides only over approximately 1.5 to 1 frequency range. By properly selecting this frequency range, it is possible to operate far enough from cutoff so that the guide parameters do not vary too rapidly and to avoid also the frequency region where other modes are possible.

One of the major differences between a waveguide and a two-wire transmission system is the existence of a cutoff frequency in the waveguide, below which energy cannot propagate. This cutoff frequency f_c, or cutoff wavelength λ_c, for the commonly used rectangular waveguide operating in the TE_{10} mode (as shown in Fig. 2–13) is given by

$$\lambda_c = 2a = \frac{v_c}{f_c} \tag{40}$$

where, as previously, $v_c =$ velocity of light.

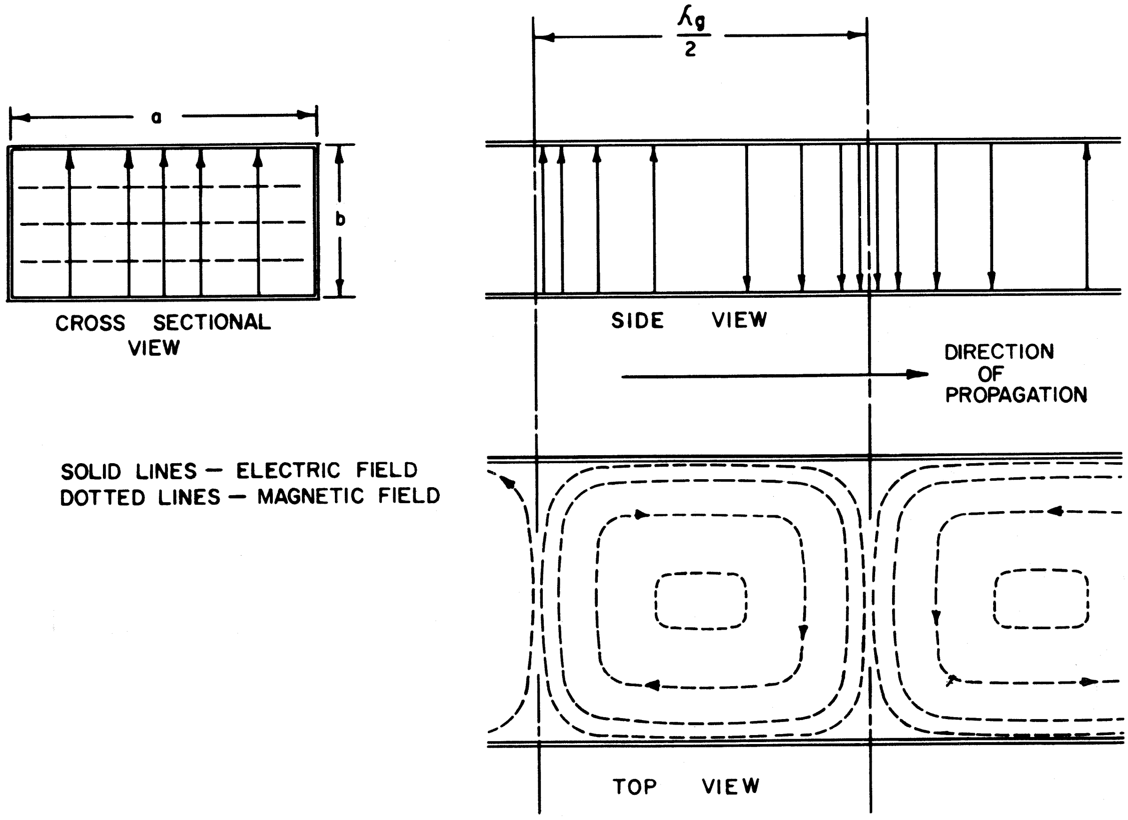


FIGURE 2–13 *Fields in a rectangular guide (TE_{10} mode).*

In any microwave transmission system, the distributed nature of the circuit must be taken into account. Thus the basic theory that has already been discussed is applicable to waveguides, but the presence of the low-frequency cutoff makes some modifications necessary. This low-frequency cutoff affects the velocity of propagation and the determination of the characteristic impedance.

For any coaxial system the velocity of propagation is constant for all frequencies. In waveguide systems the velocity of propagation changes with frequency. For normal waveguides with no dielectric inside the guide, the velocity of propagation v is given by

$$v = v_c \frac{\lambda_c}{\sqrt{\lambda_c^2 - \lambda_0^2}} \tag{41}$$

A more common notation is

$$v = \frac{v_c}{\sqrt{1 - \left(\frac{f_c}{f_0}\right)^2}} \tag{42}$$

Then, using Eqs. 8 and 9, we find

$$\lambda_g = \frac{v}{f} = \frac{v_c}{f} \cdot \frac{\lambda_c}{\sqrt{\lambda_c^2 - \lambda_0^2}} = \frac{\lambda_0}{\sqrt{1 - \left(\frac{f_c}{f_0}\right)^2}} \tag{43}$$

where

$\lambda_g =$ guide wavelength.

λ_c and $f_c =$ cutoff wavelength and frequency.

λ_0 and $f_0 =$ free space wavelength and frequency.

Now

$$\beta = \frac{2\pi}{\lambda_g} = \frac{2\pi}{\lambda_0} \sqrt{1 - \left(\frac{f_c}{f_0}\right)^2} \tag{44}$$

Characteristic impedance cannot be determined as easily for waveguide as for a coaxial line, because there are no unique currents and voltages. But this is not really a problem, since the process of normalization eliminates characteristic impedance as a requirement for calculations. The basic quantities for waveguide work are reflection coefficient, standing-wave ratio, and propagation constant. From these, the normalized impedance at any point can be determined, and the complete waveguide system can be described in terms of its performance and characteristics.

For coaxial systems the reflection coefficient was defined as the ratio of the reflected signal to the incident signal; it is the same for waveguide. Starting from this fact, and using the propagation constant as defined in Eq. 44, we find that all the equations developed earlier are applicable to waveguide.

SECTION THREE

Microwave Measurements

3-1 Microwave Equipment

Microwave parameters such as power, impedance, noise figure, attenuation, and frequency can be measured with commercially available test and measurement instruments. Also available is a wide variety of accessories and waveguide fittings, such as slotted lines, detectors, mounts, attenuators, phase shifters, directional couplers, etc. Instrumentation and accessory equipment manufactured by Hewlett-Packard is available for making all normal measurements in all microwave frequency bands to 40 gc (kmc). The specific equipment types mentioned in this publication are fully described in hp catalogs and technical literature.

Equipment designations. Each item of hp equipment is identified by a model number. For waveguide components, the model numbers are normally prefixed by a letter which designates the waveguide size and frequency band. Each hp waveguide instrument for a given band will have the same prefix to its model number. All instruments with a given prefix are compatible.

Nine band-designation prefixes are used; they are outlined in Table 3–1 on page 24. For example, a hp 370 fixed waveguide attenuator designed for use with $3 \times 1\frac{1}{2}$ in. waveguide is designated Model S370. The same instrument for use with 0.702×0.391 in. waveguide is designated Model P370.

Many Hewlett-Packard instruments also have suffix letters as part of the model number. In many cases an "A" suffix identifies the original design of an instrument, whereas "B," "C," and other letters indicate a revised, modified, or special version of the basic model. However,

TABLE 3-1

BAND	FITS WAVEGUIDE SIZE (In.)	FREQUENCY RANGE
S	3 × 1½	2.60 to 3.95 gc
G	2 × 1	3.95 to 5.85 gc
C	1.718 × 0.923	4.90 to 7.05 gc
J	1½ × ¾	5.30 to 8.20 gc
H	1¼ × ⅝	7.05 to 10.00 gc
X	1 × ½	8.20 to 12.40 gc
M	0.850 × 0.475	10.00 to 15.00 gc
P	0.702 × 0.391	12.40 to 18.00 gc
N	0.590 × 0.335	15.00 to 22.00 gc
K	0.500 × 0.250	18.00 to 26.50 gc
R	0.360 × 0.220	26.50 to 40.00 gc

in the case of certain hp microwave items, the *suffix* letter indicates specific attenuation or coupling factors. Six designator letters are used:

"A"	3 db	"D"	20 db
"B"	6 db	"E"	30 db
"C"	10 db	"F"	40 db

For example, the 20 db coupling version of the hp Model 750 cross-guide coupler is designated the Model 750D. In addition, the model of the 750 built for 1 × ½ in. waveguide systems will, of course, have the size prefix designator "X." Therefore, the complete model number of a 750 series coupler with 20 db coupling for use with 1 × ½ in. equipment is the hp X750D cross-guide coupler.

Flanges. All hp waveguide equipment items have plain cover flanges which mate with standard UG-( ) flanges. Model 290 cover-to-choke flange adapters may be used to connect individual instruments to a choke flange system.

Types of measurements. In general, the measurements made in research, design, test, and checkout of microwave equipment can be divided into five types: frequency, attenuation, impedance, power, and noise figure.

3-2 Frequency Measurement

Microwave frequency can be measured by a number of different mechanical and electronic techniques. The mechanical devices commonly use such elements as resonant cavities and slotted lines, both of which depend on physical dimensions for their operation and accuracy. The electronic devices are primarily counters and high-frequency heterodyne systems which compare harmonics of a known lower frequency with the unknown microwave frequency. Although electronic frequency-measuring systems are usually more expensive and complex than the physical methods of measurement, the electronic systems are capable of considerably higher accuracy.

Resonant reactions. Resonant devices range all the way from lumped constant L-C circuits, calibrated to be resonant at various frequencies in their band, to microwave cavities, which are also resonant at specific frequencies in their band. In general, these resonant circuits are coupled into the circuit under test; as the frequency passes through the resonant point, a reaction is evident in the circuit.

Above 1000 megacycles, mechanical frequency meters use sections of coaxial line or circular waveguide that are resonant at known calibrated frequencies. The circuit reaction caused by these resonant devices is used for the tuning indication. The most popular frequency meter in use is probably the cavity type, since the tolerances are quite easy to hold to very high accuracy, and the instrument is straightforward, simple, and rugged.

The reaction cavity meter to be studied in this book depends on the resonance at a particular frequency within the cavity. This cavity is loosely coupled to a section of straight waveguide through a small coupling slot in the guide. If the microwave power in the main line is at the proper frequency to resonate the cavity, some energy is coupled off into the cavity, and there is a slight reduction of transmitted power down the main line of the meter. One advantage in the reaction-type meter is that when the meter is operated slightly off its tuned frequency, the effect caused by the slot is negligible; thus, it may be left in the circuit at all times.

Most reaction meters are made direct-reading in frequency for convenient measurement and have high effective Q's so that good accuracy may be obtained. For instance, the hp 532B frequency meter has a Q of about 5000 and gives a reaction indication approximately 2 mc wide at 10 gc. The basic accuracy of the instrument is $\pm$ 0.1 per cent, i.e., an accuracy of one part in one thousand, or better.

The resonant frequency of a cavity wavemeter is determined primarily by its physical configuration and the dielectric constant of the medium inside the cavity. Consequently, most wavemeters are commonly affected by temperature changes (which cause differential expansions in the cavity) and humidity (which causes a slight change of the dielectric constant inside the cavity). Both of these effects are minimized in more expensive precision wavemeters by the use of sealed cavities and special temperature-compensating materials. Temperature and humidity calibration or correction charts may also be used to improve accuracy.

Null measurement technique. A less frequently used mechanical frequency-measuring technique is with a slotted line. This technique depends on the fact that a swr set up in a transmission line produces nulls (minimums) every ½ wavelength. If these nulls are detected and the distance between them is measured, the frequency may be determined. It is necessary, however, to make a correction for the guide wavelength. Accuracies obtainable with this technique are usually limited to approximately 1 per cent, because the guide wavelength depends primarily upon very accurate waveguide dimensions. (Frequency measurement in coaxial systems is not dependent upon the physical dimensions of the system.)

3-3 Attenuation Measurement

Attenuation measurements are made by a number of different methods; among them are power ratio, and either rf or IF substitution.

In the power ratio method, the signal source is connected to a detector, and an output-indicator reading is obtained. The attenuating device is then inserted between the signal source and the detector, and a new output-indicator reading is obtained. The power ratio shown at the output indicator is a measure of the test-device attenuation. This measurement requires that the detection law of the detector be known over the complete frequency range of the measurement, and that reflection effects in the system be negligible both with and without the test device.

The type of detecting equipment used will depend on the range of the attenuation measurement. A power-monitoring combination, such as the hp 430C microwave power meter and a bolometer mount, will allow attenuation measurement over approximately 20 db. A wider range of attenuation measurement, up to 30 or 40 db, can be achieved with a detector mount employing a barretter and a hp 415B standing-wave indicator (high-sensitivity, tuned voltmeter). In the latter case, the signal source must be modulated, and the rf power level must be kept below 200 microwatts for square-law detector characteristics. The attenuation in decibels may be read directly from the Model 415B.

To eliminate effects of reflections between generator and attenuator, and attenuator and load, it is desirable to use attenuating pads. The pads should be well matched to the transmission system.

RF substitution depends on substituting an rf attenuator of known characteristic for the unknown. For instance, a signal-generator attenuator may be used. When this method is employed, the output of the signal generator is fed to the detector. The setting of the signal-generator attenuator is noted. The unknown attenuator is then inserted, and the output of the signal generator is adjusted to obtain the same reading as before. The difference between the signal-generator attenuator settings is the attenuation of the attenuator in db. Since the detector is always operated at the same level, detector law is no problem. The attenuator measurement may similarly be performed with a hp 382A precision attenuator and a signal source.

The IF substitution method offers the highest accuracy in attenuation measurements, since its substitution standard is a precise 30 mc cutoff attenuator. The power change caused by removing the unknown rf attenuator is replaced by change of the precision IF cutoff attenuator in the IF stage of the detecting microwave receiver.

The range of attenuation measurement depends upon the sensitivity of the detector. A video (square-law) detector, such as a crystal, is able to detect rf power down to about — 50 dbm. Microwave superheterodyne receivers, on the other hand, utilize linear detection and have common sensitivities down to — 80 or — 90 dbm.

Receiver techniques are more complex, since they require two signal generators (a local oscillator and the input signal source). There must be a high degree of frequency stability between the two generators so that the difference frequency remains within the IF amplifier pass band.

The homodyne method is a variation of a linear receiver wherein the local oscillator and the test signal are derived from the same source; the frequency problem of common receivers is thus eliminated.

The homodyne method permits the measuring of attenuation as high as 100 db. In this system a signal generator furnishes local oscillator power to a mixer and at the same time drives a traveling wave tube amplifier which is modulated to produce an offset frequency. The offset frequency is fed through the attenuator to be measured and combined with the local oscillator frequency. The difference frequency is amplified in a tuned amplifier and applied to an indicating meter. Because the traveling wave tube amplifier is serrodyne-modulated, the difference frequency from the mixer is constant, and a narrow-band tuned amplifier, such as hp 415B, may be used, even though the signal-generator frequency drifts.

3-4 Impedance Measurement

Of all the possible measurements to be made in design and production, probably the most important is the measurement of impedance. With distributed parameters, impedance varies with the position of measurement. Hence, all impedance measurements must be referred to some reference plane. Since impedance determines energy reflected by the load, information concerning a load can often be obtained by determining the magnitude of the reflection coefficient.

The value of the reflection coefficient can be determined by using a slotted section of transmission line and measuring the swr (the ratio of maximum to minimum voltage in the system feeding the load). In many cases, it is necessary to know only the magnitude of the reflection coefficient. This can be measured directly with a reflectometer by sampling the incident and reflected waves and obtaining their ratio, which is equal to the reflection coefficient. The reflectometer method will be explained after the discussion of the slotted line.

Slotted-line measurements. A typical setup for making slotted-line measurements is shown in Fig. 3–1. The transmission system contains the incident wave and a reflected wave which is proportional to the mismatch of the load. These two waves will alternately cancel and add, setting up a standing-wave pattern along the line. By inserting a probe into the slotted section and sliding it along the line, the resultant voltage pattern can be measured. The usual practice is to amplitude-modulate the signal source and to use a crystal or bolometer to detect the rf at the probe. The detected output of the probe is connected to a high-sensitivity, tuned voltmeter, such as the hp 415B standing-wave indicator. If this procedure is used, the swr and the position of maximum and minimum of the load can be determined. The load is then replaced by a short circuit, and the shift of the minimum is recorded. By entering this information on a Smith chart, the measured impedance can be transformed back to the point of interest. In this way, one can determine the value of the impedance and the reflection coefficient in magnitude and phase.

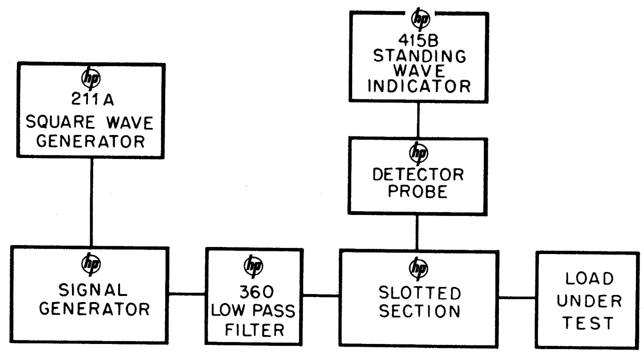


FIGURE 3–1 *Typical setup for impedance measurements.*

Slotted-line techniques. In measuring with slotted lines, there are several places where errors may occur. A proper operating technique will eliminate or minimize these errors. Among the sources of error are probe loading, generator mismatch, detector characteristics, harmonics, fm, and other spurious signals

Harmonics and spurious signals can be minimized by use of low-pass filters, such as the hp 360 and 362A series. Of special importance is the fact that modulation using very short pulses or poor-quality square waves should not be attempted. When klystrons are modulated in such a manner, the resulting fm tends to obscure the nulls of the standing waves. To avoid fm, modulation of klystron signal sources should be by square wave.

Since the ratios of different voltage levels are being measured with slotted lines, it is essential that the detector follow the same law for all levels. If barretters are operated at levels less than 200 μw and crystals at power levels of less than 20 μw, the characteristics are essentially square law. It is for this condition that the hp 415B meter scale is calibrated. This condition will be adequately met in the setup shown in Fig. 3–1 (for swr of 10 to 1 or less) if the probe coupling is reduced to a point where the minimum is 5 to 10 db above the system noise level.

The sampling probe will extract some power from the line to supply the indicating devices, and, in addition, the probe itself will set up reflections in the line. Both errors become greater as the probe insertion is increased. It is, therefore, important in a slotted-line measurement to keep the probe penetration at a minimum.

The power extraction by the probe can be explained by considering the probe as admittance shunting the line. This admittance is kept small by coupling as loosely as possible (small penetrations) and by using a high-sensitivity detector in conjunction with a source output of 1 mw or more. If the coupling between the probe and the line is not small, shunt admittance introduced by the probe will cause the measured swr to be lower than the true swr (as shown in Fig. 3–2) and will shift both the maximum and the minimum from their natural position. An exception to this minimum-penetration rule occurs when it is desired to examine in detail the minimum point on the swr pattern. For this work a greater probe penetration can be tolerated, because the voltage minimum corresponds to the lowest impedance point on the line.

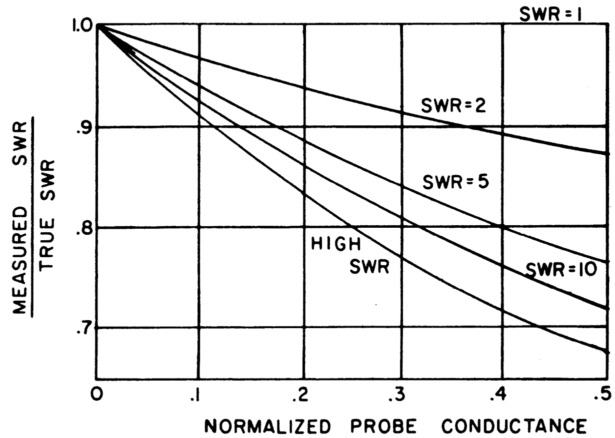


FIGURE 3–2 *Effect of probe penetration on measured swr.*

In addition to extracting power from the line, the penetration of the sampling probe into the slotted section gives rise to reflections from the probe itself. These reflections travel back towards the generator. If the generator is mismatched, the reflections are re-reflected. When the probe is moved under these conditions, the phase of the reflection is changed and errors result. However, reflections from the generator are a second-order effect, important only when one is measuring low swr (2 or less). In this case, a moderately good match between the generator and load is desirable. In general, the match of an hp signal generator is sufficient for this purpose, if the cables and connectors do not introduce additional reflection. However, when klystrons are used directly to feed a waveguide network, the match is poor. Therefore, the klystron should always be followed by a pad or an isolator.

Various methods of measuring swr have specific advantages for different swr ranges. Straightforward measurement of swr by conventional methods is generally preferred for swr in the range of 10 to 1 or less. But, when the swr is high, coupling to the probe must also be high in order to obtain readings at the minimum, and deformation of pattern may result when the maximum is measured. There is also a possibility of error caused by a change in detector characteristics because of rf level changes.

To measure swr greater than 10 to 1 within 1 per cent accuracy, the twice-minimum-power (or double-minimum-power) method is recommended. Here it is necessary only to establish the electrical distance between the points that are twice the amplitude of the minimum. The swr can be obtained by substituting this distance into the following expression, as shown in Fig. 3–3:

$$\sigma_L = \frac{\lambda_g}{\pi \Delta x}$$

σ_L = swr of load.
λ_g = guide wavelength (cm).
Δx = distance between "twice-minimum-power" points (cm).

The value referred to in this method is the twice-power value. Therefore, if a linear voltage indicator is used with a square-law detector, the voltage indication of the twice-power point will be twice that of the minimum. If a standing-wave indicator (calibrated for use on a square-law detector, such as the hp 415B or a linear receiver) is used, the voltage ratio of the two readings will be 1.4 to 1 (a 3 db difference in reading).

Reflectometer measurements. The reflectometer is the most useful impedance-measuring technique for fast, comprehensive production measurements. The reflectometer indicates magnitude of impedance, but it does not provide phase information as a slotted line measurement does. However, in the typical production situation a swr measurement alone is quite adequate, and phase information is not needed.

A typical reflectometer setup is shown in Fig. 3–4. This arrangement determines the magnitude of the reflection coefficient by using two directional couplers to sample the incident power to a load and the reflected power. The couplers drive detectors which are connected to a 1000 cycle ratio meter (hp 416A), where a ratio measurement is made. Since the hp 416A is calibrated for square-law detectors, the resultant ratio of the two sampled powers is indicated directly as reflection coefficient on a front panel meter. (The 416A also provides a d-c output to an X-Y recorder for making permanent records.) For best results in reflectometer operation, the input power should be kept to a low level by means of input attenuators so that the power at the forward detector is on the order of — 20 dbm. At the reverse detector it should be on the order of — 10 dbm at the calibration point. This power level will more nearly insure square-law operation of the crystal.

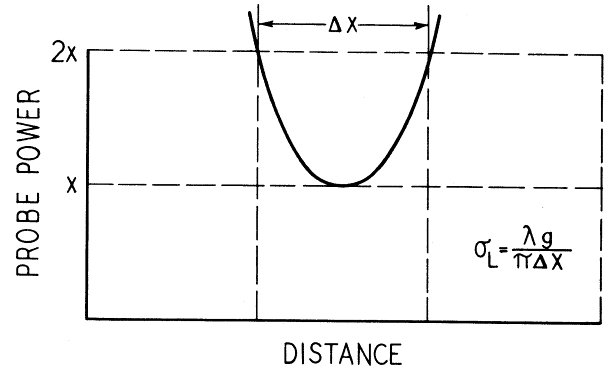


FIGURE 3–3 *Twice minimum power method for measuring swr.*

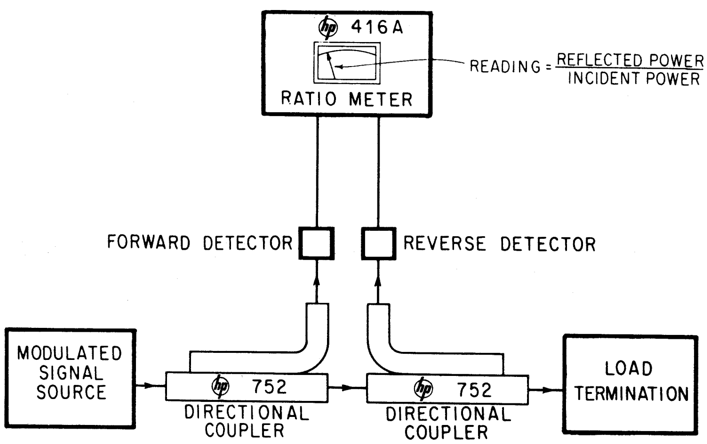


FIGURE 3–4 *Typical relectometer setup.*

The reflectometer method is most practical for measuring reflection coefficients up to approximately 0.5 (swr 3.0). When the reflectometer is used with swept-frequency techniques and is calibrated with a fixed short circuit at 100 per cent reflection, accuracies of approximately ± 0.02 may be obtained for reflection coefficients of 0.1 (swr 1.22). For reflection coefficients of 0.4 (swr 2.3), accuracies of approximately ± 0.04 may be obtained. The potential accuracy of the reflectometer, however, is greatest at low swr, when it is used at a fixed frequency. A rather simple calibration procedure, using a slide-screw tuner, a moving load, and a sliding short circuit, cancels out ambiguity caused by the reverse-coupler directivity. Under ideal conditions, errors of less than ± 0.005 in reflection coefficient are attainable; this figure is equivalent to a slotted-line measurement in a line with a residual swr of 1.01.

Although electronically swept rf sources provide faster measurements, they are not an absolute necessity. Very satisfactory measurements may be made with any manually-tuned signal source and an X-Y recorder at the output of the ratiometer. It is necessary only to sweep manually through the entire range to get a plot of reflection coefficient.

Impedance measurements with vhf bridge. Below 500 mc, slotted sections become exceedingly long and other techniques for impedance measurements are more desirable. For these frequencies, the hp Model 803A vhf bridge is ideal (see Fig. 3–5). The vhf bridge provides a convenient means of measuring impedances, showing both magnitude and phase angle. The bridge is operated simply by tuning two controls until a sharp null is obtained. At the null, one dial reads unknown impedance in ohms and the other dial shows phase angle.

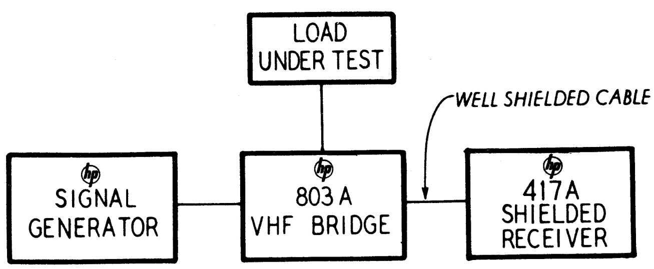


FIGURE 3–5 *Impedance measurements for frequencies below 500 mc.*

Because of the nulling nature of the measurement, the voltages measured are very small. Therefore, to avoid any effects from extraneous voltages, lines connected to the bridge should be adequately shielded. The signal source supplying the bridge should be capable of delivering several mw of power for a well-defined sharp null to be observed. The detecting equipment should have high sensitivity, as does the hp Model 417A vhf detector, which is designed primarily to be used with the hp Model 803A bridge.

The bridge is basically an unbalanced device, and in many cases it is desirable to measure balanced systems. Such measurement can be accomplished by the use of a balun, a simple form of which is shown in Fig. 3–6. A half-wavelength balun is equivalent to a 4 to 1 impedance transformer. Hence, impedances measured at the input of the balun should be multiplied by 4 in order that the actual impedance may be obtained.

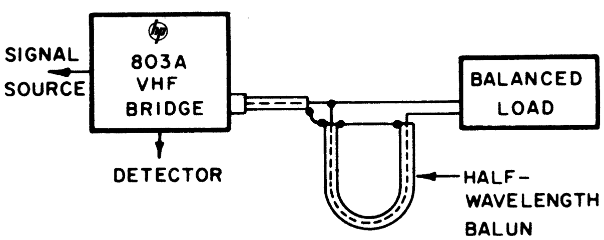


FIGURE 3–6 *Measurement setup using balun with balanced load.*

3-5 Power Measurement

In the microwave region, power measurements are considered to be more basic than current or voltage measurements, because power is invariant with position of measurement, whereas current and voltage (because of the distributed nature of the transmission system at these frequencies) are not.

In the range of 0.01 to 10 mw, power measurements are customarily made with a bolometer, which operates in a bridge circuit and changes rf energy into heat energy. This conversion causes the resistance of the bolometer to change, unbalancing the bridge. The audio power which is substituted to rebalance the bridge and keep the bolometer resistance constant is then measured. A typical bolometer arrangement is shown in Fig. 3–7.

In the range above 10 w, power measurements are generally made using calorimeter techniques. Either a dry calorimeter or a waterflow calorimeter is usually used.

Between these low- and high-power ranges, measurements can be made by using attenuators and low-power bolometers. However, these measurements are somewhat clumsy and inaccurate. The hp 434A calorimetric power meter makes direct, convenient, and accurate measurements in the 10 mw to 10 w range. This unique oil-flow calorimeter fills the need for a convenient measuring device having high accuracy and wide bandwidth.

Conventional bridge techniques. Bolometers used for microwave measurements are of two general types: barretters—metallic wire or film in which the temperature coefficient of resistance is positive, and thermistors—semiconductor material in which the temperature coefficient is negative. Both barretters and instrument fuses are used as positive temperature-coefficient bolometers. A barretter consists of a short length of very fine platinum wire suitably capsulated. A negative temperature-coefficient bolometer (thermistor) consists of a small bit of semiconductive material suspended between two fine wires.

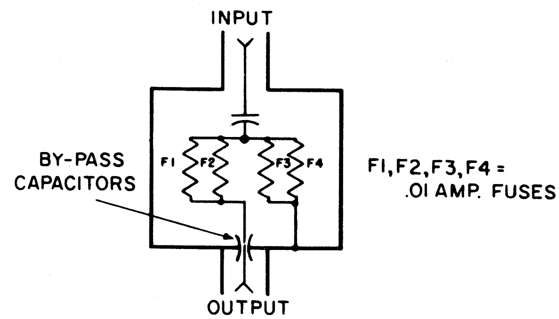


FIGURE 3–7 *Arrangement for using four instrument fuses in series parallel combination in* hp *476A bolometer mount.*

In general, barretters are delicate, and readily burned out by too much power. Even if the overload is insufficient to burn out a barretter, it may still increase the cold resistance to the point at which a self-balancing bridge meter cannot be zero set. Thermistors are much more rugged. Although they are rated at 25 mw maximum, they usually burn out at about 400 mw, and their characteristics change only slightly, if at all, upon overload.

The bolometer element is used in conjunction with a power meter, such as the hp Model 430C. This power meter is designed to operate with bolometer impedances of either 100 or 200 ohms.

The bolometer element itself must be mounted and well matched to the rf transmission system and to the power meter. The hp bolometer mounts feature low swr throughout their operating range and are available for coaxial and waveguide systems. Barretters are usually operated at 200 ohms, whereas thermistors usually operate at 100 or 200 ohm levels. Series-parallel combinations of the bolometer elements are used in hp coaxial mounts. The hp 476A bolometer mount, for example, uses four instrument fuses, each operating at 200 ohms and arranged to present 200 ohms to a microwave power meter but only 50 ohms to the rf energy.

The power measured by a bolometer mount also depends upon the relationship between the load and the source impedance. In order to obtain maximum available power, the load should present a conjugate match to source impedance. This match can be achieved by properly adjusting a double-stub tuner, an E-H tuner, or a slide-screw tuner. These tuners transform the magnitude and phase of the source impedance in order to conjugate-match it to the load impedance.

The hp 430C microwave power meter gives direct instantaneous readings of microwave power when it is connected with a suitable bolometer mount. The bias current necessary to bring the bolometer to the correct operating resistance is furnished by the 430C power meter. This power meter circuit includes a self-balancing bridge and an audio voltmeter to indicate the magnitude of the bridge amplifier output (Fig. 3–8). The self-balancing bridge uses the external bolometer element (a nonlinear resistor) as one of the bridge arms. A high-gain amplifier is connected across the bridge as a detector, and the output of the same amplifier is connected as the driving source for the bridge. Then, if there is sufficient gain, the circuit oscillates and audio power is furnished at an amplitude such that the bridge is almost balanced. When the rf power is applied to the element, the amplitude of oscillation decreases the amount necessary to maintain the element's resistance constant. This audio power decrease is equal to that power added by the rf source and can be read on the voltmeter, which is calibrated in power units.

hp bolometer mounts have been designed for both coaxial and waveguide systems at frequencies between 10 mc and 40.0 gc (kmc). These mounts are extremely simple to use, have low swr, and may be

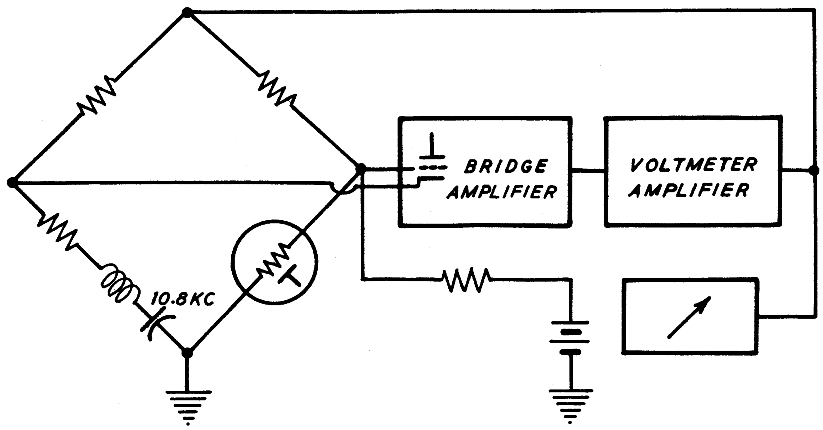


FIGURE 3–8 *Basic circuit of power meter.*

used with the hp 430C power meter to provide direct reading measurements. hp bolometer mounts may be classified according to the type of bolometer element employed—thermistor or barretter—and whether the mount is untuned (broadbanded) or tunable.

The hp untuned thermistor mounts are exceptionally broadbanded bolometers. Model 477B coaxial thermistor mount covers the frequency range of 10 mc to 10 gc, whereas the hp 487B thermistor mounts (waveguide series) are available from 2.6 to 40.0 gc. No tuning is required, and an extremely low swr is maintained throughout all frequency bands.

The Model 485B detector mount employs a single tuning control to match the applicable waveguide to a barretter power-detection element. In general, the swr is less than 1.25 over the rated frequency range when a barretter is used. This arrangement provides an excellent match to the rf line and very low mismatch losses.

The hp 476A universal bolometer mount is a fixed broadbanded bolometer in the frequency range from 10 to 1000 mc. The bolometer element consists of 8.25 ma fuses.

In general, square-wave- or pulse-modulated power can be measured accurately with a barretter, a fuse, or a thermistor, subject to certain limitations which depend upon the characteristics of the bolometer elements in conjunction with the bridge oscillator. However, in the hp 430C power meter, these limitations are not serious.

When barretters or fuses are used, precautions should be taken if the modulation frequency is below about 200 cps, since the heating time constant allows the element resistance to follow the modulation. For sine- and square-wave-modulated power, the meter reading will tend to increase at such low modulated frequencies. For use with thermistors, precautions should be taken for frequencies of less than 100 cps.

Furthermore, with barretters or fuses, care should be taken to avoid modulating frequencies approaching the bridge frequency (approximately 10.8 kc) or its submultiples. At pulse frequencies near submultiples of the bridge frequency, beats are produced which show on the meter. At modulation frequencies which are exact submultiples of the oscillator frequency, the oscillator may lock in with the modulation frequency, causing the meter pointer to dip to a low value. In either case, the effect can be avoided by changing the repetition frequency slightly. This solution can be used down to frequencies at least as low as 200 cps.

A tabulation of hp equipment to be used with the Model 430C power meter for a specific transmission system, frequency range, and power level is given in Fig. 3–9. Power levels greater than the highest range of the 430C power meter can be measured by attenuating the power (with pads or directional couplers) to the range of the Model 430C.

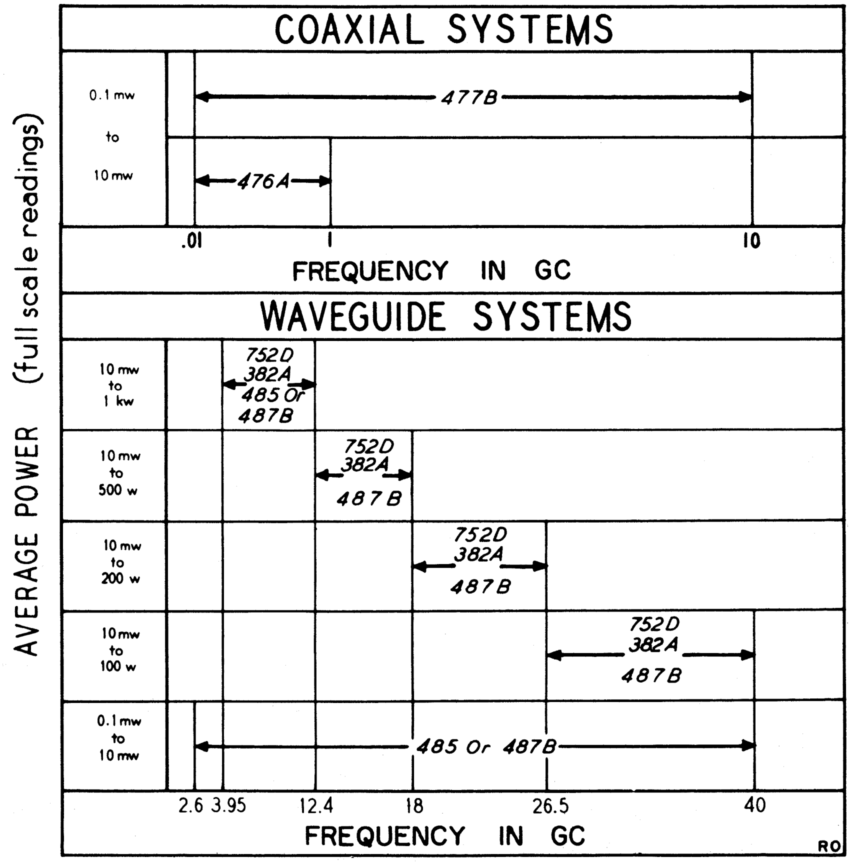


FIGURE 3–9 *hp equipment for use with hp 430C microwave power meter.*

New temperature-compensated power bridge. Conventional bolometer bridge techniques have a serious limitation in the lower-power sensitivity regions because of thermal drift in the mount itself. Since the bolometer is a temperature-sensitive element, power of all types, including ambient temperature change, causes a resistance change in the mount. In fact, typical power sensitivity of a thermistor mount to temperature change is such that a 0.005°C change is approximately equivalent to 1 microwatt. Such a high sensitivity to temperature seriously limits the low-power sensitivity unless special techniques are employed.

The new hp 431A temperature-compensated power meter represents a significant advance in stability and accuracy in power measurements. The power meter utilizes a dual bridge, temperature-compensated circuit arrangement which allows power measurements to be made down to a full-scale sensitivity of 10 microwatts. Operation is essentially drift free. Under laboratory conditions, for instance, long-term drift on the most sensitive range of 10 microwatts full scale is typically 1 or 2 microwatts per 4-hr period. This extreme stability provides a truly satisfying power measurement in the high-sensitivity region and opens a new area of convenience to power-meter users. Furthermore, power up to 10 mw may be measured directly on the same bridge. Time savings on zero setting alone are appreciable, and further time savings are made by providing that the zero setting be carried over for all power ranges. Previous conventional bridge techniques required that new zeroing be done whenever the power-range switch was turned to a different range.

Operation of the temperature-compensated power meter depends on the use of two identical thermistor bridges. Thermally sensitive elements, the thermistors, are located in close thermal proximity to each other, but one element is placed so that it absorbs rf power from a transmission line, whereas the other element is free from any rf influence. Special mount design and thermistor mounting procedures provide the nearly identical thermal environment, the coupling of rf power to one thermistor, and the shielding of the other.

An increase or decrease of ambient temperature tends to change the operating resistance of both thermistors. The unbalance, sensed by the temperature-compensating thermistor bridge, is amplified and, in turn, reduces the audio power applied to both bridges, keeping both in balance. Thus, so long as the temperature sensitivity of the two thermistors tracks with temperature and they are both maintained in close thermal proximity, temperature effects are essentially cancelled out. The rf power applied to the rf mount, however, reduces the audio power drive to both bridges, and, to maintain the temperature-compensating bridge in balance, a d-c power is supplied from the electronics of the power meter. It is this d-c power which is metered to indicate the rf power input. One feature of this bridging system is that both bolometer elements are maintained in a balanced condition, and, since they are identical thermally, there is a 1-to-1 translation between the rf power supplied to the rf bridge and the d-c rebalance power supplied to the compensating bridge. Thus, both elements are maintained at the same impedance over all power ranges. One of the advantages obtained from operating both bridges in a balanced condition is that 10 mw of rf power can be measured. The Model 431A features a "zero carry-over," which allows the power meter to be balanced on the lowest range of 10 microwatts; as the range is switched upward to 10 mw, no rezeroing need be done.

High accuracy is realized in the Model 431A by careful attention to switching resistors and bridge-determining resistors. In addition, terminals are provided on the rear of the unit for a d-c calibration input to calibrate the system with precise d-c standards for even higher accuracy. For better readability commensurate with the increased accuracy that can be obtained with precise d-c calibration, a recorder output is provided. A 3- or 4-digit digital d-c voltmeter may be connected to the recorder output and, in combination with precise d-c calibration, improves resolution and accuracy of power measurements. Model 431A provides an extremely convenient automatic balancing bridge with excellent readability for use in standards laboratories.

The special bridge-balancing arrangement in Model 431A achieves a new convenience in pulsed-power measurements. Conventional bridges respond to the audio drive voltage and mathematically convert this to indicate power. At low repetition rates the bridge tends to follow the rf modulation, with the meter responding to average voltage rather than to average power. Since Model 431A meters a d-c rebalance voltage proportional to power, true power averaging is obtained on amplitude-modulated rf waves.

10 mw to 10 w. The Model 434A calorimetric power meter automatically measures average power from 10 mw to 10 w. The instrument operates over the frequency range from d-c to 12.4 gc. The operator simply connects the source to the 434A and reads the power. Power above 10 w may be measured by reducing it to the range of the 434A with calibrated attenuators or directional couplers.

Model 434A is ideally suited for highly accurate measurements, because its over-all accuracy is 5 per cent, including rf efficiency and substitution error. Also, it allows direct measurement of intermediate powers and thus eliminates the error in the power-reducing attenuator, which is required in bolometer techniques.

Model 434A, shown simplified in Fig. 3–10, consists of a self-balancing bridge which has identical temperature-sensitive resistors (gauges) in two legs, an indicating meter, and two load resistors, one for the unknown input power and one for the comparison power. The input load resistor and one gauge are in close thermal proximity so that heat generated in the input load resistor heats the gauge and unbalances the bridge. The unbalance signal is amplified and applied to the comparison load resistor, which is in close thermal proximity to the other gauge so that the heat generated in the comparison load resistor is transferred to its gauge and nearly rebalances the bridge. The meter measures the power supplied to the comparison load to rebalance the bridge. The characteristics of the gauges are the same and the heat transfer characteristics from each load are the same, so the power dissipated in each load is the same, and the meter may be calibrated directly in input power.

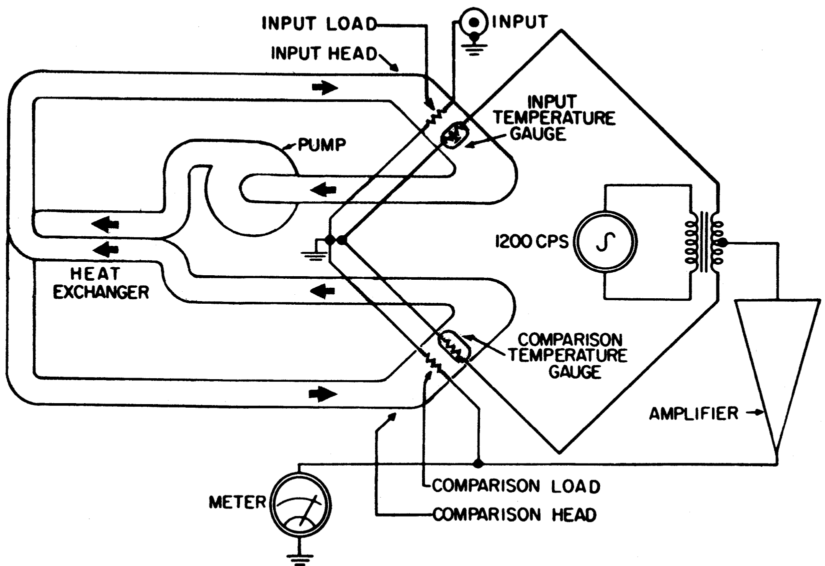


FIGURE 3–10 *Simplified diagram, hp 434A calorimetric power meter.*

To provide swift balancing, an efficient heat transfer from the loads to the temperature gauges is accomplished by immersing the components in an oil stream. This arrangement gives full-scale deflection in less than 5 sec.

The power measurement is accurate, because the flow rates through the two heads and the head characteristics are the same. To insure constant temperature, and to bring the streams to nearly the same temperature, the streams are passed through a parallel-flow heat exchanger just before they enter the heads. Identical flow rates are obtained by placing all the elements of the oil system in series, as shown in Fig. 3–10.

The accuracy of Model 434A is one of its unique attributes. Since the new power meter represents a most accurate method for measuring high-frequency power, the 434A may find much use as a laboratory standard power meter. Nominal accuracy is 5 per cent. However, higher accuracies can be achieved by employing techniques to minimize frequency and impedance mismatch effects. For example, accuracy can be improved by applying an efficiency correction to compensate for the internal power loss in the rf termination. It can also be improved by accurately matching the 434A to the source. For this purpose it is desirable that the power be carried in a waveguide rather than a coaxial cable. Use of waveguide not only reduces line loss, but also permits a waveguide slide-screw tuner to be used ahead of the waveguide-coaxial-cable transition at the instrument connector. Such waveguide tuners normally give less loss than coaxial tuners.

3-6 Noise Figure

In microwave communications, the weakest signal that can be detected is usually determined by the amount of noise added by the receiving system. Consequently, any decrease in the amount of noise generated in the receiving system will produce an increase in the output signal-to-noise ratio equivalent to a corresponding increase in received signal. From a standpoint of performance, an increase in the signal-to-noise ratio by reducing the amount of noise in the receiver is more economical than increasing the received-signal level by raising the power of the transmitter. For example, a decrease of 5 db in receiver noise is equivalent to increasing the transmitter power by 3:1.

The noise at the output of a receiver or an amplifier is the sum of the noise arising from the input termination (source) and the noise contributed by the receiver or amplifier itself. The noise factor is the ratio of the actual output noise power of the device to the noise power which would be available if the device were perfect and merely amplified the thermal noise of the input termination without contributing any noise of its own. The noise figure is the noise factor expressed in decibels.

The noise figure of a receiver may be measured by using a signal-generator input and an output-power (square-law) detector. However, this method is time consuming and has the added disadvantage that the effective power gain-bandwidth characteristics of the device must be determined. Moreover, the available signal power may be difficult to determine accurately at the low levels involved.

Automatic noise-figure measurements utilizing standard broadband noise sources which supply a noise spectrum of known power, flat with frequency, overcome the drawbacks of the signal-generator method. At intermediate and low radio frequencies, temperature-limited diodes are suitable as excess noise sources, whereas at microwave frequencies gas discharge tubes in suitable waveguide sections are both accurate and reliable. Hewlett-Packard noise-figure meters utilize the noise-source technique.

Automatic noise-figure measurements with hp noise-figure meters depend upon the periodic insertion of a known excess noise power at the input of the device under test. Subsequent detection of the noise power in later IF stages of the device results in a pulse train of two power levels. The power ratio of these two levels contains the desired noise-figure information. For instance, in the simplified diagram of Fig. 3–11, the various contributions of noise power to the output-pulse ratio are shown.

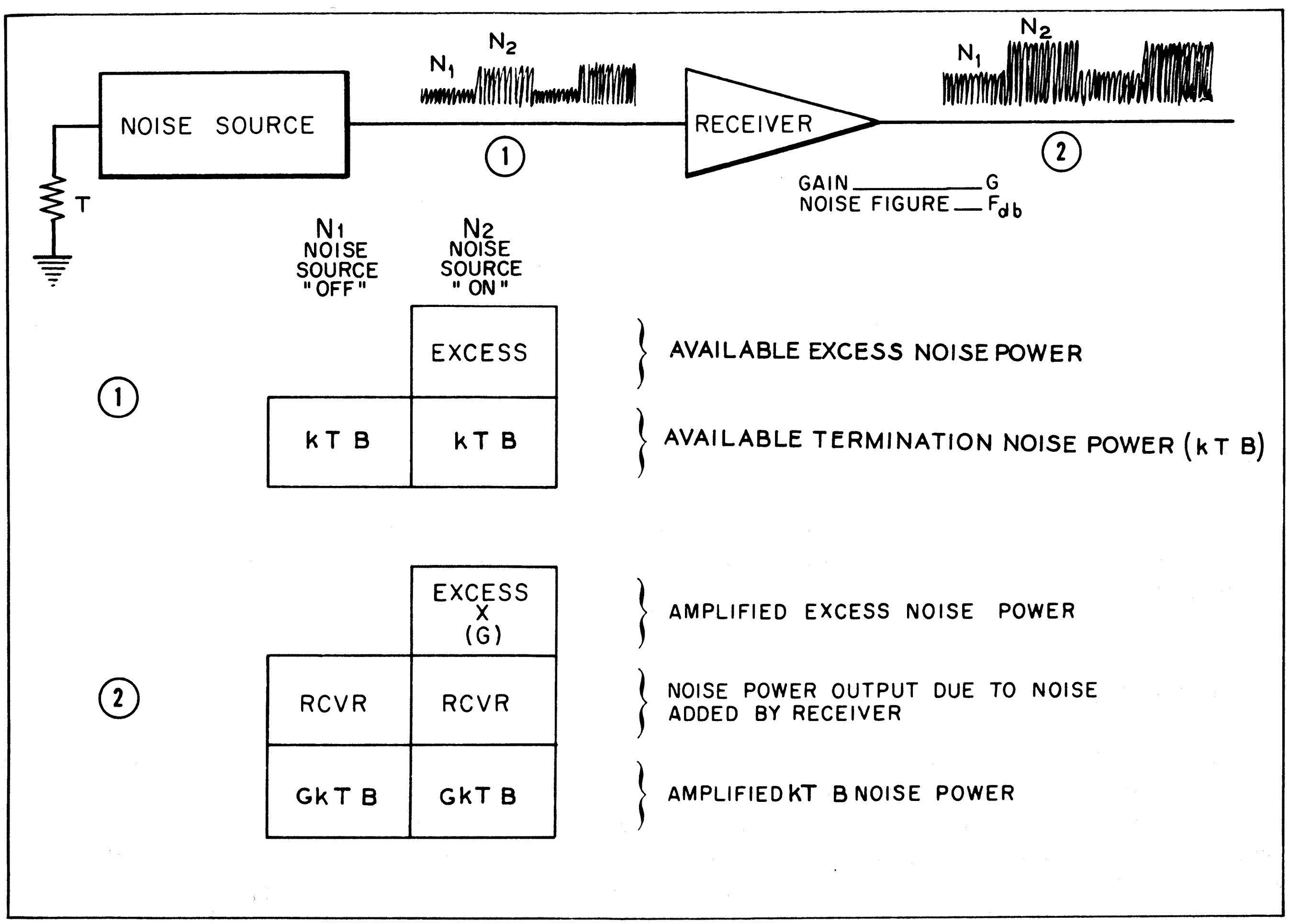


FIGURE 3–11 *Automatic noise figure measurement of microwave device (composition of noise power).*

kTB is the available noise power from the reference load, where:

k = Boltzmann's constant.
T = temperature of reference load in degrees Kelvin.
B = bandwidth of measuring system.

Excess noise power added by the noise source is based on the effective fired temperature of the source. An argon gas discharge, for instance, is 15.2 db above the reference temperature power. Then the total noise-power output of the receiver with noise source "OFF" is

$$N_1 = GkTB + \mathrm{RCVR} \tag{3–1}$$

where G is receiver power gain. Total noise-power output of the receiver with noise source "ON" is

$$N_2 = GkTB + \mathrm{RCVR} + \mathrm{Excess} \times (G) \tag{3–2}$$

Noise factor as defined above (where T_0 equals 290°K) is:

$$F = \frac{GkT_0B + \mathrm{RCVR}}{GkT_0B}, \quad \text{or} \quad \frac{\text{(Total noise output from device)}}{\text{(Output power if noiselessly amplified)}} \tag{3–3}$$

so

$$\mathrm{RCVR} = (F - 1)\, GkT_0B \tag{3–4}$$

which is noise output power contributed by the RCVR. Also, the excess noise power from the gas discharge at the input is

$$\mathrm{Excess} = \left(\frac{T_2 - T_0}{T_0}\right) kT_0B \tag{3–5}$$

where T_2 is the effective fired temperature of the noise source. Then the ratio at the output is

$$\frac{N_2}{N_1} = \frac{GkT_0B + (F-1)\, GkT_0B + \left(\frac{T_2 - T_0}{T_0}\right) GkT_0B}{GkT_0B + (F-1)\, GkT_0B} \tag{6}$$

by substitution from Eqs. 1, 2, 4, and 5.

$$F = \frac{\left(\frac{T_2 - T_0}{T_0}\right)}{\left(\frac{N_2 - N_1}{N_1}\right)} \tag{7}$$

Note that the gain-bandwidth factor (GB) has disappeared. Finally,

$$F_{\mathrm{db}} = 10 \log \left(\frac{T_2}{T_0} - 1\right) - 10 \log \left(\frac{N_2}{N_1} - 1\right) \tag{8}$$

The first term is a known quantity, expressed in decibels of excess-noise ratio. For an argon discharge, the excess-noise ratio is 15.2 db; then

$$F_{\mathrm{db}} = 15.2 - 10 \log \left(\frac{N_2}{N_1} - 1\right) \tag{9}$$

Thus the ratio N_2/N_1 contains the desired noise-figure information. hp Models 340B, 342A, and 344A noise-figure meters measure noise figure as a function of this ratio. To make noise-figure measurements, the 340B or the 342A, the appropriate noise source, and the receiver or amplifier under test are connected as shown in Fig. 3–12. The noise-figure meter square-wave-modulates the noise source at a rate of about 500 cps and measures noise figure by comparing the noise output of the device under test when the noise source is "OFF" to the noise output when the noise source is "ON."

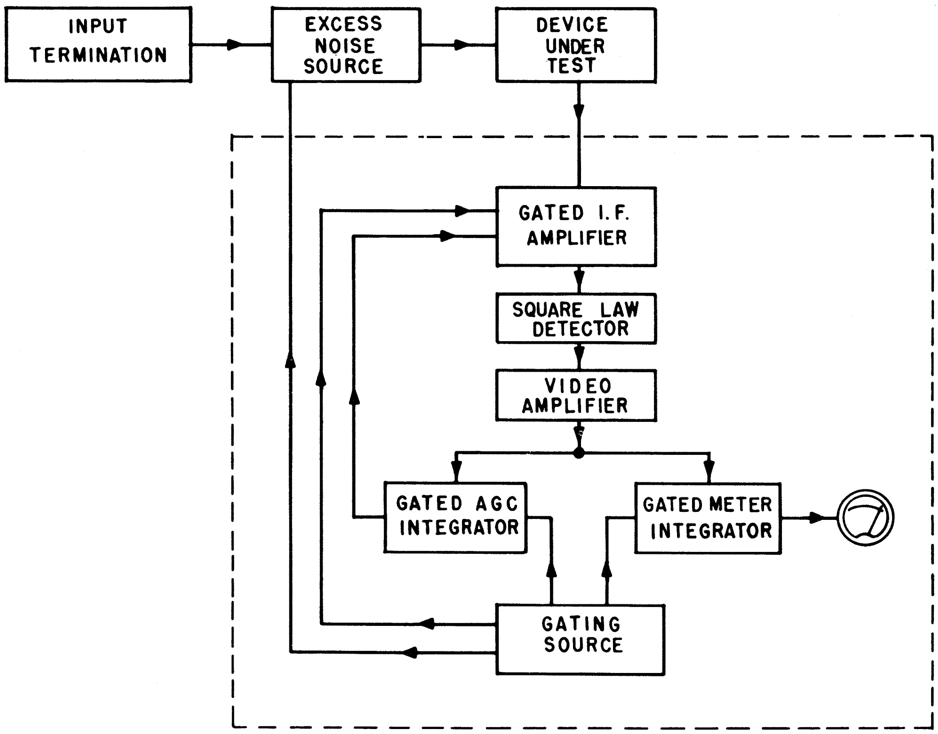


FIGURE 3–12 *Simplified block diagram of hp 340B or 342A noise figure meter.*

The input circuitry of Model 340B consists of a gated-tuned amplifier which operates at either of two frequencies, 30 or 60 mc, selected by a front panel switch. The input circuitry of Model 342A consists of a 30 mc gated-tuned amplifier preceded by a 4-channel mixer-local oscillator combination which, depending upon the position of the front panel switch, will convert four frequencies—60, 70, 105, and 200 mc—to 30 mc. The output from the 340B/342A tuned amplifier is detected, amplified, and alternately applied to two gated integrators. When the noise source is "ON," the combined noise power from the noise source and the device under test is amplified by the tuned amplifier, detected, and passed through the AGC integrator. The time constant of the AGC voltage applied to the amplifier is long enough to hold the gain of the amplifier the same whether the noise source is "ON" or "OFF."

When the noise source is turned off, the combined noise power from the source impedance (load) and the device under test is amplified, detected, passed through the meter integrator, and displayed on the meter. Because of the AGC action, the meter deflection is proportional to the ratio of the noise powers (source "ON" and source "OFF"), and, since the additional noise from the noise source (excess noise) is accurately known, the meter face is calibrated directly in decibels of noise figure.

The AGC action, in addition to establishing a reference against which noise-figure measurements can be made, provides a wide (50 db) input operating range and also eliminates the necessity for periodic recalibration of the noise-figure meter. AGC voltages appear on a pair of terminals at the rear of the instrument to facilitate measurements which require an indication of the gain of the system in relation to changes in noise figure.

The meter face is provided with two noise-figure scales—a "noise diode" scale for use with the hp 343A vhf noise source and 345B IF noise source, and a "gas tube" scale for use with 347A waveguide noise sources. Current scales indicate the current supplied to the noise sources. Special circuitry is included to enable the offsetting of the noise-figure meter scale so that low values of noise figure can be read on a more sensitive external meter. A phone jack is provided on the rear of the noise-figure meter to drive a remote meter or galvanometer recorder.

The new hp 344A transistorized noise-figure meter has been specifically designed for radar-system applications, in which time-shared noise-figure measurements are extremely important to assure that radar sets are operating at peak performance. Sensitivity has been made very high in order to permit noise sources to be decoupled by as much as 20 db from the main transmitter line. Alarm circuitry for remote indication of excessive noise figure, as well as for remote metering of noise figure, is available in the 344A.

Hewlett-Packard noise sources are available for all frequencies between 10 mc and 18,000 mc to provide measurements on all rf devices in this range. Sources have been specifically designed for very low-fired and unfired swr to lower coupling ambiguities of the excess-noise ratio into the device under test. Waveguide sources have been loaded with resonance-suppressing polyiron loads in order to eliminate high swr at points in the band caused by the insertion of the noise tube into the waveguide.

SECTION FOUR

Microwave Experiments

The series of fifteen experiments on the following pages has been planned to accomplish two objectives: first, to provide experience in making the various types of measurements required in microwave work; second, to familiarize the reader with the more common types of waveguide and test equipment as they are used in making these measurements.

Equipment for experiments. The degree to which the experiments can be utilized will depend, of course, on the equipment available for making them. Basic measurements of frequency, attenuation, and swr can be made with a minimum investment in specialized test apparatus. Some additional equipment is needed for making basic power measurements. If considerable work is to be done, an even greater investment can be justified for other equipment which simplifies the experimental setup or extends the range of measurement possible.

All experiments utilize Hewlett-Packard "X" band (8.2 to 12.4 gc) equipment. These equipment items are of convenient size and are more economical to purchase than other brands. The specific items required are listed in each experiment and are summarized in Table I. They are described in more detail in Appendix B. Since training operations often have budget limitations, the equipment required for the experiments has been divided into a basic group and two optional groups. Only the basic group is needed for many of the experiments. Individual equipment items in each group can be ordered from Hewlett-Packard as they become necessary to meet individual training needs.

The experimenter is encouraged to exercise ingenuity, both in substitution of equipment items to utilize test equipment which may already be available, and in modification of the experiments to achieve specific training objectives.

Experimental data collection

In performing each experiment, the most logical sequence usually involves calculating data and plotting curves as observed data is recorded. Often, however, the time available for using the equipment is limited, and curves must be plotted, calculations made, and questions answered after the physical experiment has ended.

To permit convenient collection of data, a detachable collection sheet (called *Results*) has been included at the end of each experiment. If time is limited, only the "observed" data need be recorded initially. The remaining data and information can be recorded at some later time.

RF power source

In order to keep power-source expenditures at a relatively low level, the Varian Associates X-13 reflex klystron has been recommended throughout the experiments. Naturally, any suitable signal generator or signal source may be substituted for the reflex klystron and the klystron power supply.

Safety precautions

Because this section is concerned with actively making microwave measurements previously discussed in theory only, the following safety precautions must be considered before any experiments are performed.

1. *Do not look into the klystron output or connecting waveguide while the klystron is energized!* Although the rf power levels available from the reflex klystron used in these experiments are generally believed not dangerous to most body tissues, the eye is particularly susceptible to permanent damage when it is subjected to microwave radiation.*
2. *Before making any connections to the klystron power supply, always ensure that no voltages are present!* The voltages required for the klystron are dangerous, and can be lethal. Note that the external modulation terminals of the Model 715A klystron power supply may also have high voltage impressed across them. It is always good practice to insulate all exposed voltage terminals from accidental contact with the body.

Equipment handling

Employment of the following common-sense rules will help preserve equipment usefulness.

1. Unless the klystron tube is forced-air-cooled, its life will be seriously reduced. Always air-cool the klystron with a fan or blower.
2. Use stands to support long strings of waveguide equipment. In this way, the flanges will not be unduly strained, and reflections from misaligned flanges will be minimized.
3. Protect the flanges from nicks and scratches. It is always good practice to keep unused pieces of waveguide capped with flange covers; this procedure also reduces the amount of dust that will gather on interior surfaces. (If waveguide interiors do become dusty, use extreme caution in cleaning them. *Never* use compressed air on waveguide items, such as attenuators, having fragile mica cards in them.)
4. Protect waveguide walls from indentations. Small projections inside the waveguide will cause unwanted reflections.
5. Use *extreme caution* when connecting or disconnecting bolometers to or from the microwave power meter. Care is especially important with barretters, which are particularly susceptible to "burnout."

* W. W. Mumford, "Some Technical Aspects of Microwave Radiation Hazards," *Proceedings of the IRE*, February, 1961, pp. 427–447.

TABLE 4–1 EQUIPMENT FOR MICROWAVE EXPERIMENTS

EQUIPMENT	QUANTITY	MODEL NUMBER	DESCRIPTION	UNIT PRICE*	EXPERIMENTS IN WHICH USED (AND QUANTITY)														
					1	2	3	4	5	6	7	8	9	10	11	12	13	14	15
	2	24	Waveguide stand	$ 3.00	2	1						1						2	2
	2	X25	Waveguide clamp	2.50	2	1						1						2	2
	1	120B	450 kc general-purpose oscilloscope	475.00	1	1		1	1			1	1	1	1	1	1	1	1
	1	X375A	Variable flap attenuator	100.00	1	1		1	1			1			1	1	1	1	1
GROUP I	1	X382A	Precision variable attenuator	275.00			1	1	1			1	1	1	1		1	1	1
Basic equipment for measuring	1	415B	Standing-wave indicator	225.00		1		1	1			1	1		1	1	1	1	1
frequency, attenuation, and swr	1	X421A	Crystal detector	75.00	1	1		1	1†					1	1			1	1
	1	444A	Untuned broadband probe with crystal	55.00		1			1			1	1		1	1	1		1
	1	X532B	Direct-reading frequency meter	200.00	1	1	1	1	1			1	1			1	1	1	1
	1	715A	Klystron power supply	325.00	1	1	1	1	1			1	1	1	1	1	1	1	1
	1	809B	Universal probe carriage	175.00		1			1			1	1		1	1	1		1
	1	X810B	Slotted section (for 809B)	90.00		1			1			1	1		1	1	1		1
	1	AC-16A	Cable assembly (two dual banana plugs)	4.50	1	1		1	1			1	1	1	1	1		1	1
	1	AC-16B	Cable assembly (dual banana plug to BNC)	5.50	1	1		1	1			1	1	1	1	1		1	1
	2	AC-16K	Cable assembly (BNC to BNC)	6.50		1	1	1	2			2	2	1	2	1	2	1	1
	1	X-13	Micrometer tuned reflex klystron	‡	1	1	1	1	1			1	1	1	1	1	1	1	1
	1		Blower or fan to cool klystron		1	1	1	1	1			1	1	1	1	1	1	1	1
GROUP II																			
Additional basic equipment for	1	430C	Microwave power meter	250.00			1		1			1	1	1	1		1		
power measurement	1	X487B	Thermistor mount	75.00			1		1			1	1	1	1		1		
	1	200CD	Wide-range oscillator	195.00										1					
	1	X281A	Waveguide-to-coaxial adapter	25.00												1			
	1	X485B	Detector mount for barretter	75.00									1	1					
GROUP III	1	X752C	Multi-hole directional coupler	110.00														1	
Optional equipment for more	1	803A-76G	Shorting plug, type N	4.50												1			
detailed measurements of atten-	1	X870A	Slide-screw tuner	130.00									1						1
uation, impedance, and power	1	X914B	Moving load	60.00					1†									1	
	1	X920A	Adjustable short	75.00		1													
	1	AC-16C	Cable assembly (type N male to female)	13.00												1			
	1	AC-16S	Cable assembly (dual banana to test leads)	7.50										1					

* Subject to change without notice.

† For optional use with this experiment.

‡ Manufactured by Varian Associates—$295.00.

EXPERIMENT 1

Reflex Klystron Characteristics

Object

To measure and observe some of the important characteristics of the reflex klystron, and to become familiar with its operation.

Theory

Many excellent discussions of detailed klystron theory are available in standard electronic textbooks. Therefore, the following information serves primarily as a review of reflex-klystron operation.

Figure 1 shows a schematic diagram of a reflex klystron and the voltages required for operation. The various elements that compose the tube are the cathode, a focusing electrode at cathode potential, a resonator which also serves as an anode, and a reflector (repeller) which is at a negative potential with respect to the cathode. The combination of the cathode, the focusing electrode, and the anode beams the electrons through the resonator gap and out toward the reflector. Since the reflector element is negative with respect to the anode, the electrons are turned back toward the anode, where they pass through the gap a second time.

When the klystron is oscillating, an alternating voltage appears across the gap of the resonator. As electrons pass through the gap, they are either accelerated or decelerated as the voltage across it changes in magnitude with time. Accelerated electrons leave the gap at an increased velocity, and decelerated electrons leave at a reduced velocity. Because of the difference in velocity, electrons leaving the gap at different parts of the gap-voltage cycle take different lengths of time to return to the gap (i.e., have different transit times). As a result, the electrons group together in bunches as they return through the gap. This variation in velocity of the electrons is called velocity modulation.

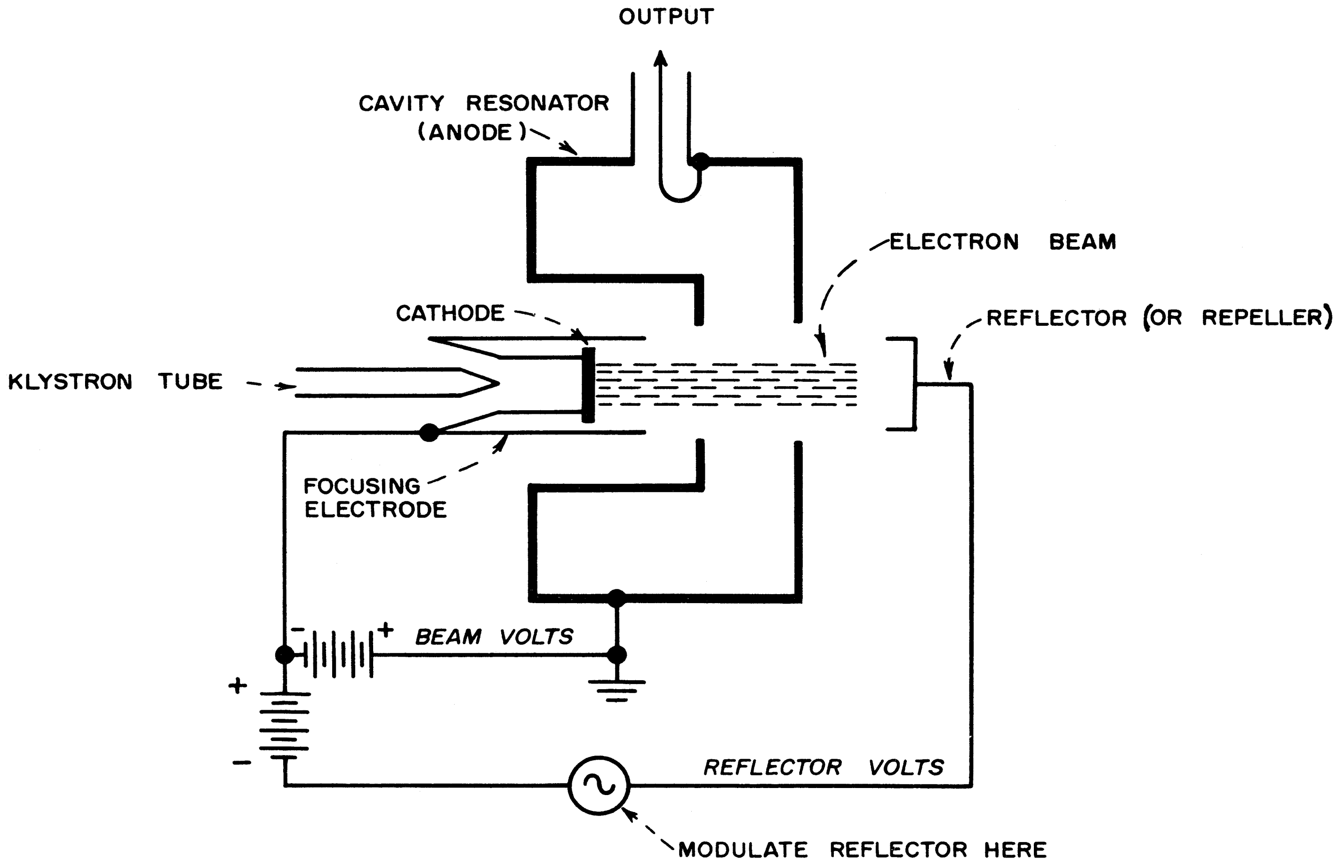


FIGURE 1

As the bunches pass back through the gap, they react with the voltage appearing across the gap. If the bunches pass through the gap at a time in the gap-voltage cycle such that the electrons are slowed down, then energy will be delivered to the resonator and oscillations will be sustained. Strongest oscillations will occur when the transit time in the gap (reflector-anode region) is $n + 3/4$ cycles of the resonator frequency, where n is an integer, including zero. If, however, the time of transit in the anode-reflector region is such that the bunches are caused to arrive at a time when they will be accelerated by the gap voltage, then energy is removed from the resonator and oscillation will tend to stop.

To generate oscillations at a given frequency and a fixed anode voltage, it is necessary to vary the transit time in the anode-reflector space to a suitable value by means of adjusting the reflector voltage. The more negative the reflector voltage, the shorter the transit time. Figure 2 shows the effect of reflector voltage on both the level and the frequency of the output of a reflex klystron. Although frequency of operation is determined primarily by the resonator dimensions, a small change in frequency may be obtained by adjusting either the reflector voltage or the anode voltage; this adjustment is known as electronic tuning. A frequency change of several per cent is possible in some tubes.

For maximum convenience, bench klystrons such as the X-13 are commonly driven with a self-contained power-supply package which provides voltages and controls for each klystron element. When such a power supply is used, the following rule should be followed to avoid damaging the klystron. *Always apply reflector voltage to the klystron before applying beam voltage.*

NOTE: The ⓗⓟ Model 715A klystron power supply has a protective diode to prevent the reflector from going more positive than the electron beam, with resultant heavy current damage to the reflector. Even so, it is good practice to use the following procedure:

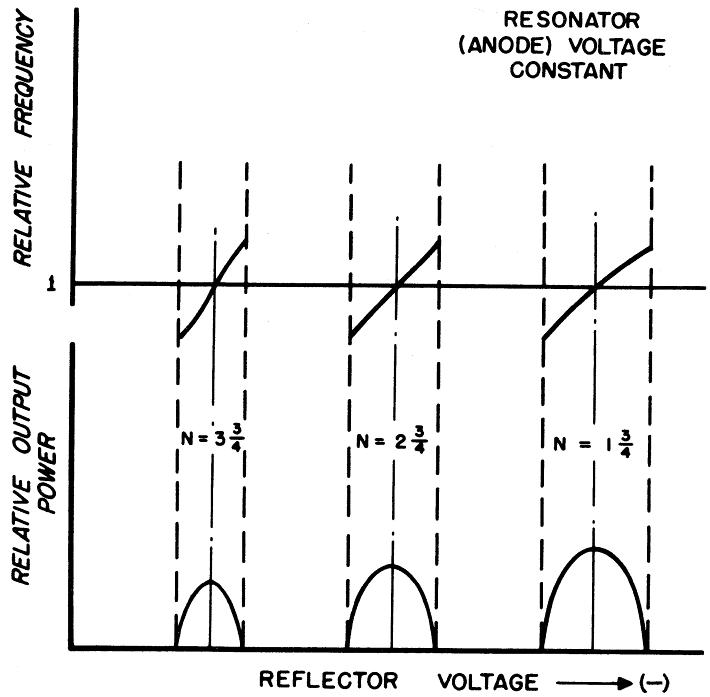


FIGURE 2

1. Turn the power switch on, and allow the instrument tubes and the klystron-tube heater to reach operating temperature.
2. Set the REFLECTOR RANGE switch to the desired voltage range and adjust the REFLECTOR VOLTS control to the desired voltage.
3. Set the MOD. SELECTOR to CW and adjust the BEAM VOLTS control for the desired klystron beam voltage. After this step, the MOD. SELECTOR can be set to the desired type of modulation.

References. Additional information to supplement this experiment may be found in the following texts (listed in Appendix C).

Ginzton, Chapter 1: 1.1–1.4, 1.6.
King, Chapter 5: 5-1–5-3.
Reich, Chapter 13.
Wind, Section 17: 17.01–17.04.

Equipment

QUANTITY	TYPE
1	ⓗⓟ 715A klystron power supply (with klystron cable)
1	Varian X-13 reflex klystron
1	Cooling fan or blower
1	ⓗⓟ X375A variable flap attenuator
1	ⓗⓟ X421A crystal detector
1	ⓗⓟ X532B frequency meter
1	ⓗⓟ 120B oscilloscope
1	ⓗⓟ AC-16A cable (dual banana to dual banana)
1	ⓗⓟ AC-16B cable (dual banana to BNC)
2	ⓗⓟ Model 24 waveguide stand with Model X25 Waveguide clamp

Procedure

Section 1—General

1-1 Review the safety precautions on page 42, and the operating instructions for each equipment item.

1-2 Set up the equipment as shown in Fig. 3. The oscilloscope horizontal input should be *a-c coupled* to the 60 cps frequency on the power line (by using the low-voltage terminals of a filament transformer, for example). The oscilloscope vertical input should be *d-c coupled* to the crystal detector output. The variable flap attenuator should be set to approximately 15 db of attenuation (to protect the crystal detector from excessive klystron power).

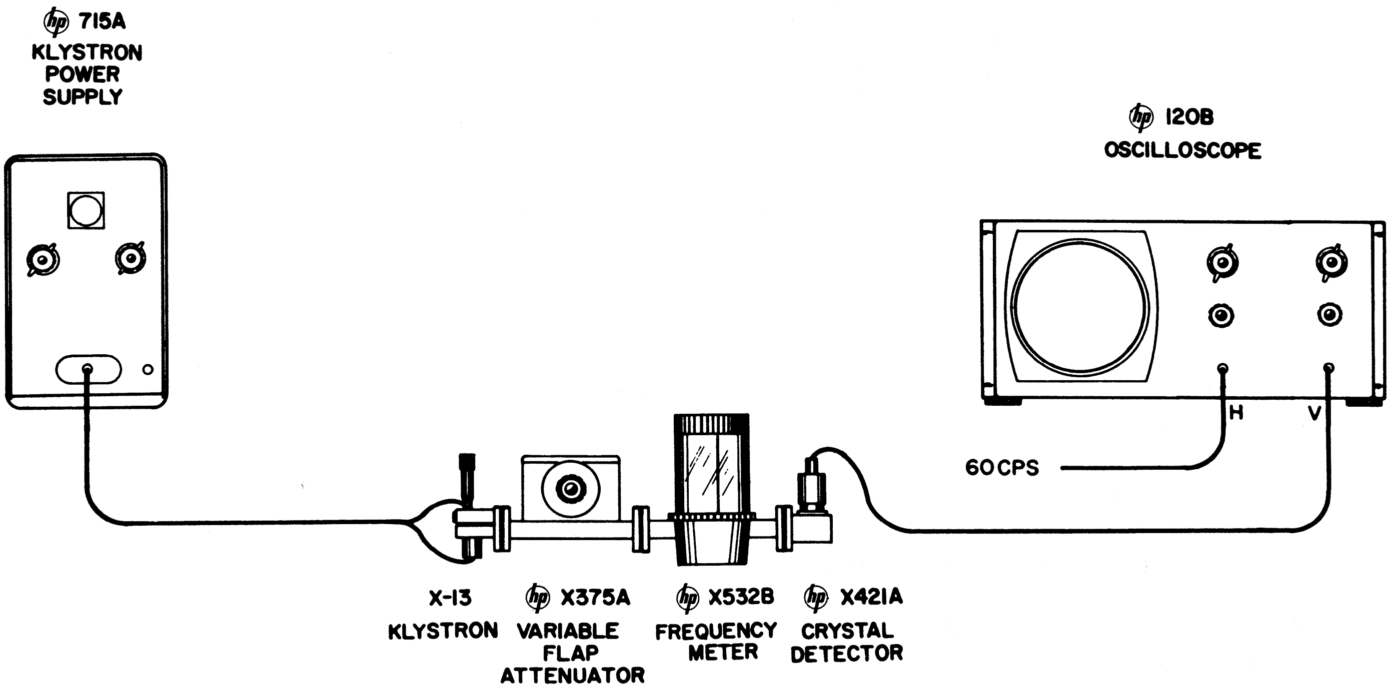


FIGURE 3

1-3 Turn on the klystron power supply, setting the reflector (repeller) voltage to approximately 300. Apply 400 beam volts.

1-4 Modulate the klystron with the internal 60 cps modulation frequency supplied by the klystron power supply, and set the modulation amplitude to its maximum value.

NOTE: The oscilloscope horizontal input above could have been driven with the 60 cps modulation supplied by the klystron power supply. This scheme, however, does not permit phase variations to be made in the oscilloscope presentation.

Section 2—Mode Characteristics

2-1 Adjust the horizontal sensitivity and position the controls on the oscilloscope so that a trace exactly 10 cm wide is centered on the oscilloscope tube.

2-2 Adjust the vertical sensitivity and position the controls on the oscilloscope until a pattern similar to Fig. 4 is observed. (Usually, two or three reflector modes will be visible.) Then adjust the vertical sensitivity so that the largest mode is from 4 to 6 cm high.

2-3 Tune the frequency meter until a "pip" is observed on the mode pattern. Record the frequency-meter reading on the *Results* sheet.

NOTE: Frequency measurement will be discussed in detail in the next experiment.

2-4 Tune the klystron micrometer while "tracking" it with the frequency meter until the klystron is set at 8.5 gc.

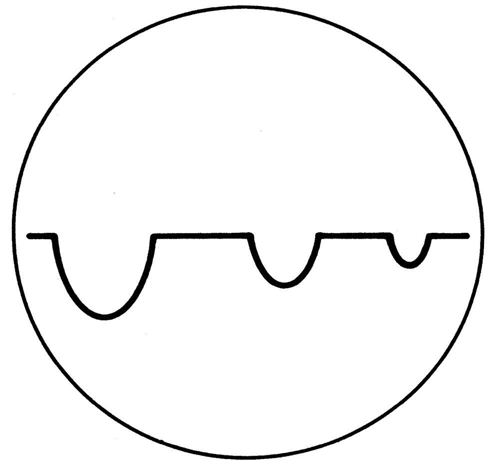


FIGURE 4

2-5 Adjust the reflector voltage, the klystron tuning, and the 60 cycle modulation amplitude and phase until the mode having the maximum amplitude is centered on the oscilloscope presentation, with the frequency meter "pip" also centered, as shown in Fig. 5. (This relationship usually occurs when the reflector voltage is between 600 and 900 v.) Note the setting of the reflector-voltage dial, and record this in Table I at the end of this experiment. Also record the klystron-micrometer reading for use later in the experiment.

NOTE 1: For greater accuracy in measuring reflector voltages, a d-c voltmeter may be used. See the hp Model 715A klystron power supply manual for technique and precautions.

NOTE 2: Although each mode has a specific numerical designation, for the purposes of this experiment the largest mode will be designated "first mode," the next largest mode will be designated "second mode," etc.

2-6 Without changing the frequency-meter setting of 8.5 gc, readjust the reflector voltage and the 60 cycle modulation amplitude until the second largest mode appears to be centered (with the frequency meter "pip"), as in Step 2-5. Record the reflector voltage in the "second mode" column of Table I.

2-7 Repeat the procedure of Step 2-5 for the other modes in succession, and record the reflector voltage for each mode on Table I.

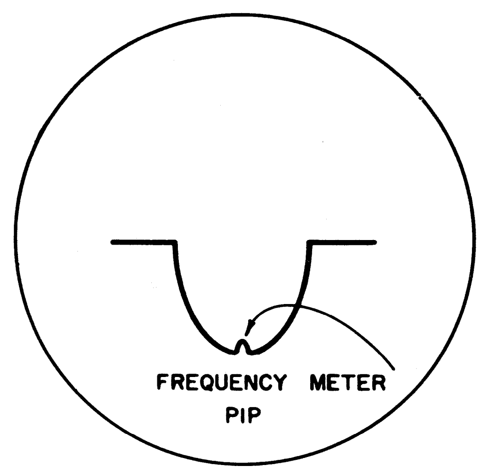


FIGURE 5

2-8 Having completed reflector-voltage measurements for three or four modes at 8.5 gc, tune the klystron and the frequency meter to 9.5 gc. Repeat the procedure of Steps 2-5 through 2-7 for this new frequency, and record the reflector-voltage data in Table I.

2-9 Perform similar mode-reflector-voltage measurements for klystron frequencies of 10.5 and 11.5 gc. (At these higher frequencies, the largest mode may be unattainable with a reflector-voltage limit of 900 v.)

Section 3—Electronic Tuning

3-1 Select the largest mode that was present for all frequencies, and tune the klystron and the frequency meter to 8.5 gc so that this mode appears as shown in Fig. 5. (To find the approximate klystron micrometer setting quickly, refer to your data recorded in Table I.)

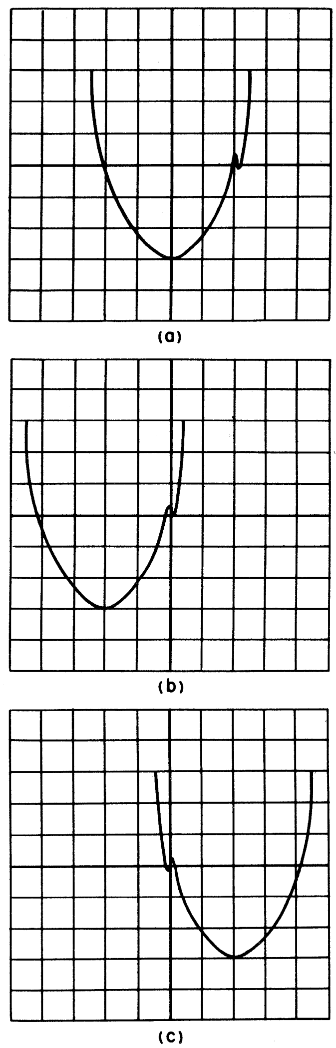


FIGURE 6

3-2 Adjust the oscilloscope vertical sensitivity so that the mode displayed is exactly 6 cm in amplitude, and adjust the frequency meter until the "pip" is exactly at the half-amplitude point (3 cm) of the mode, as shown in Fig. 6(a).

3-3 Change the reflector voltage until the mode shifts as shown in Fig. 6(b), putting the "pip" on the vertical center line. Record the frequency-meter reading (f_1) and the reflector voltage (V_1) for this condition in Table II.

3-4 Using the reflector-voltage control and the frequency meter, shift the mode presentation, as shown in Fig. 6(c), and record the frequency (f_2) and reflector voltage (V_2) in Table II.

3-5 Repeat the procedure of Steps 3-1 through 3-4 for frequencies of 9.5, 10.5, and 11.5 gc, always using the same mode. (Approximate reflector voltages for this mode at the various frequencies are recorded in Table I.)

3-6 Calculate the electronic-tuning bandwidth and the electronic-tuning sensitivity in Table II.

Name ______________________________
Course ______________________________
Date performed ______________________
Date turned in ______________________

Results (Experiment 1)

Observed		Calculated	
Step		*Step*	
2-3	________ (gc)		
2-5	Table I		
2-6	Table I		
2-7	Table I		
2-8	Table I		
2-9	Table I		
3-3	Table II		
3-4	Table II		
3-5	Table II	3-6	Table II

Questions

After plotting the data from Table I in Fig. 7 (labeling each mode) and plotting the data from Table II in Figs. 8 and 9, answer the following questions.

1. How many klystron modes did you find at a frequency of 10.5 gc for various reflector voltage settings?

2. If the reflector voltage were maintained at 300 volts and the klystron frequency were varied from 8.5 gc to 11.5 gc, through how many modes would you pass?

3. In order to use the "second" mode from 8.5 gc to 9.5 gc, what range of reflector voltages would be necessary?

______________ to ______________

4. As the klystron frequency is increased, how does the electronic tuning bandwidth react?

5. As the klystron frequency is increased, how does the electronic-tuning sensitivity react?

6. From the data plotted in Fig. 7, calculate the mode number of that mode used in Section 3. Use the constant frequency relation

$$\frac{V_1}{V_2} = \frac{N_2}{N_1}$$

where V is reflector voltage and N is the mode number ($\frac{3}{4}$, $1\frac{3}{4}$, $2\frac{3}{4}$, etc.). ______________________________

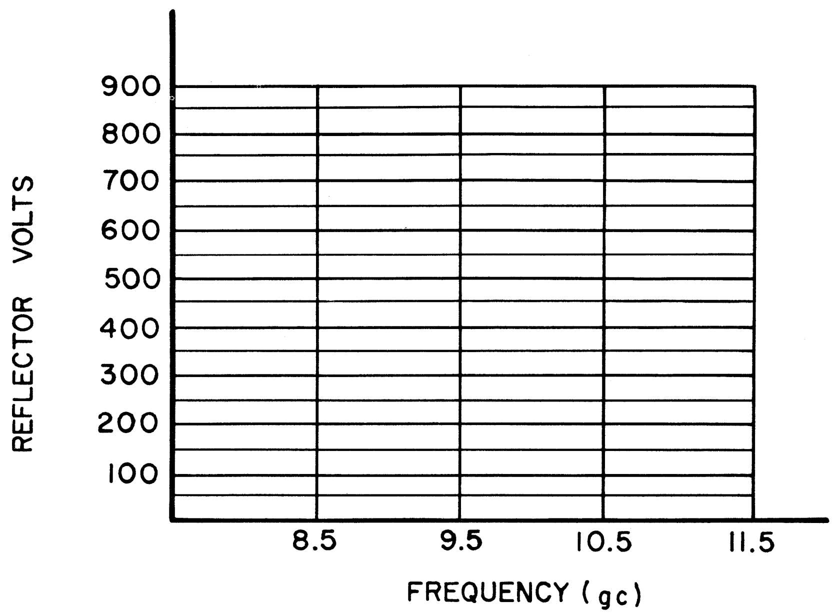

REFLECTOR VOLTS
900
800
700
600
500
400
300
200
100
8.5
9.5
10.5
11.5
FREQUENCY (gc)

FIGURE 7

TABLE I

STEP →		2-5	2-6	2-7	
		FIRST MODE	SECOND MODE	THIRD MODE	FOURTH MODE
Frequency (gc)	Klystron micrometer	Reflector volts	Reflector volts	Reflector volts	Reflector volts
8.5					
9.5					
10.5					
11.5					

Cathode (beam) volts ____________.

TABLE II

STEP →	3-3	3-3	3-4	3-4	3-6	3-6	
Frequency (gc)	f_1	f_2	$f_1 - f_2$	V_1	V_2	$V_1 - V_2$	$\frac{f_1 - f_2}{V_1 - V_2}$
8.5							
9.5							
10.5							
11.5							

$f_1 - f_2 =$ electronic tuning bandwidth (mode width to half-power points, in mc).

$\frac{f_1 - f_2}{V_1 - V_2} =$ electronic tuning sensitivity (mc/v).

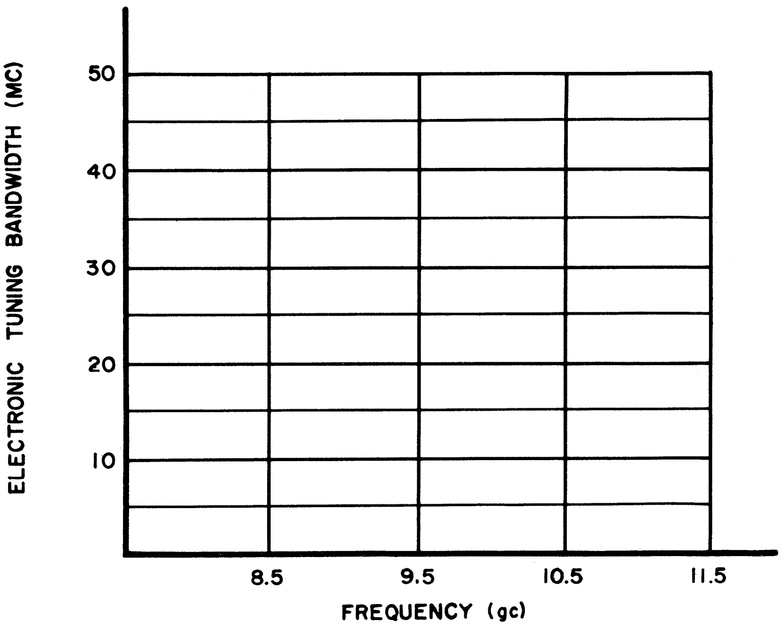


FIGURE 8

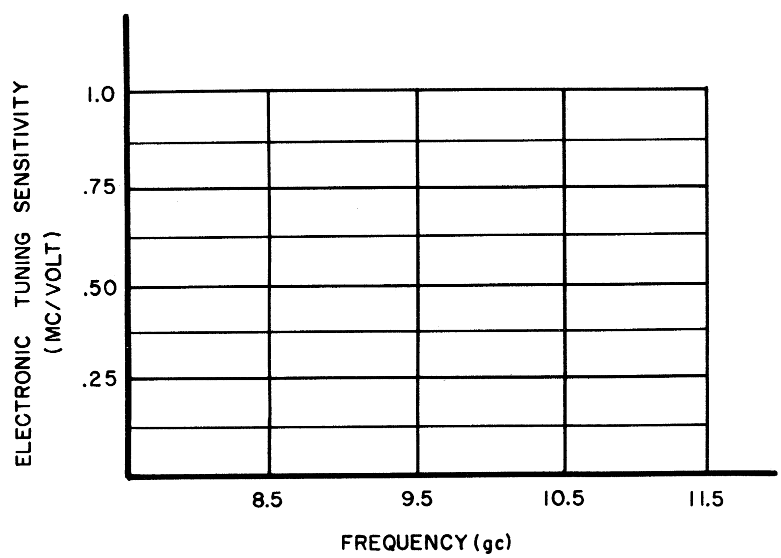


FIGURE 9

Discussion

EXPERIMENT 2

Frequency Measurement

Object

To examine further the frequency characteristics of klystrons and to become familiar with typical microwave frequency measurements. In addition, to study 1000 cps amplitude modulation of klystrons.

Theory

A discussion of frequency measurement is presented in Chapter III. This discussion is applicable to the cavity type of reaction frequency meter and the slotted-line (null) techniques used in this experiment.

References. Additional information to supplement this experiment may be found in the following texts (listed in Appendix C).

Ginzton, Chapter 7.
King, Chapter 4: 4.1, 4.5, 4.7.
Reich, Chapter 8: 8-9.
Wind, Section I.

Equipment

QUANTITY	TYPE
1	hp 715A klystron power supply (with cable)
1	Varian X-13 reflex klystron
1	Cooling fan or blower
1	hp X375A variable flap attenuator
1	hp X421A crystal detector
1	hp X532B frequency meter
1	hp 415B standing-wave indicator
1	hp 120B oscilloscope
1	hp 809B probe carriage
1	hp 444A broadband probe
1	hp X810B slotted section
1	hp X920A adjustable short
1	hp AC-16A cable (dual banana to dual banana)
1	hp AC-16B cable (dual banana to BNC)
1	hp AC-16K cable (BNC to BNC)
1	hp Model 24 waveguide stand with Model X25 waveguide clamp

Procedure

Section 1—General

1-1 Review the safety precautions on page 42, and the operating instructions for each equipment item.

1-2 Set up the equipment as shown in Fig. 1. The oscilloscope horizontal input should be *a-c coupled* to the external-modulation terminals of the klystron power supply. The oscilloscope vertical input should be *d-c coupled* to the crystal detector output. The variable flap attenuator should be set to approximately 15 db of attenuation.

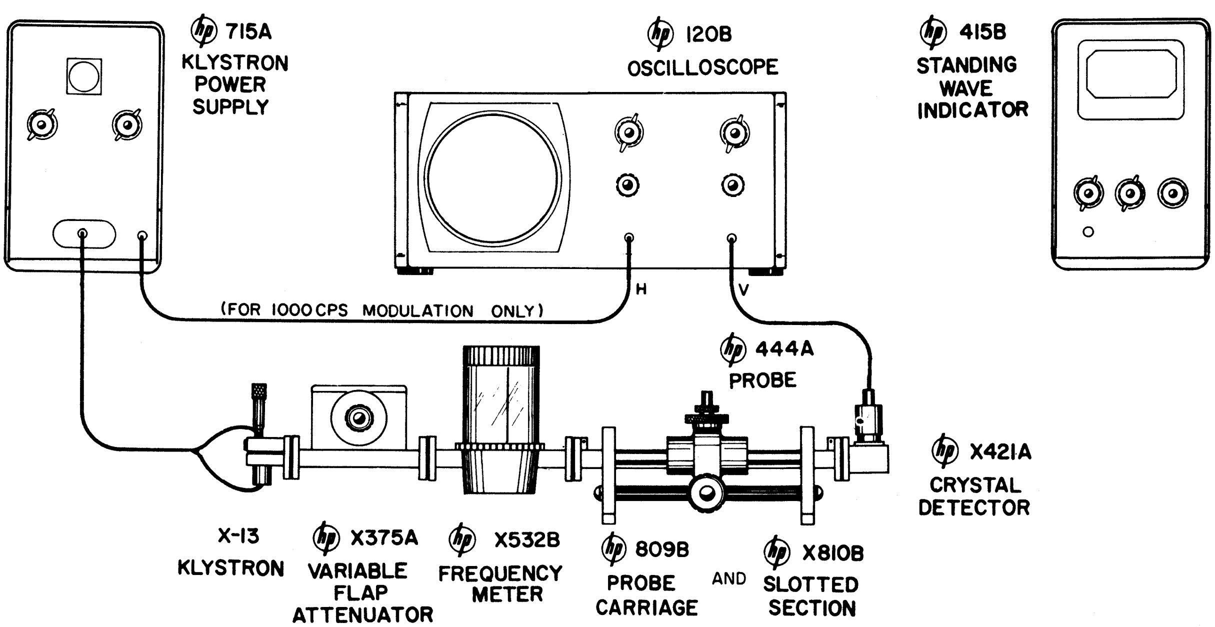
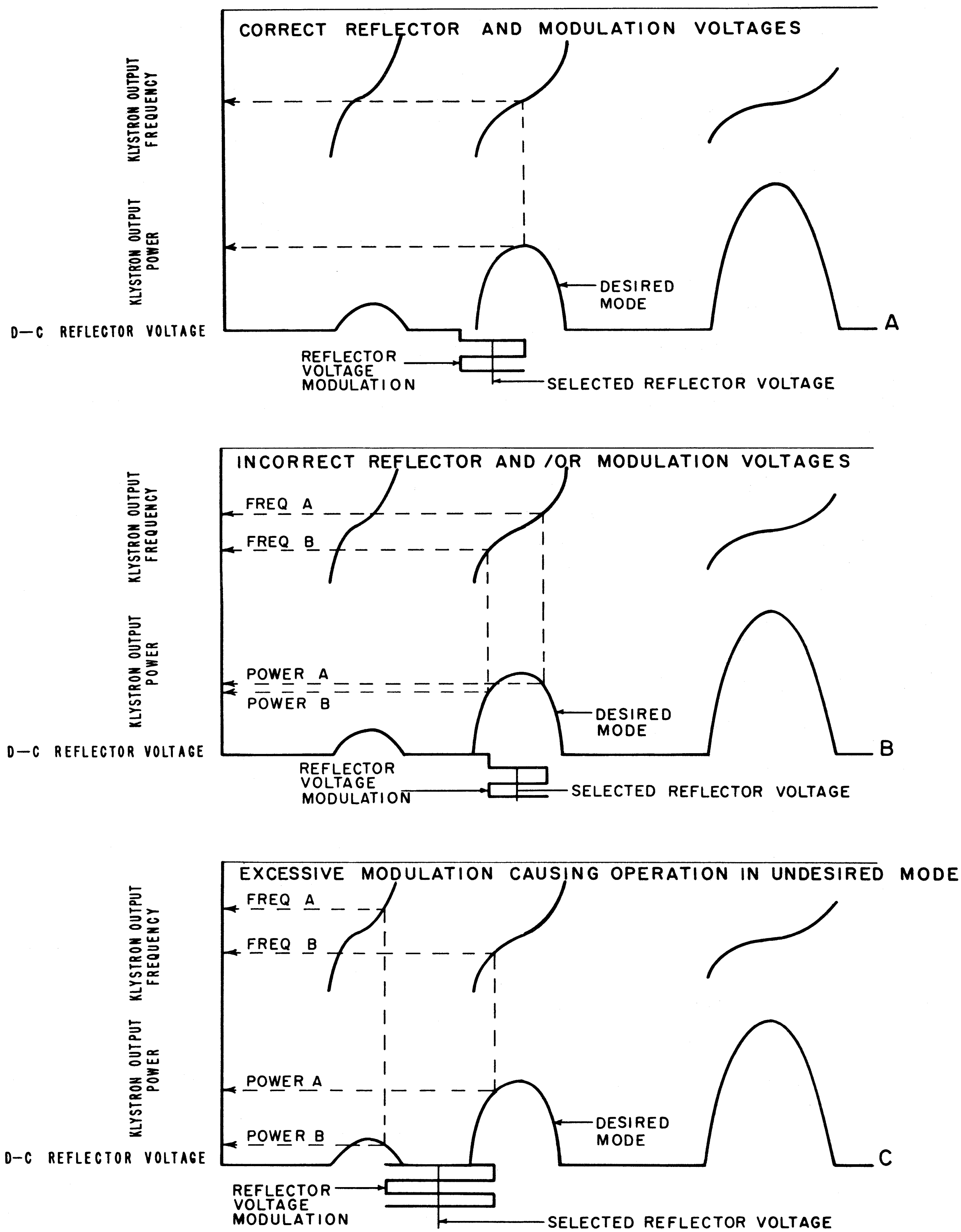


FIGURE 1



CORRECT REFLECTOR AND MODULATION VOLTAGES
KLYSTRON OUTPUT FREQUENCY
KLYSTRON OUTPUT POWER
DESIRED MODE
D—C REFLECTOR VOLTAGE
A
REFLECTOR VOLTAGE MODULATION
SELECTED REFLECTOR VOLTAGE
INCORRECT REFLECTOR AND /OR MODULATION VOLTAGES
KLYSTRON OUTPUT FREQUENCY
FREQ A
FREQ B
KLYSTRON OUTPUT POWER
POWER A
POWER B
DESIRED MODE
D—C REFLECTOR VOLTAGE
B
REFLECTOR VOLTAGE MODULATION
SELECTED REFLECTOR VOLTAGE
EXCESSIVE MODULATION CAUSING OPERATION IN UNDESIRED MODE
KLYSTRON OUTPUT FREQUENCY
FREQ A
FREQ B
KLYSTRON OUTPUT POWER
POWER A
POWER B
DESIRED MODE
D—C REFLECTOR VOLTAGE
C
REFLECTOR VOLTAGE MODULATION
SELECTED REFLECTOR VOLTAGE

FIGURE 2

1-3 Turn on the klystron power supply, setting the reflector voltage to approximately 300 v. With the klystron-power-modulation selector switch in CW, apply enough beam voltage for a 30 ma indication of beam current on the power-supply meter.

1-4 Adjust the reflector voltage until a vertical deflection is noted on the oscilloscope. (No horizontal signal need be applied at this time.) It will probably be necessary to use the most sensitive range (10 mv/cm) of the oscilloscope vertical amplifier. The deflection will be downward, because the crystal polarity is such that negative output voltage results.

1-5 Adjust the reflector voltage through its various ranges, noting the maximum negative deflection of the various modes. Select the reflector voltage that yields the maximum mode power (deflection).

Section 2—1000 Cycle Modulation

2-1 For operation of the Model 415B standing-wave indicator, it is necessary to develop a 1000 cps square-wave modulation envelope on the rf signal. This is done by applying a square wave of voltage to the reflector element of the klystron, thus causing it to move into and out of a given mode of oscillation (turning on and off at a 1000 cps rate).

2-2 Because it is desired to go from "FULL ON" to "FULL OFF" during the modulation cycle, the amplitude of the square wave and the setting of the reflector voltage is particularly critical. See Fig. 2 for examples of correct and incorrect modulation.

2-3 Modulate the klystron with a 1000 cps square wave and increase the modulation voltage (thus increasing the amplitude of the square wave applied to the reflector).

2-4 Adjust the oscilloscope horizontal sensitivity so that a horizontal deflection is visible.

2-5 Adjust the reflector voltage and the modulation voltage to obtain an oscilloscope presentation similar to Fig. 3. In this figure, the end points of the reflector square-wave swing are, in one case, at zero power outside the mode and, in the other case, at full mode power.

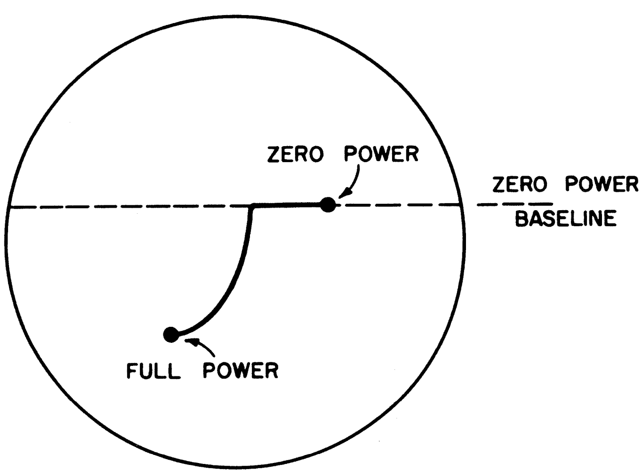


FIGURE 3

2-6 Disconnect the oscilloscope, and connect the standing-wave indicator to the output of the crystal detector. The output of the detector should now be the 1000 cps demodulation envelope. Tune the modulation frequency to peak up the standing-wave-indicator reading.

2-7 Tune the frequency meter while observing the standing-wave indicator. Watch for a dip in the indication. The frequency at which the klystron is operating can then be read directly from the dial of the frequency meter. Record this frequency on the "RESULTS" sheet. After recording the frequency, detune the frequency meter slightly to remove its reaction from the system.

2-8 The power into the crystal detector should be less than 0.1 mw to ensure square-law detector response (rf power "in" proportional to video voltage "out"). This power level can be approximated by adjusting the variable flap attenuator until the standing-wave-indicator reading is down within the 30 db range.

2-9 Using the gain control, set a reference of 0 db on the standing-wave-indicator meter scale. Tune the frequency meter until a dip is indicated, and note and record the depth of the dip (or the change of power) at the resonant point.

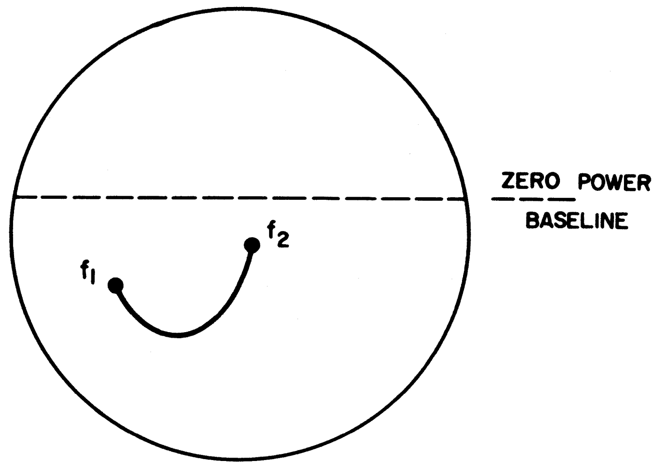


FIGURE 4

2-10 To illustrate the desirability of using an oscilloscope to interpret reflector-voltage settings, disconnect the standing-wave indicator and reconnect the oscilloscope to the crystal detector. After you have detuned the frequency meter from its setting in Step 2-9, adjust the reflector voltage for an oscilloscope presentation similar to Fig. 4. Replace the crystal-detector connection to the oscilloscope with the connection to the standing-wave indicator. Tune the frequency meter for the klystron frequency, and observe the standing-wave indicator as the meter passes through resonance. There will be a rise and a dip in indication (depending upon the direction of approach by the meter) several megacycles away from the true frequency.

The reason for this behavior can be explained with the waveform shown in Fig. 5. This waveform represents the modulated rf related to Fig. 4. If the frequency meter is tuned to f_1, the modulation envelope is reduced and the standing-wave indicator dips. As the cavity resonates at f_2, the a-c component of the envelope increases, with a similar increase on the standing-wave indicator. This situation can be eliminated by using an oscilloscope during "setup" to ensure "full on—full off" operation of the klystron.

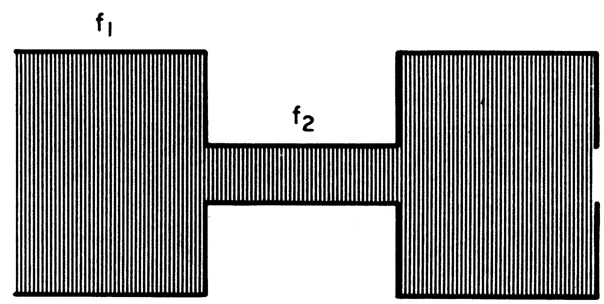


FIGURE 5

Section 3—Klystron Frequency-Tuning Log

3-1 With the oscilloscope connected to the crystal detector, modulate the klystron with a 60 cps sine wave. (Obtain the 60 cps signal for the oscilloscope horizontal input from the power line rather than from the klystron power supply. In this way, changing the modulation phase at the klystron power supply will affect only the klystron.) Adjust the oscilloscope horizontal sensitivity and the modulation voltage to obtain several modes on the oscilloscope.

3-2 Using the frequency meter "pip" indication at the mode maximum-power point, tune the klystron to 8.2 gc (as in Experiment 1).

3-3 Record the klystron-micrometer reading at 8.2 gc in Table I. Repeat this procedure for the other frequencies listed, and retain this log for approximate klystron frequencies vs. micrometer readings for use in future experiments.

3-4 If the klystron has a serial number, record the number in Table I.

Section 4—Frequency Meter Cavity Q

4-1 Tune the klystron to 10.0 gc. Continue to use 60 cps modulation, and observe the crystal detector output with the oscilloscope.

4-2 Tune the frequency meter until the "pip" appears at the maximum-power point of the largest mode. Reduce the modulation voltage until the frequency meter "pip" occupies approximately one quarter of the over-all sweep width, as shown in Fig. 6.

NOTE: It may be necessary to adjust the reflector voltage and the 60 cycle phase to keep the "pip" visible. It may also be necessary to increase the oscilloscope horizontal sensitivity so that the desired portion of the sweep occupies most of the presentation.

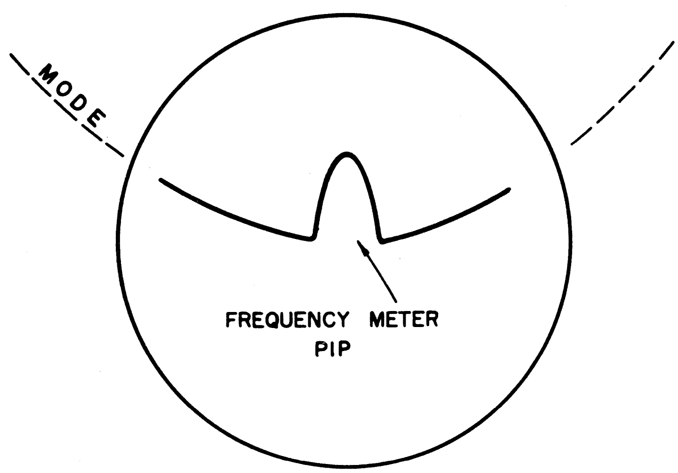


FIGURE 6

4-3 Adjust the horizontal sweep width for an arbitrary horizontal trace length of 10 cm.

4-4 Adjust the frequency meter so that the peak of its "pip" occurs at the left edge of the trace. Record this frequency reading (to nearest mc).

4-5 Adjust the frequency meter so that its "pip" occurs at the right edge of the trace. Record this frequency reading.

4-6 Using these two known points on the horizontal scale, calculate a horizontal calibration factor (megacycle per centimeter).

NOTE: The horizontal oscilloscope scale is now calibrated. Do not change the horizontal sensitivity of the oscilloscope.

4-7 Adust the frequency meter "pip" to a convenient point on the screen, and measure the width between its half-power points.

4-8 Calculate the bandwidth of the frequency-meter resonance, using the calibration factor from Step 4-6.

4-9 Calculate the "Q" of the resonant cavity by using the formula: f_0/bandwidth, where f_0 is the center frequency of the frequency meter, as set in Step 4-1. A typical frequency meter of this type should exhibit a Q of approximately 5000 in the middle of X band; however, fairly wide variations are possible.

Section 5—Slotted-Line Frequency-Measuring Technique

5-1 Using the klystron setting from Section 4 (10.0 gc), modulate the klystron with a 1000 cps square wave. Follow the procedure of Steps 2-3 through 2-5 to obtain an oscilloscope presentation similar to Fig. 3. Connect the standing-wave indicator to the broadband probe mounted in the probe carriage.

5-2 Replace the crystal detector with the sliding short. Connect the standing-wave indicator to the broadband probe mounted in the probe carriage, and adjust the probe penetration so that an indication is obtained on the standing-wave indicator.

5-3 Tune the frequency meter for a dip in the indication on the standing-wave indicator, and record the frequency.

5-4 Using the probe carriage, move the probe along the slotted section until a minimum (null) is indicated on the standing-wave indicator. It may be necessary to use the maximum sensitivity of the standing-wave indicator to obtain a precise null position. Record the horizontal position of the probe with respect to the centimeter scale on the probe carriage.

5-5 Move the probe along the slotted section until a new minimum is obtained. Record this new horizontal probe position.

5-6 From the knowledge that minimums are spaced one-half wavelength apart, calculate the guide wavelength (λ_g).

5-7 If you were to use guide wavelength and the following equation in the calculation of frequency, what would be your answer?

$$f = \frac{\text{velocity of light}}{\text{guide wavelength}}$$

5-8 Convert from guide wavelength to free-space wavelength (λ), using the equation:

$$\lambda = \frac{\lambda_g}{\sqrt{1 + \left(\frac{\lambda_g}{2a}\right)^2}}$$

where a is the width (long dimension) of the waveguide. For X-band waveguide, this width is 0.900 in. (2.285 cm).

5-9 From the free-space wavelength, calculate the frequency.

Name ____________________
Course ____________________
Date performed ____________________
Date turned in ____________________

Results (Experiment 2)

Observed		Calculated	
Step		*Step*	
2-7	________gc		
2-9	________db		
3-3	Table I		
4-4	________mc		
4-5	________mc	4-6	________mc/cm
4-7	________cm	4-8	________mc
		4-9	________
5-3	________gc		
5-4	________cm		
5-5	________cm	5-6	________cm
		5-7	________gc
		5-8	________cm
		5-9	________gc

Questions

1. Does the frequency calculated in Step 5-7 differ from the frequency measured in Step 5-3? Why?

2. Does the frequency calculated in Step 5-9 more closely agree with the frequency measured in Step 5-3? Explain.

3. What is the effect on klystron frequency if reflector voltage is increased?

4. How would you use a frequency meter to tune a klystron to some desired frequency?

5. Why must frequency-meter cavities be free of spurious responses?

6. After being used for a frequency reading, why should the frequency meter always be turned off frequency?

7. What effect might you expect on the cavity Q if you cut a larger coupling slot in the waveguide?

Would a larger coupling slot cause a larger frequency meter "pip"? ______________

Discussion

TABLE I

RF FREQUENCY (gc)	KLYSTRON MICROMETER READING
8.2	
8.5	
9.0	
9.5	
10.0	
10.5	
11.0	
11.5	
12.0	
12.4	

KLYSTRON SERIAL NUMBER ____________________

EXPERIMENT 3

Power Measurement

Object

To become familiar with the bolometer method of measuring power at microwave frequencies.

Theory

The discussion of power-measurement theory in Chapter 3 should be reviewed before this experiment is begun.

There are several possible errors associated with making power measurements. Some of them are

1. Mismatch error. The load does not present a conjugate match to the source impedance, and maximum power transfer does not take place. (Experiment 13 will discuss this in greater detail.)
2. Mount-efficiency error. This error is caused by the absorption of power in the mount between the power-sensitive bolometer and the input connector (or flange). This absorption may be due to lossy conductors, connections, or dielectrics. Mount efficiency is expressed as:

$$\frac{\text{Power dissipated in bolometer}}{\text{Total power dissipated in mount}}$$

and is typically 97 or 98 per cent for waveguide mounts. In coaxial mounts, efficiency may be about 90 per cent at 10 gc, although it may approach 100 per cent at lower frequencies.

3. Power-substitution error. Such an error is caused by d-c and microwave currents not having identical heating effects (primarily because of skin depth effects). Although heating effect is not normally significant when barretter elements are used, it will cause about a 1 per cent error with thermistors at 10 gc.
4. Instrument error. Independent of the mount or the bolometer element used, this error results from the inherent inaccuracies within the power meter. For the Model 430C it is a maximum of 5 per cent; for the Model 431A it is a maximum of 3 per cent.
5. Thermal error. This error is caused by a temperature change in the bolometer mount, which changes the calibration made for the bolometer element at its initial temperature.

References. Additional information to supplement this experiment may be found in the following texts (listed in Appendix C).

Ginzton, Chapter 3: 3.1–3.2.
King, Chapter 3: 3.1.
Reich, Chapter 8: 8-2.
Wind, Section 4.

Equipment

QUANTITY	TYPE
1	ⓗⓟ 715A klystron power supply (with cable)
1	Varian X-13 reflex klystron
1	Cooling fan or blower
1	ⓗⓟ X382A precision variable attenuator
1	ⓗⓟ 430C microwave power meter
1	ⓗⓟ X487B thermistor mount
1	ⓗⓟ X532B frequency meter
1	ⓗⓟ AC-16K cable (BNC to BNC)

Procedure

Section 1—General

1-1 Review the safety precautions on page 42, and the operating instructions for each equipment item.

1-2 Set up the equipment as shown in Fig. 1. Set the precision variable attenuator to 20 db.

1-3 Connect the thermistor mount (100 ohms resistance, negative temperature coefficient) to the microwave power meter, and apply only enough bias current to permit an indication on the power meter, controllable by the ZERO SET. In this condition, the thermistor is at its proper operating resistance.

Section 2—Thermal Effects

2-1 Zero the microwave power meter on the 1.0 mw range, and then place your hand on the thermistor mount.

2-2 Record the drift in mw.

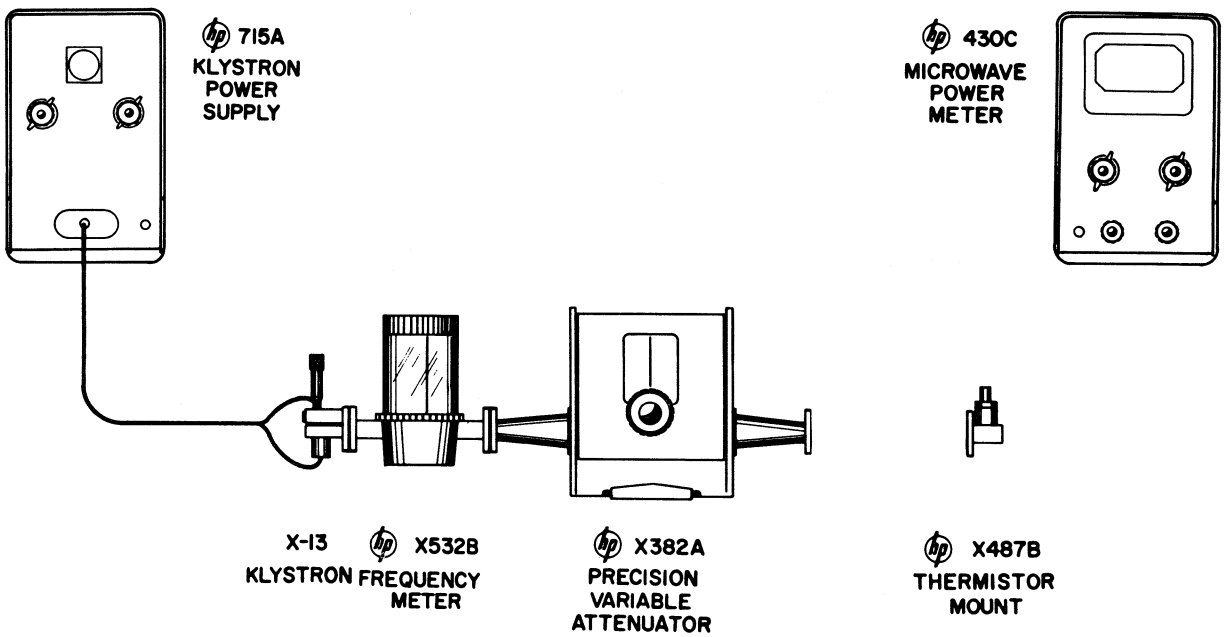


FIGURE 1

2-3 If the drift coefficient for the mount is approximately 0.15 mw per Centigrade degree, how many degrees did you warm the mount?

Section 3—Power Measurement

3-1 Attach the thermistor mount to the precision variable attenuator, allow sufficient time for the mount temperature to stabilize, and rezero the power meter.

3-2 Energize the klystron (CW), and tune the klystron for any convenient frequency.

3-3 Vary the beam voltage and the reflector voltage to the klystron to find a point of maximum output power. Record the power-meter *reading* in mw and dbm.

3-4 By increasing attenuation in the transmission system, the power level at the thermistor mount can be reduced to a suitable low level to permit zeroing the power meter. Is the 50 db attenuation of the precision variable attenuator sufficient to permit zeroing the meter?

3-5 In Experiment 1, data was taken of mode reflector voltage vs. frequency. Using a beam voltage of 400 v, measure and record the maximum klystron output power at the same frequencies (use Table I). If you are using the same klystron as in Experiment 1 or Experiment 2, you may refer to the recorded klystron micrometer settings for convenience. Otherwise, tune the klystron with the frequency meter (remembering to detune the frequency meter after setting each frequency). When the precision variable attenuator is set at 20 db to ensure efficient thermistor operation, it will also provide a convenient multiplier ($\times$ 100) for converting power-meter readings into true klystron output power levels ahead of the attenuator. In addition to making maximum-power-level measurements, record the cathode (beam) current and the reflector voltage for each frequency and mode.

3-6 Using Fig. 2, plot maximum power vs. frequency and draw lines through the points for each mode. If there is any question in identifying each mode, your data from Experiment 1 should help you.

3-7 Power into the klystron may be roughly calculated by multiplying beam voltage (400 v) by the cathode (beam) current.

An approximation of klystron efficiency can be calculated from:

$$\frac{\text{rf power out}}{\text{d-c power in}} \times 100 = \text{efficiency}$$

Calculate the efficiency at each frequency and plot your information, using Fig. 3.

Name ______________________

Course ______________________

Date performed ______________________

Date turned in ______________________

Results (Experiment 3)

OBSERVED		CALCULATED	
Step		*Step*	
2-2	____________mw	2-3	____________°C
3-3	____________mw		
	____________dbm		
3-4	____________		
3-5	Table I	3-6	Figure 2
		3-7	Figure 3

Questions

1. Is the power measured in Step 3-3 the true output power of the klystron?

2. Why is the quantity power, rather than voltage or current, more often measured at microwave frequencies?

3. To determine the zero-power reference level before a measurement is made, the bolometer is often removed from the rest of the system. If it is removed, what precautions are necessary (especially on lower-power ranges)?

4. What are two methods (other than the method in Question 3) for removing microwave power from the bolometer so that the instrument can be zeroed? When would you use each method?

5. Thermistor and barretter mounts are absorption-type devices. Can you suggest a method for monitoring system power level without absorbing all of the available power in the system?

6. From your data plotted in Fig. 3, would you consider the reflex klystron to be an efficient device?

TABLE I

	FIRST MODE		SECOND MODE		THIRD MODE		FOURTH MODE		
Frequency (gc)	P_{max}	V_R	P_{max}	V_R	P_{max}	V_R	P_{max}	V_R	I_C
8.5									
9.5									
10.5									
11.5									

P_{max} = maximum power in mw.
V_R = reflector voltage in v.
I_C = cathode (beam) current in ma.

Beam voltage = 400 v.

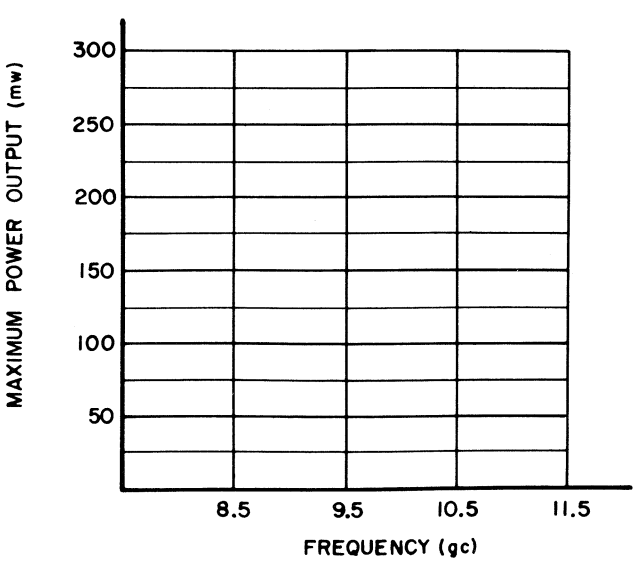


FIGURE 2

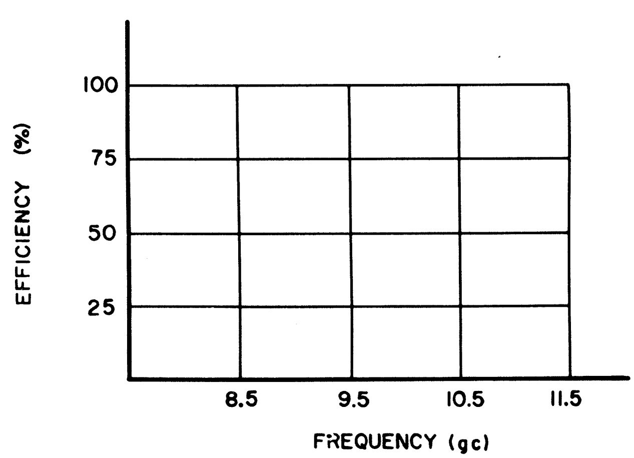


FIGURE 3

Discussion

EXPERIMENT 4

Attenuation Measurement

Object

To demonstrate the techniques for measuring attenuation in a component part of a waveguide system, and to become familiar with typical equipment used for these measurements.

Theory

Attenuation of component parts of a waveguide system can be measured by either the "power-ratio" method, or the "rf-substitution" method. In the power-ratio method, the output of the system is measured first with the unknown part in place, and then with it removed. The ratio of the two output powers, as used in the following formula, yields attenuation.

$$\text{attenuation} = 10 \log_{10} \frac{P_1}{P_2}$$

where

$P_1 =$ mw reading with unknown part out of system.
$P_2 =$ mw reading with unknown part in system.

(If the power readings are taken in db, the attenuation is merely the difference in the two readings.)

An inherent disadvantage of the power-ratio method is that the detector operates at different power levels, and any variation of the detection law must be considered in the attenuation value determined. Detector characteristics will be covered in Experiment 11; however, it should be pointed out at this time that the detector error can equal several tenths of a db in the 20 db range, as used in this experiment, even if power at the detector is kept below 0.1 mw. If the power is allowed to increase to 1 mw, the error can become as great as 1 db in 20 db.

The rf-substitution method eliminates the inherent detector error of the power-ratio procedure. In this method, the output power is held constant by replacing the attenuation to be measured with a precisely calibrated variable attenuator. The unknown attenuation is then equal to the value of the attenuation which must be inserted to obtain the original reading on the power indicator.

Both methods, as described, ignore errors caused by mismatch of various components in the system, including the changing reflection coefficient at the component being tested. Although errors of this type are discussed more fully in Experiment 13, some examples can be given here.

The hp Model X421A crystal detector used in this experiment has a maximum allowable swr of 1.5:1. If the unknown attenuation also has a 1.5:1 swr, the error caused by putting this attenuation into the line directly in front of the detector can be as great as 0.5 db. If the swr of the unknown attenuation is not greater than 1.15 (this experiment uses the hp Model X375A variable flap attenuator, which has that specification), the maximum error is reduced to 0.15 db.

The remaining errors are those associated with the precision variable attenuator (attenuation standard) and the standing-wave indicator. Specified calibration accuracy of the hp Model X382A precision variable attenuator is $\pm$ 2 per cent of the dial reading, while the hp Model 415B has an allowable cumulative range-error specification of $\pm$ 0.2 db. Error of the attenuation standard is significant in the substitution method, whereas the standing-wave-indicator range error and the crystal-detection law must be considered in the power-ratio method.

Equipment

QUANTITY	TYPE
1	hp 715A klystron power supply (with cable)
1	Varian X-13 reflex klystron
1	Cooling fan or blower
1	hp X375A variable flap attenuator
1	hp X382A precision variable attenuator
1	hp 415B standing-wave indicator
1	hp X421A crystal detector
1	hp X532B frequency meter
1	hp 120B oscilloscope
1	hp AC-16A cable (dual banana to dual banana)
1	hp AC-16B cable (dual banana to BNC)
1	hp AC-16K cable (BNC to BNC)

References. Additional information to supplement this experiment may be found in the following texts (listed in Appendix C).

Ginzton, Chapter 11.
King, Chapter 3: 3.14.
Wind, Section 3.

Procedure

Section 1—General

1-1 Review the safety precautions on page 42, and the operating instructions for each equipment item.

1-2 Set up the equipment as shown in Fig. 1. The oscilloscope horizontal input should be *a-c coupled* to the external-modulation terminals of the klystron power supply. The oscilloscope vertical input should be *d-c coupled* to the crystal detector output. The precision variable attenuator should be set to approximately 15 db of attenuation, to act as a pad isolating the klystron from the system and to protect the crystal detector from excessive power.

NOTE: A second pad of about 10 db could be placed between the "unknown" variable flap attenuator and the crystal detector. Using both pads would provide a nearly matched source and a nearly matched termination, eliminating most mismatch errors.

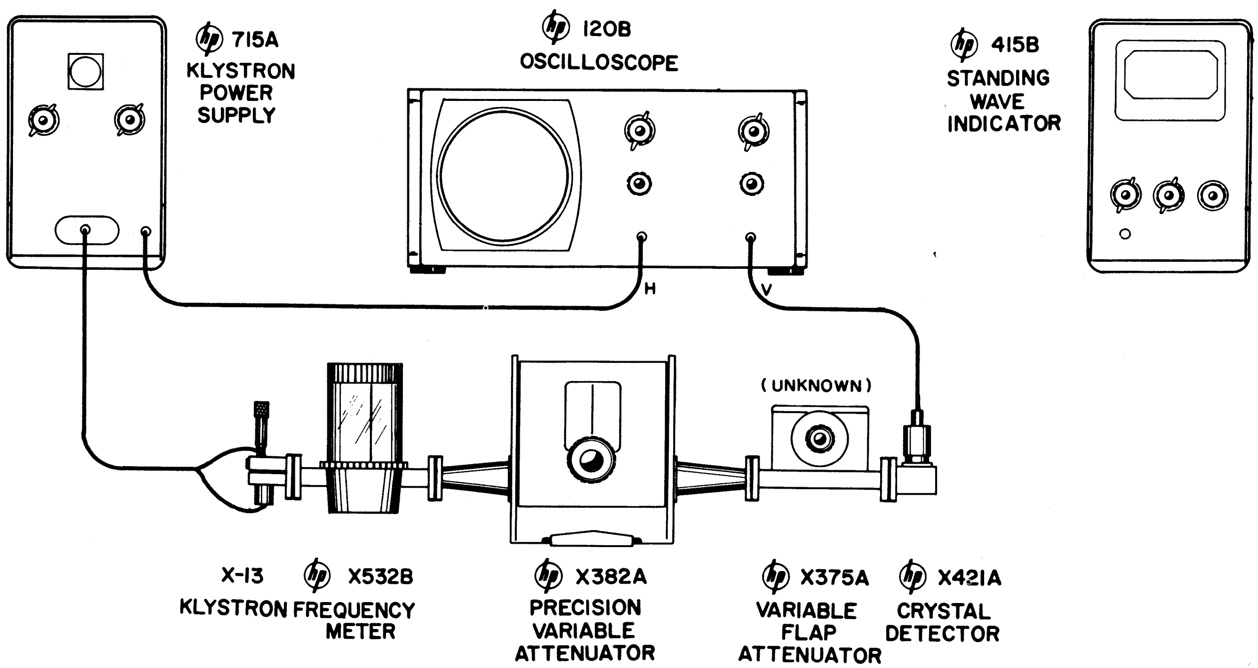


FIGURE 1

1-3 Energize the klystron at 10.0 gc and modulate it with a 1000 cps square wave. Use the oscilloscope technique of Experiment 2 to set the reflector and modulating voltages properly.

1-4 Disconnect the oscilloscope connection to the crystal detector, and, instead, monitor the detector with the standing-wave indicator. Adjust the 1000 cps modulation frequency to get a peak indication on the standing-wave indicator.

Section 2—Power-Ratio Method of Attenuation Measurement

2-1 For reliable operation, the detector must operate within its "square-law" detection range. This requirement necessitates that signal power be 0.1 mw or less at the crystal. (This signal power is comparable to a standing-wave-indicator reading on the 30 db range.) If necessary, the precision variable attenuator can be adjusted for additional attenuation to bring power to the required level.

2-2 Set the "unknown" variable flap attenuator dial to 0 db. Adjust the standing-wave-indicator gain control and the precision variable attenuator for a 0 db reading on the 30 db range of the standing-wave indicator.

2-3 Increase the attenuation dial setting of the "unknown" attenuator in 5 db steps to 20 db. Record the standing-wave-indicator reading for each step, and calculate the actual attenuation of the "unknown" for each step (Table I).

2-4 Repeat Steps 2-2 and 2-3 for klystron frequencies of 9.0, 10.0, 11.0, and 12.4 gc.

Section 3—RF-Substitution Method of Attenuation Measurement

3-1 Set the dial of the variable flap attenuator (unknown) to 0 db, and adjust the precision variable attenuator (standard) to at least 40 db. Adjust the standing-wave indicator RANGE switch (and then the precision variable attenuator) for a 0 db reading on the standing-wave indicator. Record the precision-variable-attenuator (standard) dial reading (Table II).

3-2 Set the "unknown" attenuator dial to 5 db. Reset the standing-wave-indicator reading to 0 db by decreasing the attenuation of the "standard" attenuator. *Do not* change any control on the standing-wave indicator. Record the new dial reading on the "standard" attenuator. Calculate the actual attenuation of the "unknown" from the amount of attenuation change in the "standard."

3-3 Repeat the procedure in Step 3-2 for each 5 db step on the "unknown" attenuator, at this frequency.

3-4 Repeat the procedure of Steps 3-1 through 3-3 for the other frequencies shown in Table II.

Name ______________________

Course ______________________

Date performed ______________________

Date turned in ______________________

Results (Experiment 4)

OBSERVED		CALCULATED	
Step		*Step*	
2-3	Table I	2-3	Table I
2-4	Table I	2-4	Table I
3-1	Table II	3-2	Table II
3-2	Table II	3-3	Table II
3-3	Table II	3-4	Table II
3-4	Table II		

Questions

1. If the error in resetting the "unknown" dial to the same point each time is disregarded, what was the maximum dial-calibration deviation from 0 to 10 db?

2. What was the maximum dial-calibration error from 10 to 20 db?

3. If the "unknown" dial-calibration specification is ± 1 db from 0 to 10 db, and ± 2 db from 10 to 20 db, is the "unknown" attenuator within specifications?

4. Which method of measurement would you use for a large value of attenuation, such as 40 db?

TABLE I

"Unknown" dial	8.2 gc		9.0 gc		10.0 gc		11.0 gc		12.4 gc	
	Reading	Actual	Reading	Actual	Reading	Actual	Reading	Actual	Reading	Actual
0 db										
5 db										
10 db										
15 db										
20 db										

Reading = standing-wave-indicator reading (in db).
Actual = calculated attenuation of the "unknown" (in db).

TABLE II

"Unknown" dial	8.2 gc		9.0 gc		10.0 gc		11.0 gc		12.4 gc	
	S dial	Actual	S dial	Actual	S dial	Actual	S dial	Actual	S dial	Actual
0 db										
5 db										
10 db										
15 db										
20 db										

S dial = dial reading of the standard attenuator (in db).
Actual = calculated attenuation of the "unknown" (in db).

Discussion

EXPERIMENT 5

Measuring SWR

Object

To become familiar with the basic techniques for measuring voltage standing-wave ratio.

Theory

Measurement of voltage standing-wave ratio (commonly abbreviated to swr) is a basic requirement for determining impedance at any point in a waveguide system. By measuring the swr of a load with a slotted section of waveguide, and by determining the distance between the maximum and minimum points of the standing wave, both the impedance and the reflection coefficient of the load can be determined. Relatively simple techniques can then be used to determine the impedance at points in the waveguide system other than the load.

There are several methods for measuring swr with a slotted waveguide. In the most straightforward method, the swr can be read directly on the hp Model 415B standing-wave indicator. For swr higher than about 10, the straightforward method of measurement can result in erroneous readings. When the swr is high, probe coupling must be increased if a reading is to be obtained at the voltage minimum. However, at the voltage maximum, this high coupling may result in a deformation of the pattern, with consequent error in reading. In addition to the error caused by probe loading, there is also danger of error resulting from the change in detector characteristics at higher rf levels. Two other techniques, the double-minimum method and the calibrated-attenuator method, can be used to obtain greater accuracy.

In the double-minimum method, it is necessary to establish the electrical distance between the points at which the output is double the minimum (see Fig. 1). In the calibrated-attenuator method, a calibrated variable rf attenuator is used between the signal source and the slotted sec-

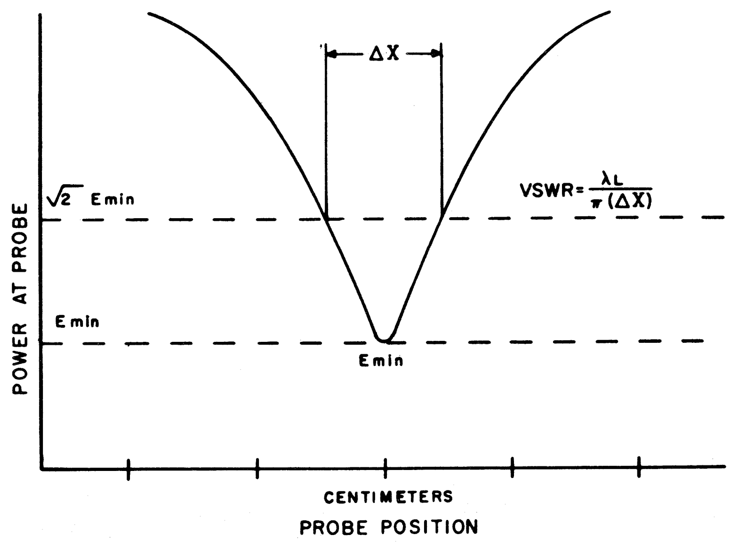


FIGURE 1

tion, and is adjusted to keep the rectified output of the crystal detector equal at the voltage-minimum and the voltage-maximum points. The swr in db is then merely the difference in the attenuator settings.*

Signal sources can introduce at least three undesirable characteristics which will affect slotted line measurements. These include presence of rf harmonics, fm, and spurious signals. Signal sources used for standing-wave measurements should have relatively low harmonic content in their output. The swr at a harmonic frequency may be considerably higher than at the fundamental. Spurious frequencies in the signal source are also undesirable, for, unless they are very slight, they will obscure the minimum points at high swr values. Figure 2 shows the plot of a swr pattern made with the signal source producing unwanted fm.

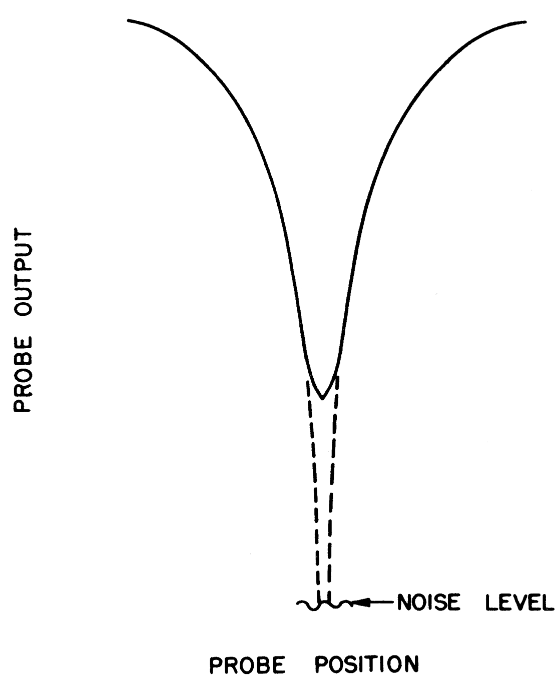


FIGURE 2

Instances are common in which the presence of rf harmonics has led to very serious errors in swr measurements. Such harmonics are usually present to an excessive degree only in signal sources that have coaxial outputs. Coaxial pickups of a broadband type will often pass harmonic frequencies with greater efficiency than the fundamental. In waveguide systems, signal sources such as internal-cavity klystrons' have a more or less fixed coupling and, in addition, do not have pickups extending into the tuned cavity to cause agitations of the cavity fields. Consequently, the harmonic problem is generally limited to coaxial systems. Harmonics become especially troublesome when the reflection coefficient of a load at a harmonic frequency is much larger than it is at the fundamental frequency—a common condition. When the harmonic content of the signal source is high, the large reflection coefficient of the load at the harmonic frequency can cause the harmonic standing-wave fields to be of the same order of magnitude as the fields at the funda-

* These methods are briefly described in the *Model 415B Operating and Service Manual.*

mental frequency. Thus, a device having a swr of 2.0 at the fundamental frequency will often have a swr of 20 or more at the second harmonic frequency. If such a device is driven from a signal source having, say, 15 per cent second harmonic content, the peaks of the standing waves of second harmonic will be about one-fourth the amplitude of the peaks at the fundamental frequency. Figure 3 shows a typical swr pattern obtained when the rf signal contains harmonics.

References. Additional information to supplement this experiment may be found in the following texts (listed in Appendix C).

Ginzton, Chapter 5.
King, Chapter 6.
Reich, Chapter 8: 8-7.
Wind, Section 2.

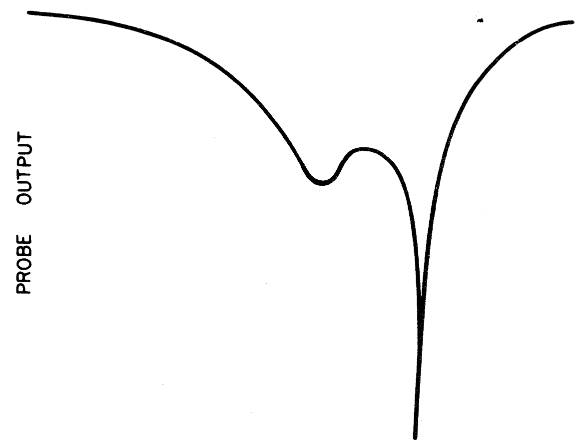


FIGURE 3

Equipment

QUANTITY	TYPE
1	hp 715A klystron power supply (with cable)
1	Varian X-13 reflex klystron
1	Cooling fan or blower
1	hp X375A variable flap attenuator
1	hp X382A precision variable attenuator
1	hp 415B standing-wave indicator
1	hp 430C microwave power meter
1	hp X487B thermistor mount
1	hp X532B frequency meter
1	hp 809B probe carriage
1	hp 444A broadband probe
1	hp X810B slotted section
1	hp 120B oscilloscope
1*	hp X421A crystal detector
1*	hp X914B moving load
1	hp AC-16A cable (dual banana to dual banana)
1	hp AC-16B cable (dual banana to BNC)
2	hp AC-16K cable (BNC to BNC)

* Optional.

Procedure

Section 1—General

1-1 Review the safety precautions on page 42, and the operating instructions for each equipment item.

1-2 Set up the equipment as shown in Fig. 4. The oscilloscope horizontal input should be *a-c coupled* and the vertical input should be *d-c coupled.* The variable flap attenuator should be set at 10 db and the precision variable attenuator should be set at 10 db.

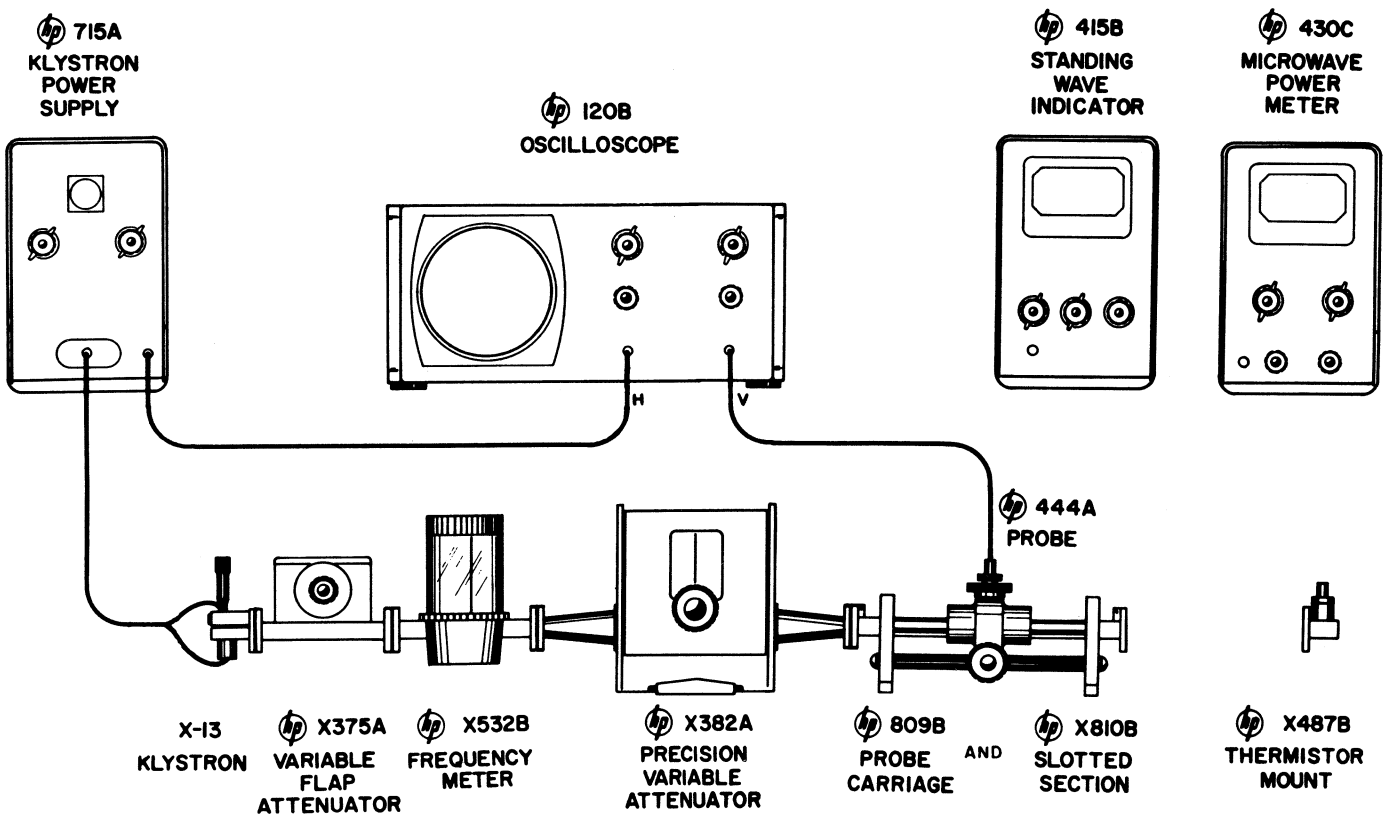


FIGURE 4

1-3 Connect the thermistor mount (100 ohms resistance, negative temperature coefficient) to the microwave power meter, and apply only enough bias current to permit an indication on the power meter, controllable by the ZERO SET. Attach the thermistor mount to the slotted section.

1-4 Energize the klystron at 10.0 gc and modulate it with a 1000 cps square wave. Use the oscilloscope technique of Experiment 2 to set the reflector and modulating voltages properly.

Section 2—Measuring Low SWR

2-1 Connect the standing-wave indicator to the broadband probe in place of the oscilloscope connection. Adjust the 1000 cps modulation frequency to get a peak indication on the standing-wave indicator. Move the probe along the slotted section until the standing-wave indicator reveals the voltage-maximum point. (It may be necessary to adjust the probe penetration to keep the standing-wave indicator "on scale" on either the 30 or 40 db range.)

2-2 Adjust the standing-wave-indicator gain until its meter indicates a swr of 1 (full scale).

2-3 Move the probe along the slotted section to obtain a voltage-minimum reading on the standing-wave indicator.

NOTE: Do not change the range switch or gain controls on the standing-wave indicator.

2-4 Read and record the swr, which is indicated directly on the standing-wave indicator.

2-5 Using the procedure of Steps 2-1 through 2-4, measure the swr at 8.5 and 11.5 gc. (For accurate measurements, it is good practice to take several readings with different amounts of probe penetration to detect any probe-loading error in the standing-wave pattern.)

NOTE: A graph of swr in db vs. voltage swr is shown in Fig. 5.

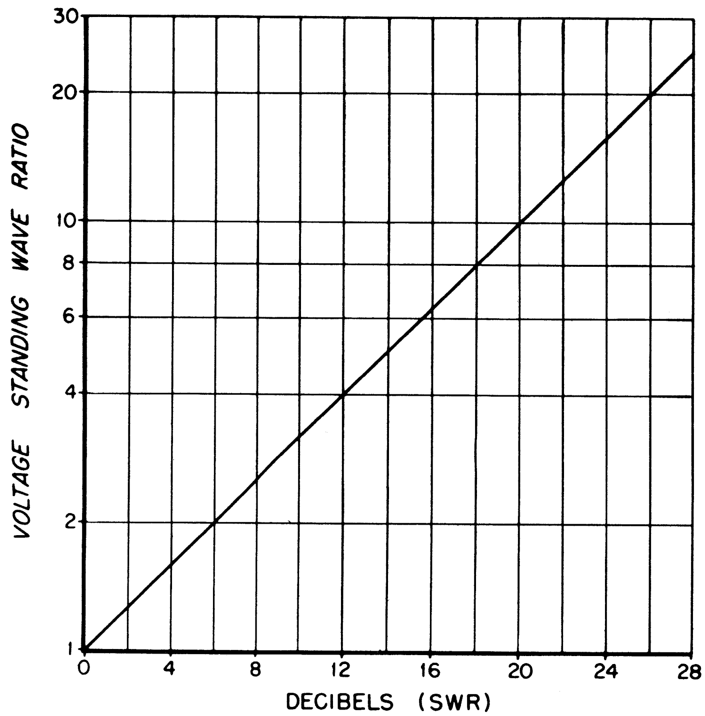


FIGURE 5

Section 3—Measuring High SWR; Calibrated-Attenuator Method

3-1 With the klystron set at 10.0 gc, remove all bias current from the thermistor mount. In this condition, the thermistor is not properly matched to the system and causes a high swr to exist.

3-2 Move the broadband probe along the slotted section until the standing-wave indicator reveals a voltage-minimum point. Adjust the precision variable attenuator for a convenient reference indication on the standing-wave indicator, and record the attenuator dial setting.

3-3 Move the probe along the slotted section until a voltage maximum is indicated. Adjust the precision variable attenuator until the standing-wave indicator is at the same reference used in Step 3-2. Record this new attenuator dial setting.

3-4 The swr (in db) is merely the difference in the attenuator settings of Step 3-2 and Step 3-3. Record this swr.

NOTE: Although this method overcomes the effect of detector variations from a square-law characteristic, the effect of probe loading still remains. Be careful; always use minimum probe penetration.

3-5 Using the procedure of Steps 3-1 through 3-4, measure and record the swr at 8.5 and 11.5 gc.

Section 4—Measuring High SWR; Double-Minimum Method

4-1 At 11.5 gc, with all bias current removed from the thermistor mount (as in Section 3), move the probe along the slotted section until a voltage minimum is indicated. Note and record the probe position with respect to the centimeter scale on the probe carriage (Table I).

4-2 For reference, adjust the standing-wave-indicator gain to obtain a reading of 3.0 on the db meter scale.

4-3 Move the probe along the slotted section to each side of the reference minimum, and obtain a full-scale reading (0) on the db meter scale. Record the probe positions at these two equal readings as d_1 and d_2.

4-4 Remove the termination (thermistor mount), short the slotted section, and measure the distance between successive minima. Twice this distance is λ_g, the guide wavelength.

NOTE: Short circuits may be obtained by using a brass plate to short the end of the guide. (Open circuits, although they reflect some power, are not as reliable or repeatable because of fringing fields. They should not be used for reference planes.)

4-5 The swr can then be obtained by substituting the distance λ_g into the expression:

$$\text{swr} = \frac{\lambda_g}{\pi(d_1 - d_2)} = \frac{\lambda_g}{\pi(\Delta x)}$$

where λ_g is the guide wavelength; d_1 and d_2 are the locations of the double-minimum points.

NOTE: This method overcomes the effect of probe loading, since the probe is always set around a voltage minimum where larger probe loading can be tolerated. However, it does not overcome the effect of detector characteristics.

4-6 Using the procedure of Steps 4-1 through 4-5, measure the swr at 8.5 and 10.0 gc.

Section 5—SWR of Different Terminations (Optional)

5-1 Measure the swr with the system terminated by a crystal detector (X421A).

5-2 Measure the swr with the system terminated by a moving load (X914A).

Name ______________________
Course ______________________
Date performed ______________________
Date turned in ______________________

Results (Experiment 5)

Observed			Calculated		
Step			*Step*		
2-4	________:1				
	________db				
2-5	________:1	(8.5 gc)			
	________db				
	________:1	(11.5 gc)			
	________db				
3-2	________db				
3-3	________db		3-4	________db	
				________:1	
3-5	________db	(8.5 gc)		________db	
	________db			________:1	
	________db	(11.5 gc)		________db	
	________db			________:1	
4-1	Table I				
4-3	Table I				
4-4	Table I		4-5	Table I	
			4-6	Table I	

Questions

1. Does your experimental data indicate that the swr due to a termination varies with frequency?

2. If you were measuring a high swr and you wished to minimize error due to square-law variations in the detector, what method of measurement would you use?

3. What type of error *can* enter into the method selected in Question 2?

4. List at least three conditions that could cause unwanted reflection of incident power in your experiments.

5. Can you suggest a means of measuring the swr of only the load element in the moving load used in Step 5-2?

TABLE I

STEP ⟶	4-1	4-3		4-4			4-5	
frequency (gc)	REFERENCE MINIMUM (cm)	d_1 (cm)	d_2 (cm)	FIRST MINIMUM (cm)	SECOND MINIMUM (cm)	λ_g (cm)	SWR :1	db
8.5								
10.0								
11.5								

Discussion

EXPERIMENT 6

Introduction to the Smith Chart

Object

To become familiar with the theory of the Smith Chart, and to learn how it can be used for microwave computations.

Theory

The Smith Chart is just as important a tool for determining the characteristics of a waveguide system as are the various items of equipment used to measure frequency, attenuation, and swr. In fact, because impedance cannot be measured directly, it is only by using the Smith Chart that measurements of swr made with a slotted section can be translated into needed information on impedance without resorting to lengthy and laborious calculations.

A uniform, infinitely long line presents a uniform impedance (Z_0) at all points along the line. This impedance causes the ratio between the impressed voltage and the current flowing at any point to be related by that Z_0. If a wave moving in a direction opposite to the normal transmitted wave is excited in the line, the interaction of the two waves causes current and voltage distributions which are no longer directly related to the characteristic impedance. Instead, current and voltage distributions vary at all points along the line.

The primary purpose of transmission lines is to transmit power efficiently from one point to another. For this reason, most of the effort in microwave design and measurement is concerned with the impedance relationships occurring on a transmission line, with respect to the generators and loads placed on each end of that line. These impedances have been developed mathematically in Section 2. Various equations point out the relation between the impedance at a particular point on the line as a function of the characteristic impedance, the load impedance, and the position on the line itself.

The equation for finding impedance at a particular point (see Eq. 26, page 16) is not easy to use, nor does it present a good physical picture of the impedance changes along the line. Early attempts to provide some sort of graphic approach to the transmission-line relations were concerned with the well-known reactance charts (Fig. 1).

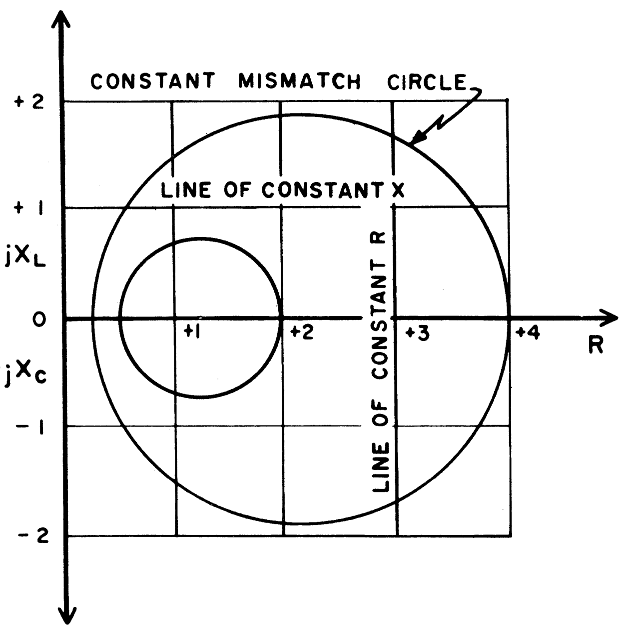


FIGURE 1

In Fig. 1, the reactance axis is vertical, whereas the resistance axis is horizontal. In the chart shown, the scales are normalized so that a characteristic impedance of 1 occurs at the 0, 1 point. Several constant-mismatch circles are plotted on this chart by substitution in the preceding equation. For instance, the inside circle is based on a mismatch of 2 to 1 where the load resistance (Z_L) would be 2. This, when applied to the preceding equation, results in the circle as shown. The outer circle is that of a mismatch of 4 to 1.

Although the latter approach is more graphic than the equation method, it still has several distinct drawbacks:

1. All values of impedance out to infinity are not presented on the chart.
2. It is hard to interpolate between constant swr circles, since the pattern is in an ever-expanding-circle array.
3. Phase-angle indications are not radial.

The contribution made by Phillip H. Smith in the late 1930's was a chart which avoids all three of the reactance-chart disadvantages. The Smith Chart, a conformal transformation of the reactance chart just mentioned, simply pulls the infinite points of reactance around until they join at a common "infinity" point on the right-hand side of the chart. Described another way, the chart is a special kind of impedance co-ordinate system, mechanically arranged so that the relationships of impedances can be determined at any point along a transmission line. Only the swr and the positions of the voltage maximum and voltage minimum need be known to make use of the Smith Chart.

All values of resistance and reactance on a Smith Chart are normalized. Multiplication by the characteristic impedance of the line will convert these normalized values to actual values.

Circles tangent to the right-hand side of the chart are circles of constant resistance. The constant resistance circles for normalized resistances of 0, 0.3, 1.0, 3.0 and 10.0 are emphasized in Fig. 2. Note that the resistance values are given along the center line of the chart.

Curved lines starting from the right-hand (infinity) side of the chart and going above and below the center are the reactance co-ordinates. The horizontal center line represents zero reactance. Curved lines going above center represent positive reactance; lines going below represent negative reactance. Note that the normalized-reactance values are given along the outer limits of the reactance lines. Positive and negative reactance co-ordinates are emphasized in Fig. 3 for values of 0.3, 1.0, 3.0, and 10.0.

Any specific value of impedance can be located on the Smith Chart by locating the proper co-ordinate positions. For example, the normalized impedance of a line terminated in its characteristic impedance would be $1.0 + j0$. On the Smith Chart this point would be at 1.0 on the horizontal center line. A nor-

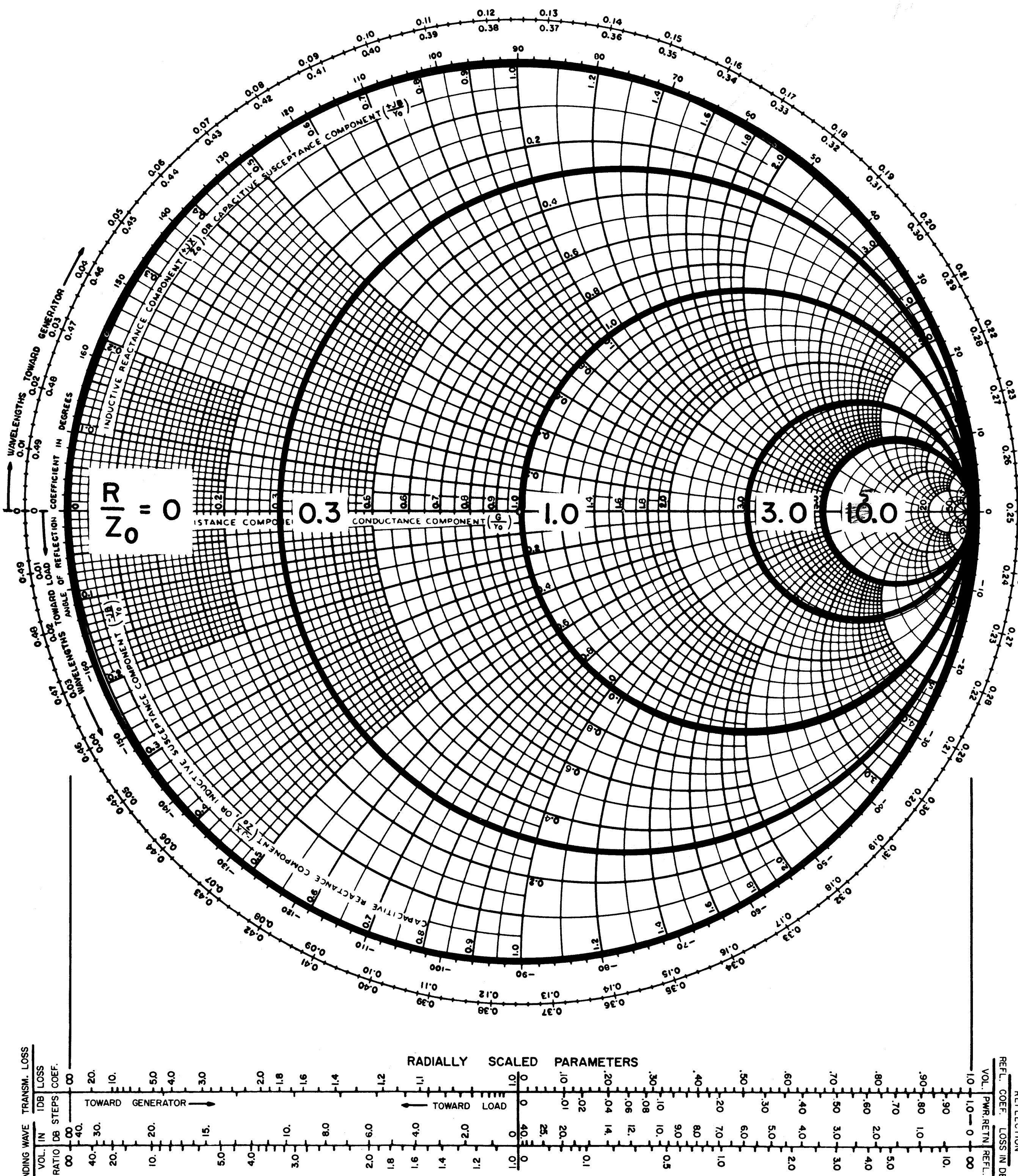

WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE COMPONENT (+jB/Yo)
CAPACITIVE REACTANCE COMPONENT (-jX/Zo), OR INDUCTIVE SUSCEPTANCE COMPONENT (-jB/Yo)
CONDUCTANCE COMPONENT (G/Yo)
R/Zo = 0
0.3
1.0
3.0
10.0
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
CENTER
STANDING WAVE VOL. RATIO
IN DB
TRANSM. LOSS 1DB STEPS
LOSS COEF.
REFLECTION
REFL. COEF. VOL.
PWR.
RETN LOSS IN DB
REFL. LOSS IN DB

FIGURE 2

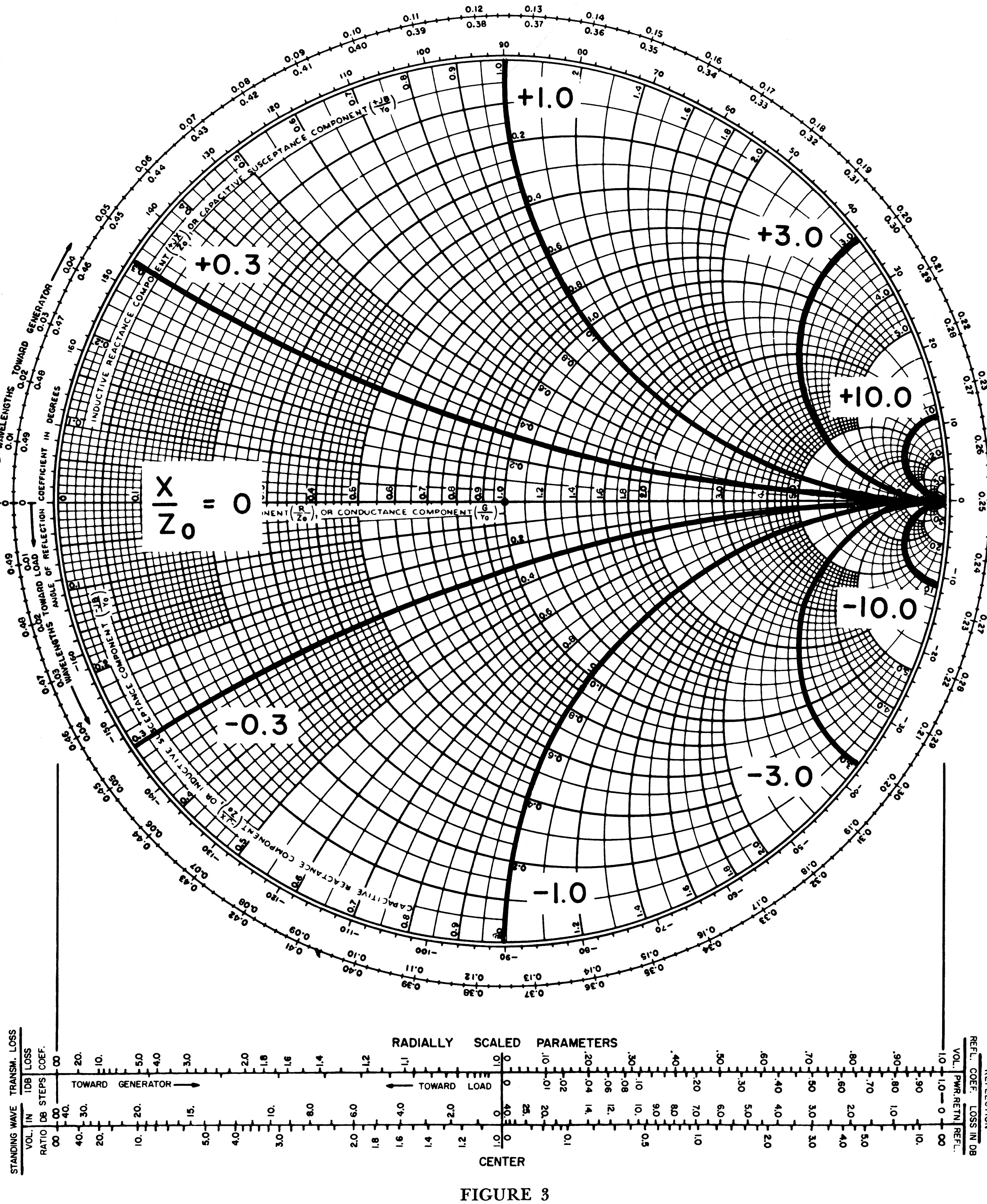

+1.0
+3.0
+0.3
+10.0
X/Z0 = 0
-10.0
-0.3
-3.0
-1.0
WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
INDUCTIVE REACTANCE COMPONENT (+jX/Z0), OR CAPACITIVE SUSCEPTANCE COMPONENT (+jB/Y0)
CAPACITIVE REACTANCE COMPONENT (-jX/Z0), OR INDUCTIVE SUSCEPTANCE COMPONENT (-jB/Y0)
RESISTANCE COMPONENT (R/Z0), OR CONDUCTANCE COMPONENT (G/Y0)
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
STANDING WAVE
TRANSM. LOSS
REFLECTION
CENTER

FIGURE 3

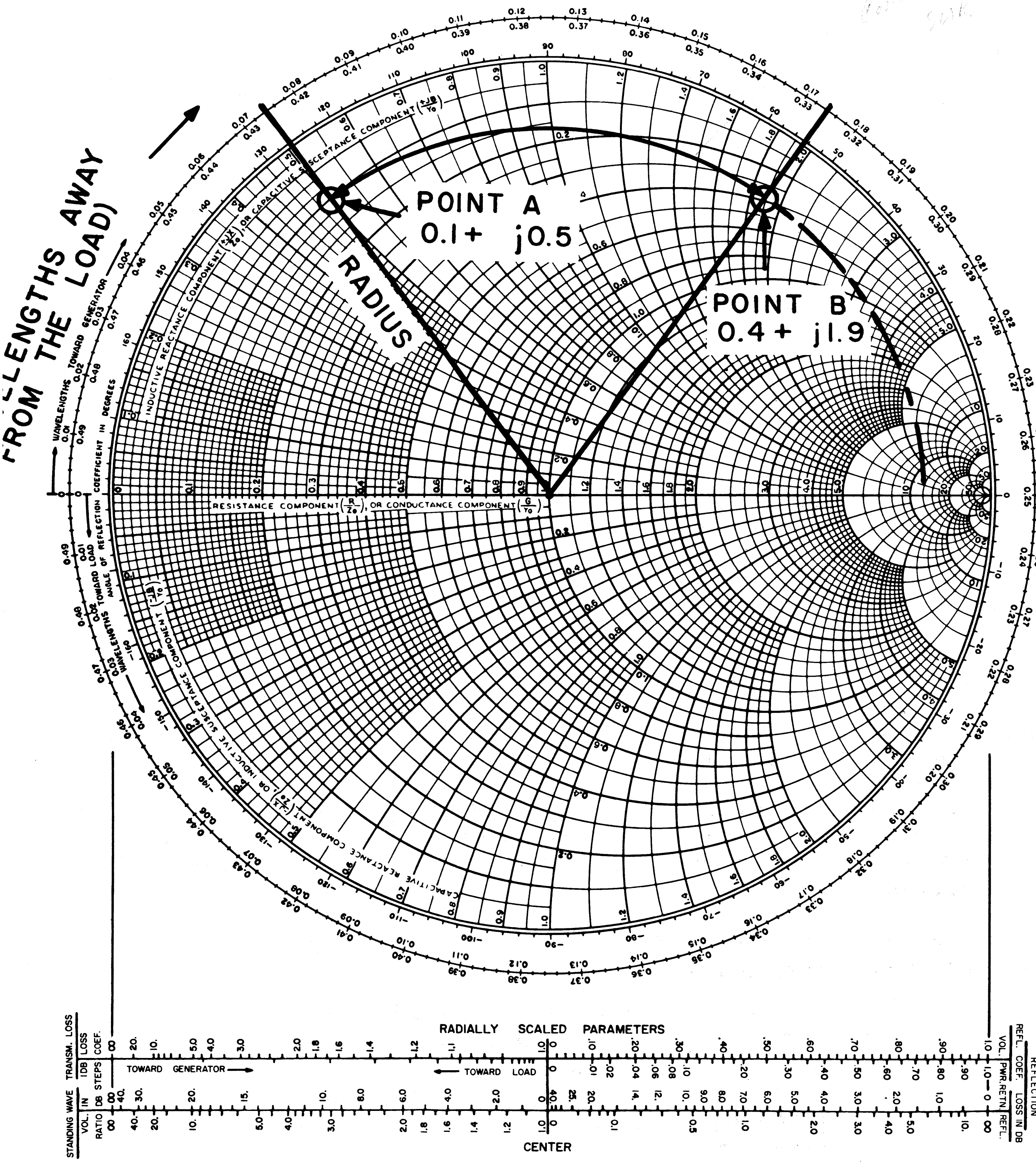

LENGTHS AWAY
FROM THE LOAD)
POINT A
0.1+ j0.5
RADIUS
POINT B
0.4 + j1.9
RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
CENTER

FIGURE 4

malized impedance of $0.1 + j0.5$ is shown at point A in Fig. 4. (Note that the chart values of resistance are always positive, and that the positive reactance values are plotted on the upper half of the chart, whereas negative reactance values are plotted on the lower half of the chart.)

One of the inherent characteristics of the Smith Chart is that any given lossless uniform line is plotted as a concentric circle about the point $1 + j0$, and the radius of the circle is determined by the known amount of impedance at any point in the line. (Thus, a single impedance value is all that is required to enter the chart.) The phase relationship along a transmission line is represented by movement around the Smith Chart on a concentric circle, and the electrical distance on the line is also directly related to the distance around the circle. To allow convenient measurement of circle movement, the Smith Chart has two circular scales on its outer edge. One is calibrated in fractional wavelength; the other in degrees.

The wavelength scale shows that a complete revolution on the chart is equivalent to a half wavelength. Thus, 180 electrical degrees on the transmission line are represented by 360 degrees of revolution on the chart. The degree scale shows that, in a complete revolution of the chart, the reflection coefficient goes through a complete cycle of 180 degrees positive and 180 degrees negative. These scales are important, because the impedance varies cyclically along a line terminated in other than its characteristic impedance. They are used to determine the impedance at various points along a line after the impedance is determined for any one specific point. The entire variation in impedance on a line is repeated cyclically in each half wavelength along the line.

Example 1. Consider that point A in Fig. 4 ($0.1 + j0.5$) represents the impedance of the load and that impedance information is desired for a point 0.1 wavelength away from that load. This information can be obtained by swinging an arc from the center of the chart (1.0 on the center line) through point A in a clockwise direction. (Note that "wavelengths away from the load" is equivalent to "wavelengths toward the generator.") Radius of the arc is the distance from the center to point A. A line is then drawn from the center to the point on the outer scale 0.1 wavelength away from point A. The intersection of this line and the arc at point B in Fig. 4 represents the desired impedance. It can be read from the chart as $0.4 + j1.9$.

The arc drawn in this example illustrates another fundamental and important property of the Smith Chart. Any circle drawn around the chart center is a circle of constant swr. The value of the swr is determined by the point at which the circle crosses the center line. The swr can be read directly on the right half of the line, or reciprocal of swr can be read on the left half. Extension of the arc in Fig. 4 shows a swr of about 12:1.

Another useful feature of the Smith Chart is that *admittance* can be determined at any point for which the impedance is known. It is necessary only to locate the point on the swr circle ¼ wavelength away from the *impedance* position. The normalized admittance can then be read directly from the conductance and susceptance co-ordinates.

Practice Problem 1. A uniform coaxial line with a normalized impedance of 1 has a normalized load of $0.2 - j0.2$ (Fig. 5).

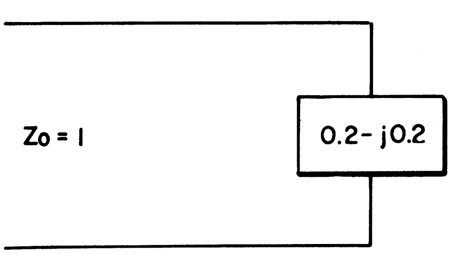


FIGURE 5

1. What is the normalized impedance 0.25 wavelength *toward* the generator?
2. What is the swr on the line?
3. How far from the load (in wavelengths) is the first maximum on the standing-wave pattern?

Practice Problem 2. This problem involves normalizing as well as calculating wavelengths; in the wavelength calculation it is assumed that the line is a coaxial air-dielectric line which has the velocity of propagation equal to the speed of light. (Coaxial lines with dielectric material between conductors or any type of waveguide have different velocities.)

A 50 ohm coaxial air line is loaded with 100 ohms of resistance in series with 40 ohms of inductive reactance. The frequency is 3000 mc.

1. What is the wavelength?
2. What is the normalized impedance of the load?
3. What is the swr on the line?
4. What is the impedance 2.5 cm toward the generator? At 12.5 cm toward the generator?

Practice Problem 3. The coaxial air line of Problem 2 is open circuited (maximum impedance). Enter at this point on the Smith Chart, and go around the constant swr circle until you reach an impedance minimum.

1. What electrical distance is this?
2. What is the swr, as read off the horizontal center line?
3. Does the impedance of this line ever have a real component of resistance?
4. How far down the line (from the open) is the point at which the normalized reactance is $+j1.0$?

Use with slotted-line technique. In most practical cases, the actual impedance values are not known. Usually, the impedance is measured in an indirect way by sampling the standing-wave pattern appearing on a transmission line (by using a sliding probe to sample part of the magnetic or electric field in the line). This sampling results in a measurement of the swr and, of course, immediately defines a constant swr circle. For the case of a lossless line, the Smith Chart has a repeating quality in that the impedance is repeated every ½ wavelength. Therefore, the slotted-line probe need not be inserted exactly at the load, but instead may be inserted down the line some distance.

A simple measurement of swr is not enough to define an impedance, since impedance is directly related to its position on the line. Solution of impedance problems also requires measuring the distance between voltage minima, first with the load connected and then with the plane of the load shorted. Figure 6 shows the standing waves which exist on a line for the conditions: (A) normal load and (B) a short in place of the load. The voltage-minimum point for the shorted condition has a different position from the minimum with load. This minimum shift (from loaded to shorted line) contains the desired phase information of the load impedance.

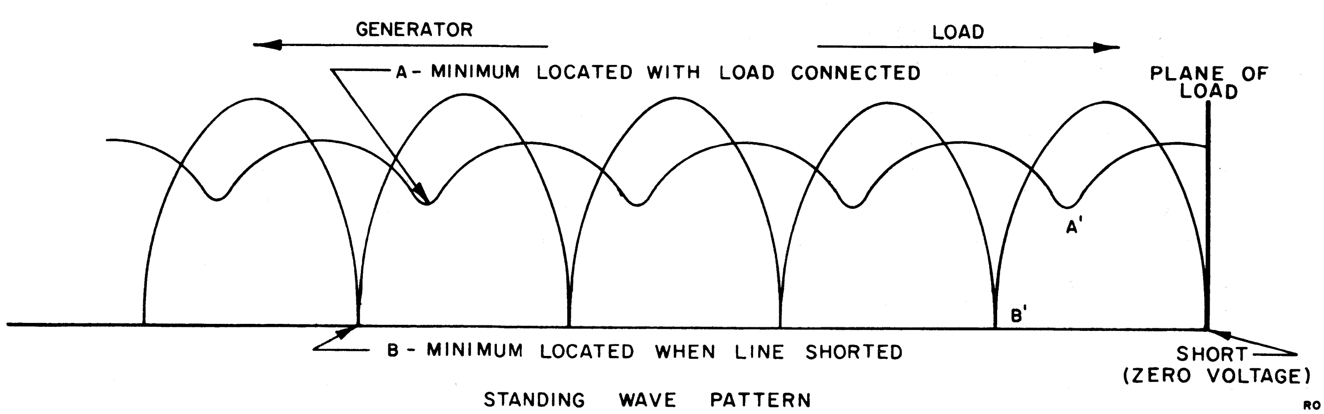


FIGURE 6

There are a number of rules which are helpful in interpreting slotted-line measurements and in using the measured data in Smith Chart plots. The rules are stated below and summarized in Fig. 7.

1. The shift in the minimum after the load has been shorted is never more than $\pm \frac{\lambda}{4}$.
2. If shorting the load causes the minimum to move toward the load, a capacitive component exists in the load.
3. If shorting the load causes the minimum to shift toward the generator, an inductive component exists in the load.
4. If shorting the load causes no shift in the minimum, a completely resistive load exists equal to Z_0/swr.
5. If shorting the load causes the minimum to shift exactly $\lambda/4$, the load is completely resistive and has a value of (Z_0)(swr).
6. When the load is shorted, the minimum will always be a multiple of $\lambda/2$ from the load.

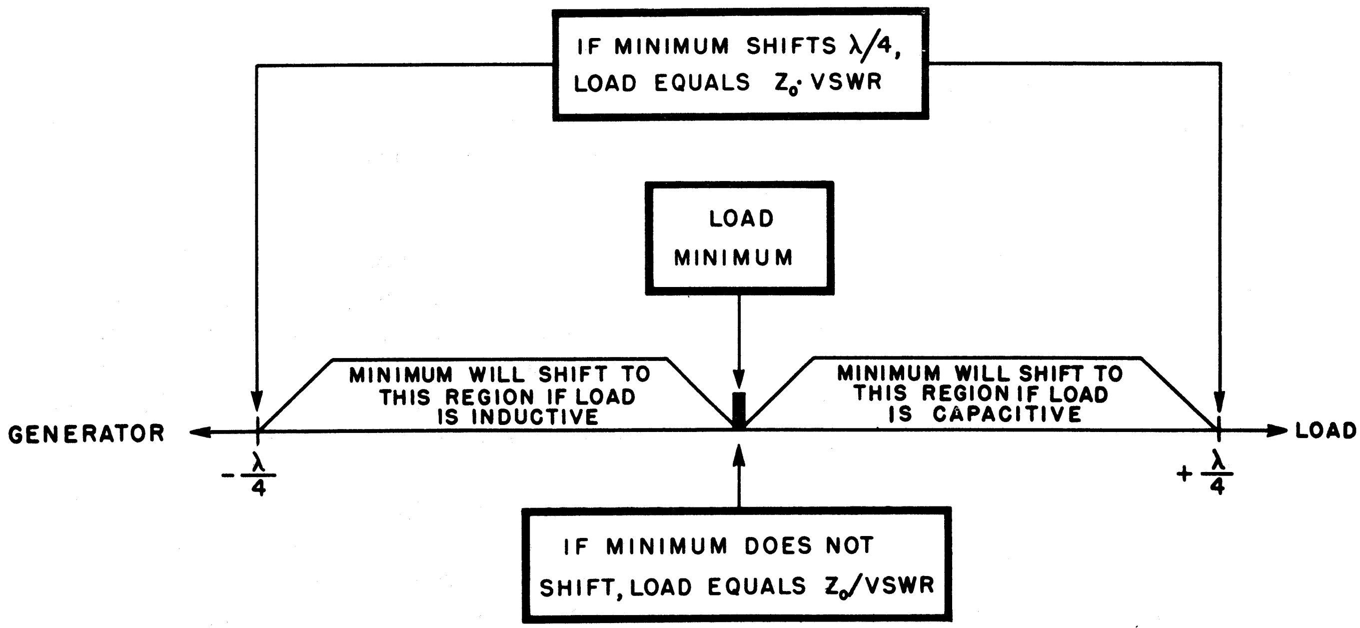


FIGURE 7

Example 2. A coaxial air-dielectric line is used in measuring an unknown load at some undetermined distance from the load. The swr is 3.3, and the source frequency is 3000 mc. Replacing the load with a short causes the voltage minimum (null) to move 1.0 cm toward the generator. What is the impedance of the load? For the solution, refer to Fig. 8.

(a) Draw the swr circle (swr = 3.3).
(b) Determine the number of fractional wavelengths represented by 1 cm. (Wavelength at 3000 mc is 10 cm if coaxial line is used; hence, 1 cm = 0.1 wavelength.)
(c) Draw a line from the chart center to the point on the outer scale representing 0.1 wavelength toward the generator (use rules of Fig. 7).
(d) Read the impedance of the load as $0.44 + j0.63$.

Practice Problem 4. A slotted-line measurement of a load shows the swr to be 1.6, and exhibits a null shift of 3 cm toward the load (loaded to shorted condition). The source frequency is 1000 mc on the coaxial air line.

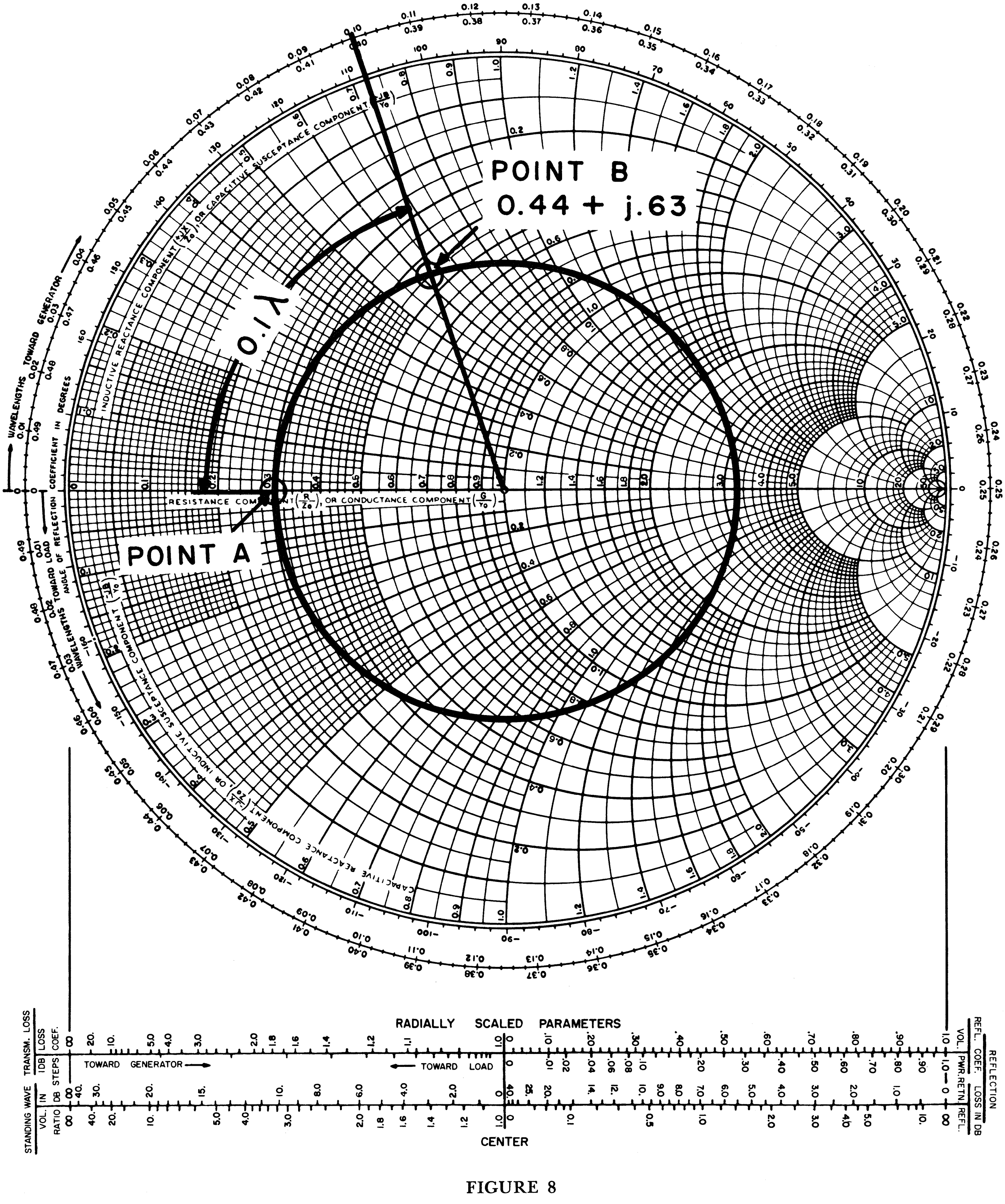

POINT B
0.44 + j.63
POINT A
0.1λ
WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
INDUCTIVE REACTANCE COMPONENT
OR CAPACITIVE SUSCEPTANCE COMPONENT
RESISTANCE COMPONENT
OR CONDUCTANCE COMPONENT
OR INDUCTIVE SUSCEPTANCE COMPONENT
CAPACITIVE REACTANCE COMPONENT
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
CENTER
STANDING WAVE
VOL. RATIO
IN DB
TRANSM. LOSS
1DB STEPS
LOSS COEF.
REFLECTION
REFL. COEF.
VOL.
PWR.
RETN. LOSS IN DB
REFL.

FIGURE 8

1. What is the impedance of the unknown at the shorted reference plane?
2. If you changed frequency and found that the swr remained at 1.6, what would you expect with regard to the nature of the unknown load beyond the reference plane?
3. In Question 2, what are the two possible normalized values of that load?

In all waveguides and some coaxial lines, the wavelength does not equal the free-space wavelength. In such a situation, it is common procedure to obtain $\lambda/2$ by physically measuring the distance between two adjacent nulls. There results a wavelength measurement for that particular line without any need for velocity of propagation calculations.

Practice Problem 5. Using a Model X810B waveguide slotted section for a slotted-line measurement results in a swr of 1.9, and a "load" minimum located at 13.0 cm (on the Model 809B probe carriage scale). The minimums measured with a short are located at 11.0 cm and 16.0 cm.

NOTE: On the Model 809B, the centimeter scale is calibrated from the right or "load" end.

1. What is the value of $\lambda/2$?
2. How much did the null shift (in fractional wavelengths, 0.25λ or less)? In which direction?
3. What is the impedance of the unknown at the reference plane?

Practice Problem 6. A slotted-line measurement yields a swr of 5.1. The "shorted" minimums occur at 9.2 cm and 12.4 cm. The "loaded" minimum occurs at 11.6. What is the normalized impedance?

References. Additional information to supplement this experiment may be found in the following texts (listed in Appendix C).

Ginzton, Chapter 4: 4.9.
Reich, Chapter 3: 3-22–3-27.
Wind, Appendix A.

Name ____________________
Course ____________________
Date performed ____________________
Date turned in ____________________

Results (Experiment 6)

Practice Problem 1

1. ____________
2. ____________:1
3. ____________

Practice Problem 2

1. ____________cm
2. ____________
3. ____________:1
4. ____________ (at 2.5 cm)
   ____________ (at 12.5 cm)

Practice Problem 3

1. ____________
2. ____________:1
3. ____________
4. ____________

Practice Problem 4

1. ____________
2. ____________
3. ____________
   ____

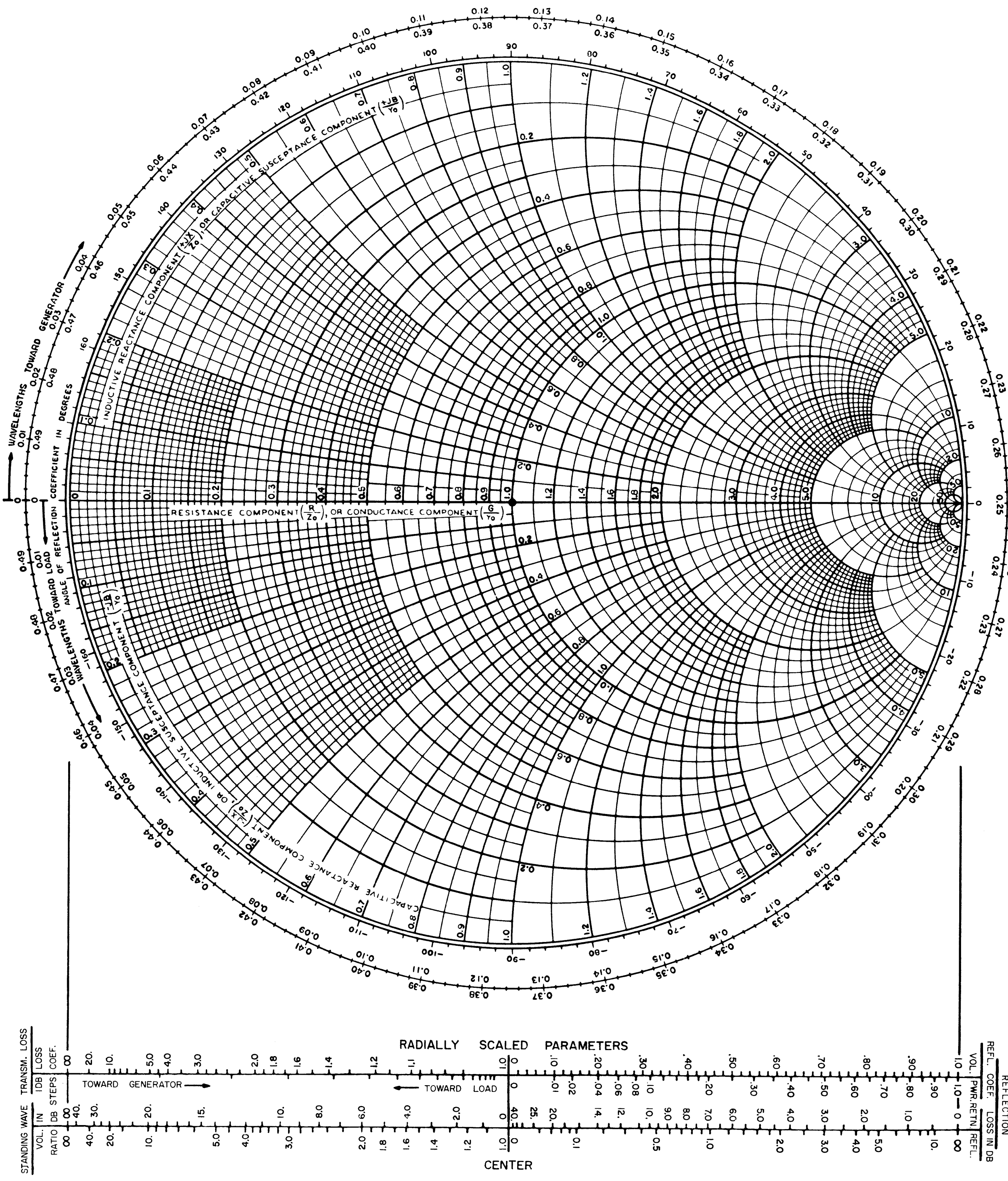
Practice Problem 5

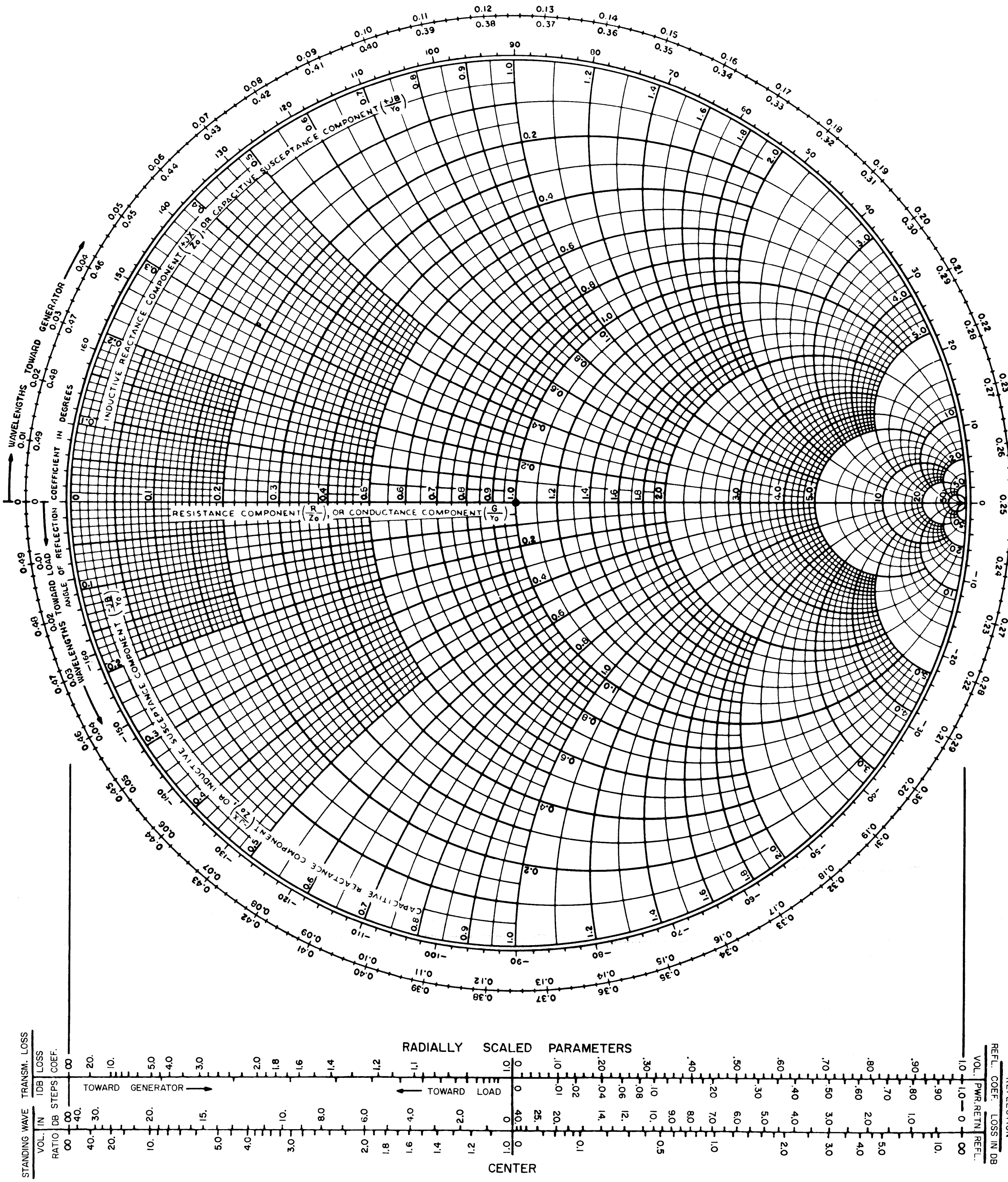
1. ____________cm
2. ____________
   ____
3. ____________

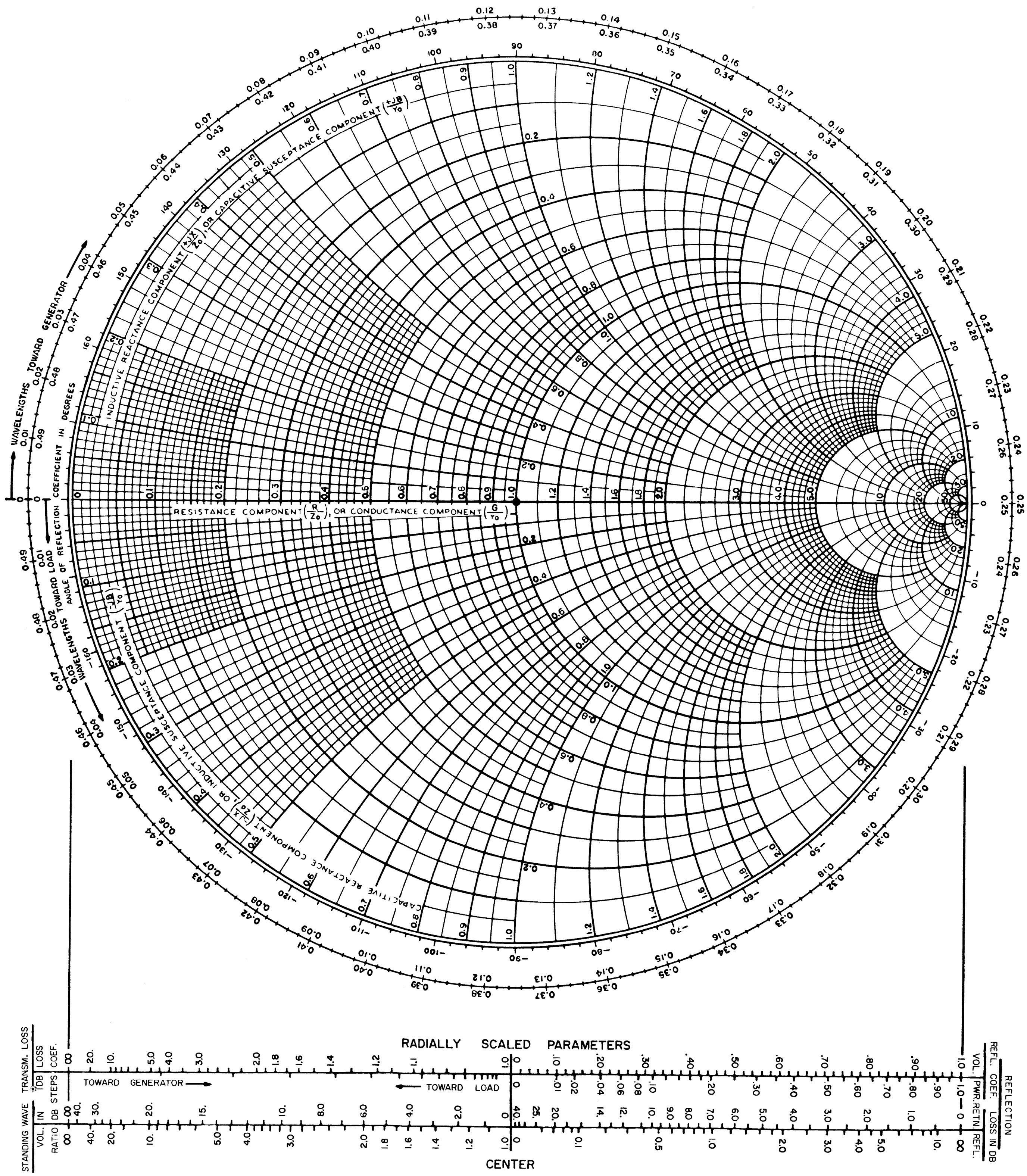
Practice Problem 6

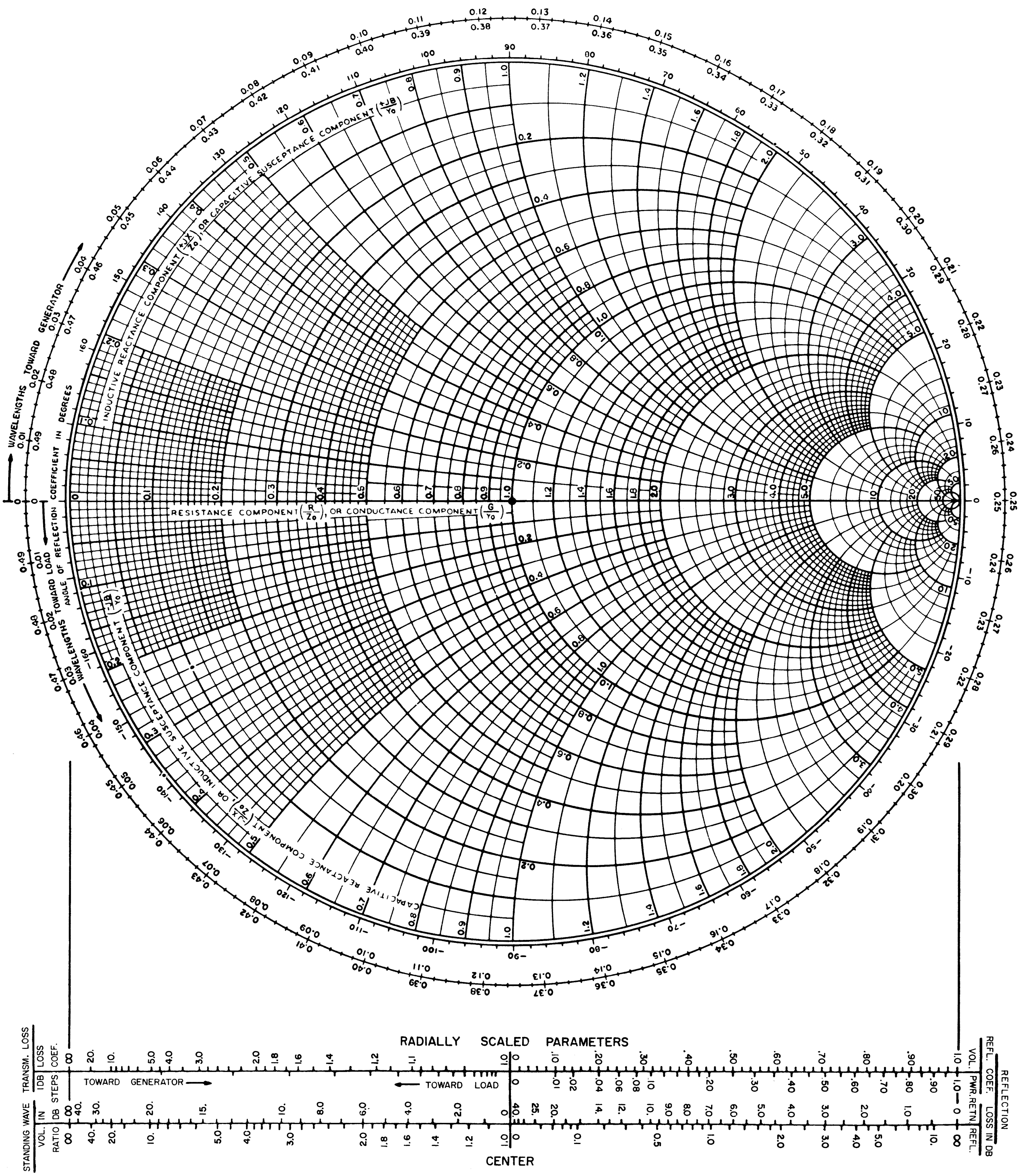
____

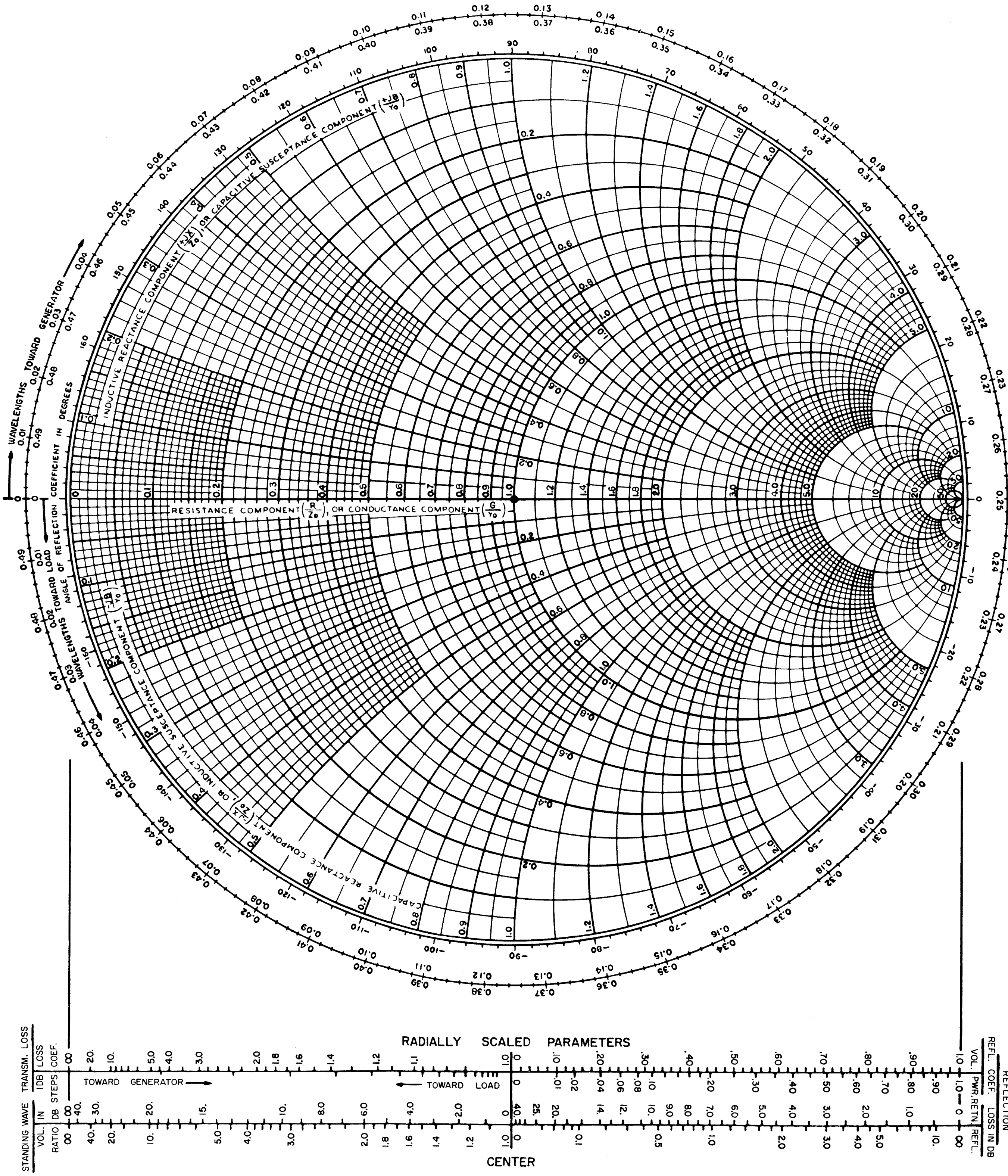
Discussion

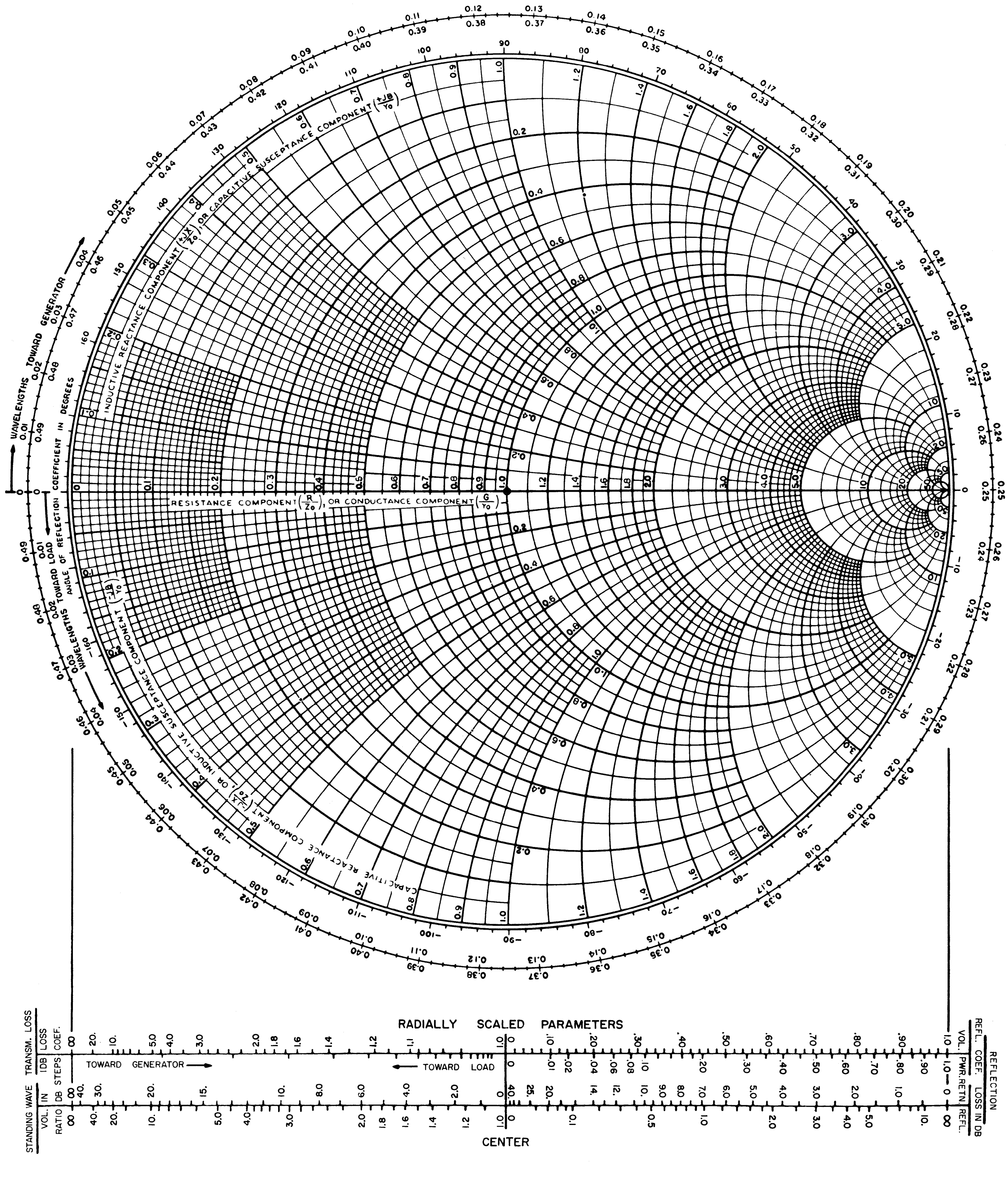

RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE COMPONENT (+jB/Yo)
CAPACITIVE REACTANCE COMPONENT (-jX/Zo), OR INDUCTIVE SUSCEPTANCE COMPONENT (-jB/Yo)
WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
CENTER
STANDING WAVE
TRANSM. LOSS
REFLECTION


WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE COMPONENT (+jB/Yo)
RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
CAPACITIVE REACTANCE COMPONENT (-jX/Zo), OR INDUCTIVE SUSCEPTANCE COMPONENT (-jB/Yo)
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
CENTER
STANDING WAVE
VOL. RATIO
IN DB
TRANSM. LOSS
1DB STEPS
LOSS COEF.
REFLECTION
REFL. COEF. VOL.
PWR.
RETN. LOSS IN DB
REFL. COEF.


WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE COMPONENT (+jB/Yo)
RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
CAPACITIVE REACTANCE COMPONENT (-jX/Zo), OR INDUCTIVE SUSCEPTANCE COMPONENT (-jB/Yo)
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
CENTER
STANDING WAVE VOL. RATIO
IN DB
TRANSM. LOSS 1DB STEPS
LOSS COEF.
REFL. COEF. VOL.
PWR.
RETN. LOSS IN DB
REFL. LOSS IN DB
REFLECTION


WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE COMPONENT (+jB/Yo)
RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
CAPACITIVE REACTANCE COMPONENT (-jX/Zo), OR INDUCTIVE SUSCEPTANCE COMPONENT (-jB/Yo)
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
TRANSM. LOSS
1DB STEPS
LOSS COEF.
STANDING WAVE
IN DB
VOL. RATIO
REFLECTION
REFL. COEF.
VOL.
PWR.
RETN. LOSS IN DB
REFL.
CENTER


WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE COMPONENT (+jB/Yo)
CAPACITIVE REACTANCE COMPONENT (−jX/Zo), OR INDUCTIVE SUSCEPTANCE COMPONENT (−jB/Yo)
RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
TRANSM. LOSS
1DB STEPS
LOSS COEF.
STANDING WAVE
VOL. RATIO
IN DB
REFLECTION
REFL. COEF.
VOL.
PWR.
RETN LOSS IN DB
REFL.
CENTER


WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE COMPONENT (+jB/Yo)
CAPACITIVE REACTANCE COMPONENT (-jX/Zo), OR INDUCTIVE SUSCEPTANCE COMPONENT (-jB/Yo)
RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
STANDING WAVE
VOL. RATIO
IN DB
TRANSM. LOSS
1DB STEPS
LOSS COEF.
REFLECTION
REFL. COEF.
VOL.
PWR.
RETN. LOSS IN DB
REFL. LOSS IN DB
CENTER

EXPERIMENT 7

More Characteristics of the Smith Chart

Object

To study some of the additional characteristics of the Smith Chart, and to work associated practical problems.

Theory

In Experiment 6 it was seen that the slotted line (section) and the Smith Chart are inseparable for impedance measurements. In the development of a microwave device, measurements normally are made to ensure that the impedance of the device is well matched into the transmission line. In the early stages of development, the cut-and-try process is quite often used, so the Smith Chart is invaluable for interpreting the characteristics of a particular design.

Impedance-admittance conversion. One of the useful properties of the Smith Chart is that a normalized impedance value may be graphically converted to admittance simply by plotting the image point $\lambda/4$ around the constant swr circle from the known impedance point (Fig. 1). This characteristic of the chart is extremely useful in design work with parallel impedances, because the admittance values are far more convenient for parallel-circuit considerations.

Example 1. Consider the circuit shown in Fig. 2. The load is an antenna which, because of design considerations, can be matched to only a normalized impedance of $0.4 - j0.4$. It is

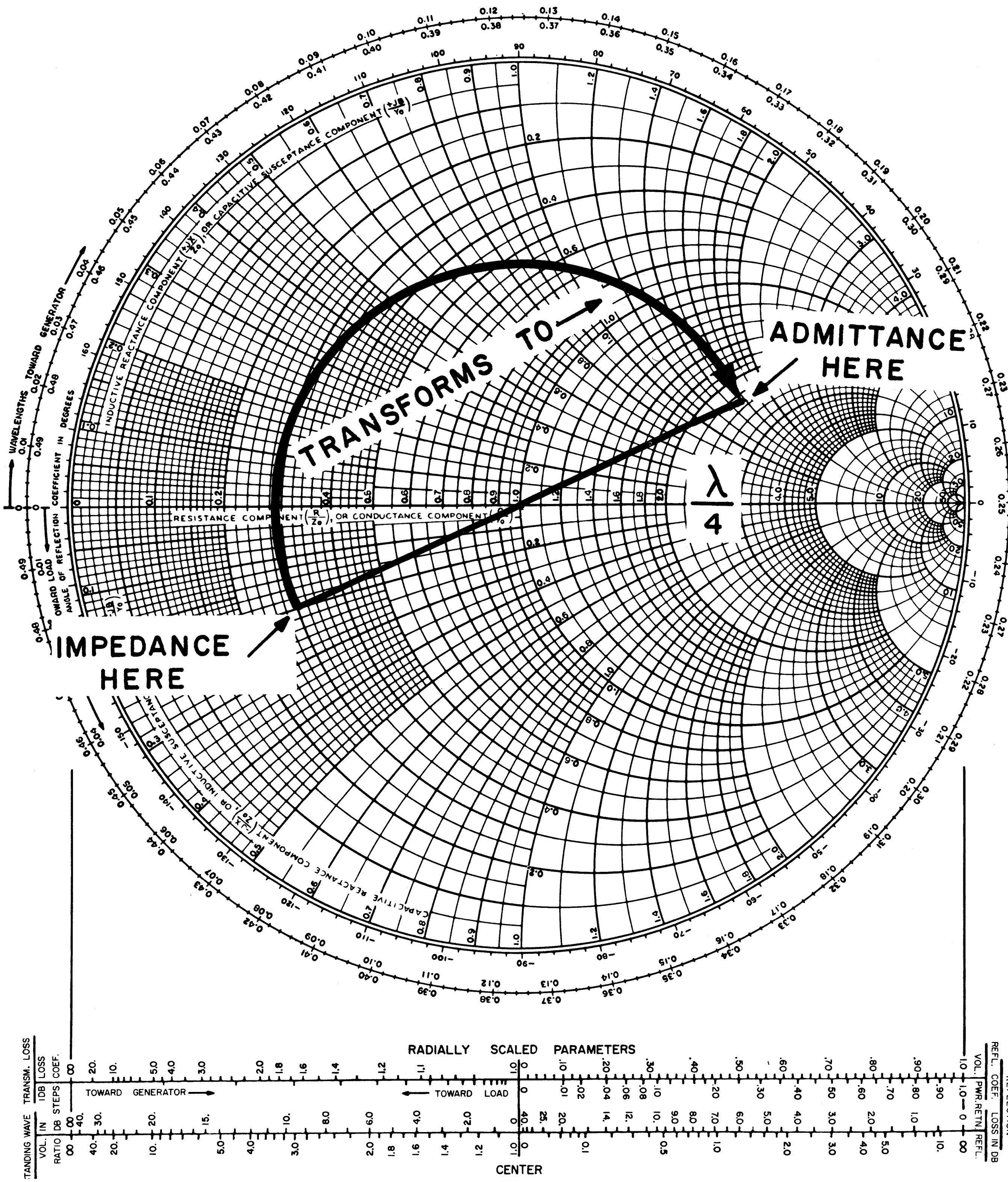

TRANSFORMS TO
ADMITTANCE HERE
IMPEDANCE HERE
λ/4
RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE COMPONENT (+jB/Yo)
CAPACITIVE REACTANCE COMPONENT (-jX/Zo), OR INDUCTIVE SUSCEPTANCE
WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
CENTER
STANDING WAVE VOL. RATIO
IN DB
TRANSM. LOSS 1DB STEPS
LOSS COEF.
REFL. COEF. VOL.
PWR.
RETN. LOSS IN DB
REFL.
REFLECTION

FIGURE 1

desired to place a matching stub on the transmission line at distance Y (physically removed from the antenna) and to use this stub to match or "flatten" the transmission line. The frequency is 3000 mc, and the line is a coaxial air line. The stub of length X is a purely reactive element having an adjustable, sliding short on its end. What is the design distance Y from the antenna to the stub, and what is the length X of the stub for proper matching?

The solution, shown in Fig. 4, is detailed as follows:

(a) Plot the antenna impedance $0.4 - j0.4$ (point A) and the swr circle, and convert the antenna (load) impedance to admittance (point B). This admittance value is $1.23 + j1.23$ mhos.

(b) Move around the swr circle toward the generator to the constant-conductance circle which passes through the center of the chart (point C). At point C the admittance of the line looking back toward the antenna is $1.0 - 1.15$ mhos. If a susceptance of $+ j1.15$ mhos is added in parallel at this point, the combination will equal a resistive component of 1.0 and the line will be perfectly flat from that point toward the generator. This means the input admittance of stub X should look like $+ j1.15$ capacitive susceptance.

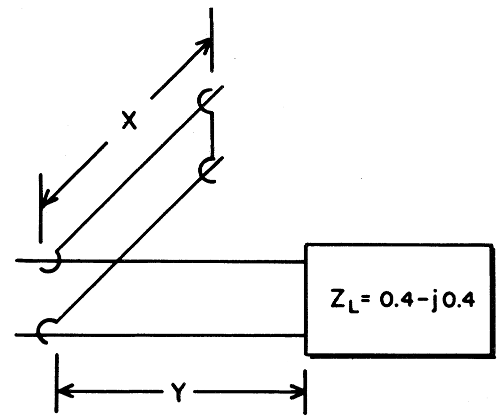


FIGURE 2

(c) The distance Y, which can be easily computed from the phase angle between point B and point C, is found to be 0.153λ, or, in this case, 1.53 cm from the antenna to the matching stub.

(d) To determine the length of stub X, it is necessary only to enter the Smith Chart with the desired capacitive susceptance of $+ j1.15$ (point D). Note that this point is on an infinite swr circle precisely at the circumference of the chart, because the matching stub is completely reactive and has no resistive components in it. From point D, move around the Smith Chart toward the load until a shorted point is obtained at point E. This movement requires 0.386, or, in the case of 3000 mc, 3.86 cm.

By using a shorted matching stub 3.86 cm long placed 1.53 cm from the mismatched antenna, the entire system can be made to appear "flat," minimizing losses in the line caused by high swr.

Double-stub tuners. Because of the relative ease in varying stub length, compared with changing the distance from the stub to the load, tuning is often accomplished by using a double-stub tuner. Operation of coaxial double-stub tuners may be easily explained and visualized by using the Smith Chart.

Consider the following example (Fig. 3) of a double-stub tuner used to match out a mismatched barretter mount. The assembly represents an instrument in which it is desired to match a 200 ohm barretter or instrument fuse into a 50 ohm transmission line for maximum power transfer in a power-measuring circuit. As shown in Fig. 3, stub 1 is located a short fixed distance from the 200 ohm load and has an adjustable shorting element. Stub 2, attached still further down the line, also has an adjustable shorting element. Figure 5 illustrates a typical operation of the double-stub tuner in matching the 200 ohm load to the 50 ohm line at a particular frequency.

The 200 ohm resistive load represents a normalized impedance of 4.0, which is equivalent to a normalized admittance of 0.25 (plotted as point A on the Smith Chart). The admittance is transformed around the constant swr circle from point A

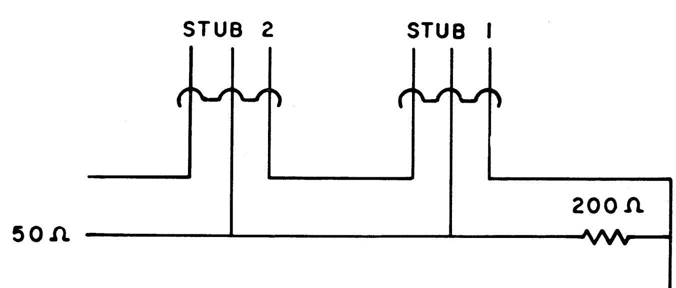


FIGURE 3

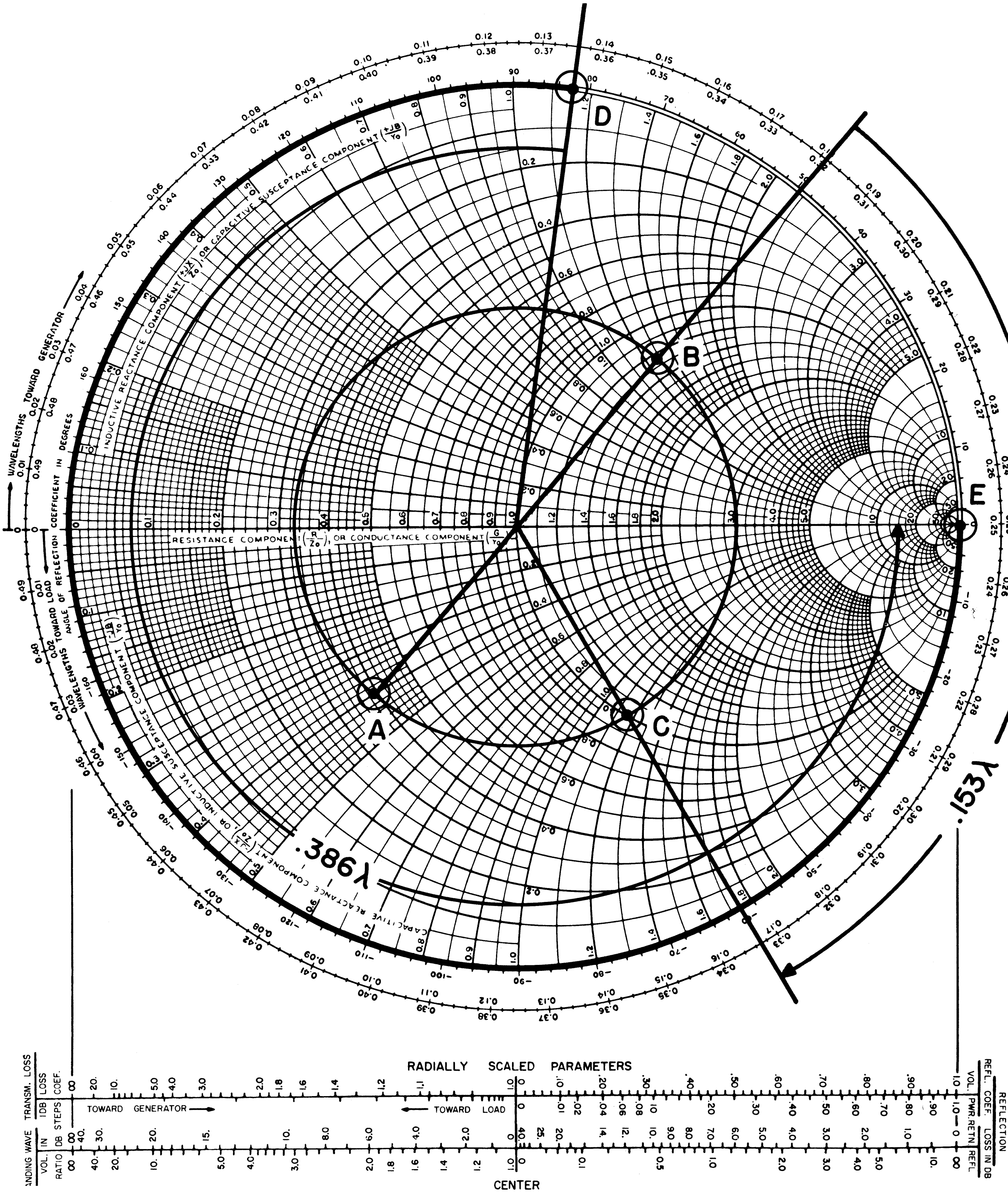

D
B
E
A
C
.386λ
.153λ
RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE COMPONENT (+jB/Yo)
CAPACITIVE REACTANCE COMPONENT (-jX/Zo), OR INDUCTIVE SUSCEPTANCE COMPONENT (-jB/Yo)
WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
CENTER
TRANSM. LOSS
1DB STEPS
LOSS COEF.
STANDING WAVE
VOL. RATIO
IN DB
REFLECTION
REFL. COEF. VOL.
PWR.
RETN. LOSS IN DB
REFL.

FIGURE 4

to point *B*. At point *B*, parallel susceptance is added from the first stub. The susceptance represented is purely reactive, since it is assumed that there are no losses in the stub. This parallel combination results in movement along a constant-conductance circle on the chart to point *C*, since no dissipative element is present in tuning stub 1. From point *C*, the movement is toward the generator on a new constant swr circle (existing in the center portion of the 50 ohm line toward point *D*). At point *D* another parallel susceptance is added by stub 2, which again moves the admittance along a constant-conductance circle into the center of the diagram, or a 1.0 normalized admittance.

The addition of the two susceptances from the double stubs provides a transformation from a badly mismatched line to a line which is matched and limited only by the precision of the tuning procedure. Mismatch loads of any impedance or phase could likewise be tuned out by using a single tuner having two types of adjustment (stub length and position). Another method could use two fixed-length stubs that are shorted on the ends and movable along the line.

NOTE: As the frequency increases and the spacing between the two stubs comes nearer $\lambda/2$, a limitation of tuning range is encountered.

Other Smith Chart scales. A number of radially scaled parameters are provided on Smith Charts on pages 123 and 125. All of these parameters are scaled in such a manner that they may be radially set off from the center point of the Smith Chart (by using a pair of dividers, for instance). Application of the more useful parameters will be briefly discussed.

1. *Standing wave.* The voltage-standing-wave ratio (swr), shown as the lowest left-hand scale, is a translation of the scale on the horizontal center line of the Smith Chart [Fig. 6(A)]. Note that the lower scale progresses from a reading of 1 at the center to infinity at the left-hand margin, and agrees with all the axis crossings on the horizontal center line. Voltage-standing-wave ratio is commonly expressed in db; conversion to db can be conveniently made on the adjacent "in db" scale. The relation can be expressed as

$$\text{swr}_{\text{db}} = 20 \log_{10} \text{swr}_{\text{numerical}}$$

For example, a swr of 2.0 occurs opposite the 6.0 db point on the adjacent scales. Consequently, 6 db equals 20 $\log_{10}$ (2.0), where 2.0 is the ratio of voltages.

2. *Reflection coefficient.* The voltage reflection coefficient (ρ), shown as the upper right-hand scale [in Fig. 6(B)], is related to the swr by the simple equation:

$$\text{swr} = \frac{1+\rho}{1-\rho}$$

Note that a reflection coefficient of 0.333 is very nearly equal to a swr of 2.0.

The adjacent power-reflection-coefficient scale is little used; mathematically, it is simply a squared function of adjacent voltage reflection coefficient.

3. *Loss in db (return).* This very important scale is located near the lower right-hand corner. Return loss is the relation between the power returning down the line from a mismatched load and the power incident to that load. Thus, a return loss of 10 db means that the reverse power traveling in a line from a mismatched load is 10 db below the reference power incident on that mismatched load. For example, the swr with a certain mismatched load is 2.0, which is equal to a return loss of about 9.5 db (i.e., the reflected power is 9.5 db down from the incident power). To confirm that quantity for the same swr, we reason as follows: the voltage vector of the backward-flowing power is 0.333 (if incident voltage was 1.0). Power returning down the line is then $(0.333)^2$, or 0.111. Note that 0.111 is approximately $-$ 9.5 db. Mathematically, return loss is equal to $-$ 20 $\log_{10}\rho$ (where ρ is the voltage reflection coefficient).

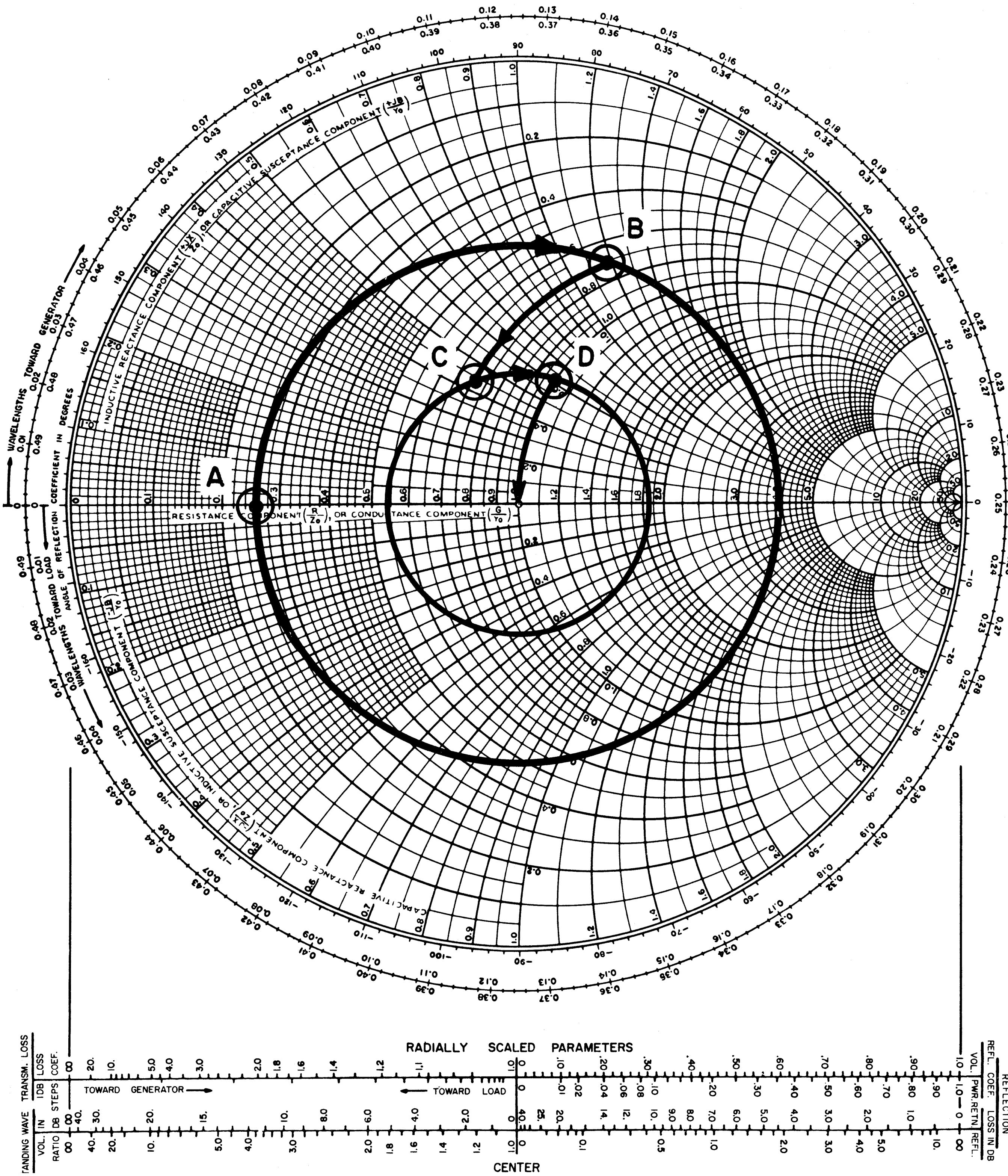

A
B
C
D
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
CENTER

FIGURE 5

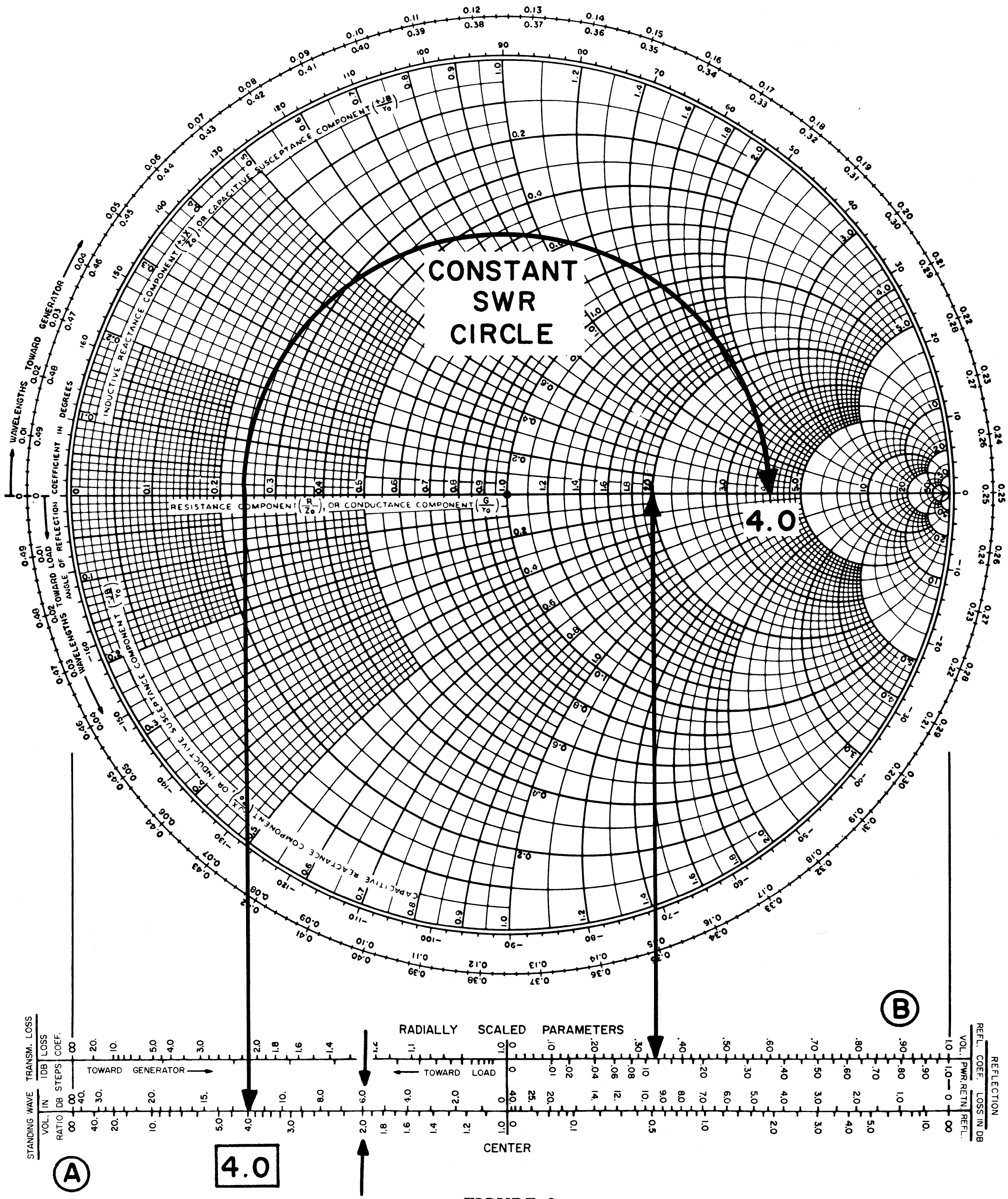

CONSTANT SWR CIRCLE
4.0
RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE COMPONENT (+jB/Yo)
CAPACITIVE REACTANCE COMPONENT (-jX/Zo), OR INDUCTIVE SUSCEPTANCE COMPONENT (-jB/Yo)
WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
CENTER
A
B
4.0

FIGURE 6

4. *Loss in db (reflected).* The reflection loss in db, also commonly known as *mismatch* loss, is the relation of the power being transferred in the forward direction to the incident power on a mismatched plane. Consequently, mismatch loss is the power which actually arrives at the load and is dissipated, or is the power which is transmitted in an antenna, etc.

When the swr is 2.0, for example, it is seen that the reflected loss (scale adjacent to return-loss scale) is 0.5 db. In other words, the power actually being dissipated in a 2-to-1 normalized load is 0.5 db lower than the incident reference power. Mathematically, reflection loss is equal to

$$-10 \log_{10} (1 - \rho^2)$$

Both the return-loss and the reflected-loss scales are extremely convenient, because they describe the amount of power which is being transferred forward and that which is being reflected from the given mismatched load. Surprisingly, even large mismatches, such as 3 to 1 (which cause quite large standing waves on a transmission line), result in only slightly more than 1 db of power loss in the transmitted forward-direction power (as read on the reflected-loss scale).

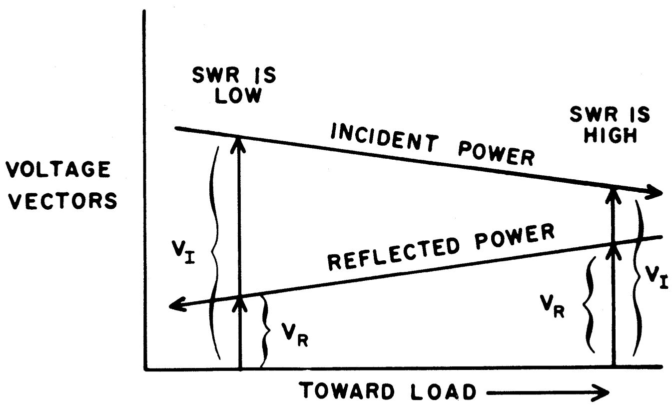


FIGURE 7

5. *Transmission loss (1 db steps).* So far, all of our analysis has considered only lossless lines, where a given line is described by a constant-radius swr circle. In the case of a line with finite attenuation, however, a correction must be made, because the power (and hence the voltage) does not remain constant in amplitude as it travels along a transmission line; the forward and reverse interaction results in swr no longer constant with position along the line (Fig. 7).

Graphically, a lossy line is represented by a spiral on the Smith Chart, because the power being transferred toward the load in a lossy system and thereupon being reflected back toward the generator experiences an attenuation in both directions. Therefore, the nearer the observer moves to the generator, the smaller is the ratio between the reflected voltage and the incident voltage. The 1 db transmission-loss scale is used for swr measurement as shown in the following example.

Example 2. A slotted line is used to measure the swr in a system which consists of an unknown load at the end of a remote cable having 2 db of insertion loss (Fig. 8). The swr obtained is 2.0, and the null shift caused by going from the unknown load to a short at the remote reference plane is 0.2λ toward the load. Calculate the true impedance of the unknown load at the remote reference plane.

The solution is shown on the Smith Chart in Fig. 9. The procedure is to plot the measured swr and to determine the phase angle from the null shift. However, the intersection of the radial line and the swr circle does not yet result in the remote impedance. To obtain the remote impedance, it is necessary to apply a 2 db correction to compensate for the cable loss. The correction is made by marking point *A* on the 1 db transmission-loss scale to correspond with the 2.0 swr circle. A 2 db correction is made toward the load to point *B*. The radius related to point *B* is that of the swr circle at the remote reference point. By swinging this radius around, it is seen that the new swr circle has a radius equiva-

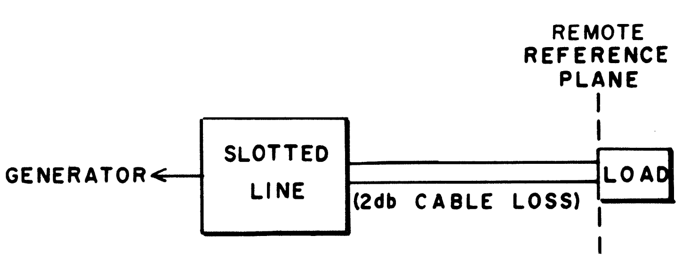


FIGURE 8

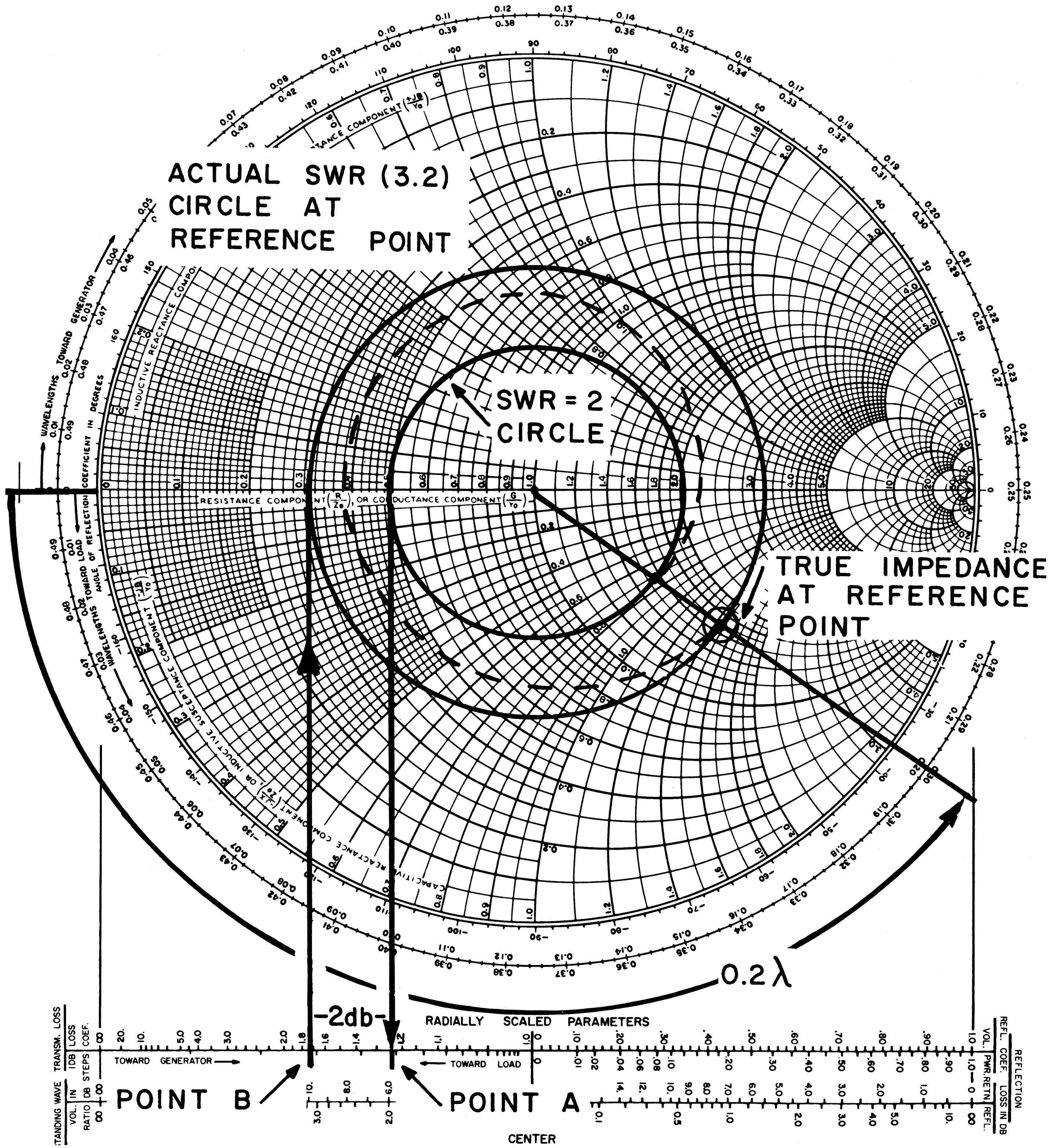

ACTUAL SWR (3.2)
CIRCLE AT
REFERENCE POINT
SWR = 2
CIRCLE
TRUE IMPEDANCE
AT REFERENCE
POINT
0.2 λ
-2db-
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
POINT B
POINT A
CENTER

FIGURE 9

lent to a swr of 3.2. Note that the actual impedance relationships occurring along the line are now described by the approximate spiral shown.

Incidentally, it can be noted that the "transmission loss in 1 db steps" scale is related to the "return loss in db" scale by a factor of 2. This relationship can be justified by considering that the return loss includes the two-way attenuation through a given piece of cable, whereas the transmission loss is defined as merely a one-way attenuation loss.

6. *Transmission loss* (*coefficient*). This scale is a correction factor relating to the additional line losses in a transmission line in which the swr is greater than unity. The simplified reason for such a situation is that standing waves cause increases of current at certain points on the line and increases of voltage at other points. Since the resistive losses are related to the current squared, and the dielectric losses are related to the voltage squared, the average line losses of a line with a large swr are significantly larger at large swr. Therefore, the line losses should be used to correct calculated attenuation factor when a very large swr is encountered.

Practice Problem 1. An equipment arrangement similar to that in Fig. 8 is used to measure the characteristics of a remote antenna system. The antenna causes the swr to be 3.6, as measured on the slotted line, and the null shifts 0.2λ toward the load when the load (antenna) is replaced with a short at the remote reference plane. With the remote reference plane shorted, a swr of 8.5 is obtained with the slotted line.

1. What is the loss on the cable? *Hint:* Calculate this loss by considering that the short is transformed through the lossy cable to the "shorted" swr reading given above.
2. What is the load (antenna) swr at the reference plane?
3. What is the antenna impedance at the reference plane?
4. What is the voltage-reflection coefficient at the reference plane?
5. What is the power-reflection coefficient at the reference plane?
6. If there is a 1 mw forward signal at the slotted line (0 dbm), what is the forward power level at the reference plane with the antenna in place?
7. From your knowledge of the swr at the reference plane, what is the amount of power entering the antenna to be transmitted?
8. What is the power returning from the reference plane?
9. What is the return power at the slotted line with the antenna in place?
10. As a check on your power transmission and reflection calculations, consider the ratio between the forward power in the slotted line and the return power. What is this ratio? Use this ratio in db and enter the "return loss" radial scale. This ratio in db should, of course, result in a radius equal to the swr measured originally with the slotted line. If your calculations do not check, go back through questions 6–9.

Practice Problem 2. With an equipment arrangement similar to that shown in Fig. 8, the "unshorted" swr is 2.5. Shorting the system at the reference plane causes the null to shift 0.2

toward the generator, and the resulting swr is 7.0. For this system, find answers to the questions asked in Practice Problem 1.

Practice Problem 3. If you were using a Model 420A crystal detector (swr of 3) and you wanted to use an attenuator to give match with a swr of 1.2, how much attenuation (one-way cable loss) would be required to reduce the apparent swr of this crystal?

Name ______________________

Course ______________________

Date performed ______________________

Date turned in ______________________

Results (Experiment 7)

Practice Problem 1

1. ______________ db
2. ______________
3. ______________
4. ______________
5. ______________
6. ______________ dbm
7. ______________ dbm
8. ______________ dbm
9. ______________ dbm
10. ______________

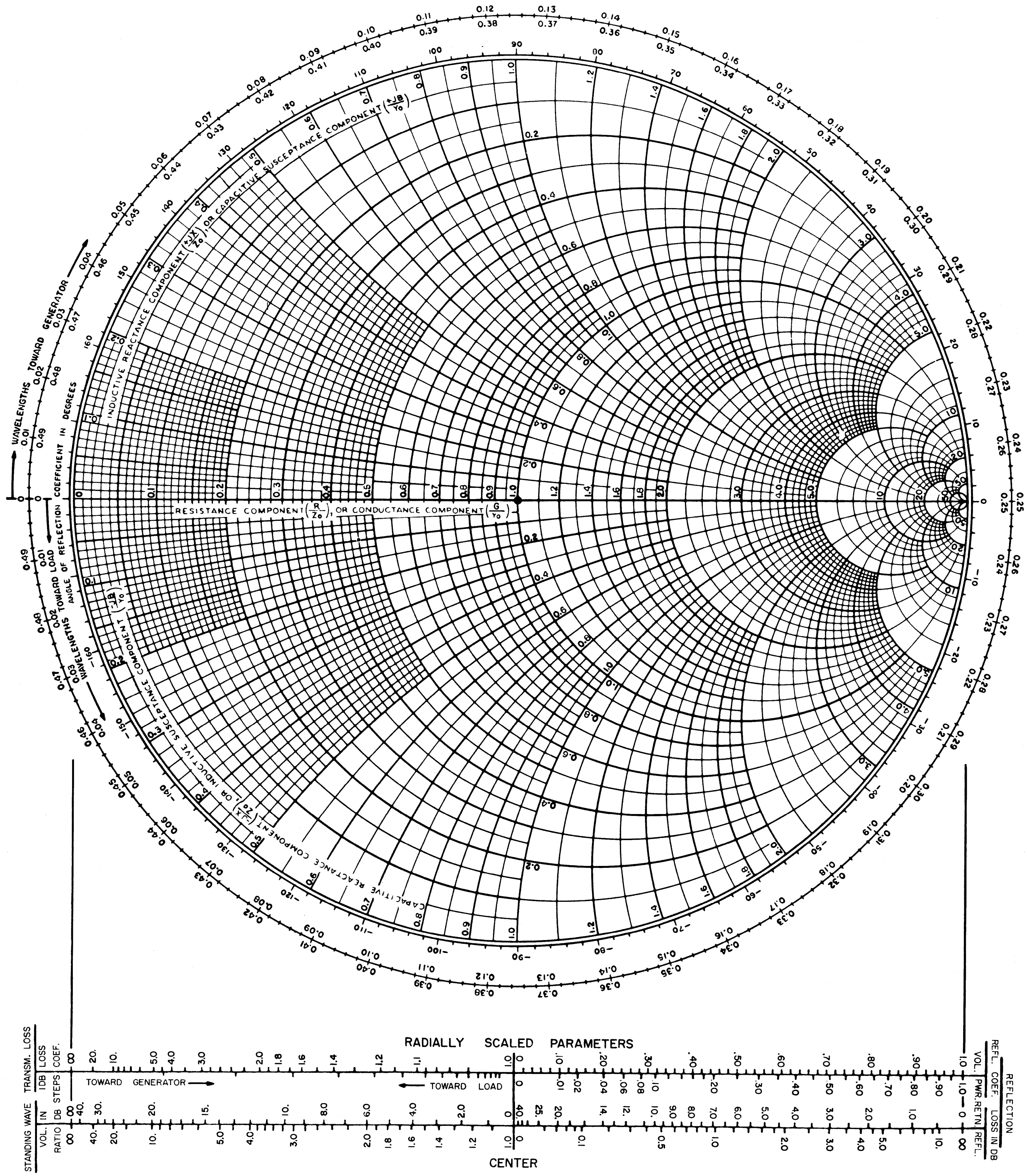
Practice Problem 2

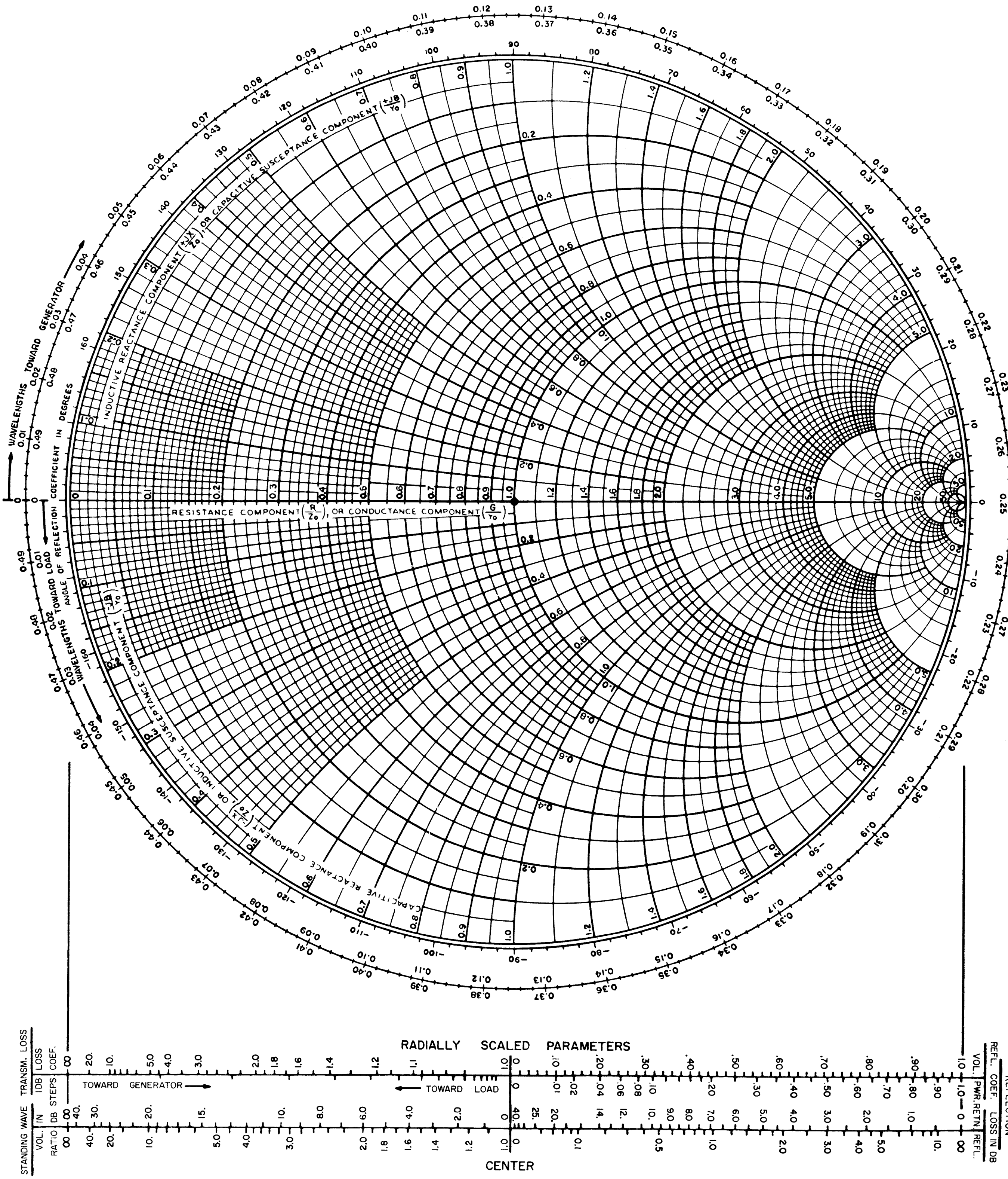
1. ______________ db
2. ______________
3. ______________
4. ______________
5. ______________
6. ______________ dbm
7. ______________ dbm
8. ______________ dbm
9. ______________ dbm
10. ______________

Practice Problem 3

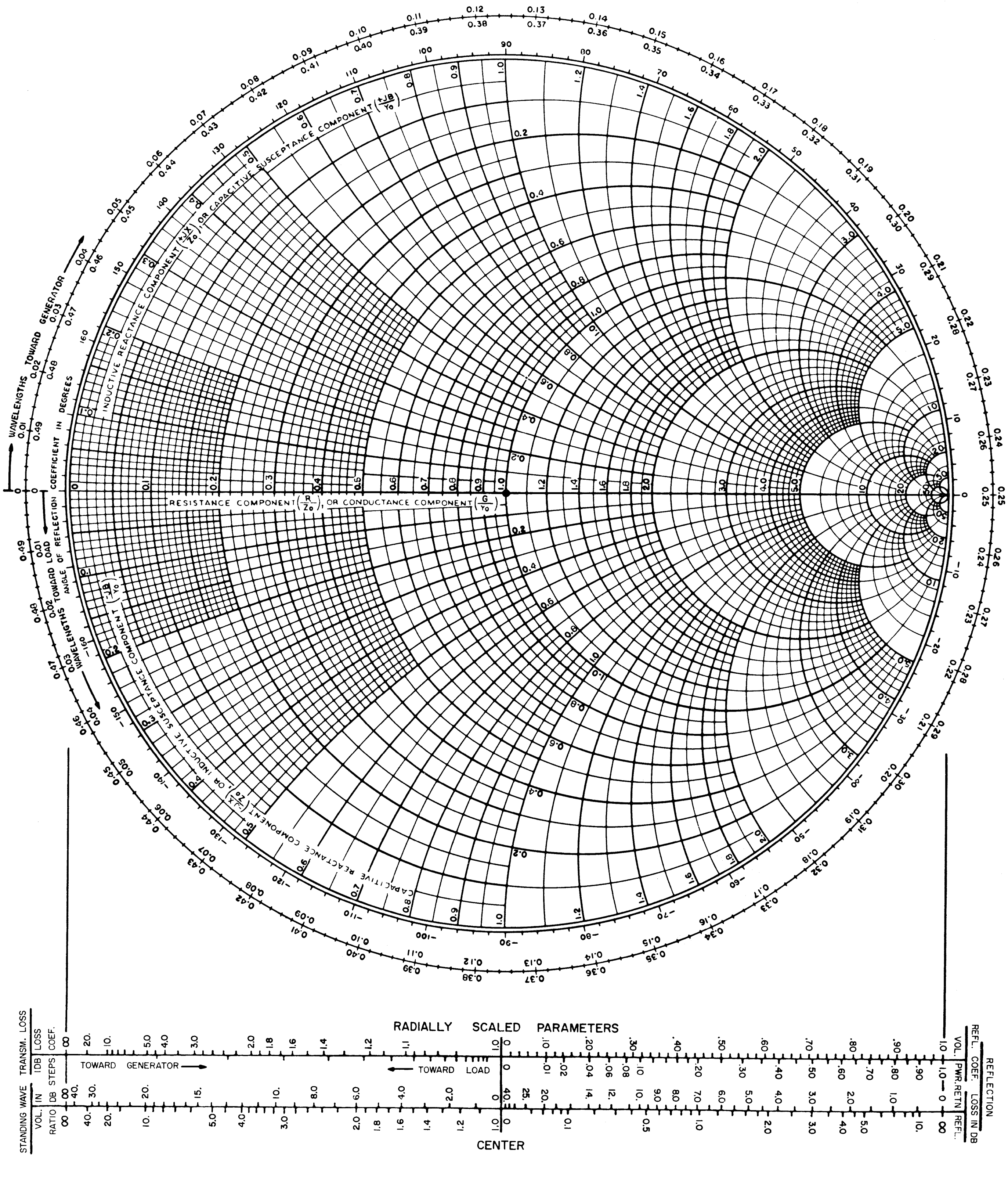
______________ db

Discussion


WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE COMPONENT (+jB/Yo)
CAPACITIVE REACTANCE COMPONENT (-jX/Zo), OR INDUCTIVE SUSCEPTANCE COMPONENT (-jB/Yo)
RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
CENTER
STANDING WAVE VOL. RATIO
IN DB
TRANSM. LOSS 1DB STEPS
LOSS COEF.
REFLECTION COEF. VOL.
PWR.
RETN. LOSS IN DB
REFL. LOSS IN DB



RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE COMPONENT (+jB/Yo)
CAPACITIVE REACTANCE COMPONENT (-jX/Zo), OR INDUCTIVE SUSCEPTANCE COMPONENT (-jB/Yo)
WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
STANDING WAVE VOL. RATIO
IN DB
TRANSM. LOSS 1DB STEPS
LOSS COEF.
REFLECTION COEF. VOL. PWR.
RETN LOSS IN DB
REFL.
CENTER



RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE COMPONENT (+jB/Yo)
CAPACITIVE REACTANCE COMPONENT (-jX/Zo), OR INDUCTIVE SUSCEPTANCE COMPONENT (-jB/Yo)
WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
STANDING WAVE VOL. RATIO
IN DB
TRANSM. LOSS 1DB STEPS
LOSS COEF.
REFLECTION COEF. VOL.
PWR.
REFLECTION LOSS IN DB
RETN
REFL.
CENTER

EXPERIMENT 8

Impedance Measurement Using the Smith Chart

Object

To gain additional experience in using the Smith Chart and in determining impedance through measurements of swr.

Theory

An important characteristic of any microwave apparatus is the manner in which the impedance varies with frequency. Since many transmission systems are presently designed to carry a broad band of frequencies, variations in the component impedances due to frequency could cause undesirable reflections at several points in the band. Typical of these broadband systems are radar sets designed to operate over bandwidths of more than 500 mc, with random programming of frequency and pulse width (to counteract jamming).

In this experiment, the techniques of frequency measurement introduced in Experiment 2, of swr measurement covered in Experiment 5, and of Smith Chart plotting discussed in Experiments 6 and 7 will be used. The relevant experiments may be reviewed if necessary.

Equipment

QUANTITY	TYPE
1	hp 715A klystron power supply (with cable)
1	Varian X-13 reflex klystron
1	Cooling fan or blower
1	hp X375A variable flap attenuator
1	hp X532B frequency meter
1	hp 415B standing-wave indicator
1	hp 830C microwave power meter
1	hp X487B thermistor mount
1	hp 444A broadband probe
1	hp 809B probe carriage
1	hp X810B slotted section
1	hp 120B oscilloscope
1	hp AC-16A cable (dual banana to dual banana)
1	hp AC-16B cable (dual banana to BNC)
2	hp AC-16K cable (BNC to BNC)
1	hp Model 24 waveguide stand with Model X25 waveguide clamp

Procedure

Section 1—General

1-1 Review the safety precautions on page 42, and the operating instructions for each equipment item.

1-2 Set up the equipment as shown in Fig. 1. The oscilloscope horizontal input should be *a-c coupled* and the vertical input should be *d-c coupled*. The variable flap attenuator should be set to approximately 10 db.

1-3 Connect the thermistor mount (100 ohms resistance, negative temperature coefficient) to the microwave power meter, and apply only enough bias current to permit an indication on the power meter, controllable by the ZERO SET. Attach the thermistor mount to the slotted section.

1-4 Energize the klystron at 8.5 gc and modulate it with a 1000 cps square wave. Use the oscilloscope technique of Experiment 2 to set the reflector and modulating voltages properly.

Section 2—Thermistor Mount Matched to System

2-1 Connect the standing-wave indicator to the broadband probe in place of the oscilloscope connection. Adjust the 1000 cps modulation frequency to get a peak indication on the standing-wave indicator. For the various frequencies shown in Table I, obtain and record the data when: (1) the thermistor mount is attached to the system, and (2) the thermistor mount is removed and a short is attached instead. Use the swr measuring techniques covered in Experiment 5 and the slotted-line-minimum (null) technique discussed in Experiment 6. Short circuits may be obtained by using a brass plate to short the end of the waveguide.

2-2 Calculate the null-shift information from the data collected, and locate and plot the complex-impedance point on the Smith Chart for each frequency. Use the plotting rules from Experiment 6, and identify each point by frequency as it is plotted.

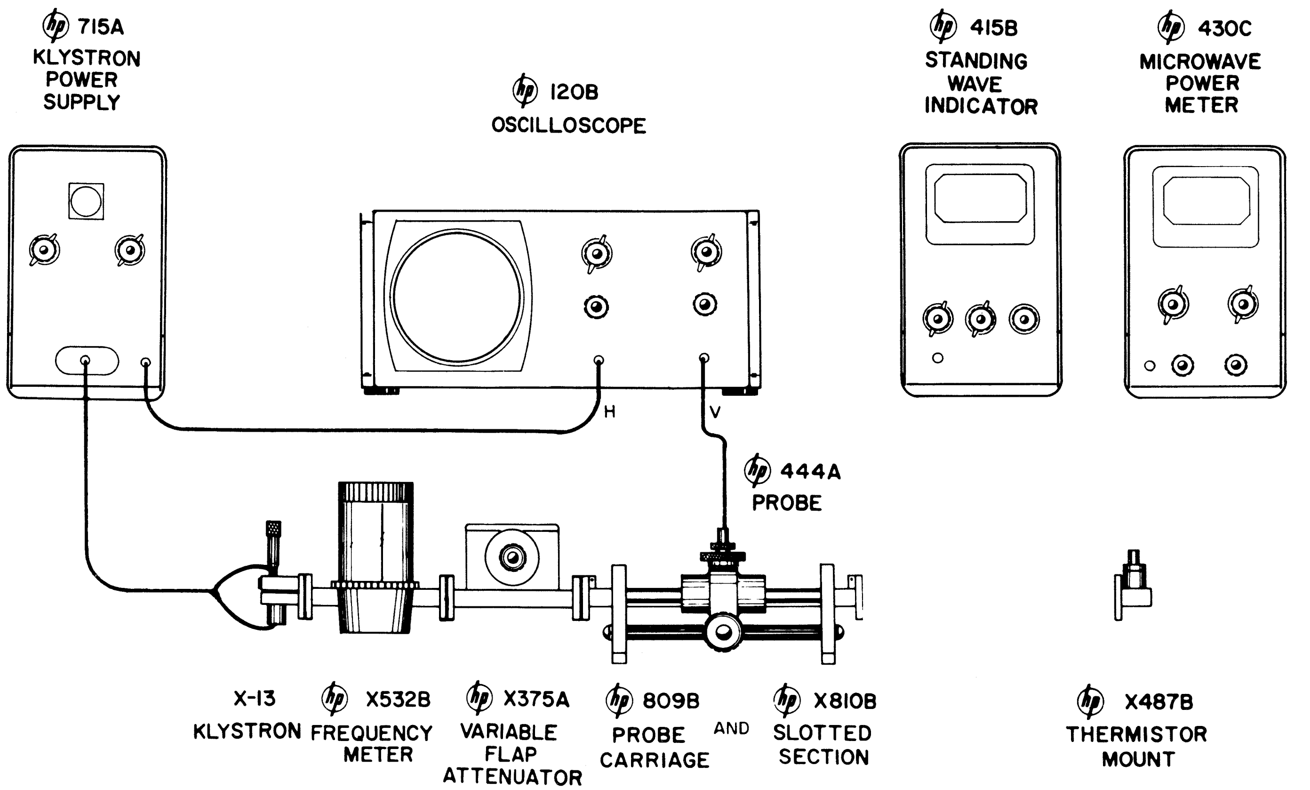


FIGURE 1

2-3 Does the thermistor mount meet its swr specification? (For the hp Model X487B the specification is 1.5:1 from 8.2 to 12.4 gc.) *Hint:* Draw a specification swr circle on the Smith Chart.

Section 3—Thermistor Mount Not Matched to System

3-1 Remove all bias current from the thermistor mount (as in Experiment 5).

3-2 Using the procedure of Steps 2-1 and 2-2, complete Table II and plot the points on the Smith Chart.

Name ____________________
Course ____________________
Date performed ________________
Date turned in ________________

Results (Experiment 8)

OBSERVED	CALCULATED
Step	*Step*
2-1 Table I	2-2 Table I 2-3 __________
3-2 Table II	3-2 Table II

Questions

1. The hp Model X487B thermistor mount has the thermistor element mounted 1.4 cm from the reference input flange. How would you transfer the impedances of Steps 2-2 and 3-2 down the guide to the mounting plane of the thermistor element? Do so for several plotted points.

2. If the slotted section used has a residual swr, each plotted point will have a circle of ambiguity around it caused by a random phase of the residual reflection compared to the load reflection. For a residual swr of 1.01:1, how large is the circle of ambiguity?

3. When would you express swr in units of db?

Discussion

TABLE I

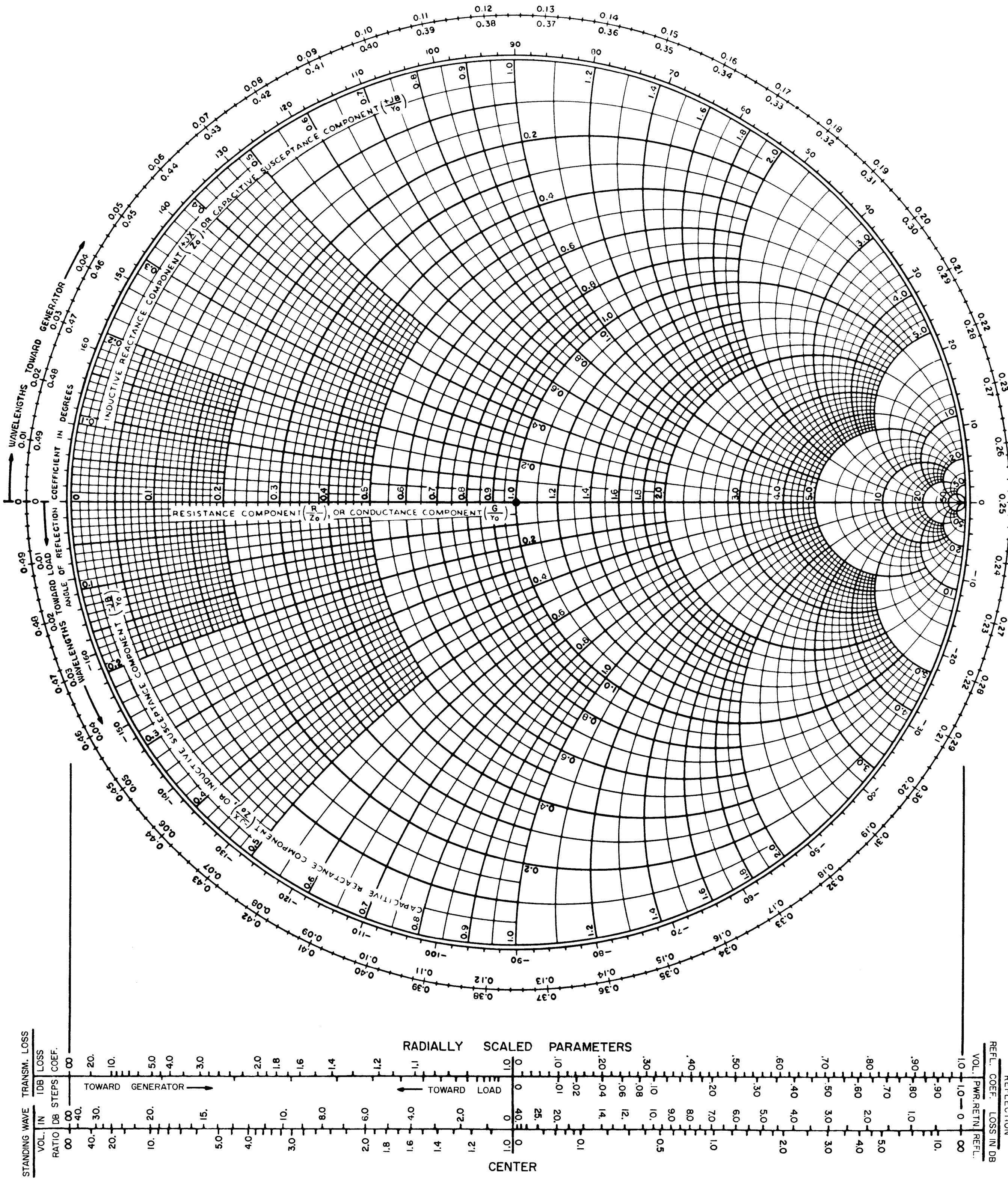
Frequency (gc)	WITH LOAD		WITH SHORT			NULL SHIFT		
	SWR	Null position N_L	Null position N_{S_1}	Null position N_{S_2}	Guide wave-length (λ_g) . . . $2(N_{S_1} - N_{S_2})$	ΔN (cm) . . . Load to short	Direction . . . Toward load or generator?	$\frac{\Delta N}{\lambda_g}$ . . . (Must be less than $0.25\lambda_g$)
8.5								
9.5								
10.5								
11.5								

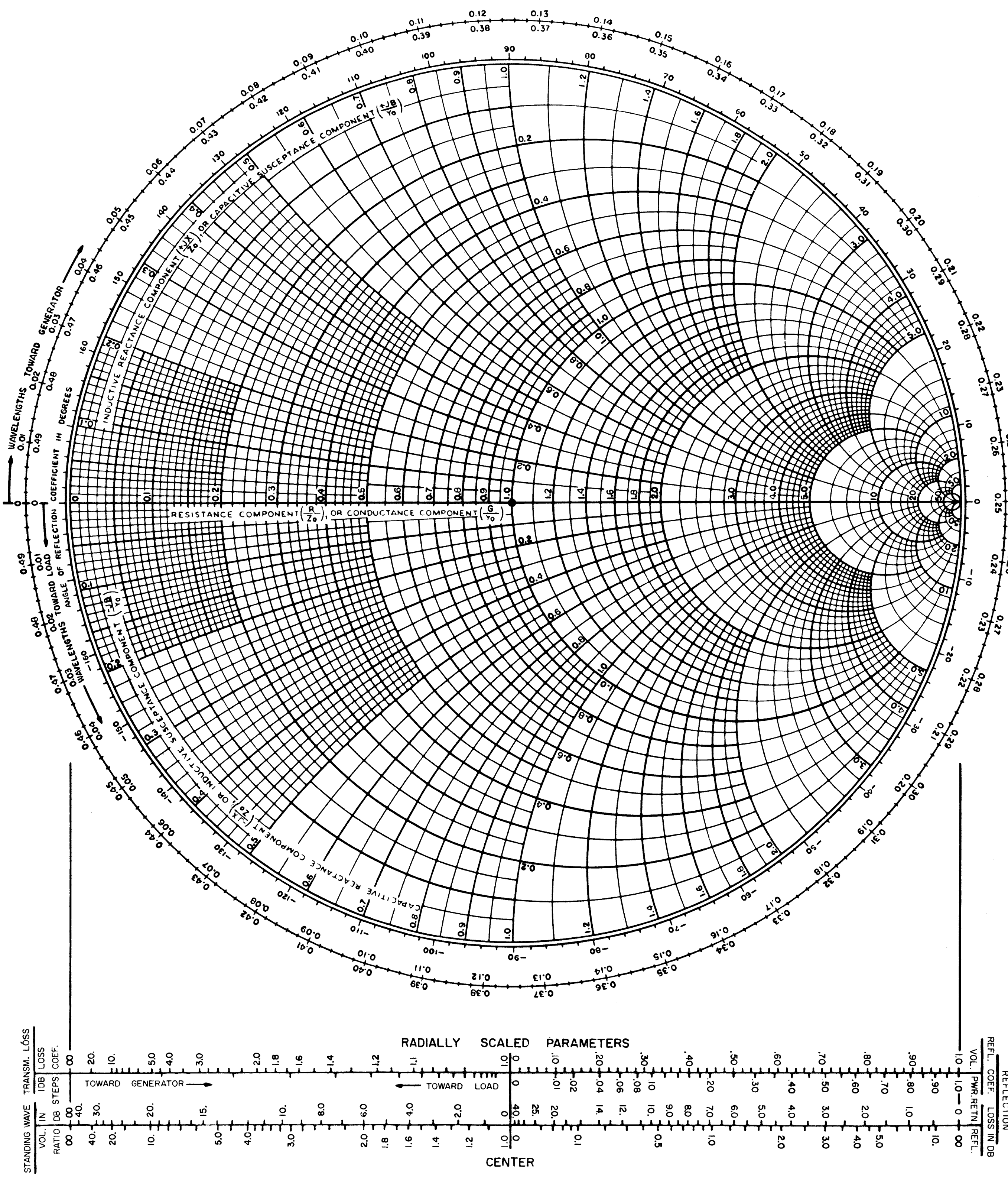
NOTE: The cm scale readings on the hp Model 809B probe carriage increase toward the left (or generator) end.

TABLE II

Frequency (gc)	WITH LOAD		WITH SHORT			NULL SHIFT		
	SWR	Null position N_L	Null position N_{S_1}	Null position N_{S_2}	Guide wave-length (λ_g) . . . $2(N_{S_1} - N_{S_2})$	ΔN (cm) . . . Load to short	Direction . . . Toward load or generator?	$\frac{\Delta N}{\lambda_g}$ . . . (Must be less than $0.25\lambda_g$)
8.5								
9.5								
10.5								
11.5								

NOTE: The cm scale readings on the hp Model 809B probe carriage increase toward the left (or generator) end.


WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE COMPONENT (+jB/Yo)
CAPACITIVE REACTANCE COMPONENT (-jX/Zo), OR INDUCTIVE SUSCEPTANCE COMPONENT (-jB/Yo)
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
CENTER
TRANSM. LOSS
1DB STEPS
LOSS COEF.
STANDING WAVE
VOL. RATIO
IN DB
REFLECTION
REFL. COEF. VOL.
PWR.
RETN. LOSS IN DB
REFL.


RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE COMPONENT (+jB/Yo)
CAPACITIVE REACTANCE COMPONENT (-jX/Zo), OR INDUCTIVE SUSCEPTANCE COMPONENT (-jB/Yo)
WAVELENGTHS TOWARD GENERATOR
WAVELENGTHS TOWARD LOAD
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
RADIALLY SCALED PARAMETERS
TOWARD GENERATOR
TOWARD LOAD
STANDING WAVE
VOL. RATIO
IN DB
TRANSM. LOSS
1DB STEPS
LOSS COEF.
REFLECTION
REFL. COEF. VOL.
PWR.
RETN. LOSS IN DB
REFL. LOSS IN DB
CENTER

EXPERIMENT 9

Bolometer Mounts for Microwave Measurements

Object

To compare barretters and thermistors and their applications in the measurement of microwave power. To examine the use of the barretter as a 1000 cps detector.

Theory

Power measurements are usually considered more basic than voltage or current measurements at microwave frequencies, because power does not vary with the position of measurement along a distributed-type transmission line as voltage and current do. By "power measurement" we mean the determination of the *absolute* power in the system. This type of measurement should be differentiated from the *detection* of microwave power, in which only the relative power is of interest or the characteristics of the modulation envelope are desired.

The measurement of rf power at high frequencies is accomplished by means of a device called a bolometer. "Bolometer" is a very general term that is applied to several different devices whose resistances change with application of power (converted to heat). For rf measurements, the bolometer is mounted in a suitable holder so that the bolometer absorbs all available rf power. The rf energy that is applied to the bolometer causes it to heat, and, in turn, the heat causes the d-c resistance of the bolometer to change.

Bolometer types. The two general classifications of bolometers are

1. *Conductors* (positive temperature coefficient). Conductors are exemplified by the barretter, which consists of a very fine platinum wire mounted in a holder to permit easy measurement of resistance changes. The most common type of barretter is constructed of Wollaston wire, which consists of a platinum core plated with silver, drawn to an extremely small diameter, and soldered in place. The silver is then etched away, leaving the platinum core of about 50×10^{-6} in. in diameter.

Fuses, of the type used to protect sensitive meters, are also useful as positive-temperature-coefficient bolometers; they feature lower replacement cost.

2. *Semiconductors* (negative temperature coefficient). The most popular semiconductor type of bolometer is the thermistor. The thermistor is constructed in the form of a small bead of semiconducting material suspended between two fine lead-in wires. The tiny bead, about 0.04 cm in diameter, is composed of a mixture of manganese, cobalt, nickel, and copper oxides. The bead is coated with a film of glass to prevent oxidation and to improve stability. It may be mounted directly in waveguide, or it may be encapsulated to provide physical protection.

Comparison of bolometers. 1. CW *power measurement.* Although both the thermistor and the barretter can be made to operate over the same range of power, each has certain specific attributes which make it more useful in some applications. Both can be made in very small sizes and can be mounted easily in waveguide transmission systems. The thermistor is basically more sensitive than the barretter, so it is more desirable in some low-power applications. It is, however, much more sensitive to changes in ambient temperature than the barretter, and it requires very special mount design in order to achieve its maximum sensitivity.

Thermistors are less delicate than barretters and may be mounted directly in waveguide. They also have the protective property of decreasing their resistance with increasing power. Consequently, high rf powers drastically lower the resistance and reflect much of the power back down the line. Barretters, on the other hand, are operated near their maximum-dissipation point and are very susceptible to burnout. Barretters can be made with remarkable uniformity with regard to their rf and d-c characteristics. A defective unit can be replaced by a new one without readjusting any of the rf matching sections.

2. *Modulated power measurements.* Pulsed power, as well as cw power, can be measured with bolometers. Since bolometers have time constants in the order of 100 microseconds or longer, the measurement of the usual pulse train provides an automatic averaging process. However, the measurement of certain pulsed power by means of bolometers can introduce some special factors which can lead to considerable error. Errors can arise because the bolometer resistance during the measurement is not constant; evaluation of them can become quite complicated. These errors are most serious in barretters, which have a relatively short time constant. Since thermistors have longer time constants (on the order of 1 sec) than barretters, they are more suitable for application in pulse measurements.

Detection with a barretter. Because a barretter has a time constant on the order of 100 μsec, its resistance can change rapidly enough to follow the rf power change of a 1000 cps amplitude-modulated wave. Barretters are better suited to wide dynamic range attenuation measurements than are crystals, since their resistance change depends upon rf heat dissipated (and thus depends upon rf power). Hence, barretters are better square-law detection elements.

In 1000 cps detection, a constant-bias current is applied to the barretter element. As rf power is applied, the element resistance changes slightly; the constant current thus provides a slight voltage change across the element. This change is usually transformer-coupled to the input of 1000 cps amplifiers. The sensitivity of a typical Sperry Type 821 barretter element is 4.5 ohms per mw. If we assume a bias current of 8.75 ma, this sensitivity is approximately 40 mv per mw of rf.

In this experiment, power will be measured with two types of waveguide bolometer mounts. The ⓗⓟ Model X485B detector mount is tunable and has a maximum swr of 1.25. The usual barretter element is a Sperry Type 821. The ⓗⓟ Model X487B thermistor mount is not tunable and has a maximum swr of 1.5.

References. Additional information to supplement this experiment may be found in the following texts (listed in Appendix C).

Ginzton, Chapter 3: 3-3–3-6.
King, Chapter 3: 3.2.
Reich, Chapter 8: 8-3–8-4.
Wind, Section 4: 4.02.

Equipment

QUANTITY	TYPE
1	ⓗⓟ 715A klystron power supply (with cable)
1	Varian X-13 reflex klystron
1	Cooling fan or blower
1	ⓗⓟ X382A precision variable attenuator
1	ⓗⓟ 415B standing-wave indicator
1	ⓗⓟ 430C microwave power meter
1	ⓗⓟ X485B detector mount
1	ⓗⓟ X487B thermistor mount
1	ⓗⓟ 444A broadband probe
1	ⓗⓟ 809B probe carriage
1	ⓗⓟ X810B slotted section
1	ⓗⓟ X532B frequency meter
1	ⓗⓟ X870A slide-screw tuner
1	ⓗⓟ 120B oscilloscope
1	ⓗⓟ AC-16A cable (dual banana to dual banana)
1	ⓗⓟ AC-16B cable (dual banana to BNC)
2	ⓗⓟ AC-16K cable (BNC to BNC)
1	BNC "tee" connector

Procedure

Section 1—General

1-1 Review the safety precautions on page 42, and the operating instructions for each equipment item.

1-2 Set up the equipment as shown in Fig. 1. The oscilloscope horizontal input should be *a-c coupled,* and the vertical input should be *d-c coupled.* The precision variable attenuator should be set to approximately 20 db.

1-3 A barretter element is an extremely fragile item. If it receives a total power greater than the safe rating of 32.5 mw (for the Sperry 821) from any source (audio, d-c, or rf), it will quickly burn out. It responds quickly to surges and is not self-protecting and rugged like the thermistor. For this reason, the following general precautions should be observed:

1. Always provide enough rf attenuation in the line to maintain rf power in known limits below 10 mw. This provision is especially important when you are using bench klystrons such as the Varian X-13, which can deliver 250 mw. In these experiments, it is good practice to keep 20 db of attenuation

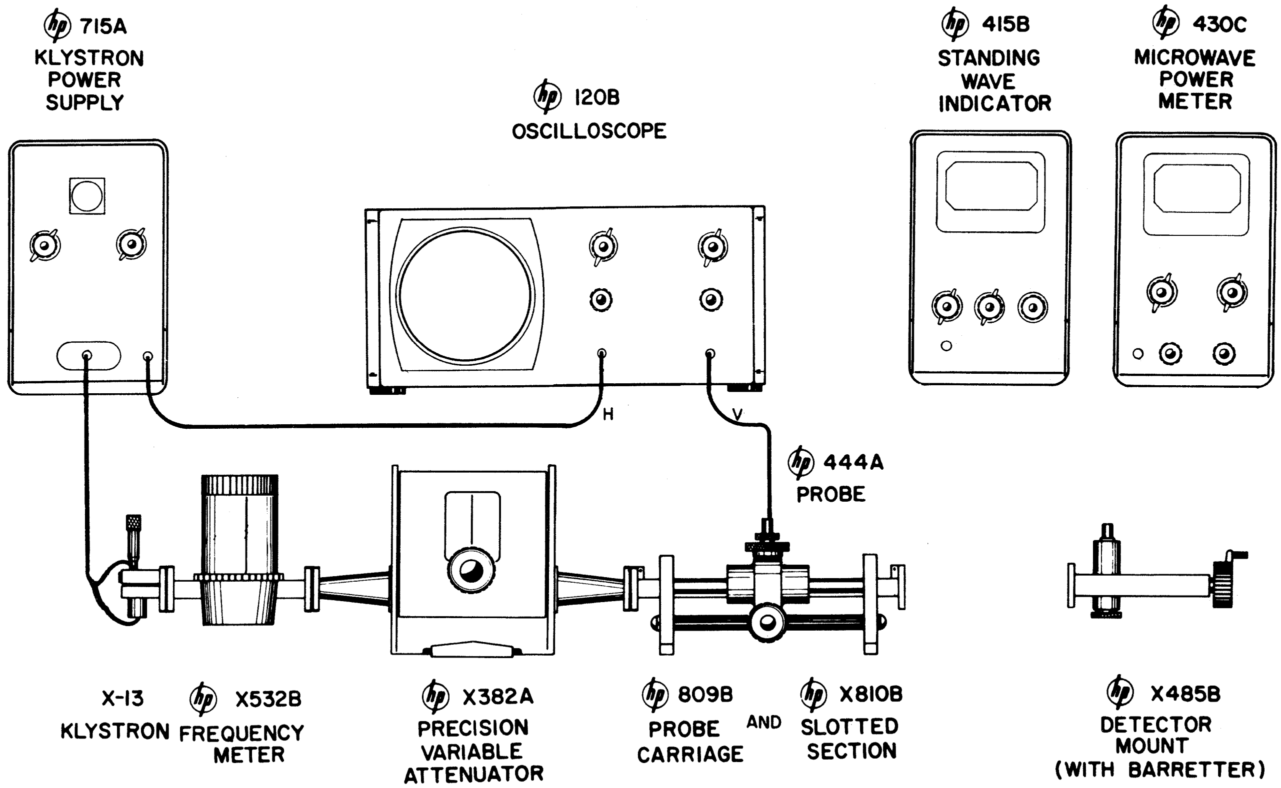


FIGURE 1

in the main line, at least until the power meter indicates that all of the system is operating properly and that the klystron is on the proper mode.

2. Always be sure that there is *no* bolometer bias current before you connect a barretter to the microwave power meter. The voltage developed during open circuit across the input connection of the power meter is sufficient to burn out a barretter—even when the bias current is set to a safe current level for the barretter in use.

3. Always set the bolometer-resistance switch and the bolometer-temperature switch to the proper values before you connect a barretter to the power meter. (For the X485B mount with Sperry 821 barretter, the operating resistance is 200 ohms, with a positive temperature coefficient.)

Connect the barretter to the power meter, and apply only enough bias current to the barretter to permit an indication on the power meter controllable by the ZERO SET. (Never turn the bolometer-bias-current switch past the 6–10 ma block.) Connect the barretter mount to the slotted section and zero the power meter.

1-4 Energize the klystron at 10.0 gc and modulate it with a 1000 cps square wave. (Remember the 20 db attenuator setting.) Use the oscilloscope technique of Experiment 2 to set the reflector and modulating voltages properly.

Section 2—The Barretter Mount

2-1 Connect the standing-wave indicator to the broadband probe in place of the oscilloscope connection. Adjust the 1000 cps modulation frequency to get a peak indication on the standing-wave indicator.

2-2 Tune the mount for maximum power output, and adjust the precision variable attenuator to provide an indicated output power of 1 mw. Record the precision-variable-attenuator dial setting in Table I.

2-3 Measure and record the swr in Table I.

2-4 Reduce the rf power on the barretter by *increasing* the attenuation of the precision variable attenuator. Remove all bolometer bias current from the barretter, and then disconnect the barretter from the power meter.

Section 3—The Thermistor Mount Only

3-1 Connect the thermistor mount (100 ohms resistance, negative temperature coefficient) to the power meter, and apply the necessary bias current. Attach the thermistor mount to the slotted section.

3-2 Repeat the procedure of Steps 2-2 and 2-3 (except for tuning the mount).

Section 4—The Thermistor Mount with Slide-Screw Tuner

4-1 Insert the slide-screw tuner between the thermistor mount and the slotted section.

4-2 Adjust the penetration and position of the slide-screw tuner for a maximum power indication.

4-3 Repeat the procedure of Steps 2-2 and 2-3.

Section 5—Thermistor Mount SWR at Different Power Levels

5-1 Measure the thermistor mount swr at power levels of 10 mw, 1 mw, and 0.1 mw (switching the power meter to the proper range for each reading).

Section 6—Thermistor Protection

6-1 Adjust the power into the thermistor mount for 10 mw.

6-2 Reduce the precision variable attenuator dial setting by 6 db to provide 40 mw of power into the thermistor mount. Measure and record the swr.

Section 7—Barretter Elements as Detectors

7-1 Set the precision variable attenuator to at least 20 db, remove all bias current from the thermistor mount, and remove the mount from the slotted section. Disconnect the standing-wave indicator from the broadband probe.

7-2 Connect the barretter (detector) mount to the standing-wave indicator, using the following procedure:

1. Switch the bolometer bias current in the standing-wave indicator to its high position for the 8.75 ma barretter element.
2. Switch the standing-wave input selector to 200 K ohms. Doing so will minimize any possible surge through the barretter element when the cable is connected.
3. Connect the oscilloscope vertical input (d-c coupled) to the input of the standing-wave indicator with a BNC-tee, and then connect the barretter mount to the indicator (in parallel with the oscilloscope).

4. Switch the standing-wave-indicator input selector to the 200 ohm-bolometer position. The barretter is now ready for use with the standing-wave indicator.

7-3 Energize and modulate the klystron with a 1000 cps square wave. With the oscilloscope internally swept and the vertical input set for maximum sensitivity, the 1000 cps square wave should be visible. It will be necessary to adjust the reflector voltage slightly in order to get 1000 cps square-wave modulation at the peak of the reflector mode. (Although you may slightly and carefully reduce the attenuation provided by the precision variable attenuator so that the waveform is visible, it should not be necessary to apply much more than 1 mw to the barretter. See Table I for a safe attenuator setting.) Note that the waveform is not a true square wave, because the time constant of the barretter is not fast enough to follow the 1 μsec period.

7-4 Find the zero-power baseline on the oscilloscope by temporarily turning off the klystron beam voltage. When the baseline position is known, return to 1000 cps modulation and estimate the barretter time constant.

Name ______________________
Course ______________________
Date performed ______________________
Date turned in ______________________

Results (Experiment 9)

OBSERVED

Step

2-2 Table I
2-3 Table I

3-2 Table I

4-3 Table I

5-1 ______________ (10 mw)
______________ (1 mw)
______________ (0.1 mw)

6-2 ______________

7-4 ______________

Questions

1. Why is there a difference between the attenuator readings of Section 3 and Section 4?

2. Why is the swr recorded for Section 4 not 1.0:1 when the power transfer is maximum?

3. From the data recorded in Table I, which method of measuring power seems to be most efficient?

4. Using the swr recorded in Step 6-2, calculate the amount of power reflected from the thermistor. Note that this reflection of power is an inherent protective quality of thermistors.

5. If you wanted to examine the shape of a radar pulse having a width of 10 μsec, would you use a crystal or barretter detector? Why?

Discussion

TABLE I

MOUNT	ATTENUATOR READING	SWR
Barretter	__________	______
Thermistor	__________	______
Thermistor with slide-screw tuner	__________	______

EXPERIMENT 10

Power Bridges for Microwave Measurements

Object

To study the operation, the use, and the limitations of a typical bolometer type of power measurement bridge, the hp Model 430C.

Theory

One of the most universally used methods of measuring microwave power is by means of a bolometer element which changes rf energy into heat energy and, in turn, changes resistance in proportion to the heat applied.

The change in resistance can be measured and used to determine the amount of rf energy. However, this method of rf-energy determination is accurate only over small variations in power, since bolometer elements have nonlinear characteristics in that the slope of resistance vs. power is not constant. Consequently, power measurements made by a system based on a simple measurement of change in resistance would be quite slow and would require extensive correction tables or charts.

The above-mentioned difficulty may be avoided by bringing the bolometer element up to a predetermined operating resistance with d-c power before any rf power is applied. In the simplest case, a circuit to translate microwave power to a meter reading can look like that in

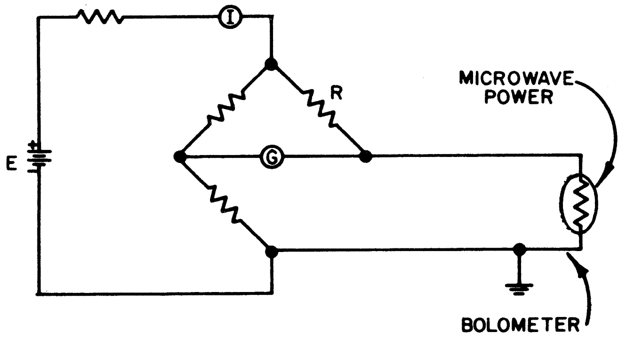


FIGURE 1

Fig. 1. With this simple Wheatstone bridge, sufficient power is supplied by the battery, so the resistance of the bolometer is made equal to its normal operating resistance. The remaining bridge resistors are chosen so that the bridge balances in this condition. When microwave power is supplied to the bridge, the bolometer heats, its resistance changes, and the bridge becomes unbalanced. The balance is restored by decreasing the battery power, and this decrease is taken as a measure of the rf power. Consequently, the bolometer element is used only to indicate specific reference levels; therefore, bolometer nonlinearity is of no concern.

To eliminate the somewhat tedious adjustments and measurements needed to determine the rf power with a simple bridge circuit, specially designed bridges such as the hp Model 430C are used (Fig. 2). In the 430C, the bolometer element is brought to its predetermined operating resistance (100 or 200 ohms) in the absence of rf power by the application of both d-c and af power. The circuit is so arranged that audio power is automatically removed as rf power is applied. The amount of af power removed is displayed by an accurate vacuum-tube voltmeter, calibrated to show a *decrease* in audio power as an *increase* in rf power. The heart of the hp 430C is a self-balancing audio bridge in which the bolometer element functions as one arm.

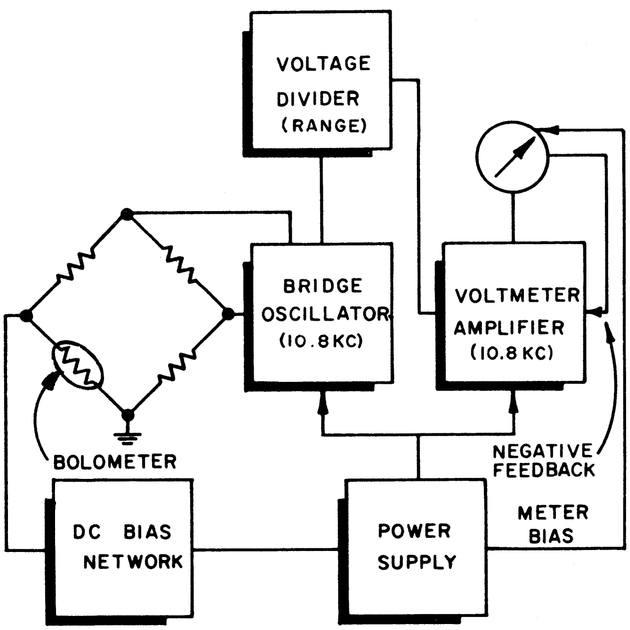


FIGURE 2

NOTE: In this bridge, as in all so-called self-balancing bridges, actual balance never occurs, since a small unbalance is necessary to provide oscillator feedback.

Bridge resistors are selected and arranged so that the bridge will be balanced when the bolometer element is brought up to a predetermined operating resistance. This balance is accomplished by supplying the bolometer with d-c power from a stable internal source, and with approximately 10.8 kc of af power from an internal RC oscillator. If, in the absence of rf power, 15 mw should be required to bring a bolometer element up to operating resistance, 5 mw would be supplied automatically by the af source if 10 mw were supplied by the d-c supply; 10 mw would be supplied automatically if the d-c supply furnished only 5 mw.

In theory, an infinite number of af–d-c power combinations might be used to bring the bolometer element up to operating resistance. In practice, the number is limited by the requirements imposed by the vacuum-tube voltmeter used to read the amount of af power removed when rf power is applied. This vacuum-tube voltmeter, which incorporates a substantial amount of negative feedback to minimize the effect of nonlinear crystal resistance in the meter circuit, is arranged to zero on a specific level of af power. For example, on a 10 mw range, the meter zeros on 12 mw of audio power and reads full scale (+ 10 mw rf) with 2 mw of audio power. Thus, with reference to the 15 mw bolometer example, only 3 mw would be supplied by the d-c source. However, on the lower rf power ranges, the amount of audio power required to zero and to give full-scale readings decreases, and, therefore, d-c power requirements increase. In the previous example, at zero set, on the 1 mw range 1.2 mw would be supplied by the audio source and 13.8 mw by the d-c supply; on the 0.1 mw range, 0.12 mw audio power and 14.88 mw d-c power would be required. Under such conditions, it is easy to see the necessity for exceptional d-c stability on the lower rf power ranges.

Detailed bridge and oscillator operation. *The bolometer bridge* is a resonant bridge, one leg of which is adjustable to either 400 ohms or 200 ohms to establish a 3:1 voltage relationship when a 200 ohm or a 100 ohm bolometer element is used, respectively. As shown in Figs. 3 and 4, the circled values 1, 3, $1 \pm 3/A$ represent the balance voltage relationships around the bridge; the boxed values show resistance ratios. If the bridge is out of balance, it will permit either an increase or a decrease in transmission through it from the oscillator, and thus an increase or decrease in the current through the bolometer element. The successful balance of the bridge is determined by the resistance value of the element. When the element reaches its operating resistance, determined by the RESISTANCE switch setting, the bridge balances, and the oscillator is forced to adjust its power output accordingly. *The oscillator circuit* is composed of *V1, V2,* and *V3* as a simple feedback oscillator.

Example 1. For purposes of illustration, assume that a 100 ohm thermistor mount is used to measure power on the 10 mw range. Also assume that 15 mw is required to drive the thermistor mount from its high cold resistance down to 100 ohms operating resistance. As soon as the mount is connected, the oscillator sees an unbalanced bridge and supplies nearly maximum output to the thermistor. As the thermistor heats, its resistance drops, and the bridge approaches a balance. However, the meter is

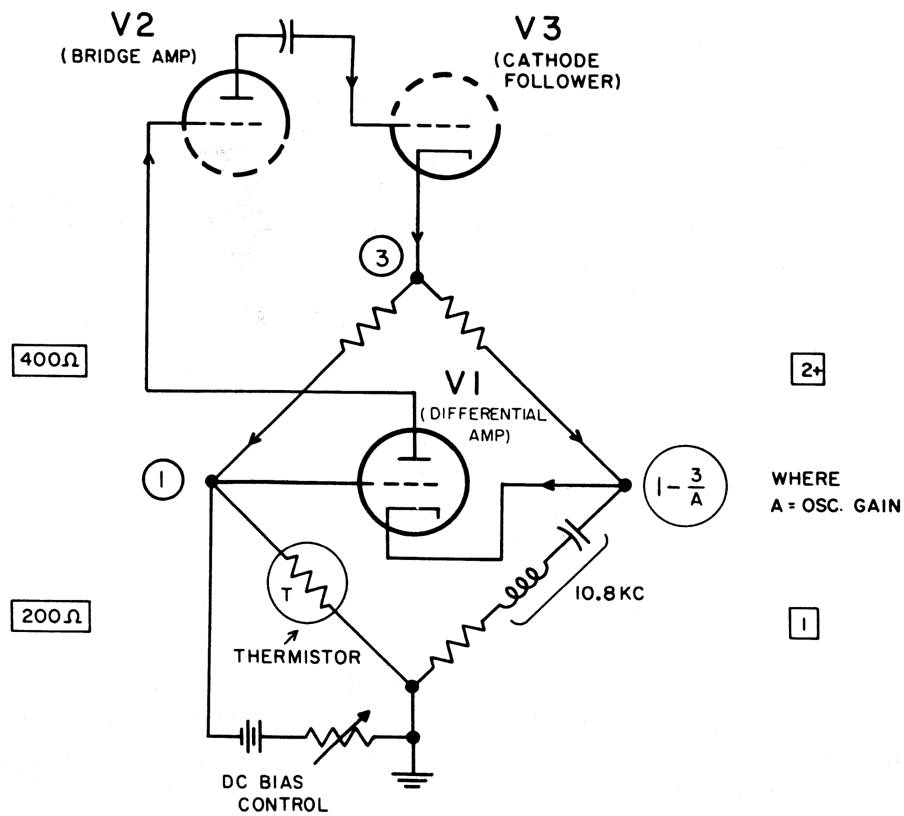


FIGURE 3

calibrated to zero-set on the 10 mw range when the oscillator is supplying 12 mw only. (It must be remembered that the meter samples the oscillator power, and does not directly sample any rf which is externally supplied to the thermistor. The oscillator, in effect, "backs off" when external rf is supplied, so the meter is a reverse-reading, or upscale, device with regard to its calibration.) On the 10 mw range, the meter zero-sets at 12 mw and reads full scale when the oscillator supplies approximately 2 mw. At this point in the example, the oscillator is supplying much more than 12 mw, because the bridge is far out of balance. To force the oscillator to supply only 12 mw and thus to zero the meter, d-c bias current is supplied to the thermistor. If the BIAS CURRENT and the ZERO SET controls are adjusted to furnish exactly 3 mw to the thermistor, the oscillator backs off to 12 mw. The bridge is balanced and the meter reads zero. If the thermistor is connected to an external rf source (10 mw, for example), the 3 mw of d-c which is being supplied to the thermistor remains constant, but the oscillator sees an unbalanced bridge in the form of a decreased grid load on *V1* (see Fig. 3). The output of *V1* decreases until its grid load is again 100 ohms. To accomplish this, it must decrease the exact amount of the external field (10 mw). It now supplies 2 mw instead of 12 mw, and the meter reads full scale, or 10 mw.

The oscillator action for a barretter mount is similar. However, as shown in Fig. 4, the bridge is rearranged to accommodate the positive temperature coefficient of the barretter.

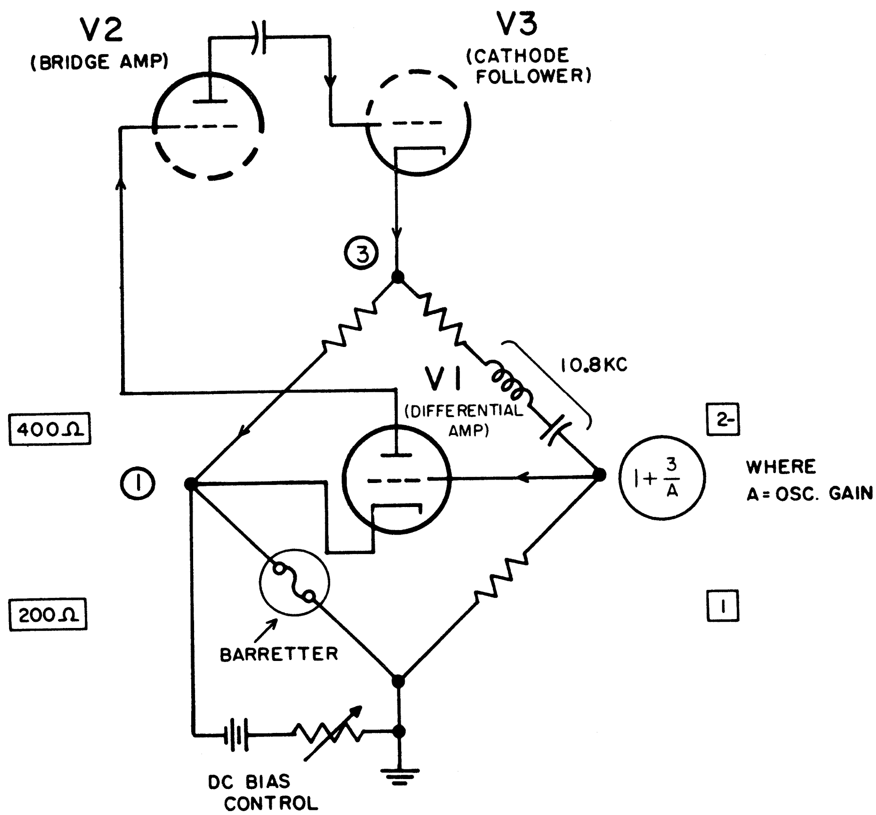


FIGURE 4

Error-producing effects. Experimental data show that, under most conditions of modulated-power measurement, the readings of the Model 430C are accurate irrespective of modulation type or bolometer type. The data also show, however, that errors can occur under certain conditions of low-frequency modulation (sine wave or square wave), as well as at certain repetition rates of pulse modulation. These errors, which are exceptions to the rule, will be discussed as envelope-tracking errors and beat-frequency errors in the following procedure.

References. Additional information to supplement this experiment may be found in the following texts (listed in Appendix C).

Ginzton, Chapter 3: 3.7.
King, Chapter 3: 3.6.
Reich, Chapter 8: 8-5.
Wind, Section 4: 4.03–4.04.

Equipment

QUANTITY	TYPE
1	hp 715A klystron power supply (with cable)
1	Varian X-13 reflex klystron
1	Cooling fan or blower
1	hp X382A precision variable attenuator
1	hp 430C microwave power meter
1	hp X421A crystal detector
1	hp X485B detector mount
1	hp X487B thermistor mount
1	hp 120B oscilloscope
1	hp 200CD wide-range oscillator
1	hp AC-16A cable (dual banana to dual banana)
1	hp AC-16B cable (dual banana to BNC)
1	hp AC-16K cable (BNC to BNC)
1	hp AC-16S cable (dual banana to alligator clips)
1	BNC "tee" connector

Procedure

Section 1—General

1-1 Review the safety precautions on page 42, and the operating instructions for each equipment item.

1-2 Set up the equipment as shown in Fig. 5. The attenuator should be set to approximately 20 db.

1-3 Using proper procedure, connect the thermistor mount to the power meter and prepare to use the power meter on the 10 mw range. Attach the thermistor mount to the attenuator, but do not yet energize the klystron.

Section 2—D-C, AF, and RF Power

2-1 If the af bridge oscillator of the power meter supplies 1.2 times the full-scale power at zero set on any given range, calculate the audio voltage across the 100 ohm element of the thermistor mount.

$$P = (1.2)10(10^{-3}) = \frac{V^2_{\text{rms}}}{100}$$

2-2 Adjust the zero set controls until the power meter is zeroed, and read the rms voltage from the oscilloscope presentation. (An a-c voltmeter may be used, if desired, to yield a direct indication of rms voltage.)

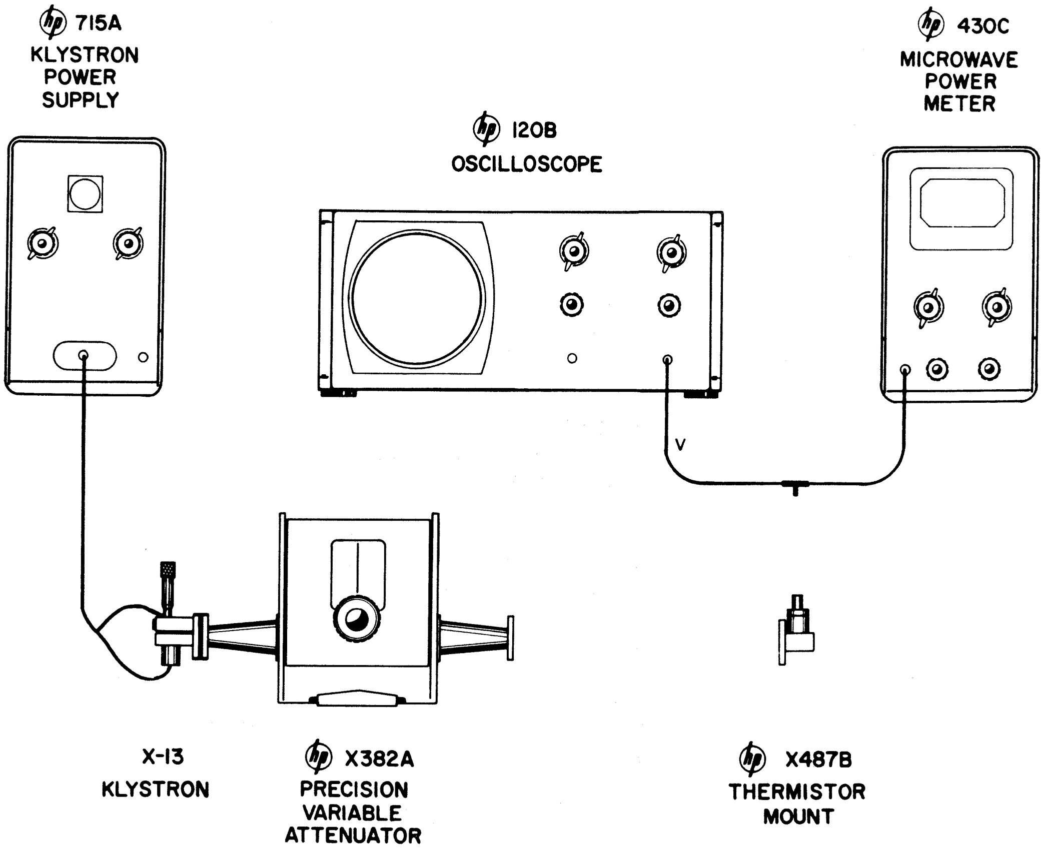


FIGURE 5

2-3 Energize the klystron (cw) for any convenient frequency, and adjust the reflector voltage for an indication on the power meter. Adjust the precision variable attenuator for a 10 mw reading on the power meter. Once again, read the rms voltage indicated on the oscilloscope.

2-4 Calculate the change of af power experienced as the power-meter reading went from 0 to 10 mw.

$$\Delta P = \frac{V_1^2 - V_2^2}{100}$$

2-5 To show that d-c and rf power are basically interchangeable in the thermistor mount, remove rf power by de-energizing the klystron (zero beam voltage). Note that adjusting the d-c bias-current control has the same effect on the af voltage as did varying rf power.

2-6 Repeat the procedure of Steps 2-2 through 2-4 for the power range of 1 mw.

2-7 Using the internal sweep of the oscilloscope, estimate the frequency of the audio signal provided by the power meter oscillator.

Section 3—Envelope-Tracking Error

3-1 At low sine- and square-wave frequencies, the bolometer heating attempts to follow the modulation envelope. Therefore, the bridge oscillator attempts to follow the envelope by adjusting its power output to accommodate the changes in resistance of the bolometer. The modulation frequency becomes impressed upon the oscillator voltage, and the average responding meter will indicate the average of the troughs and peaks of the modulation envelope. The quick response of the barretter, resulting from its small time constant, makes it particularly susceptible to this action.

The power indicated on the meter no longer corresponds to the power being measured, because the meter responds to average voltages rather than to rms voltages. Since the average value is lower than the rms value, the meter reading will be high. The amount of the error depends upon the modulation frequency and upon the Q of the bridge-oscillator circuit. The Q of the oscillator in the Model 430C has been made high to limit this effect and to lower the critical modulation frequencies at which it occurs.

The effect of envelope-tracking error is that as the modulation frequency is reduced, the meter indication is steady until a *critical modulation frequency* is reached. At this point the meter indication starts to increase. As the modulation frequency is further lowered, the meter indication continues to increase to a maximum possible error of approximately 1 db high. This condition is the same for both sine-wave and square-wave modulation.

The maximum error occurs when a barretter is used on the 10 mw range of the instrument at frequencies below 200–300 cps. When other ranges are used or when a thermistor is used, the critical frequency is below 100 cps. Above these critical frequencies, the readings are accurate for sine-wave, pulse, or square-wave operation. Remember: be especially careful when you are using barretters for modulated power measurements.

3-2 Remove the thermistor mount from the system, and attach the crystal detector to the attenuator. Connect the crystal output to the oscilloscope vertical input. Turn off all voltages available at the klystron power supply, and connect an audio oscillator to the external-modulation terminals of the power supply. A capacitor should be used to isolate the oscillator from the d-c voltage of the power supply. Connect the oscilloscope horizontal input to the oscillator output. *NOTE:* The oscillator should be capable of delivering at least 20 volts peak-to-peak.

3-3 Energize the klystron and externally modulate it with the oscillator. With the modulation-voltage control on the klystron power supply set to its maximum value, adjust the reflector voltage until the audio peaks are just driving the reflector to the peak of a mode. (The slow transition in and out of the mode causes a certain amount of fm during the envelope, but it is not critical for this measurement.)

3-4 Remove the crystal detector from the system. Using the *extreme cautions* discussed in Experiment 9, connect the barretter mount (200 ohms, positive coefficient) to the power meter. (Set bias power to zero before connection.) Before attaching the barretter mount to the attenuator, be sure that there is at least 20 db of attenuation between the barretter and the klystron.

3-5 After the barretter mount has been attached to the attenuator, set a maximum power level on the power meter. Starting at approximately 1000 cps, slowly reduce the external klystron modulation (sine-wave) frequency supplied by the audio oscillator until the power-meter reading *increases.* Record this modulation frequency.

3-6 Carefully substitute the thermistor mount for the barretter mount. (*CAUTION:* Be sure to remove all bias current from the barretter mount before disconnecting it from the power meter. Also make sure that there is at least 20 db attenuation present in the system.)

3-7 When the thermistor mount (100 ohms, negative coefficient) has been properly connected to the power meter, and has been attached to the attenuator, repeat Step 3-5. Record the modulation frequency at which the power-meter reading *increases.*

Section 4—Beat-Frequency Error

4-1 Power measurement of pulse-modulated signals is accurate, since it varies in a linear manner with both repetition rate and pulse width. However, as the pulse width is increased, a region will be approached in which square-wave behavior exists. In this square-wave region, critical frequency effects may arise.

When you are measuring pulse-modulated power, avoid repetition rates which are submultiples of the bridge-oscillator frequency (approximately 10.8 kc). On submultiple frequencies, a beating effect occurs in the bridge circuit which is reflected in the meter indications. This effect is particularly active when a barretter is used on the 0.1 mw range, where, if the pulse repetition rate is appropriate, violent beating occurs, followed by the meter's falling to a low value as the bridge oscillator locks in. Any slight variation in the pulse repetition rate will remove this difficulty, since the tuning is so sharp that it is a simple matter to set the rate between successive submultiple frequencies down to about 200 pps.

The effect described above is noticeable when thermistors are employed, but the beats are small with no oscillator lock-in, and readings are not affected. (For convenience, the sine-wave oscillator used in Section 3 to externally modulate the klystron will also be used in this section.)

4-2 Using the same equipment setup as in Step 3-7, tune the klystron to approximately 10 gc and set the power-meter reading to approximately 0.8 mw on the 1 mw range. Starting at 11 kc, slowly reduce the audio sine-wave frequency and watch the power meter for a beat note. Record the af and the pulsed-power beat amplitude in Table I.

4-3 Continue to reduce the af, recording several other beat-note frequencies and amplitudes. (Look for bridge frequency sub-harmonics.)

4-4 *Carefully* replace the thermistor mount with the barretter mount, and repeat the procedures of Steps 4-2 and 4-3.

Name ______________________
Course ______________________
Date performed ______________________
Date turned in ______________________

Results (Experiment 10)

OBSERVED		CALCULATED	
Step		*Step*	
2-2	______________	2-1	______________(v rms)
2-3	______________	2-4	______________
2-6	______________	2-6	______________
	______________		
2-7	______________		
3-5	______________		
3-7	______________		
4-2	Table I		
4-3	Table I		
4-4	Table I		

Questions

1. From your experience in Section 2, can you suggest why the primary method of calibrating microwave power bridges uses audio techniques rather than microwave techniques?

2. Based on your experimental findings, which bolometer element is more susceptible to envelope-tracking error? Why?

3. Did you find the barretter or the thermistor more likely to support beat-frequency error in pulsed-power measurements?

4. Why is stable d-c bias-power-supply operation so important in a power meter such as the hp Model 430C?

5. Why is 10.8 kc selected as the bridge operating frequency?

Discussion

TABLE I

THERMISTOR MOUNT		BARRETTER MOUNT	
Audio frequency	Beat amplitude	Audio frequency	Beat amplitude

EXPERIMENT 11

Crystal Detectors

Object

To study the characteristics of crystal detectors, especially their "square-law" behavior.

Theory

Crystal detectors have wide use in the microwave field because of their sensitivity and simplicity. They are used as "video detectors" to provide either a d-c output when unmodulated microwave energy is applied, or a low-frequency a-c output (up to tens of mc or higher) when the microwave signal is modulated. They are also used as mixers in superheterodyne systems, especially at microwave frequencies, where other mixers (vacuum tubes, for example) are inefficient.

The essential parts of a crystal detector are a semiconducting wafer and a metal "whisker" which contacts the wafer. A typical microwave crystal detector uses a silicon wafer about $\frac{1}{16}$ in. square and a pointed tungsten whisker wire about 0.003 in. in diameter. The other parts comprising a crystal detector or mount are needed simply to support the wafer and the whisker and to couple electrical energy to the detector. Although crystal detectors are successful at microwave frequencies partly because of their extremely small size, their dimensions also limit their power-handling capability; 100 mw is sufficient to damage some crystals.

This experiment will be concerned with the use of crystals as video detectors. They are very convenient for use as detectors on slotted lines and for indicating relative power levels. For these applications the user must usually know the law relating output voltage to applied microwave voltage.

One can easily deduce the law for a simple circuit, which will illustrate the general behavior in more complicated situations. Consider the idealized circuit shown in Fig. 1, where a sinusoidal microwave voltage is applied to the rf terminals. Capacitor C bypasses the rf, leaving d-c current to flow through the milliammeter. A typical crystal detector has a current-voltage characteristic similar to that of Fig. 2. Any such curve can always be approximated by a Taylor series consisting of terms involving powers of v. That is,

$$i = a_0 + a_1 v + a_2 v^2 + a_3 v^3 + \cdots \tag{1}$$

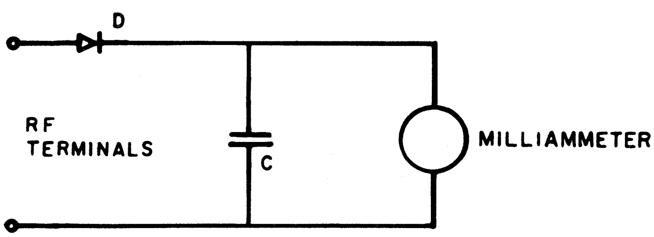


FIGURE 1

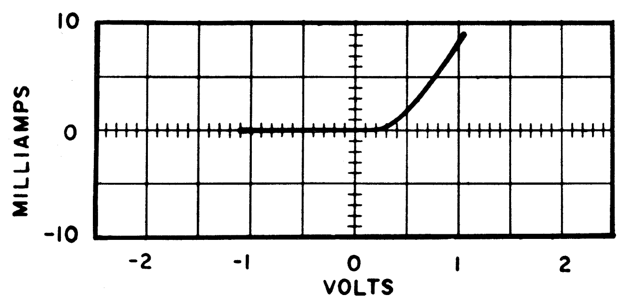


FIGURE 2

If the "operating point" is the origin ($v = 0$, $i = 0$), as it is for Fig. 1, then a_0 is zero.

Let
$$v = A \cos \omega t$$

where A is the amplitude, $\omega = 2\pi f$, and f is the microwave frequency. Substituting in Eq. 1 yields

$$i + a_1(A \cos \omega t) + a_2(A \cos \omega t)^2 + a_3(A \cos \omega t)^3 + \cdots \tag{2}$$

For extremely small signals, all terms of Eq. 2 except the first are negligible, and $i = a_1(A \cos \omega t)$. The current is simply proportional to the applied voltage, and the crystal behaves as a simple resistor, with negligible d-c current flowing through the milliammeter. However, for somewhat larger signals the second term must be included to obtain reasonable accuracy.

$$i = a_1(A \cos \omega t) + a_2(A \cos \omega t)^2$$

or
$$i = a_1(A \cos \omega t) + \frac{a_2 A^2}{2}(1 + \cos 2\omega t)$$

where a standard trigonometric identity has been used:

$$\cos^2\theta = \tfrac{1}{2} + \tfrac{1}{2} \cos 2\theta$$

The current now includes a d-c component $a_2 \frac{A^2}{2}$, which flows through the milliammeter, and a second harmonic component $\frac{(a_2 A^2)}{2} \cos (2t)$, which flows through C. Thus, the milliammeter indication is proportional to the *square* of the amplitude A of the microwave voltage. At still higher signal levels, more terms of Eq. 2 must be retained, and the crystal behavior departs from "square law." Commonly used video crystal-detector circuits are more complicated than Fig. 1. They are characterized by a square-law region limited at the low-power end by the noise level of the meter or amplifier connected to the crystal, and at the high-power end by departure from square-law behavior.

In this experiment, crystal-detector laws will be measured by observing the audio output voltage as a function of the applied 1000 cps square-wave-modulated microwave power. It will be shown that the actual deviation from square law is dependent upon the video load resistance seen by the crystal. In particular, the square-law characteristic of the crystal in the slotted-line probe will be measured with the input impedance selector of the Model 415B standing-wave indicator set to "200KΩ," and then with the selector set to "200Ω." If the particular slotted-line crystal is an "average" crystal in good condition, it will be evident that there is an "optimum" load value (somewhere between 200 and 200,000 ohms) for maximum square-law-detection range without exceeding a specified deviation.

Possible sources of error to consider include inaccuracies in setting the microwave power level and in measuring the audio voltage. The hp Model X382A waveguide attenuator is accurate within ± 2 per cent of its reading or 0.1 db, whichever is greater. The hp Model 415B standing-wave indicator is accurate within ± 0.1 db per 10 db step, with a minimum cumulative error of ± 0.2 db. The meter face is calibrated with the assumption that a square-law detector will be used; consequently, any deviation from square law as the power is changed will show up as a discrepancy between the attenuator reading and the meter indication.

References. Additional information to supplement this experiment may be found in the following texts (listed in Appendix C).

Ginzton, Chapter 2: 2.3.
King, Chapter 3: 3.3.
Reich, Chapter 6: 6-29.
Wind, Section 15.

Equipment

QUANTITY	TYPE
1	hp 715A klystron power supply (with cable)
1	Varian X-13 reflex klystron
1	Cooling fan or blower
1	hp X375A variable flap attenuator
1	hp X382A precision variable attenuator
1	hp 415B standing-wave indicator
1	hp 430C microwave power meter
1	hp X421A crystal detector
1	hp X487B thermistor mount
1	hp 444A broadband probe
1	hp 809B probe carriage
1	hp X810B slotted section
1	hp 120B oscilloscope
1	hp AC-16A cable (dual banana to dual banana)
1	hp AC-16B cable (dual banana to BNC)
2	hp AC-16K cable (BNC to BNC)

Procedure

Section 1—General

1-1 Review the safety precautions on page 42, and the operating instructions for each equipment item.

1-2 Set up the equipment as shown in Fig. 3. The variable flap attenuator should be set to at least 10 db, and the precision variable attenuator should be set to 0 db.

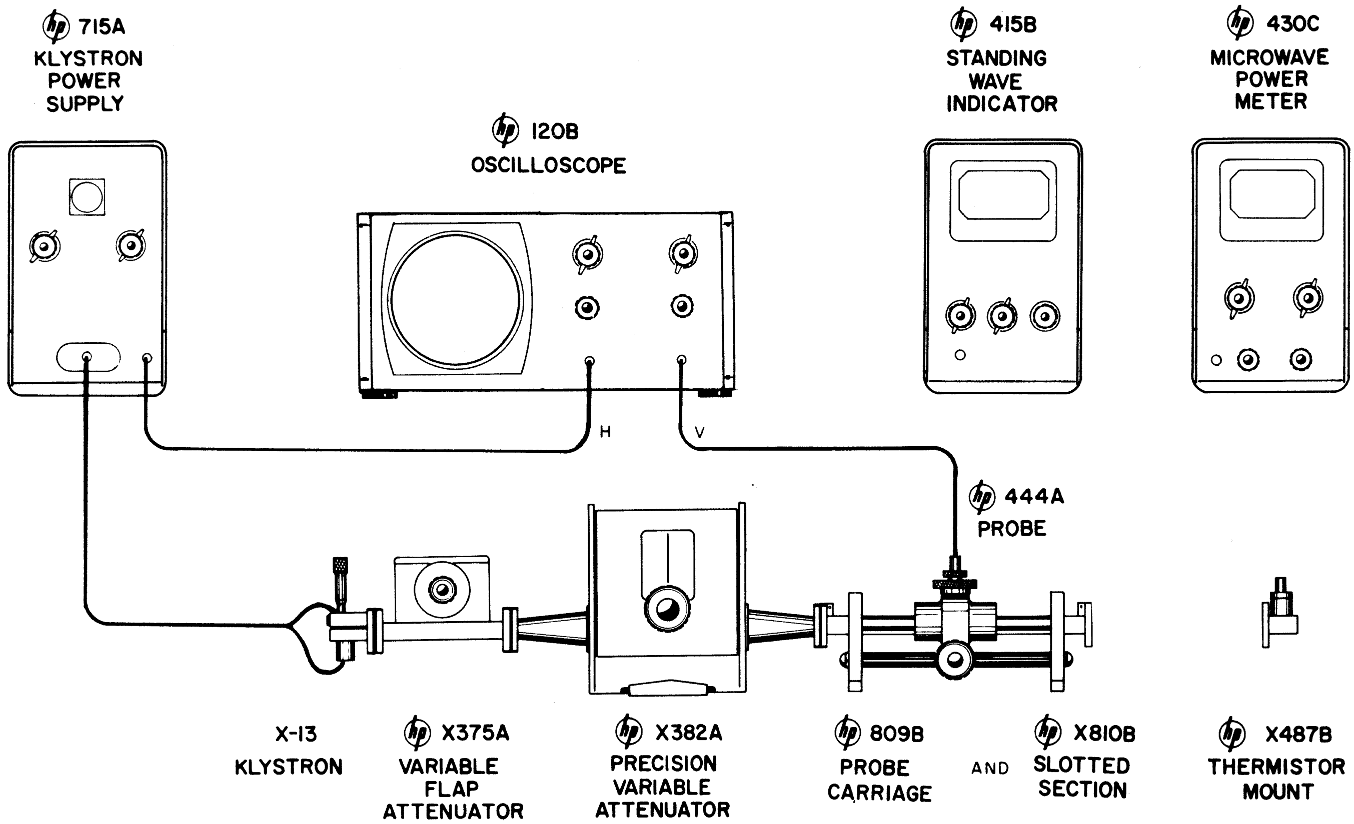


FIGURE 3

1-3 Energize the klystron at any convenient rf frequency and modulate it with a 1000 cps square wave. Use the oscilloscope technique of Experiment 2 to set the reflector and modulating voltages properly. When this is done, connect the probe to the standing-wave indicator, and adjust the 1000 cps modulation frequency for a peak indication on the standing-wave indicator.

1-4 Using proper procedure (Experiment 3), connect the thermistor mount to the power meter, and attach the mount to the slotted section.

Section 2—The Broadband Probe (with Crystal)

2-1 Using the variable flap attenuator, set the power-level reading on the power meter to 10 mw.

2-2 Move the broadband probe along the slotted section to a maximum point on the standing-wave pattern, and then insert the probe until the indicated power on the power meter drops about 0.3 db, or until the probe has maximum penetration.

NOTE: This procedure severely overcouples the probe for normal slotted-line measurements, and is used here only to study the crystal detector in the probe at higher-than-usual power levels.

2-3 With the standing-wave-indicator gain controls set to a maximum and the input selector set to XTAL-200K ohms, adjust the precision variable attenuator to 50 db. Record the standing-wave-indicator reading, using the db scales on the meter face and the range switch, on Table I. (If the standing-wave-indicator reading is too close to the noise level, adjust the precision variable attenuator for less attenuation.) Decrease the attenuation in 5 db steps, and record the precision-variable-attenuator dial reading and the standing-wave-indicator reading for each step.

NOTE: Do not change any controls during the above measurements except the precision-variable-attenuator dial and the standing-wave-indicator range switch. The standing-wave indicator is used simply as a sensitive, tuned audio voltmeter.

OPTIONAL: You may wish to connect a conventional a-c voltmeter to the crystal detector at the higher-power levels in order to measure the absolute voltage out of the detector. Bear in mind that the standing-wave indicator is calibrated with the assumption that a square-law detector (rf power proportional to voltage out) will be used. Consequently, a change in audio voltage into the standing-wave indicator by a factor of 10 produces a *10 db* change in its indication.

2-4 Plot the results of Step 2-3 on linear graph paper, with the attenuator readings plotted horizontally (numbers decreasing to the right) and the standing-wave-indicator readings plotted vertically (numbers decreasing upward).

2-5 Change the standing-wave-indicator input selector to XTAL-200 ohms, and repeat Steps 2-3 and 2-4.

Section 3—The Waveguide Crystal—The Detector Mount

3-1 Remove the slotted section, the probe carriage, and the broadband probe from the system, and connect the thermistor mount directly to the precision variable attenuator.

3-2 With the precision variable attenuator set at 0 db, adjust the power level indicated by the power meter to 0.5 mw, using the variable flap attenuator. (Excess power can be further reduced by operating the klystron at less beam voltage.) This procedure causes the "on" portion of the square-wave-modulated microwave power to be 1 mw, which is the highest power for which the hp Model X421A crystal detector was designed.

3-3 Replace the thermistor mount with the crystal-detector mount (connected to the standing-wave indicator). With the standing-wave-indicator input selector set to XTAL-200K ohms, repeat Steps 2-3 and 2-4.

3-4 Change the standing-wave-indicator input selector to XTAL-200 ohms, and repeat Steps 2-3 and 2-4.

Section 4—Crystal-Detector Load Resistor

Remove the connector cap of the Model X421A crystal-detector mount and determine the value of the load resistor installed there.

Name ______________________
Course ______________________
Date performed ______________________
Date turned in ______________________

Results (Experiment 11)

OBSERVED		CALCULATED	
Step		*Step*	
2-3	Table I	2-4	(plot)
2-5	Table I	2-5	(plot)
3-3	Table I	3-3	(plot)
3-4	Table I	3-4	(plot)
4	______________ohms		

Questions

1. Did your measurements verify that the "crystal-detector law" of the broadband probe crystal begins to deviate from square-law behavior with increasing power by "drooping" for one input-selector setting, and "rising" with the other setting?

2. At approximately what meter indications did the deviations in Question 1 become significant?
   With input selector at 200KΩ: ______________
   With input selector at 200Ω: ______________
   (For normal slotted-line measurements, it is advisable to keep the probe penetration and power levels such that the meter always indicates below these "critical" levels, unless correction factors or more elaborate setups are used.
3. Based upon your experience in Section 3 of the experiment, what is the approximate minimum signal detectable with a typical video crystal detector (Model X421A) followed by a tuned audio amplifier, such as the Model 415B standing-wave indicator?
   ______________dbm

NOTE: The Model X421A crystal-detector mount includes a disc resistor ahead of the crystal in order to improve the swr over the entire waveguide range. Without this disc resistor, which absorbs some microwave power, the crystal detector would be somewhat more sensitive, but it would have a much higher swr.

Discussion

TABLE I

STEP ⟶	2-3		2-5		3-3		3-4	
	Precision variable attenuator reading	Standing wave indicator reading	Precision variable attenuator reading	Standing wave indicator reading	Precision variable attenuator reading	Standing wave indicator reading	Precision variable attenuator reading	Standing wave indicator reading

EXPERIMENT 12 Cable Measurements

Object

To become familiar with the method of measuring coaxial-cable characteristics at microwave frequencies.

Theory

Transmission lines (cables) are characterized by two parameters: the characteristic impedance Z_0 and the propagation constant γ. The characteristic impedance is the ratio of voltage to current in a travelling wave on the line. The propagation constant is a complex number whose real part α (attenuation constant) yields the change in amplitude of the wave along the line, and whose imaginary part β (phase constant) gives the change in phase of the wave along the line.

Wave amplitude at a point on the line, a distance x from a point of reference, is expressed as $e^{-(\alpha + j\beta)x}$. Attenuation for a distance x is then 20 $\log_{10}(e^{\alpha x})$ or $(8.686\alpha x)$ db, and attenuation per unit length is 8.686α. The phase change (radians) per unit length, β, is related to frequency and velocity of propagation of the wave: $\beta = 2\pi/\lambda = 2\pi f/v$, where λ is the wavelength, f is the frequency, and v is the velocity of propagation.

In this experiment both α and v are measured directly. Characteristic impedance Z_0 is derived from an additional measurement of the total cable capacitance: $Z_0 = l/Cv$, where l is the physical length of the cable sample and C is the capacitance of the cable sample.

Figure 1 shows the setup used for measurement of α and v. The experiment is limited in accuracy by the residual reflections present at the transition from waveguide to coaxial line.

These reflections will be fairly small if the cable sample used has a characteristic impedance close to 50 ohms.

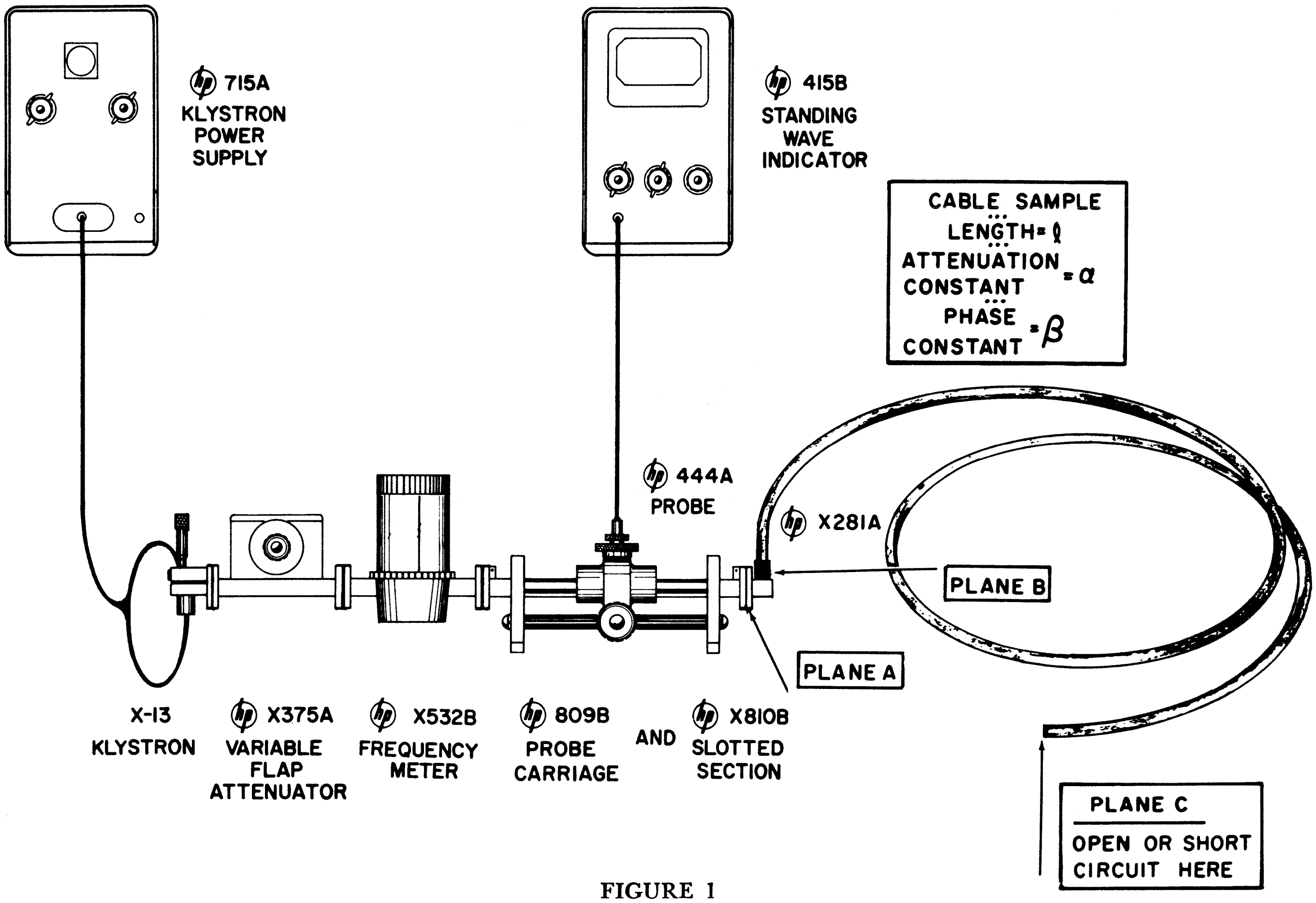


FIGURE 1

Assume for the moment that the transition is reflectionless. With the cable short-circuited at plane C, the reflection coefficient at plane B is $-e^{-2(\alpha+j\beta)l}$. At plane A (the end of the slotted section) the reflection coefficient is changed in phase by twice the transmission phase shift of the transition. A measurement of swr yields the magnitude of the reflection coefficient, which is $e^{2\alpha l}$, the two-way cable loss. A measurement of the change in phase of the reflection coefficient for a small change in frequency from F_1 to F_2 yields the value

$$\frac{4\pi l}{v}(F_1 - F_2),$$

from which v can be calculated.

The situation becomes more complicated when the waveguide-to-coaxial transition has reflections. Conditions at plane A with plane C short-circuited can be approximately represented by the vector diagram of Fig. 2(a). The incident wave amplitude is E_i, and the main reflected wave which traverses the cable twice is $-kE_ie^{-(\alpha+j\beta)l}$, where k is also a vector quantity which takes into account both the change in amplitude and the change in phase of the wave in traversing the transition twice. The vector KE_i is composed of the transition reflection and the multiple signals reflected in the cable between plane B and plane C. The magnitude of K would be on the order of 0.2 or more, and k on the order of 0.96 if the transition swr (measured with a matched load termination) was on the order of 1.5. The sum of the two vectors yields the actual reflected signal at plane *A*, which is $E_i\rho_{sc}$. If the cable is now open-circuited at plane *C*, the vector KE_i is not changed significantly. The main reflected signal is reversed in phase by 180 deg, as shown in Fig. 2(b); this reversal yields the vector $E_i\rho_{oc}$ as the net reflected wave at plane *A*. By a combination of short- and open-circuit measurements of the reflection coefficient, it is possible to reduce the error caused by transition reflections.

It can be shown by vector construction that $\rho_{oc} - \rho_{sc}/2$ is equal to $ke^{-2(\alpha + j\beta)l}$, which is within a few percent of the desired value $e^{-2(\alpha + j\beta)l}$. As long as *K* is small compared with *k*, however, the arithmetic average of the reflection coefficients' magnitudes is a good approximation for the vector difference. The attenuation calculations for this experiment use this approximation.

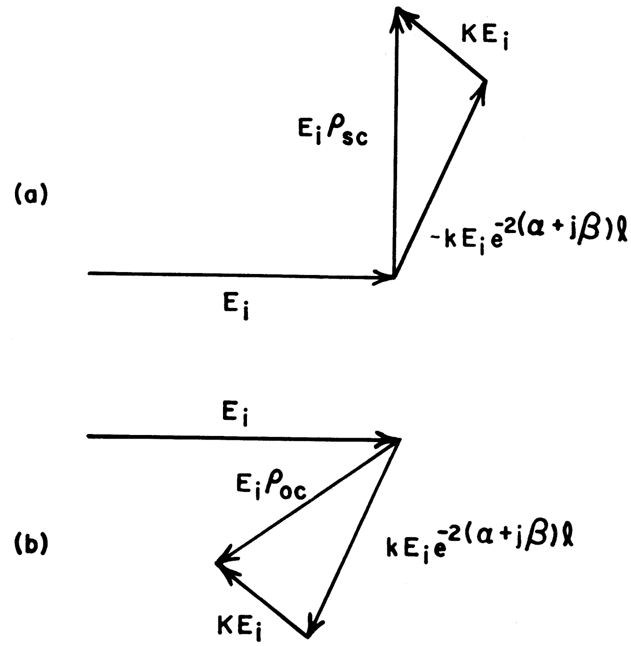


FIGURE 2

Velocity of propagation measurements depend upon finding the frequency difference required to shift the phase of the resultant reflection coefficient by 360 degrees (2π radians). This technique provides for convenient measurement and interpretation.

Equipment

QUANTITY	TYPE
1	ⓗ 715A klystron power supply (with cable)
1	Varian X-13 reflex klystron
1	Cooling fan or blower
1	ⓗ X375A variable flap attenuator
1	ⓗ X281A waveguide-to-coaxial adapter
1	ⓗ 415B standing-wave indicator
1	ⓗ X532B frequency meter
1	ⓗ 444A broadband probe
1	ⓗ 809B probe carriage
1	ⓗ X810B slotted section
1	ⓗ 120B oscilloscope
1	ⓗ AC-16A cable (dual banana to dual banana)
1	ⓗ AC-16B cable (dual banana to BNC)
1	ⓗ AC-16C cable (coaxial 50 ohm RG-9A/U, with one male and one female Type N connector
1	ⓗ AC-16K cable (BNC to BNC)
1	ⓗ 803A-76G shorting plug

Procedure

Section 1—General

1-1 Review the safety precautions on page 42, and the operating instructions for each equipment item.

1-2 Set up the equipment as shown in Fig. 1. The variable flap attenuator should be set at 10 db, and the cable sample should be left open-circuited.

1-3 Energize the klystron at approximately 10.0 gc, using 1000 cps square-wave modulation. Use the oscilloscope technique of Experiment 2 to set the reflector and modulating voltages properly. With the broadband probe connected to the standing-wave indicator, adjust the 1000 cps modulation frequency to get a peak indication on the standing-wave indicator.

Section 2—Cable Attenuation

2-1 For low values of attenuation normally found in nominal lengths of cable, a shorted-cable test can be used for determining the attenuation. This method is quite accurate as long as a reasonably accurate measurement of the swr can be made; it is, therefore, limited to cable attenuations on the order of up to 5 or 10 db.

2-2 Adjust the broadband probe penetration into the slotted section to get a standing-wave indication on the 40 db range of the standing-wave indicator.

2-3 Measure and record (in Table I) the cable swr with plane *C* open-circuited.

2-4 Short the cable with the shorting plug, and again measure and record the swr.

2-5 Repeat Steps 2-3 and 2-4 for other frequencies of Table I.

2-6 Calculate the reflection coefficients from the values of swr as read from the standing-wave indicator. Using the approximation of Fig. 2, find the average of the open- and short-circuit reflection coefficients, and record this on Table I.

2-7 Enter the appropriate parameter scale on a Smith Chart with the average reflection coefficient and read out the two-way return loss. Calculate the one-way attenuation loss in the cable sample.

2-8 Measure the length of the cable sample and record the attenuation per foot in the appropriate column. Calculate the return loss per foot. (It is also possible to use the formula for one-way attenuation, 10 $\log_{10}\rho$.)

2-9 Plot attenuation per foot in Fig. 4.

Section 3—Propagation Constant; Phase Constant

3-1 The phase constant of a transmission line (β) is related inversely to the velocity of propagation of that cable. Since the velocity of propagation is the easier parameter to measure, this measurement will be made.

The length of cable with an open circuit at plane *C* sets up a large number of half-wavelength standing waves along its length, which, of course, results in a null back at the probe location in the slotted section. If there are enough standing-wave minimums along this length, a small change of frequency (causing a small change of individual half wavelengths) is enough to result in an appreciable movement of the null position at the slotted section, a number of wavelengths away from the open end. This movement is commonly known as the long line effect.

3-2 Disconnect the standing-wave indicator from the broadband probe, and connect the oscilloscope vertical input to the probe. Connect the oscilloscope horizontal input to a 60 cps source, and modulate the klystron at 60 cps. Adjust the klystron to a frequency of approximately 8.5 gc, and move the broadband probe against the cable end of the slotted section.

3-3 If the klystron frequency were swept back and forth through a particular frequency such that the center frequency of the mode placed a null right in the center of the mode plot [Fig. 3(A)], the null pattern would actually appear as in Fig. 3(B). Move the probe slightly and note that as it moves down the slotted section, it meets the null at a slightly different frequency, as in Fig. 3(C).

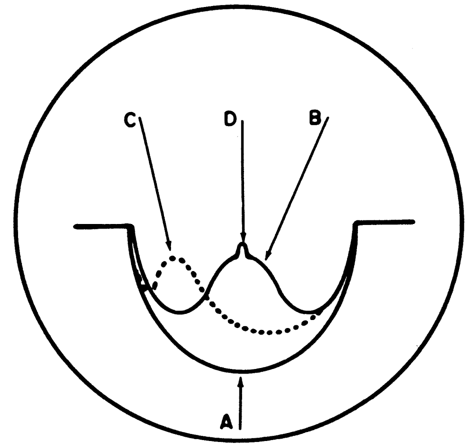


FIGURE 3

3-4 With the probe against the cable end of the slotted section and the cable open-circuited, adjust the klystron frequency until the null appears near the center of the mode plot, as shown in Fig. 3(B). Adjust the frequency meter until its "pip" appears in the center of the null, as shown in Fig. 3(D). Record the frequency as F_1 in Table II.

3-5 Slowly adjust the klystron frequency with the klystron micrometer until the *next* adjacent null moves to the center of the repeller mode pattern, as in Step 3-4. Tune the frequency meter to the center of this new null, and record the new frequency as F_2 in Table II.

3-6 Using the following equation, calculate the velocity of propagation.

$$v = 2l\,(F_2 - F_1) \text{ cm/sec}$$

where l = cable length in centimeters.

F_1 = frequency of first null.

F_2 = frequency of second null.

This calculation results in the velocity of propagation for the higher frequency of the F_2 and F_1 measurements. Record the results in Table II. This calculation ignores the electrical length of the system between the end of the coaxial cable and the point at which the probe is inserted in the line. A more accurate method would actually place a crystal detector probe in a T connector on the end of the cable itself. A somewhat closer approximation can be made with the slotted-section technique, if it is assumed that the physical length between the end of the cable and the point of insertion of the broadband probe is added to the cable-length measurement.

3-7 Measure the low-frequency cable capacity by using a standard 1 kc capacity bridge, if one is available. If a capacity bridge is not available, assume a cable capacity of 30 *pf*/ft.

3-8 Calculate characteristic impedance Z_0, using the following equation: $Z_0 = \dfrac{l}{C \times v}$
where l is cable length in centimeters, C is total capacity in farads, and v is velocity in cm/sec. Record the characteristic impedance in Table II.

3-9 Repeat Steps 3-2 through 3-8 for the other frequencies in Table II, and plot Z_0 calculations on Fig. 5.

Name ____________________
Course ____________________
Date performed ____________________
Date turned in ____________________

Results (Experiment 12)

OBSERVED		CALCULATED	
Step		*Step*	
2-3	Table I		
2-4	Table I		
2-5	Table I	2-6	Table I
		2-7	Table I
		2-8	Table I
		2-9	Figure 4
3-4	Table II		
3-5	Table II	3-6	Table II
3-7	____________pf	3-8	Table II
3-9	Table II	3-9	Table II
			Figure 5

Questions

1. An approximation for calculating velocity of propagation for a coaxial cable is to divide the velocity of light (3×10^{10} cm/sec) by $\sqrt{k}$, where k is the dielectric constant of the cable dielectric. If polyethylene ($k = 2.26$) is the dielectric of your cable sample, how does this approximation agree with your calculations in Table II?

2. How many wavelengths are there in 6 ft of cable at 10 gc?

3. Can you suggest a way of measuring the attenuation of 100 ft of RG-9A/U cable at 10 gc (about 5 db per 100 ft)?

4. Is the "open-shorted" technique of Section 2 easily used with cables having high values of attenuation? Defend your answer.

Discussion

TABLE I

STEP ⟶	2-3	2-4	2-6			2-7		2-8
Frequency (gc)	swr_{oc}	swr_{sc}	ρ_{oc}	ρ_{sc}	ρ average	**Total return loss**	**Total one-way loss**	**Attenuation per foot**
8.5								
9.0								
9.5								
10.0								
10.5								
11.0								
11.5								
12.0								

Cable length (l) = ____________ ft. = ____________ cm

TABLE II

STEP⟶	3-4	3-5	3-6	3-8
Nominal frequency (gc)	Lower frequency (F_1) (gc)	Higher frequency (F_2) (gc)	v (cm/sec)	Z_0 (ohms)
8.5				
9.0				
9.5				
10.0				
10.5				
11.0				
11.5				
12.0				

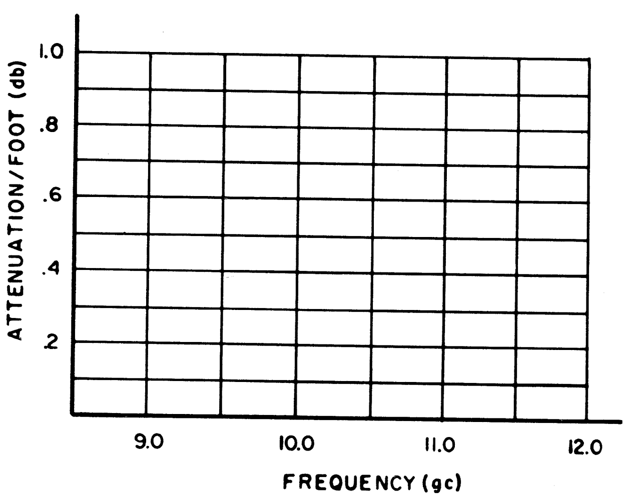


FIGURE 4

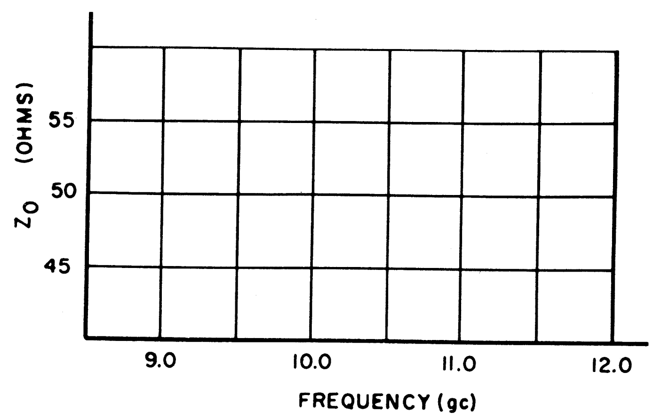


FIGURE 5

EXPERIMENT 13

Mismatch Loss and Maximum Power Transfer

Object

To study the effect of mismatch loss on transmission-line measurements and the conditions which must exist for maximum power transfer from the source to the load.

Theory

Mismatch loss is a very important factor that is frequently overlooked when measurements are made on transmission lines. Simply defined, it is a measure of the loss of transmitted power caused by reflection. If both the load and generator are mismatched, multiple reflections occur which can add in random phase determined by the electrical length of the line. This random phase produces an ambiguity in power and attenuation measurements, and causes a variation of transferred power. If the swr of both source and load are known and if attenuation in the line is neglected, the maximum and minimum values of the ambiguity can be determined.

A chart for determining the maximum and minimum power losses for various load and source swr is shown in Fig. 1. Note that for higher load and source swr the maximum power loss increases, and that as load and source swr come closer to being equal the minimum power loss approaches 0 db. In addition, note that if either the load or source swr is unity, the maximum and minimum power losses resulting from the other mismatch are equal.

POWER LOSS CURVES

(SOLID LINES INDICATE MINIMUM POWER LOSS, BROKEN LINES MAXIMUM POWER LOSS)

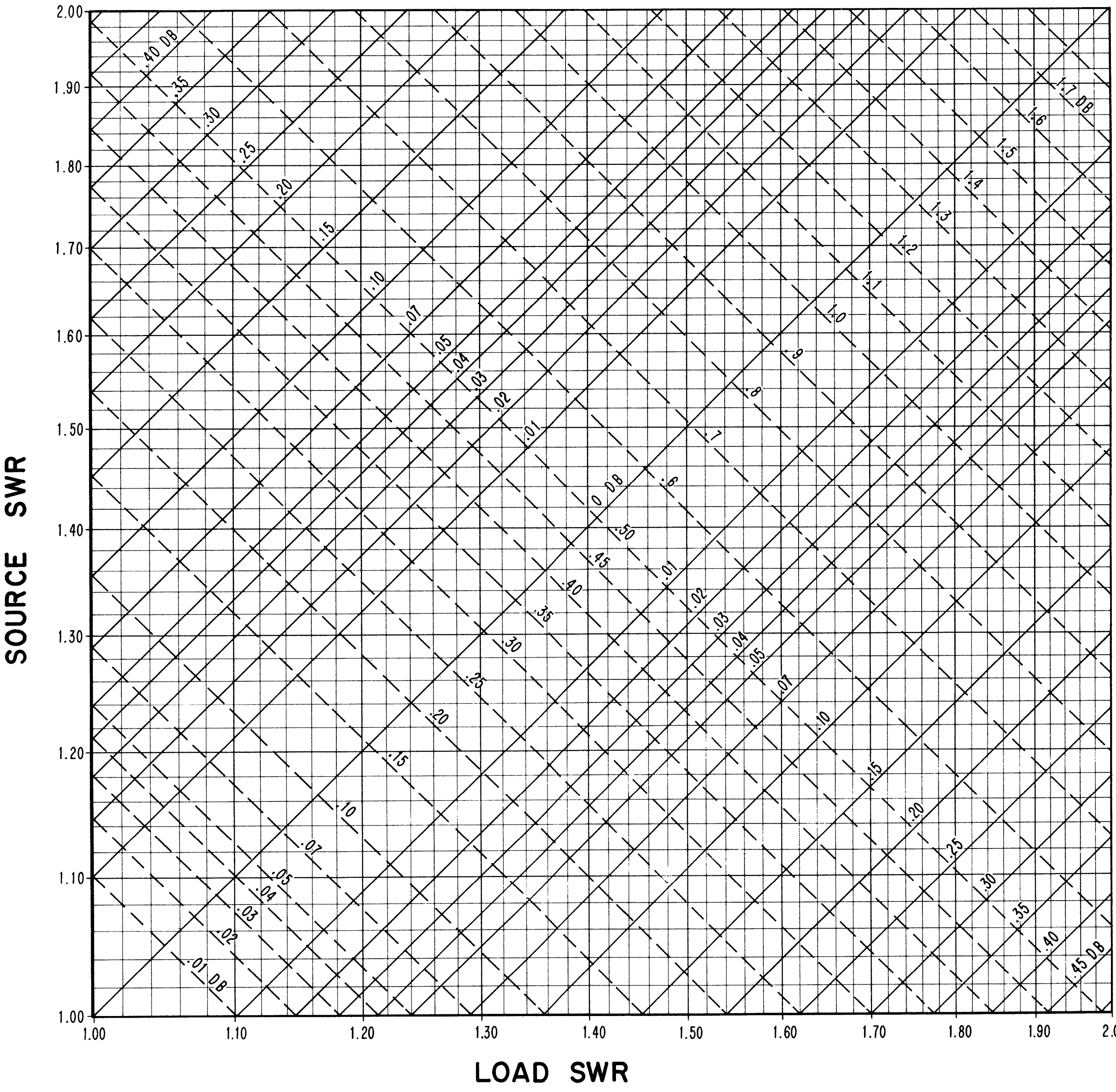


FIGURE 1

For matching, tuners may be used for single-frequency applications. If sufficient power is available and it is desired to reduce reflections on a transmission line, attenuating "pads" may be used between a high swr source (for example) and the rest of the line. The use of such pads will reduce the source swr to the swr of the attenuator. (Naturally, the swr of the attenuator being used as a pad must be lower than the mismatch that is being masked out.)

References. Additional information to supplement this experiment may be found in the following texts (listed in Appendix C).

Ginzton, Chapter 11: 11.1.
King, Chapter 2: 2.7–2.8, 2.13.
Reich, Chapter 2: 2–11.
Wind, Section 3: 3.03.

Equipment

QUANTITY	TYPE
1	hp 715A klystron power supply
1	Varian X-13 reflex klystron
1	Cooling fan or blower
1	hp X375A variable flap attenuator
1	hp X382A precision variable attenuator
1	hp 415B standing-wave indicator
1	hp 430C microwave power meter
1	hp 444A broadband probe
1	hp 487B thermistor mount
1	hp X532B frequency meter
1	hp 809B probe carriage
1	hp X810B slotted section
1	hp 120B oscilloscope
2	AC-16K cable (BNC to BNC)

Procedure

Section 1—General

1-1 Review the safety precautions on page 42, and the operating instructions for each equipment item.

1-2 Set up the equipment as shown in Fig. 2. The precision variable attenuator should be set to approximately 15 db, and the variable flap attenuator should be set to 0 db.

1-3 Using the proper procedure (Experiment 3), connect the thermistor mount to the power meter and attach the mount to the slide-screw tuner.

1-4 Energize the klystron at 10.0 gc and modulate it with a 1000 cps square wave. Use the oscilloscope technique of Experiment 2 to set the reflector and modulating voltages properly. When the broadband probe is reconnected to the standing-wave indicator, adjust the 1000 cps modulation frequency to get a peak indication on the indicator.

Section 2—Finding Power Loss

2-1 With the slide-screw tuner completely out of the waveguide, measure and record (in Table I) the swr of the thermistor mount. If the mount has a better match than 1.2, provide additional mismatch at

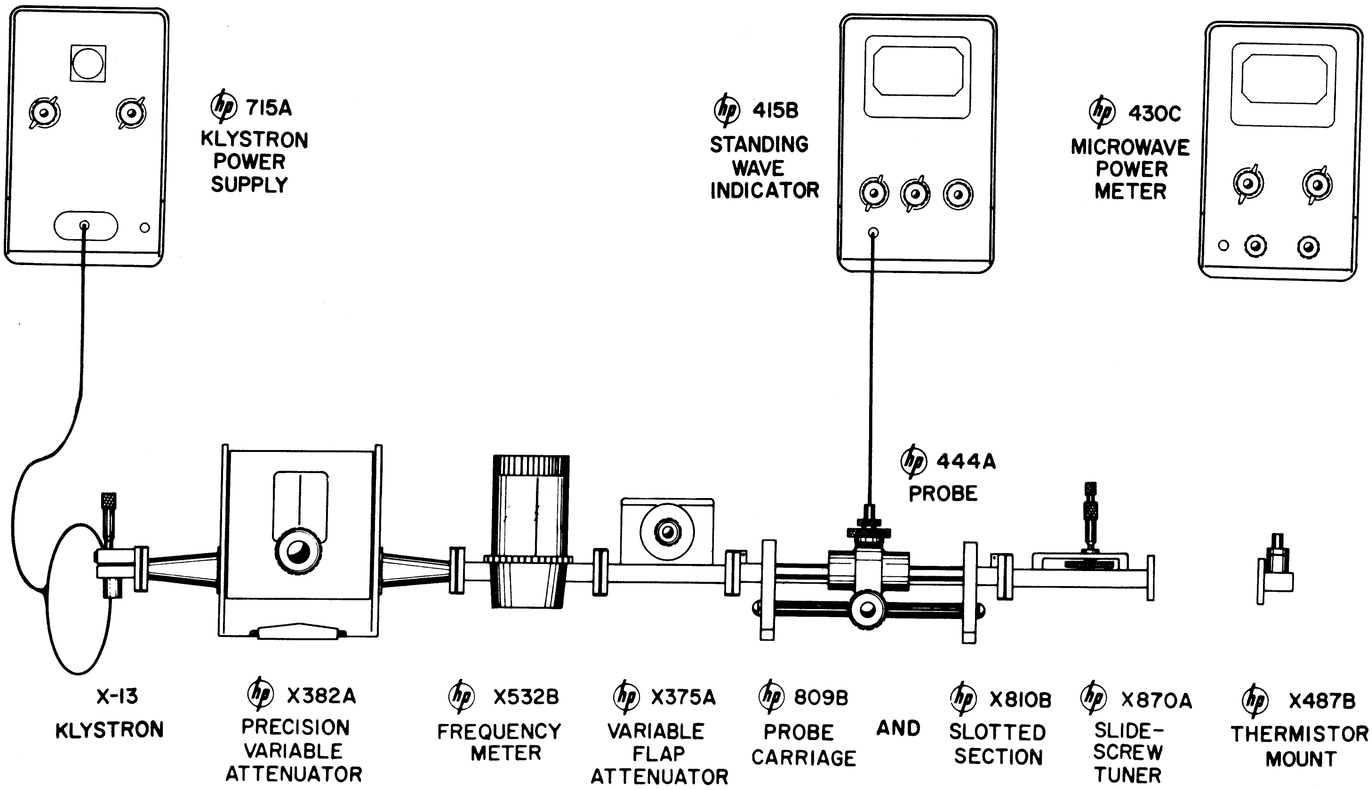


FIGURE 2

the mount by running the thermistor element at 200 ohms with the power meter. Do not provide a thermistor mismatch greater than 1.8.

2-2 Slowly increase the depth of the tuner probe until the indicated swr is 2 (caused by only the tuner probe, not the thermistor mount). Record the tuner-probe-micrometer depth setting. *NOTE:* The swr of the tuner may be conveniently set by leaving the broadband probe in the slotted section stationary, and moving the tuner probe back and forth to obtain the maximum and minimum indications on the standing-wave indicator as the tuner probe depth is increased.

2-3 Turn the string of equipment including the flap attenuator, the slotted section, and the slide-screw tuner end-for-end so that the tuner is attached to the frequency meter, and the flap attenuator is attached to the thermistor mount.

2-4 Remove the broadband probe from the slotted section, and insert the slide-screw tuner probe to the depth recorded in Step 2-2. From the standpoint of the slotted section, the source swr is now 2, and the load swr is that recorded in Step 2-3.

NOTE: Rather than use the technique just described, a second rf power source of the same frequency could have been used in measuring the swr of the source used in this experiment.

Using the load and source mismatches measured previously and the chart in Fig. 1, record the maximum and minimum possible power losses due to mismatch.

2-5 Move the tuner back and forth along the line and record the maximum and minimum power transferred to the thermistor mount. This step simulates various phase conditions of the re-reflections from the mismatched source.

2-6 Remove the tuner probe and record the measured power output. Note that the power measured is not the maximum power available to a matched load, because the thermistor has a mismatch measured in Step 2-1.

2-7 With the load mismatch (swr) of Step 2-1 and a source mismatch of 1, use Fig. 1 to find the mismatch loss. Record this mismatch loss.

2-8 Add the mismatch power of Step 2-7 to the measured power of Step 2-6, and record this maximum available power in Table I.

2-9 From the maximum available power of Step 2-8, subtract the calculated maximum and minimum losses recorded in Step 2-4. Record the answers in Table I.

2-10 Adjust the variable flap attenuator to read 6 db, and measure the swr now seen from the source end of the system looking toward the load.

2-11 Repeat the procedure of Steps 2-1 through 2-10 for the remaining frequencies in Table I.

Name ______________________

Course ______________________

Date performed ______________________

Date turned in ______________________

Results (Experiment 13)

OBSERVED	CALCULATED
Step	*Step*
2-1 Table I	
2-2 Table I	
2-4 Table I	
2-6 Table I	2-7 Table I
	2-8 Table I
	2-9 Table I
2-10 Table I	
2-11 Table I	2-11 Table I

Questions

1. How does the swr measured in Step 2-10 generally compare with the swr measured in Step 2-1?

2. Why is having a matched source and a matched load a requirement for accurate attenuation measurements?

3. What happens to the power that is not absorbed by the mismatched load?

4. Would you expect many or few power variations for a given frequency band if a mismatched source were connected to a mismatched load through a long lossless line? Why?

Discussion

TABLE I

STEP ⟶	2-1	2-2	2-4		2-5		2-6	2-7	2-8	2-9		2-10
Frequency (gc)	Thermistor (load) swr	Tuner probe insertion micrometer reading	Chart maximum loss (db)	Chart minimum loss (db)	Maximum power transfer (dbm)	Minimum power transfer (dbm)	Measured power output (dbm)	Load mismatch loss	Maximum power available (dbm)	Actual maximum power (dbm)	Actual minimum power (dbm)	Load swr (with 6 db attenuation)
9.0												
9.5												
10.0												
10.5												
11.0												

EXPERIMENT 14

Directional Couplers

Object

To determine the characteristics of a directional coupler, and to use a coupler in the measurement of swr.

Theory

Directional couplers are important tools in waveguide measurements. They may be used to monitor power, to measure reflections, to mix signals, or to isolate signal sources or frequency meters.

The directional coupler used in this experiment has two major parts, the main guide and the auxiliary guide (or arm). Ideally, power flowing in one (the forward) direction of the main guide is coupled to the output of the auxiliary guide, whereas power flowing in the other (reverse) direction is not coupled to the output of the auxiliary guide. The ratio, expressed in db, of forward power at the input of the main guide to the power out of the auxiliary guide is the "coupling factor." For example, a 20 db coupling factor means a ratio of power of 100:1.

In practice, some reverse power in the main guide is coupled to the output of the auxiliary guide and the ratio, also in db, of the powers out of the auxiliary guide from equal forward and reverse powers in the main guide is the coupler's "directivity."

Power dividers (couplers) exhibiting directional properties can be constructed in several ways. In this discussion only high-directivity (at least 40 db), broad-bandwidth (50 per cent), waveguide directional couplers will be considered. The hp Model X752C is an example of this type of coupler.

The ⓗⓟ Model X752C is an interference-type coupler. That is, directivity is achieved by producing, in the auxiliary arm of the coupler, two or more signals of such phase and magnitude that the signals travelling in one direction add, whereas the signals travelling in the opposite direction cancel each other. The simplest form of interference coupler consists of two holes spaced $\frac{1}{4}$ wavelength apart and connecting two parallel sections of waveguide (Fig. 1). As shown in the figure, the signal in the main arm propagates through both holes. At each hole, the signal splits into two components, one travelling in the forward direction (to OUTPUT 2) and one travelling in the reverse direction (to "DEAD-END"). The

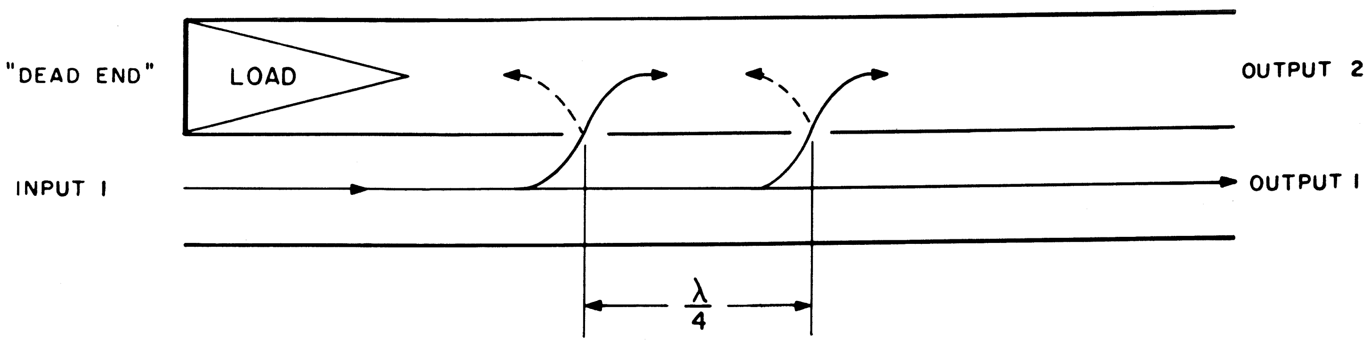


FIGURE 1

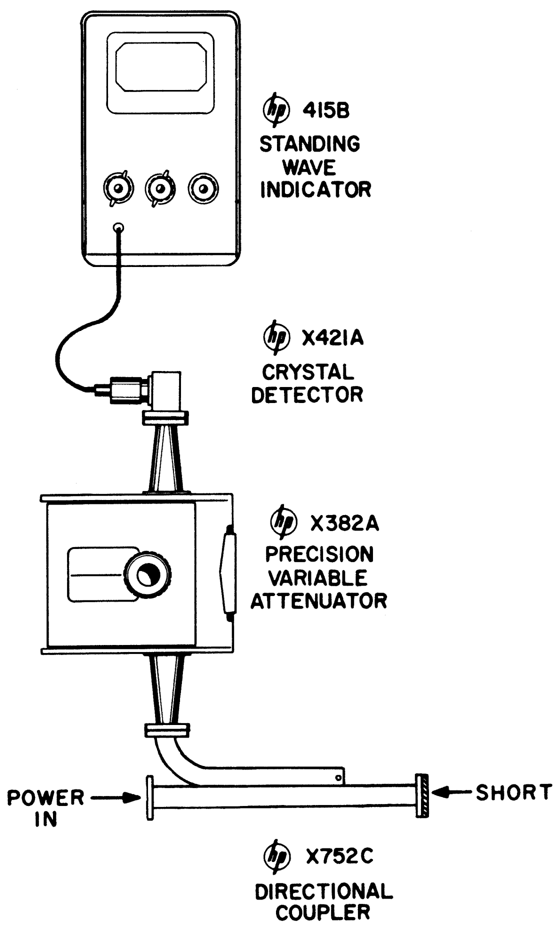


FIGURE 2

two wave components travelling from INPUT 1 to OUTPUT 2 move the same electrical distance and are in phase; hence, they add together. At the same time, the two wave components travelling from INPUT 1 toward "DEAD-END" move over paths that are electrically a half wavelength (180 deg) different; consequently, they are exactly out of phase and they cancel each other. This reasoning applies only when the coupled power is small compared with the INPUT 1 power (weak coupling), for only then are the two coupled signals of equal magnitude.

The two-hole coupler provides good directivity over a narrow bandwidth, but at wavelengths other than the design wavelength the two reverse-coupled signals are not 180 deg out of phase, so they do not completely cancel. To increase the useful bandwidth, the theory of the two-hole device is extended to the design of a multihole array. An analysis of a large array is beyond the scope of this experiment; however, the design of the multihole array is similar in theory to that used for end-fire antennas. In both cases, the radiation from a number of elements is combined to provide an equal-ripple type of response.

Because of its ability to separate forward and reverse waves in a transmission system, the multihole coupler provides an accurate means of measuring reflection coefficient without using a slotted section. If such measurements are to be made while the frequency is swept, two couplers are required—one sampling forward power and the other sampling reverse power. At a fixed frequency (as in this experiment), only one coupler and one detector are required. That coupler then samples the power reflected from the component under test (Fig. 2).

To calibrate the system, the short is assumed to reflect all the power ($\rho = 1$). Any lesser reflection can then be simulated by adding attenuation with the precision variable attenuator. The attenuation to be added is defined as return loss.

$$\text{return loss (in db)} = -20 \log_{10}\rho$$

and

$$\text{swr} = \frac{1+\rho}{1-\rho}$$

where ρ is the reflection coefficient of the simulated termination. When these formulas are used, the swr of any termination can be determined by comparison with a short and with a calibrated attenuator. In such measurements the directivity of the coupler determines the smallest reflection that can be accurately measured. The power coupled directly from the source to the detector produces an error signal, and when this signal approaches the magnitude of the reflected signal, the readings become ambiguous. For a coupler with 40 db directivity, this limit is equivalent to a $\rho = 0.01$ (swr $= 1.02$).

References. Additional information to supplement this experiment may be found in the following texts (listed in Appendix C).

King, Chapter 3: 3.12.
Reich, Chapter 6: 6.25.
Wind, Section 13.

Equipment

QUANTITY	TYPE
1	hp 715A klystron power supply (with cable)
1	Varian X-13 reflex klystron
1	Cooling fan or blower
1	hp X375A variable flap attenuator
1	hp X382A precision variable attenuator
1	hp 415B standing-wave indicator
1	hp X421A crystal detector
1	hp X532B frequency meter
1	hp X752C waveguide directional coupler (10 db coupling, at least 40 db directivity)
1	hp X914B moving load
1	hp 120B oscilloscope
1	hp AC-16A cable (dual banana to dual banana)
1	hp AC-16B cable (dual banana to BNC)
1	hp AC-16K cable (BNC to BNC)
2	hp Model 24 waveguide stand with Model X25 waveguide clamp

Procedure

Section 1—General

1-1 Review the safety precautions on page 42, and the operating instructions for each equipment item.

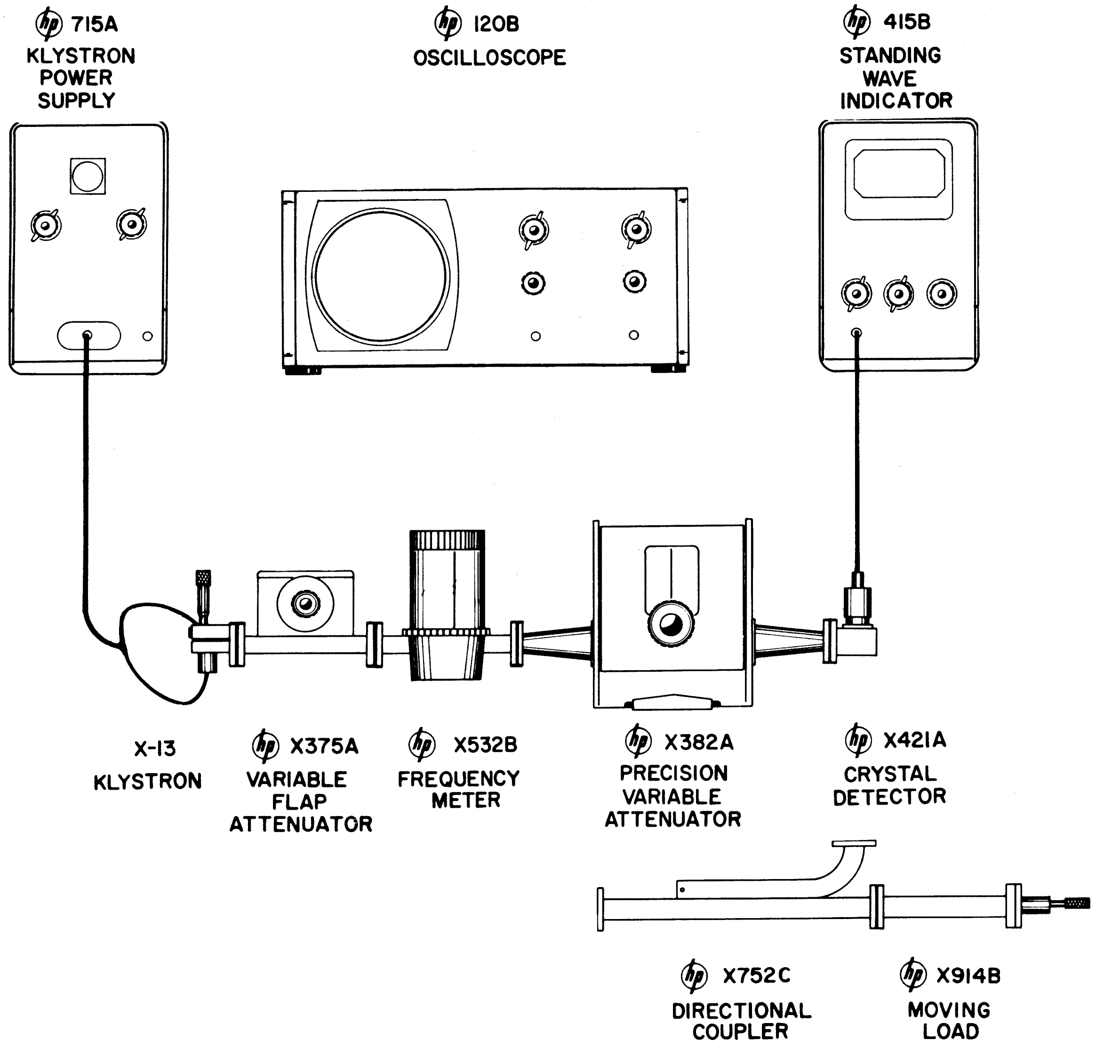


FIGURE 3

1-2 Set up the equipment as shown in Fig. 3. The variable flap attenuator should be set to approximately 10 db, and the precision variable attenuator should be set to 0 db.

1-3 Energize the klystron at 10.0 gc, and modulate it with a 1000 cps square wave. Use the oscilloscope technique of Experiment 2 to set the reflector and modulating voltages properly. When the crystal detector is reconnected to the standing-wave indicator, adjust the 1000 cps modulation frequency to get a peak indication on the indicator.

Section 2—Coupling Measurement

2-1 Coupling can be measured in a manner similar to the rf substitution method of measuring attenuation in Experiment 4. Adjust the precision variable attenuator for a reading of 20 db, and note the reference reading on the standing-wave indicator.

2-2 Attach the direction coupler and moving load to the frequency meter, and move the precision variable attenuator and the crystal detector to the auxiliary arm of the coupler.

2-3 Adjust the precision variable attenuator for the same standing-wave-indicator reference reading noted in Step 2-1. Record this new attenuator setting. Calculate and record the coupling of the directional coupler, which is equal to the change in attenuator dial settings.

2-4 Repeat the procedure of Steps 2-1 through 2-3 for klystron frequencies of 8.2 gc and 12.4 gc.

Section 3—Directivity Measurement

3-1 Remove the moving load and reverse the coupler in the system; leave the precision attenuator in the auxiliary arm and set it to 50 db. Using a brass plate or a similar object, short the exposed end of the coupler (Fig. 2). Because the short produces total reflection, the reading on the standing-wave indicator is now equivalent to a reading that would be obtained by feeding power from the shorted end (forward coupling). Note the reference reading on the standing-wave indicator.

3-2 Replace the short with the moving load. (The load body must be carefully aligned with the coupler to reduce reflections at the flange, for even small reflections will produce erroneous signals at the detector.) Slide the load element until a peak is indicated on the standing-wave indicator. Adjust the precision variable attenuator to obtain the original reference reading of Step 3-1, and record this new attenuator setting in Table I.

3-3 Slide the load element until a minimum is indicated on the standing-wave indicator. Readjust the precision variable attenuator for the original reference reading of Step 3-1, and record this attenuator setting in Table I.

3-4 Calculate and record the directivity on Table I.

NOTE: If the two attenuator readings of Steps 3-2 and 3-3 are only a few db apart, the directivity can be taken as the average change in attenuation between the shorted and matched conditions. If the two readings vary widely, the load is poor, and a correction should be applied to the average value. Because of the complexity of this correction factor, it will be neglected in this experiment, and directivity will be defined as average change of attenuation.

3-5 Repeat the procedure of Steps 3-1 through 3-3 for the remaining klystron frequencies listed in Table I.

Section 4—The Reflectometer

4-1 The equipment arrangement used in Section 3 can be used as a reflectometer. Replace the moving load with the test device (a thermistor mount that is connected to and biased by a power meter).

4-2 Set the precision variable attenuator to 0 db, and note the reference reading on the standing-wave indicator.

4-3 Remove the test device (thermistor mount) and short the exposed end of the directional coupler. Reset the attenuator for the reference reading of Step 4-2, and record this reading on Table II. Calculate the change in swr from the attenuator reading (return loss). *Hint:* Use lower scales of a Smith Chart.

4-4 Repeat the procedure of Steps 4-1 through 4-3 for the remaining klystron frequencies listed.

Name ______________________
Course ______________________
Date performed ______________________
Date turned in ______________________

Results (Experiment 14)

OBSERVED		CALCULATED	
Step		*Step*	
2-3	______________db	2-3	______________db
2-4	______________db (8.2 gc)	2-4	______________db
	______________db (12.4 gc)		______________db
3-2	Table I		
3-3	Table I	3-4	Table I
3-5	Table I	3-5	Table I
4-3	Table II	4-3	Table II
4-4	Table II	4-4	Table II

Questions

1. If a mismatch of 1.02 exists at the directional-coupler-moving load junction of Section 3, calculate the error in the directivity measurement.

2. If your swr calculation of Step 4-3 was 1.5 and the accuracy of the precision attenuator is 2 per cent of the reading in db, what are the possible limits of error in your 1.5 db calculation?

3. What ambiguity is introduced into the reflectometer measurement by the measured directivity?

4. Why is a tapered load placed inside one end of the auxiliary line?

Discussion

TABLE I

STEP ⟶		3-2	3-4	3-3	3-4	
Frequency (gc)	Original attenuator setting (db)	Attenuator setting, load at max. (db)	Attenuation change at max. (db)	Attenuator setting, load at min. (db)	Attenuation change at min. (db)	Directivity (db)
8.2	50					
10.0	50					
12.4	50					

TABLE II

Frequency (gc)	Attenuator reading	swr
8.2		
10.0		
12.4		

EXPERIMENT 15

Microwave Transmission in Air

Object

To study some basic effects of microwave transmission in an air medium.

Theory

Communications, radar, and radio astronomy are just a few of the applications of electromagnetic propagation of energy. Most of these applications either require that energy be propagated from a point with its echoes being received at the same point, or that energy be propagated from one point and received at another. In one case, the common principle of radar transmits a powerful pulse from a large antenna and, after a suitable waiting period, receives the tiny echo from the distant target; in another example, telephone messages are relayed from one point to the next across intervening distances of 50 miles or more. All such applications demand that microwave power be efficiently launched into the air, and then be received properly without loss of the signal power and without interference from other signals.

From the early days of radio, antenna development has progressed concurrently with the other advances of the art. Antenna ranges spot the hillsides around most major electronic centers, demonstrating the importance of this design art.

The primary purpose of an antenna is to provide directivity, i.e., to focus the available power from the power source in such a way that its primary radiation and power is directed toward the desired receiving point. The better the directivity, the better the antenna. The normal figure of merit used to express the quality of an antenna is the term "antenna gain." Antenna gain is defined as the ratio of the actual power arriving at a certain distant point to the power received from a simple lossless isotropic radiator (one which is radiating equally in all directions).

NOTE: Some precedence is established for referring the gain to a Hertzian dipole or a half-wave doublet, which give gains relative to an isotropic radiator of 3/2 and 1.64, respectively.

A common characteristic for most antennas is the antenna-pattern plot, which describes the basic geometry of the field-strength pattern of the antenna. This pattern is a good indication of the characteristic directivity of the antenna; it also shows the side and back lobes of the pattern. These side and back lobes actually become segments of radiation in the case of a transmitter, or areas of reception in the case of a receiver. The back lobes of the radio astronomy antenna, for instance, then become very detrimental, since part of the received signal is coming from sources on the ground behind the antenna, in addition to the sky signal. These ground sources make an undesirable contribution to the over-all background noise.

For the simple case of a combination of transmitting and receiving antennas aligned for maximum transmission along a line joining them, the following relation exists:

$$\frac{P_r}{P_t} = \frac{G_t G_r \lambda^2}{(4\pi R)^2}$$

where P_r = received power.
P_t = transmitted power.
G_t = transmitter antenna gain.
G_r = receiver antenna gain.
λ = wavelength in cm.
R = spacing of antennas in cm.

If the ratio of received power to transmitted power can be measured and the spacing and wavelength is known, it is possible to determine gain by using two identical antennas. Under these conditions it can be assumed (as an approximation) that the respective antenna gains are equal.

References. Additional information to supplement this experiment may be found in the following texts (listed in Appendix C).

King, Chapter 7.
Reich, Chapter 7.
Wind, Section 20.

Equipment

QUANTITY	TYPE
1	(hp) 715A klystron power supply (with cable)
1	Varian X-13 reflex klystron
1	Cooling fan or blower
1	(hp) X375A variable flap attenuator
1	(hp) X382A precision variable attenuator
1	(hp) X421A crystal detector
1	(hp) 415B standing-wave indicator
1	(hp) 444 broadband probe
1	(hp) X532B frequency meter
1	(hp) 809B probe carriage
1	(hp) X870A slide-screw tuner
1	(hp) X810B slotted section
1	(hp) 120B oscilloscope
1	(hp) AC-16A cable (dual banana to dual banana)
1	(hp) AC-16B cable (dual banana to BNC)
1	(hp) AC-16K cable (BNC to BNC)
2	(hp) Model 24 waveguide stand with Model X25 waveguide clamp

Procedure

Section 1—General

1-1 Review the safety precautions on page 42, *with particular attention to avoiding eye damage.* Also review the operating instructions for each equipment item.

1-2 Set up the equipment as shown in Fig. 1. The precision variable attenuator should be set to approximately 15 db, and the variable flap attenuator should be set to 0 db.

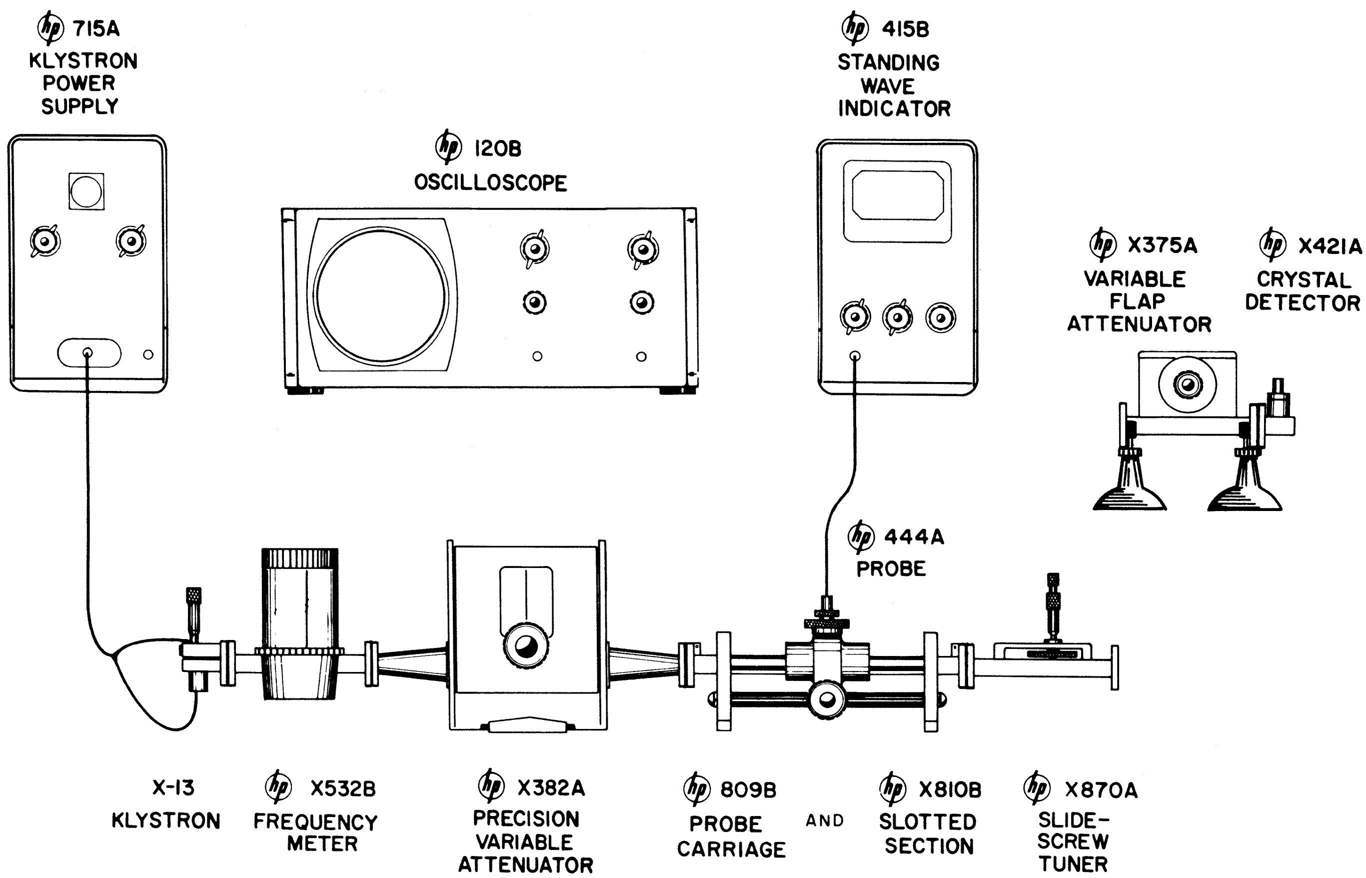


FIGURE 1

NOTE: In this experiment, microwave energy will be launched into an area where the energy will be affected by small external influences. Typical influences to be avoided are steel beams (often used in building construction) and steel tables. It is recommended that when the system is launching energy into air, the launching flange (exposed end of the slide-screw tuner) be placed over a wooden-top table. If only steel-top tables (or tables with steel understructure) are available, arrange the system so that the pattern is launched into the air off one table toward another adjacent table supporting the receiver.

1-3 Energize the klystron at 10.0 gc and modulate it with a 1000 cps square wave. Use the oscilloscope technique of Experiment 2 to set the reflector and modulating voltages properly. When the standing-wave indicator has been reconnected to the broadband probe, adjust the 1000 cps modulation frequency to get a peak indication on the standing-wave indicator.

Section 2—Launching the Wave

2-1 With the slide-screw-tuner probe completely out of the waveguide, measure and record the swr of the open-ended flange.

2-2 Using this swr and the scales on the bottom of a Smith Chart, calculate the loss of power (power not being launched) because of mismatch at the waveguide-to-air transition.

NOTE: Tapered "horns" are available that provide better transitions between waveguide and air. The plain flange is used in this experiment for convenience and for the sake of limiting the amount of equipment needed.

2-3 Adjust the slide-screw-tuner-probe penetration and horizontal position to "flatten" the line for match. Try to reduce the swr to less than 1.1.

2-4 With the swr less than 1.1, how much mismatch loss is being experienced in launching power into the air?

2-5 Attach the variable-flap-attenuator-and-crystal-detector combination to the exposed flange of the slide-screw tuner. Connect the crystal detector to the standing-wave indicator, using the 200 ohm or 200K ohm indicator input, whichever gives the maximum reading for the particular crystal.

NOTE: This attenuator-crystal-detector-indicator combination will now be called the *receiver.*

2-6 Adjust the precision variable attenuator and the gain on the standing-wave indicator until the reference reading on the indicator is 0 db on the 60 db range. If the precision attenuator goes beyond 50 db, turn in some attenuation with the flap attenuator. Record the attenuator setting.

2-7 Withdraw the receiver from the transmitter (the rest of the system) in 25 cm increments along the axial line of the system. At each incremental stop, adjust the precision variable attenuator to bring the standing-wave indicator back to its original reference reading of Step 2-6. (Do not change the gain-control setting on the standing-wave indicator.) Record each attenuator setting.

2-8 Using the initial precision variable attenuator setting as a 0 db reference level, calculate and record the relative decrease in signal level for increasing flange separation.

2-9 Using Fig. 4 (see page **208**), plot the power characteristic vs. flange distance separation.

Section 3—Antenna Pattern

3-1 Without changing any settings on the transmitter, place the receiver 100 cm from the transmitter on the axial line. Remove Fig. 2 (page 205) from the book and place it under the front waveguide stand supporting the receiver. The waveguide clamp on this stand should be as close as possible to the open flange on the flap attenuator, so that the open flange is near the center of the circle. Remove beam voltage from the klystron, and align the transmitter and receiver axially by eye. This is the 0 deg reference line. Re-energize the klystron.

3-2 Reduce the precision-variable-attenuator setting until a 0 db reference can be obtained on the standing-wave indicator's 40 db range. It may be necessary to slide the receiver closer to the transmitter to get enough received power for the sensitivity desired. *NOTE:* While setting the reference, and while making subsequent readings, step away from the receiver to eliminate interference with the receiver-antenna pattern. You may even wish to locate the standing-wave indicator at a remote point, and have an assistant take readings.

3-3 When the reference power level has been set, tape Fig. 2 to the table top and pencil an axial index mark on the front waveguide stand base (above the 0 deg line on Fig. 2).

3-4 Turn the receiver combination so that it presents a 10 deg angle to the line of transmission. Record the standing-wave-indicator reading on Fig. 3 (see page 207), as db down from the 0 db reference (40 db scale).

3-5 Repeat Step 3-4 for each 10 deg of rotation up to 180 deg. Plot the antenna pattern as a function of db on Fig. 3. During the plot do not change transmission power, but simply switch down on standing-wave-indicator ranges as the received power drops because of angular position.

Section 4—Transmission in Air

4-1 Bring the transmitting and receiving flanges back into alignment and space them about 1 meter apart to obtain a reference power-level reading on the standing-wave indicator.

4-2 Place a lossy material, such as a book about 1 in. thick, in the transmission path. The dielectric loss will show up as an apparent decrease in transmitted power. (This loss is partially due to dissipation and partially due to reflection.)

4-3 Try different orientations of the lossy material to study the reflection effects.

4-4 To observe the effect of multiple-path transmissions, move your hand slowly back and forth toward the center line connecting the transmitter and the receiver. Note that maximums and minimums of power transfer are being received at the receiver. The phasing is due to the constructive and destructive additions of power at the receiver, caused by the two signals arriving at the receiver. One signal is the direct-path transmission, whereas the second signal is that being reflected from the operator's hand. As the longer path has a slightly different phase, it adds and subtracts from the main power transmission, and hence causes the variation in the coupled power. This effect is another reason for having an antenna with high directivity, since multiple-path transmission is quite serious in many cases. For instance, the arrival time of coded transmission may jitter intolerably if there are variations in path lengths.

Section 5—Polarization

5-1 Propagation of energy in dominant-mode waveguide, such as that employed in the transmitter, consists entirely of field configurations with vertical E-fields. In general, the launched wave behaves with the same polarization.

5-2 While the transmitter and receiver are separated by 1 meter, set an arbitrary reference on the standing-wave indicator.

5-3 To demonstrate the effect of polarization, rotate the flap-attenuator-crystal-detector combination by 90 deg around its axial center line.

5-4 Record the change of signal level caused by the 90 deg rotation.

Section 6—Additional Antenna Patterns

6-1 If time permits, repeat the procedure of Section 3 for klystron frequencies of 8.2 gc and 12.4 gc.

Section 7—Calculations

7-1 Using the relative signal level recorded in Step 2-6 for the 100 cm flange separation, calculate both transmitter and receiver antenna gain, using the equation introduced under "Theory" in this experiment. If you had to use spacing closer than 100 cm, use that spacing as R. Assume both gains to be equal, and convert gain to db. At 10 gc, λ equals 3 cm. (Be sure to convert P_r/P_t from db.)

NOTE: The actual gain depends greatly upon conditions of measurement, and reflections from people, desks, etc. will alter results. The typical gain is approximately 6–10 db.

7-2 To check the calculations just made, it is possible to compute antenna gain from the antenna-pattern plot of Section 3. This computation will be made by using approximate integration of the intensity sector product, and comparing this to an equal intensity illumination plot of the surface. Figure 3 will be used.

To integrate, consider each of the eighteen 10 degree segments. Estimate the average intensity from the numerical power scale of 1000 for each 10 degree segment. Add all eighteen average intensities, and divide by 18 for an over-all average relative intensity. Compute the ratio between this average relative intensity and the 1000 figure (the forward-intensity reference). Convert this ratio to db. This antenna gain should compare within 1 or 2 db with the calculated gain in Section 1.

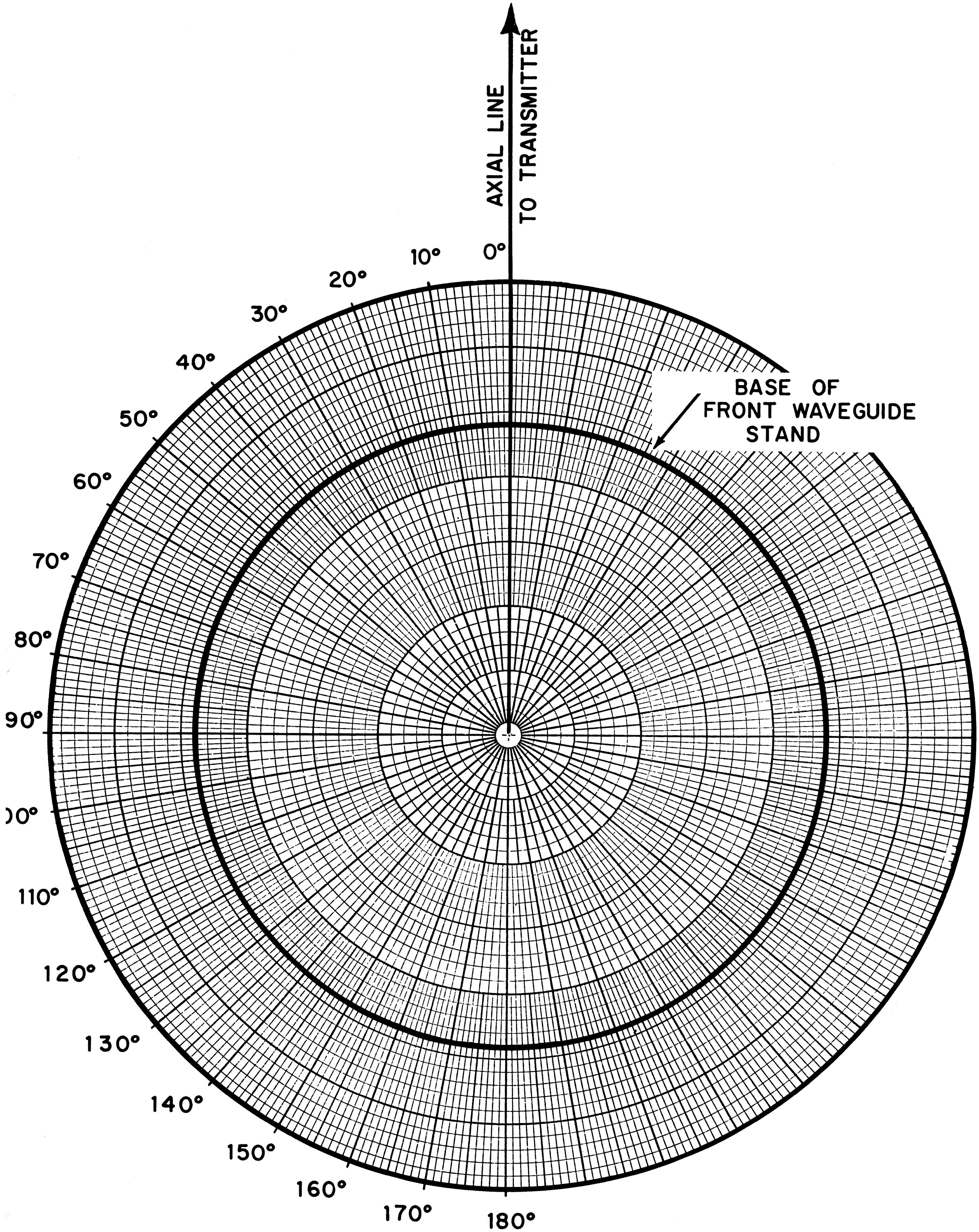

AXIAL LINE
TO TRANSMITTER
0°
10°
20°
30°
40°
50°
60°
70°
80°
90°
100°
110°
120°
130°
140°
150°
160°
170°
180°
BASE OF
FRONT WAVEGUIDE
STAND

FIGURE 2

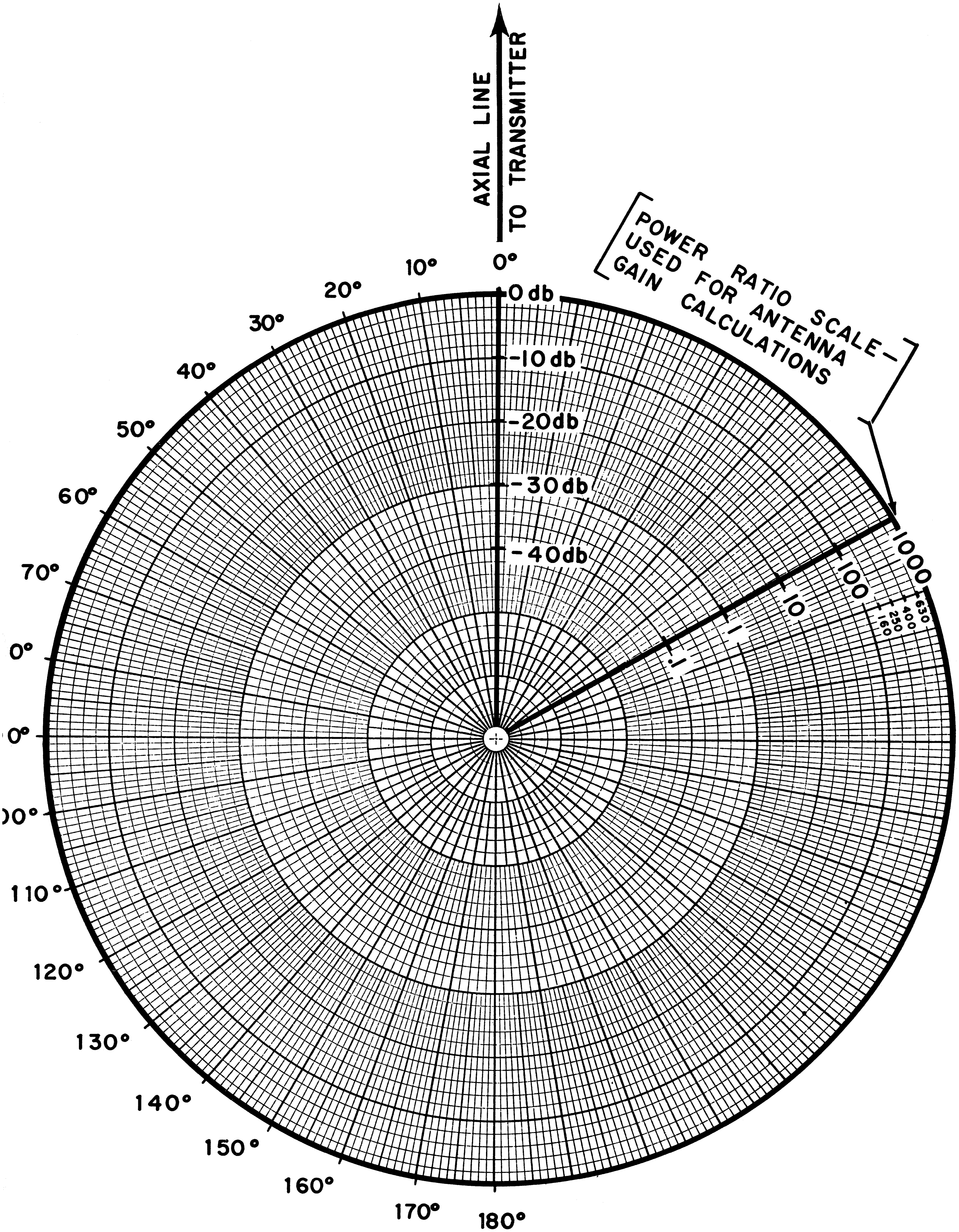

AXIAL LINE
TO TRANSMITTER
POWER RATIO SCALE—
USED FOR ANTENNA
GAIN CALCULATIONS
0°
10°
20°
30°
40°
50°
60°
70°
110°
120°
130°
140°
150°
160°
170°
180°
0 db
-10 db
-20db
-30db
-40db
1000
100
10
1
.1
630
400
250
160

FIGURE 3

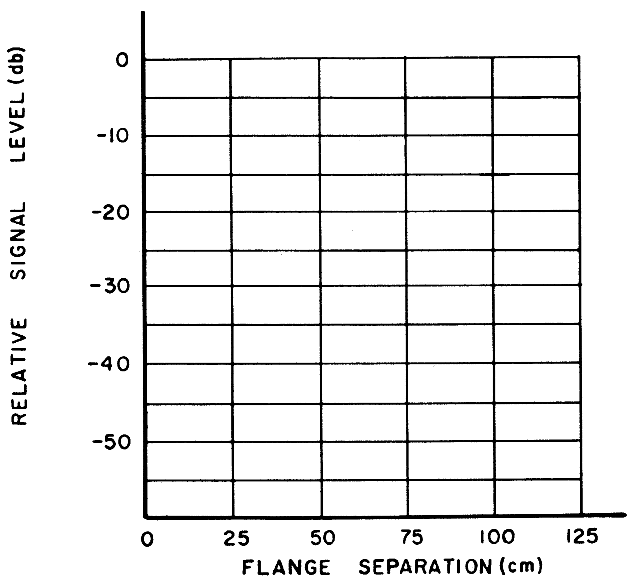

RELATIVE SIGNAL LEVEL (db)
0
-10
-20
-30
-40
-50
0
25
50
75
100
125
FLANGE SEPARATION (cm)

FIGURE 4

Name ______________________

Course ______________________

Date performed ______________________

Date turned in ______________________

Results (Experiment 15)

OBSERVED		CALCULATED	
Step		*Step*	
2-1	____________	2-2	____________
		2-4	____________
2-6	Table I		
2-7	Table I	2-8	Table I
		2-9	Figure 4
3-4	Figure 3		
3-5	Figure 3		
5-4	____________db		
		7-1	____________
		7-2	____________

Questions

1. How does the gain of your "antenna" (Fig. 4) compare with commercial "standard-gain" horns of 15 db, or with radar antennas of 40 and 50 db?

2. If you wanted to utilize the phasing effects noted in Step 4-4 in making a personnel (burglar) detector by detecting the phasing beats with an audio amplifier, would you choose a narrow-beam or broad-beam antenna? Why?

3. Suggest a way to make a traffic-type speed radar using the Doppler principle of reflection noted in Step 4-4. Would the frequency of the audio phasing be related to traffic speed?

4. If the angle of rotation in Section 5 had been 45 deg, how much decrease in signal level might you expect? *Hint:* The E-field should have a value decreased by cos θ.

   Prove your answer experimentally if time permits.

Discussion

TABLE I

STEP ⟶	2-6 and 2-7	2-8
Transmitter-receiver flange separation (cm)	Precision attenuator setting (db)	Relative signal level (db)
0		
25		
50		
75		
100		
125		

Appendices

APPENDIX A — Glossary of Microwave Terms

Attenuation. Decrease in magnitude of current, voltage, or power of a signal in transmission between points.

Attenuation constant. For a travelling plane wave of a given frequency, the rate of exponential decrease of the amplitude of a field component (or of the voltage or current) in the direction of propagation. Expressed in Nepers or db per unit length.

Attenuator, flap. A device designed to introduce attenuation into a waveguide circuit by means of a resistive material moved into the guide.

Attenuator, rotary vane. A device designed to introduce attenuation into a waveguide circuit by means of varying the angular position of a resistive material in the guide.

Backward-wave tube. A travelling wave tube in which the electrons travel in a direction opposite to that in which the wave is propagated. A microwave oscillator.

Barretter. A metallic resistor with a positive temperature coefficient of resistivity. Used for detection and power-level measurements.

Bend, E-plane. A bend in a waveguide in the plane of the electric field. ("Easy" bend.)

Bend, H-plane. A bend in a waveguide in the plane of the magnetic field. ("Hard" bend.)

Bolometer. A barretter, a thermistor, or any other device utilizing the temperature coefficient of resistivity of some resistance element.

Choke joint. A type of joint for connecting two sections of waveguide. It is so arranged that there is efficient energy transfer without the necessity of an electrical contact at the insides of the guide.

Coaxial line. A transmission line in which one conductor completely surrounds the other, the two being coaxial and separated by a continuous solid dielectric or by dielectric spacers. Such a line is characterized by no external field and by having no susceptibility to external fields from other sources.

Coupler, directional. A device consisting of two transmission lines coupled together in such a way that a wave travelling in one line in one direction excites a wave in the other guide; ideally, in one direction only.

Coupler, forward. A directional coupler used to sample incident power.

Coupler, reverse. A directional coupler used to sample reflected power.

Coupling coefficient. A ratio between the power entering the main arm of a directional coupler in one direction to the power coupled into the auxiliary arm in the same direction.

Cutoff frequency. The lowest frequency at which lossless waveguide will propagate energy in some particular mode without attenuation.

Cutoff wavelength. The longest wavelength at which lossless waveguide will propagate energy in some particular mode without attenuation.

Demodulator. A device whose output voltage is proportional to the square of its input voltage (i.e., input power).

Detector. An element which reproduces the modulation of an rf wave, usually a semiconductor crystal. Barretters are sometimes used to detect low-frequency modulation.

Directivity. The ratio of (1) power flowing out of the auxiliary arm of a directional coupler when power is flowing in the forward direction in the main arm, to (2) power flowing out of the auxiliary arm of the coupler when power is flowing in the reverse direction in the main arm (both forward and reverse powers in the main arm being equal in magnitude).

Directivity signal. A spurious signal present in the output of a coupler because the directivity of the coupler is not infinite.

Efficiency, bolometer mount. The percentage of net applied power that is absorbed by the rf termination.

EHF. Extremely high frequency. The band of frequencies between 30,000 mc (30 gc) and 300,000 mc (300 gc).

E-H tee. A junction composed of a combination of E and H plane tee junctions having a common point of intersection with the main guide.

E-H tuner. An E-H tee used for impedance transformation, having two arms terminated in adjustable plungers.

Gigacycle. 10^9 cycles (formerly kilomegacycle). Common term for expressing microwave frequencies.

Guide wavelength. The length of waveguide corresponding to one cycle of variation in the axial (transmitted) direction.

Impedance, characteristic (of a rectangular waveguide). For the dominant TE_{10} mode of a lossless rectangular waveguide at a frequency above the cutoff frequency, the ratio of the square of the rms voltage between midpoints of the two conductor faces normal to the electric vector, and the total power flowing when the guide is match-terminated.

Impedance, characteristic (of a two-conductor transmission line). For a travelling, transverse electromagnetic wave, the ratio of the complex voltage between the conductors to the complex current on the conductors.

Impedance, normalized. Any impedance of a system divided by the characteristic impedance of that system.

Incident power or signal. Power flowing from the generator to the load.

Iris. In a waveguide, a conducting plate or plates, of small thickness compared to a wavelength, occupying a part of the cross section of the waveguide. When only a single mode can be supported, an iris acts substantially as a shunt admittance.

Isolator ferrite. A microwave device which allows rf energy to pass through in one direction with very little loss while rf power in the reverse direction is absorbed.

Junction, hybrid. A waveguide arrangement with four branches which, when branches are properly terminated, has the property that energy can be transferred from any one branch into only two of the remaining three. In common usage this energy is equally divided between the two branches.

Magnetron. A high-power microwave oscillator tube with a fixed or limited frequency range. Frequency, efficiency, and power depend on magnetic field strength and anode voltage.

MASER (Microwave Amplification by Stimulated Emission of Radiation). A low-noise, microwave amplifier utilizing a change in energy level of a material to obtain signal amplification. Common materials are gases (ammonia) and crystals (ruby).

Matched termination (waveguide). A termination producing no reflected wave at any transverse section of the waveguide.

Microstrip. A microwave-transmission component utilizing a single conductor supported above a ground plane.

Microwave region. That portion of the electromagnetic spectrum lying between the far infrared and conventional rf portion. Commonly regarded as extending from 1000 megacycles (30 cm) to 300,000 megacycles (1 mm).

Millimeter waves. The band of frequencies having wavelengths shorter than 1 cm (above 30,000 mc).

Mismatch loss (reflection loss). The ratio, expressed in db, of the incident power to the transmitted power at a discontinuity. A measure of the loss caused by reflection.

Mode (of transmission propagation). A form of propagation of guided waves that is characterized by a particular field pattern in a plane transverse to the direction of propagation. The field pattern is independent of the position along the axis of the waveguide and, for uniconductor waveguide, independent of frequency.

Noise figure. A figure of merit for microwave amplifiers. A ratio in db between actual output noise power, and the output noise power which would come from a noiseless amplifier with identical gain and bandwidth.

Parametric amplifier (MAVAR—Mixer Amplification by Variable Reactance). A microwave amplifier utilizing the nonlinearity of a reactance element to obtain amplification. A low-noise amplifier.

Propagation constant. A transmission characteristic of a line which indicates the effect of the line on the wave being transmitted along the line. It is a complex quantity having a real term, the attenuation constant, and an imaginary term, the phase constant.

Rat race (hybrid ring). A hybrid junction which consists of a re-entrant line (waveguide) of proper electrical length to sustain standing waves, to which four side arms are connected. Commonly used as an equal power divider.

Reflected power or signal. Power flowing from the load back to the generator.

Reflection coefficient. A numerical ratio between the reflected voltage and the incident voltage.

Reflectometer. A microwave system arranged to measure the incident and reflected voltages and to indicate their ratio (swr).

Reflex klystron. A low-power microwave oscillator tube which depends primarily on the physical size of a cavity resonator for its frequency. Normally has a wider frequency range than a magnetron.

Reike diagram. A polar-coordinate load diagram for microwave oscillators, particularly klystrons and magnetrons.

Return loss. The ratio, expressed in db, between the power incident upon a discontinuity and the power reflected from the discontinuity. (The number of db reflected power is down from incident power.)

Rotator. In waveguides, a means of rotating the plane of polarization. In rectangular waveguide, rotation is accomplished simply by twisting the guide itself.

SHF. Super high frequency. The band of frequencies between 3000 and 30,000 mc.

Slotted section. A length of waveguide in the wall of which is cut a nonradiating slot used for standing-wave measurements.

Smith Diagram or Chart. A diagram with polar co-ordinates; developed to aid in the solution of transmission-line and waveguide problems.

Thermistor. A resistance element made of a semiconducting material which exhibits a high negative temperature coefficient of resistivity.

Travelling-wave tube. A broadband, microwave tube which depends for its characteristics upon the interaction between the field of a wave propagated along a waveguide and the beam of electrons travelling with the wave. A microwave amplifier.

Tuning screw (slide-screw tuner). A screw or probe inserted into the top or bottom of a waveguide (parallel to the E field) to develop susceptance, the magnitude and sign of which is controlled by the depth of penetration of the screw.

UHF. Ultra high frequency, the band of frequencies between 300 and 3000 mc.

VHF. Very high frequency, the band of frequencies between 30 and 300 mc.

Voltage standing-wave ratio (swr). The measured ratio of the field strength at a voltage maximum to that at an adjacent minimum.

Wave circuits, slow. A microwave circuit designed to have a phase velocity considerably below the speed of light. The general application for such waves is in travelling-wave tubes.

Wave, dominant. The guided wave having the lowest cutoff frequency. It is the only wave which will carry energy when the excitation is between the lowest cutoff frequency and the next higher frequency of a waveguide.

Waveguide phase shifter. A device for adjusting the phase of a particular field component at the output of the device relative to the phase of that field component at the input.

Waveguide tee. A junction used for the purpose of connecting a branch section of waveguide in series with or parallel with the main transmission line.

Waveguide tuner. An adjustable device added to a waveguide for the purpose of an impedance transformation.

Waveguide wavelength. For a travelling plane wave at a given frequency, the distance along the waveguide between points at which a field component (or the voltage or current) differs in phase by 2π rad.

Wave, phase velocity. The velocity with which a point of constant phase is propagated in a progressive sinusoidal wave.

Wave, group velocity. The velocity with which the envelope of a group of waves of neighboring frequencies travels in a medium; usually identified with the velocity of energy propogation.

Wave, transverse electric (TE wave). In a homogeneous isotropic medium, an electromagnetic wave in which the electric field vectors are everywhere perpendicular to the direction of propagation.

Wave, transverse electromagnetic (TEM wave). In a homogeneous isotropic medium, an electromagnetic wave in which both the electric and magnetic field vectors are everywhere perpendicular to the direction of propagation. Generally dominant mode of coaxial lines.

Wave, transverse magnetic (TM wave). In a homogeneous istropic medium, an electromagnetic wave in which the magnetic field vector is everywhere perpendicular to the direction of propagation.

Wave, TE_{mn} (in rectangular waveguide). In a hollow, rectangular, metal cylinder, the transverse electric wave for which $_m$ is the number of half-period variations of the electric field along the longer transverse dimension, and $_n$ is the number of half-period variations of the magnetic field along the shorter transverse dimensions.

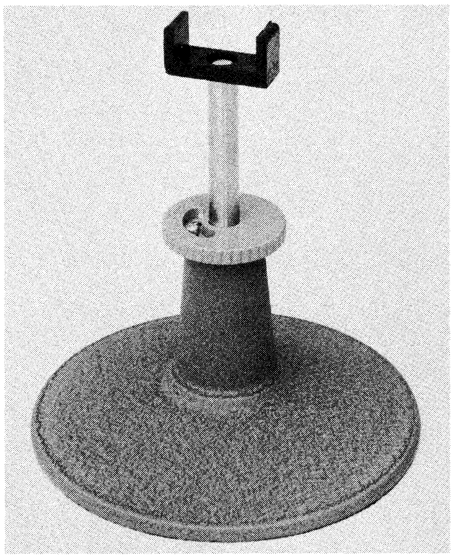
Wave, TM_{mn} (in rectangular waveguide). In a hollow, rectangular, metal cylinder, the transverse magnetic wave for which $_m$ is the number of half-period variations of the magnetic field along the longer transverse dimension, and $_n$ is the number of half-period variations of the magnetic field along the shorter transverse dimensions.

Wavemeter, absorption. A device which utilizes the characteristics of a resonator, which cause it to absorb maximum energy at its resonant frequency when loosely coupled to a source.

APPENDIX B

Microwave Equipment Data Sheets

WAVEGUIDE ACCESSORIES



hp MODEL 24 WAVEGUIDE STAND

Model 24 Waveguide Stands are cast and machined from zinc alloy. They are designed for hp Model 25 Waveguide Clamps and locks the clamp at any height from 2-3/4" to 5-1/4". Model 24 is 2-1/2" high and its base measures 4-3/4" in diameter. $3.00 each.

hp MODEL 25 WAVEGUIDE CLAMPS

These Clamps consists of a rubber molding with a steel insert. They are offered in 8 sizes to fit waveguide equipment covering frequencies from 2.6 to 40.0 kmc. They are designed for use with hp Model 24 Waveguide Stand, and when mounted in the Stand can be adjusted upward or downward to conform with a waveguide set-up. When ordering specify waveguide size.

Model	Waveguide Size (O. D.)	Price
hp Model S25	3" x 1-1/2"	$2.50 each
hp Model G25	2" x 1"	$2.50 each
hp Model J25	1-1/2" x 3/4"	$2.50 each
hp Model H25	1-1/4" x 5/8"	$2.50 each
hp Model X25	1" x 1/2"	$2.50 each
► hp Model M25	.850" x .475"	$2.50 each
hp Model P25	.702" x .391"	$2.50 each
hp Model K25	.500" x .250"	$2.50 each
hp Model R25	.360" x .220"	$2.50 each

Prices f.o.b. factory

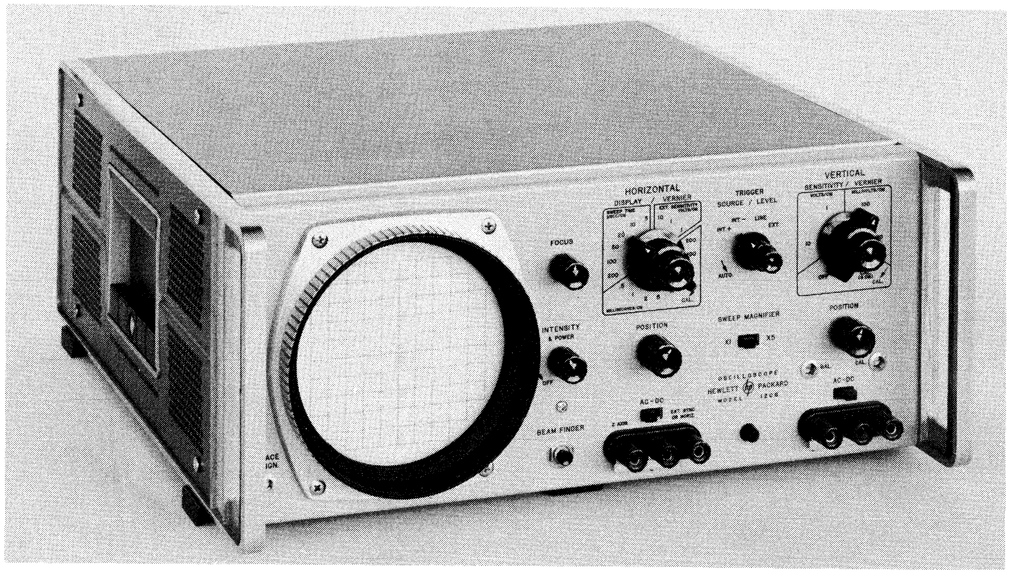
DATA SUBJECT TO CHANGE WITHOUT NOTICE

3/1/59
11/15/61

TENTATIVE DATA

HEWLETT-PACKARD COMPANY • 1501 PAGE MILL ROAD • PALO ALTO, CALIFORNIA, U. S. A.
CABLE "HEWPACK" TELEPHONE DAVENPORT 6-7000

ⓗⓟ MODEL 120B 450 KC GENERAL PURPOSE OSCILLOSCOPE



ADVANTAGES
- Parallax-free internal graticule increases accuracy
- CRT face eliminates glare, improves contrast
- New modular cabinet quickly convertible from bench to rack model
- Easy access to circuitry
- Beam finder, automatic triggering simplify operation
- Logically arranged controls simplify operation

USES
- General purpose for lab and production
- View complex waveforms
- Monitor transducer outputs

DESCRIPTION
ⓗⓟ Model 120B is an easy to use general purpose oscilloscope whose bandwidth extends from dc to 450 kc. It combines the precision characteristics of calibrated horizontal sweeps and calibrated vertical sensitivity with a new CRT which eliminates parallax. In addition, its construction provides a logical front panel layout, easy circuit accessibility and quick convertibility from a rack-mounting to a bench model configuration.

Parallax-Free Cathode Ray Tube
The internal graticule of ⓗⓟ Model 120B is the same plane as the phosphor and the trace (see figure 1).

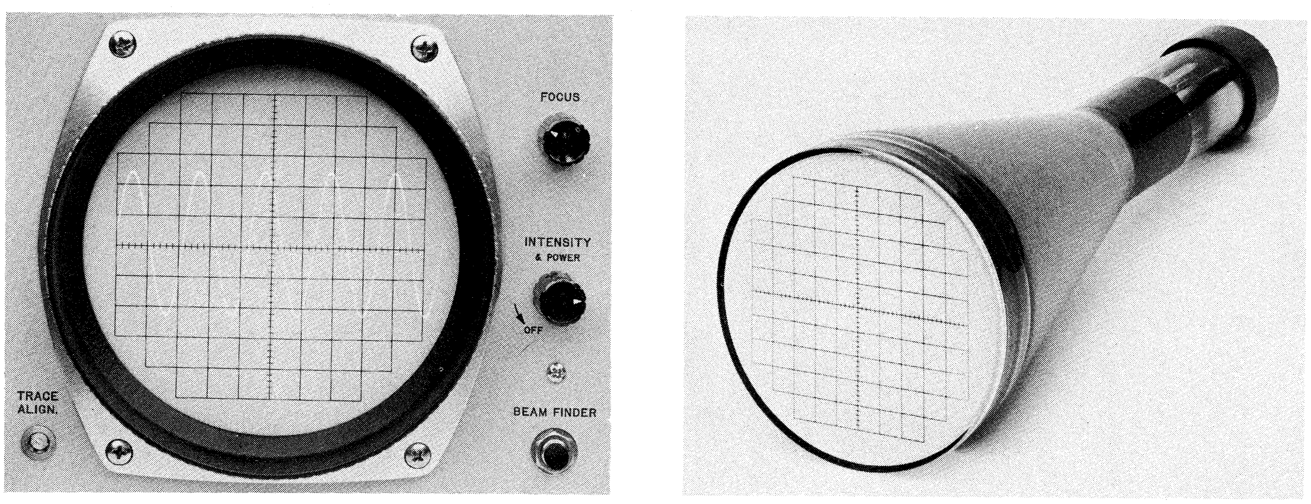


Figure 1. Parallax-Free Cathode Ray Tube of ⓗⓟ Model 120B

ⓗⓟ 120B EEM 2900

In this way the usual vertical and horizontal parallax error which is inherent in conventional CRT's is avoided. Your waveform measurements are easier, quicker and more accurate since the ambiguity caused by the parallax is eliminated.

In addition, the face plate on the crt of the 120B is made of a non-reflective glass which eliminates the filter so that you observe the trace without the distracting reflections which are common in conventional crt filter and face plates. The face plate, sealed with epoxy, is implosion proof -- and safer to handle.

Calibrated Amplifiers and Sweeps

For fast, reliable measurements, both the sweep speeds and the vertical amplifier sensitivities are accurately calibrated. A single sweep time control provides 15 direct reading sweeps from 5 microseconds/cm to 200 milliseconds/cm, while a concentric vernier control permits continuous adjustment between calibrated ranges and extends the slowest sweep range to approximately 0.5 sec/cm. This sweep time control also controls the sensitivity of the calibrated horizontal amplifier.

The vertical amplifier has calibrated ranges which extend from 10 mv/cm to 10 volts/cm with an attenuator accuracy of ±3%. Overall accuracy may be verified at any time with the built-in vertical calibrator.

Low Capacity Probes

Two pen-sized test probes with miniature alligator jaws are available for use with Model 120B. One probe, the AC-21A, has 10 pf capacitance, 10 megohm impedance and provides a 10:1 voltage division. The other, the AC-21C, has 2.5 pf capacitance, 9 megohm impedance and provides a 50:1 voltage division. Compensating capacity is adjustable by rotating one portion of the Nylon barrels. These probes are recommended for minimizing circuit loading.

Current Probe

Also available is ⓗⓟ Model AC-21F Current Probe for observing current waveforms without the necessity for inserting resistors into the circuit. Model AC-21F clamps around a wire forming a transformer with a one-turn primary and providing a 1 mv output for 1 ma flowing through the wire. Maximum current is 10 amperes rms for frequencies above 20 kc. Below 20 kc current capability is reduced proportional to frequency and is 1/2 ampere at 1 kc. Two 100 ohm terminations have been designed for the AC-21F. The AC-67B Feed-Thru Termination provides a bandpass of 2.5 kc to 30 mc while the AC-67C Compensated Termination provides a bandpass from approximately 1400 cps to 30 mc.

Convenience Features

Trigger adjustments are not necessary with the 120B, for an "automatic" trigger feature provides a bright clear base line even in the absence of a synchronizing signal. This feature combines with the "Beam Finder" to eliminate the annoying time-wasting problem of "finding the spot". Should the horizontal or vertical position or trace intensity controls be misset, the beam is brought back on screen simply by depressing the Beam Finder Control. Then, you may easily position the trace.

The trigger level adjustment is located on the front panel so that the "Automatic" triggering may easily be locked out if desired and a trigger level in the range of ±10 volts may be established. An input for intensity modulation of the beam is also available at the front panel.

Modular Cabinet

The new modular cabinet of the 120B is attractive, rugged, and convenient. The top and bottom cabinet covers may be quickly removed giving you complete accessibility to all components and adjustments. You may convert ⓗⓟ Model 120B from a rack-mounting unit to a bench-type unit in a matter of minutes. When rack-mounted the dimensions are standard (7 x 19 in.); when used on the bench, other instruments may be stacked on the louver-free top surface.

SPECIFICATIONS

SWEEP

Sweep Range:
1 μsec/cm to at least 0.5 sec/cm. 15 calibrated sweeps accurate to within ±5%, in a 1, 2, 5, 10... sequence, 5 μsec/cm to 200 millisec/cm. Vernier permits continuous adjustment of sweep time between calibrated steps and extends the 200 millisec/cm step to at least 0.5 sec/cm.

Sweep Expand:
X5 sweep expansion may be used on all ranges and expands fastest sweep to 1 μsec/cm. Expansion is about the center of the crt and expanded sweep accuracy is ±10%.

► Automatic Synchronization:
Internal: From signals 50 cps to 250 kc causing 1/2 cm vertical deflection
From signals 250 to 450 kc causing 1 cm vertical deflection
From line voltage
External: From signals 50 cps to 450 kc, 1.5 v peak-to-peak

Trigger Point:
Zero crossing, negative slope of external sync signals, zero crossing, positive or negative slope of vertical deflection signals. Front panel control overrides automatic and permits the trigger point to be set between -10 to +10 volts. Turning fully counterclockwise into auto restores automatic operation.

VERTICAL AMPLIFIER

►Bandwidth:
DC coupled: dc to 450 kc
AC coupled: 2 cps to 450 kc
Bandwidth is independent of sensitivity setting.

SPECIFICATIONS (Cont'd)

VERTICAL AMPLIFIER (Cont'd)

Sensitivity:
10 millivolts/cm to 100 volts/cm. 4 calibrated steps with attenuator accuracy of ±3%, 10 mv/cm, 100 mv/cm, 1 v/cm, and 10 v/cm. Vernier permits continuous adjustment of sensitivity between steps and extends 10 v/cm step to at least 100 v/cm.

Internal Calibrator:
Calibrating signal automatically connected to vertical amplifier for standardizing of gain, accuracy ±2%.

Input Impedance:
1 megohm, approximately 50 pf shunt

Balanced Input:
On 10 mv/cm range. Input impedance, 2 megohms shunted by approximately 25 pf.

Common Mode Rejection:
Rejection at least 40 db. Common mode signal must not exceed ±3 volts peak.

Phase Shift:
Vertical and horizontal amplifiers have same phase characteristics within $\pm 2^{o}$ to 100 kc when verniers are in CAL.

HORIZONTAL AMPLIFIER

► Bandwidth:
DC coupled: dc to 300 kc
AC coupled: 2 cps to 300 kc
Bandwidth is independent of attenuator setting.

Sensitivity:
0.1 volt/cm to 100 volts/cm. 3 calibrated steps, accurate within ±5%, .1 v/cm, 1 v/cm, and 10 v/cm. Vernier permits continuous adjustment of sensitivity between steps and extends 10 v/cm step to at least 100 v/cm.

Input Impedance:
1 megohm, nominal, shunted by approximately 100 pf.

Phase Shift:
Horizontal and vertical amplifiers have same phase characteristics within $\pm 2^{o}$ to 100 kc when verniers are in CAL.

► GENERAL

Cathode Ray Tube:
G203A (P31) internal graticule, mono-accelerator normally supplied; 2700-volt accelerating potential. Face plate eliminates glare and reduces hazard of implosion. P2, P7, and P11 phosphors are available.

Internal Graticule:
10 cm x 10 cm marked in cm squares. Major horizontal and vertical axes have 2 mm subdivisions. Eliminates parallax error.

Intensity Modulated:
Terminals on front panel. +20 volts to blank trace of normal intensity.

Dimensions:
16-3/4 in. wide, 7-1/2 in. high, 18-3/8 in. deep overall; hardware furnished for quick conversion to 7 in. x 19 in. rack mount.

Weight:
Net 32 lbs

Power:
115 or 230 volts ±10%, 50 to 1000 cps. Approximately 95 watts.

► Accessories Available:
AC-21A Probe, 10:1 division, $30.00
AC-21C Probe, 50:1 division, $30.00
AC-21F Current Probe, $100.00
AC-21J Low Frequency Probe, $9.00
AC-67B Termination for AC-21F, $17.50
AC-67C Compensated Termination for AC-21F $30.00
AC-83A Viewing Hood; face-fitting molded rubber, $5.00
H10-196A Oscilloscope Camera, $480.00
456A AC Current Probe, $190.00

Price:
hp Model 120B, $475.00

Options (no extra charge):
2. CRT with Internal Graticule and P2 phosphor installed
7. CRT with Internal Graticule and P7 phosphor installed
11. CRT with Internal Graticule and P11 phosphor installed

Prices f. o. b. factory

DATA SUBJECT TO CHANGE WITHOUT NOTICE

8/15/61
11/1/61

TECHNICAL DATA

HEWLETT-PACKARD COMPANY • 1501 PAGE MILL ROAD • PALO ALTO, CALIFORNIA, U. S. A.
CABLE "HEWPACK" TELEPHONE DAVENPORT 6-7000

hp MODEL 200CD WIDE RANGE OSCILLATOR
5 cps to 600 kc

USES

This new wide range laboratory oscillator spans the range from sub-sonic to radio frequencies. Its compact design, large, easily-read dial and flexible output circuit make it an ideal signal source for testing servo and vibration systems, medical and geophysical equipment, audio amplifiers, circuits and transducers, sonar and ultrasonic apparatus, carrier telephone systems, video frequency circuits, low radio frequency equipment, etc.

VERSATILE

The 200CD[1] frequency range of 5 cps to 600 kc is covered in five overlapping decade bands. Accurate frequency setting is provided by 85 dial divisions and an effective scale length of 78 inches. A vernier drive allows precise adjustment.

The 200CD provides a maximum of at least 10 volts across its rated load of 600 ohms and at least 20 volts open circuit. Its distortion rating is very low, being less than 0.5% below 500 kc. A special feature of the 200CD is that its waveform purity does not depend on load. Specified output waveform will be obtained even with loads of only a few ohms, although available output voltage will be decreased when using low value loads.

The output circuit of the 200CD has a nominal source impedance of 600 ohms so as to be suitable for use with audio equipment as well as carrier applications. The output transformers are balanced within .1% at the lower frequencies and within approximately 1% at the higher frequencies. A convenient panel grounding terminal is provided to ground one of the output terminals when single ended operation is desired. A simple bridged T Attenuator is provided to control output power. Where a well-balanced variable output source is desired, the hp Model AC-60A Line Matching Transformer can be used.

DESIGN

The convenient cabinet form and size allows the 200CD to occupy a minimum of bench space, while its panel arrangement aids in swift and straight-forward operation.

The improved resistance-capacity oscillator circuit is similar to the basic hp design which has made Hewlett-Packard the leading producer of this type of equipment. Highest quality components are used. Ceramic insulation at all high impedance points assures trouble-free operation even under humid or dusty conditions.

[1] Brunton Bauer and B. M. Oliver, "Those New hp Oscillators," Hewlett-Packard Journal Vol. 4 No. 4.

SPECIFICATIONS

Frequency Range:
5 cps to 600 kc covered in five ranges

Ranges:

X1	5 cps to	60 cps
X10	50 cps to	600 cps
X100	500 cps to	6 kc
X1,000	5 kc to	60 kc
X10,000	50 kc to	600 kc

Accuracy:
±2% including calibration error, warmup, changes due to aging of components, tubes, etc.

Dial:
Six-inch diameter calibrated over 300° of arc. 85 divisions. Total scale length, 78 inches.

Frequency Response:
±1 db entire frequency range (reference 1 kc)

Output:
160 milliwatts (10 volts) into 600-ohm rated load, 20 volts open circuit.

Output Balance:
Better than .1% at lower frequencies and approximately 1% at higher frequencies.

Internal Impedance:
600 ohms. Output is balanced to ground for zero attenuation. (May be operated with one side grounded if desired.)

Distortion:
Less than 0.5% below 500 kc; less than 1% 500 kc and above. Independent of load impedance.

Hum Voltage:
Less than 0.1% of rated output.
Decreases as output is attenuated.

Power:
115/230 volts ±10%, 50/1000 cps, 75 watts.

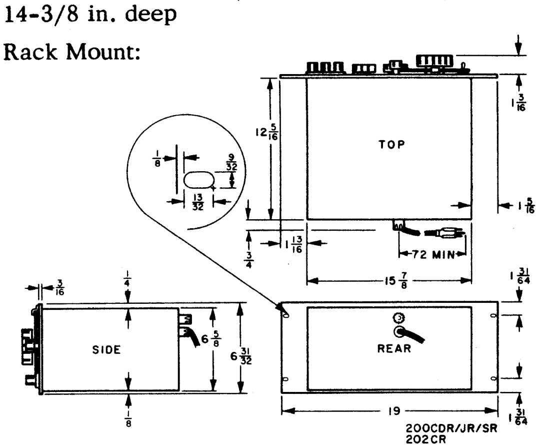
► Accessories Available:
hp AC-60A Line Matching Transformer, $60.00 (provides balanced output at any attenuator setting at 135 and 600 ohms.)

hp AC-16A Cable Assembly, 44 in. long, terminated each end with dual banana plugs, $4.50.

hp AC-16B Cable Assembly, 45 in. long, with one dual banana plug and one BNC male connector, $5.50

► Dimensions:
Cabinet Mount: 7-3/8 in. wide, 11-1/2 in. high, 14-3/8 in. deep

Rack Mount:



Weight:
Cabinet Mount: Net 22 lbs, shipping 27 lbs.
Rack Mount: Net 27 lbs, shipping 37 lbs.

► Price:
hp Model 200CD Wide Range Oscillator, Cabinet Mount, $195.00

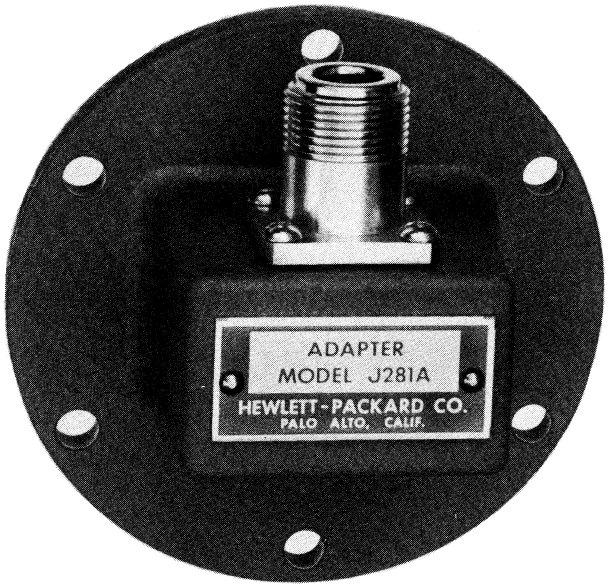
hp Model 200CDR Wide Range Oscillator, Rack Mount, $200.00

Prices f.o.b. factory
DATA SUBJECT TO CHANGE WITHOUT NOTICE

4/1/60
2/1/61

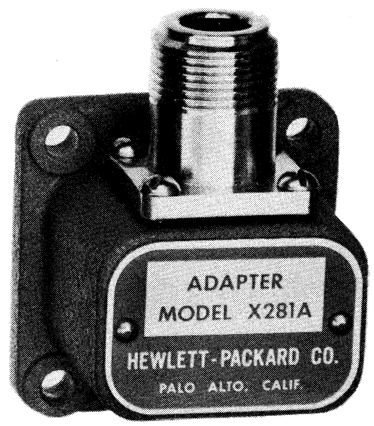
00283-2

MODEL 281A WAVEGUIDE TO COAXIAL ADAPTERS



These instruments use a probe with a low-loss dielectric sheath to transform waveguide impedance into coaxial cable impedance. They are fitted with a standard Type N female connector to a coaxial cable and a plain AN flange for connection to waveguide.

An rf gasket minimizes leakage around the Type N fitting so that the 281A may be used with sensitive receivers.



DESCRIPTION

These adapters provide a convenient coupling between waveguide and coaxial systems. Power may be transmitted in either direction, and each adapter covers the full frequency range of its waveguide size with swr of less than 1.25:1.

SPECIFICATIONS
(281A Adapters)

Model	Length (in.)	Cover Flange	Frequency Range GC	Fits Waveguide Size (in.)	► Shipping Weight (lbs.)	Price
S281A	2-1/2	UG53/U	2.60 - 3.95	3 x 1-1/2	3	$50.00
G281A	2-1/8	UG149A/U	3.95 - 5.85	2 x 1	2	40.00
J281A	2	UG344/U	5.30 - 8.20	1-1/2 x 3/4	2	35.00
H281A	1-5/8	UG51/U	7.05 - 10.00	1-1/4 x 5/8	1	30.00
X281A	1-3/8	UG39/U	8.20 - 12.40	1 x 1/2	1	25.00

281A EEM 2100

► Maximum SWR: 1.25 over entire frequency range, except J281A which has maximum SWR of 1.30 from 5.30 to 5.5 GC.

00300-2

Prices f.o.b. factory

DATA SUBJECT TO CHANGE WITHOUT NOTICE

9/30/59
5/15/61

TECHNICAL DATA

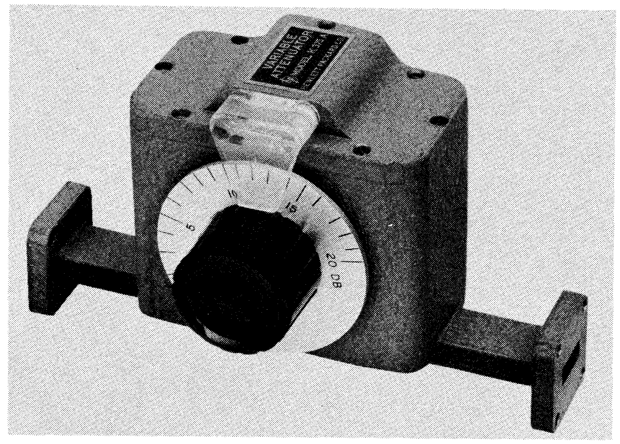
HEWLETT-PACKARD COMPANY • 275 PAGE MILL ROAD • PALO ALTO, CALIFORNIA, U. S. A.
CABLE "HEWPACK" TELEPHONE DAVENPORT 5-4451

hp MODEL 375A

VARIABLE FLAP ATTENUATORS

DESCRIPTION

Variable flap attenuators provide a simple convenient means of adjusting waveguide power level, or isolating source and load. They consist of a single slotted section in which a matched resistive strip is inserted a variable amount. The degree of strip penetration determines attenuation. A dial shows average reading over the frequency band, and a dust cover with shielded braid reduces external radiation and eliminates hand capacity effects.



Model 375 Attenuators have a maximum swr of less than 1.15 over the guide frequency range. Attenuation is variable 0 to 20 db; the equipment dissipates average power of 2 watts (except small waveguides where maximum dissipation is 0.5 or 1 watt). Dial calibration is accurate within ±1 db from 0 to 10 db, ±2 db from 10 to 20 db. K and R band models are available having either precision cover flanges or circular flanges; all other bands have precision cover flanges. Models with circular flanges are designated by following the standard model number with a "C".

SPECIFICATIONS

Model	Freq. Range kmc	Fits Waveguide Size (Inches)	Flange Type	► Power Dissipation	► Length (Inches)	Price
S375A	2.60 - 3.95	3 x 1-1/2	UG-53/U	2 watt	14-1/8	$165.00
G375A	3.95 - 5.85	2 x 1	UG-149A/U	2 watt	13	145.00
J375A	5.20 - 8.20	1-1/2 x 3/4	UG-344/U	2 watt	13	135.00
H375A	7.05 - 10.0	1-1/4 x 5/8	UG-51/U	2 watt	8-1/4	125.00
X375A	8.20 - 12.4	1 x 1/2	UG-39/U	2 watt	7-3/16	100.00
► M375A	10.0 - 15.0	.850 x .475	--------	1 watt	6-1/4	190.00
P375A	12.4 - 18.0	.702 x .391	UG-419/U	1 watt	7-1/4	135.00
K375A	18.0 - 26.5	.500 x .250	UG-595/U	0.5 watt	4-1/2	185.00
K375AC	18.0 - 26.5	.500 x .250	UG-425/U	0.5 watt	4-3/4	255.00
R375A	26.5 - 40.0	.360 x .220	UG-599/U	0.5 watt	4-3/8	200.00
R375AC	26.5 - 40.0	.360 x .220	UG-381/U	0.5 watt	4-1/2	280.00

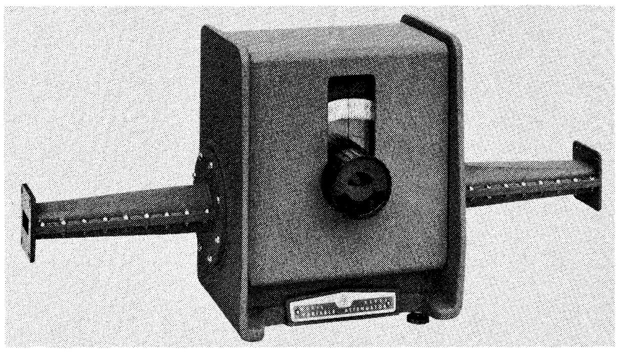
Prices f.o.b. factory
DATA SUBJECT TO CHANGE WITHOUT NOTICE

4/30/59
9/30/61

TECHNICAL DATA

HEWLETT-PACKARD COMPANY • 1501 PAGE MILL ROAD • PALO ALTO, CALIFORNIA, U.S.A.
CABLE "HEWPACK" TELEPHONE DAVENPORT 6-7000

MODEL 382A BROADBAND PRECISION WAVEGUIDE ATTENUATORS
3.95 gc to 40.0 gc



G, J, H, X, and P382A

ADVANTAGES

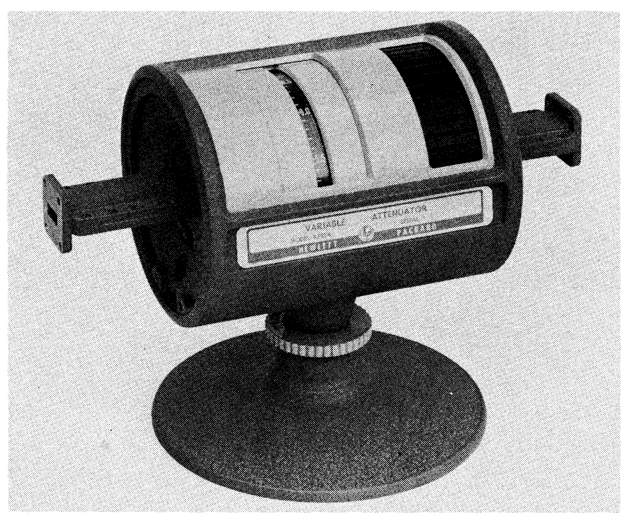
- Accurate attenuation over full guide frequency range
- Stable under varying temperature and humidity conditions
- High power-handling capacity
- Less swr from longer rectangular to circular transitions
- Superior repeatability and rigidity
- Large, easily read dial provides fast accurate setting, and direct reading

OPERATION

The operation of these precision attenuators depends on a mathematical law rather than the resistivity of an attenuating material. They provide a thoroughly reliable, true standard of attenuation for use as precision calibrators in your laboratory, or for direct comparison measurements. Since attenuation is almost completely independent of film resistivity characteristics, no frequency correction is required. The law of attenuation is readily predictable so that production techniques can be applied to their manufacture, making prices reasonable.

DESCRIPTION

The attenuators consist of three sections of round waveguide with a resistive film stretched across the diameter of each. The films at each end are in line with each other and fixed at right angles to the E-field of an incoming wave. The center section is free to rotate axially. When all three films are in line, there is no current flow in the films and no attenuation. When the center section is rotated with respect to the plane of the other two films, attenuation increases according to a cosine-squared law. The attenuation does not depend on the specific resistance of the films, but is dependent on the orientation of the central film. Hence the attenuation is independent of frequency, and variations due to temperature and humidity are minimized. Superior repeatability and long life are achieved by the use of precision ball bearings to support the center section of the attenuator. Extended length rectangular to circular transitions are provided (e.g. 4 inches minimum in the X band model) to insure optimum matching and minimum swr. K and R band models are available having either precision cover flanges or circular flanges; all other bands have precision cover flanges. Models with circular flánges are designated by a "C" following the standard model number.



00321-2

SPECIFICATIONS

► Model:	G382A	C382A	J382A	H382A	M382A	X382A	P382A	K382A[1]	R382A[2]
Frequency Range, KMC:	3.95 - 5.85	4.9 - 7.05	5.3 - 8.2	7.0 - 10.0	10.0 - 15.0	8.2 - 12.4	12.4 - 18.0	18.0 - 26.5	26.5 - 40.0
Waveguide Size (in):	2 x 1	1.718 x .923	1-1/2 x 3/4	1-1/4 x 5/8	3/4 x 3/8	1 x 1/2	.702 x .391	1/2 x 1/4	.360 x .220
Flange UG- :	149/U		344/U	51/U		39/U	419/U	595/U	599/U
Power handling capacity, watts, average continuous duty:	15	10	10	10	10	10	5	2	1
Size Length: (in)	31-5/8	23-1/8	25	19-15/16	13-7/32	15-5/8	12-1/2	7-5/8	7-1/2
Height:	9-3/4	9-11/16	8	7-15/16	5-1/2	7-1/4	7-3/4	5-1/2	5-1/2
Depth:	7-3/16	7-11/16	6-1/2	6-15/16	5-1/2	4-3/8	4-3/4	3-5/8	3-5/8
Weight (lbs) Net:	25	18	12	10	4	5	5	4	4
Shipping:	75	28	32	30	6	16	16	6	6
► Price:	$500.00	$800.00	$375.00	$350.00	$400.00	$275.00	$300.00	$475.00	$500.00

1. Available with UG-425/U flange, specify K382AC (length = 8-1/2 inch)
2. Available with UG-425/U flange, specify R382AC (length = 8-1/4 inch)

For All Models

Dial Calibration Range:	0 - 50 db (above insertion loss at zero setting)
Phase Shift:	Less than 3° variation from 0 to 50 db
Insertion loss at Zero Setting:	Less than 1 db
SWR:	Less than 1.15 entire range of attenuation and frequency.
Accuracy:	±2% of the reading in db, or 0.1 db, whichever is the greater. Includes calibration error plus frequency error.

Prices f. o. b. factory

DATA SUBJECT TO CHANGE WITHOUT NOTICE

5/30/60
7/15/61

TECHNICAL DATA

HEWLETT-PACKARD COMPANY · 1501 PAGE MILL ROAD · PALO ALTO, CALIFORNIA, U. S. A.
CABLE "HEWPACK" TELEPHONE DAVENPORT 6-7000

MODEL 415B STANDING WAVE INDICATOR

The 415B[1] is a sensitive tuned voltmeter designed to make swr measurements with the slotted lines and detector mounts. It may also be used as a null indicator, for bridge measurements, and has a 200,000 ohm input circuit for this application. Outstanding features of the 415B include an expanded swr scale, a meter shift attenuator for increased reading accuracy, recorder output and a vernier gain control.

The 415B consists of a high gain amplifier with very low noise level, operating at a fixed audio frequency. Amplifier output is measured with a square-law calibrated vacuum tube voltmeter. This meter reads direct in swr, and in db. A gain control adjusts the instrument to a convenient level.

Three scales of swr - 1 to 4, 3 to 10, and an expanded scale of 1 to 1.3 - provide exceptional readability for all types of swr measurements. A 60 db attenuator adjustable in 10 db range steps provides a calibrated range of 70 db. The unit is therefore ideally suited for measuring both high and low swr's with exceptional accuracy.

The recorder output is designed for a recording milliammeter having a 1 ma full scale deflection and an internal resistance of approximately 1500 ohms.

A logarithmic meter scale, such as the db scale of the 415B, is more accurately read on the upper half than on the lower half of the scale (see figure 1). In order to make all readings occur on the upper half of the scale, a minus 5 db attenuator has been incorporated in the 415B. If an indication appears on the lower half scale merely dropping one range step and switching in the 5 db attenuator will bring the meter indication within the upper half of the scale.

[1] Don Carmean, "A New Standing Wave Indicator with an Expanded VSWR Scale," Hewlett-Packard Journal Vol. 6 No. 7.

USE IT FOR

- SWR measurement in conjunction with slotted lines
- Determining reflection coefficient in db in conjunction with reflectometer
- Sensitive modulated rf detector for crystals or barretters
- Audio-frequency null indicator

ADVANTAGES

- Expanded scale for full scale indication of 1.3 to 1.0
- Makes all measurements in accurate, upper portion of indicating meter scale
- 70 db calibrated range
- Operates with crystals and high or low current bolometers
- Optional high impedance input for null detection
- Direct reading - eliminates tedious computations
- Output provided to drive recorder

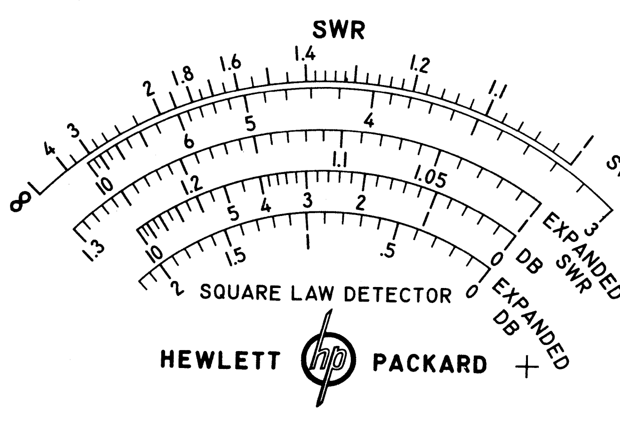


Figure 1. Detail of Meter Face

00333-2

415B EEM 2900

Three input arrangements are provided. A switch selects a 200-ohm termination, with bias of 4.3 or 8.7 ma, or unbiased for crystals; or a 200,000-ohm load for null measurements. A jack and monitor cable are provided for connecting an external milliammeter to measure bolometer current.

The instrument is normally arranged for operation at a filter frequency of 1000 cps and a power line frequency of 60 cps. On special order Model 415B is available equipped for operation at any filter frequency from 315 to 2020 cps. However, the filter frequency should not be harmonically related to the power line frequency. The frequency determining network is a plug-in unit. Units for converting the 415B to operate at frequencies in the above range can be obtained at nominal charge and installed in the field.

APPLICATION

The ⓗⓟ Model 415B Standing Wave Indicator has been specifically designed for use with the ⓗⓟ series of Slotted Lines and Detector Mounts. A complete series of Slotted Line equipment for impedance and swr measurements in coaxial and waveguide transmission systems is available to cover the entire frequency range from 500 mc to 40 gc. Also available is a convenient line of waveguide and coaxial detector mounts for the range from 10 mc to 40 gc.

SPECIFICATIONS

Frequency:
1000 cps ±2%. Other frequencies 315 to 2020 cps available on special order. Should not be harmonically related to power line frequency.

Sensitivity:
0.1 μvolt at a 200-ohm level for full scale deflection.

Noise Level:
Less than 0.03 μvolt referred to input operated from a 200-ohm resistor at room temperature.

Amplifier Q:
30 (nominal)

Calibration:
Square law. Meter reads swr, db.

Range:
70 db. Input attenuator provides 60 db in 10 db steps. Accuracy ±0.1 db per 10 db step. Maximum cumulative error ±0.2 db.

Scale Selector:
"Normal", "Expand", and "-5 db".

Meter Scales:
SWR 1-4, swr 3-10, expanded swr 1-1.3, db 0-10, expanded db 0-2.

Gain Control:
Adjusts to convenient reference level. Range at least 10 db.

Input:
"Bolo" (200 ohms). Bias provided for 8.7 ma bolometer or 1/100 amp fuse; or 4.3 ma low current bolometer.

"Crystal". 200 ohms for crystal rectifier.

"200,000 ohms". High impedance for crystal rectifier as null detector.

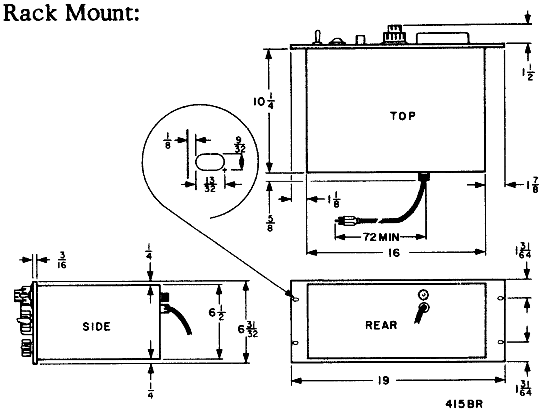
Recorder Output:
Jack provided for recording milliammeter having 1 ma full scale deflection, internal resistance of 1500 ohms or less.

Input Connector:
BNC

Power:
115/230 volts ±10%, 60 cps, 55 watts. Other frequencies on special order.

Dimensions:
Cabinet Mount: 7-1/2 in. wide, 11-3/4 in. high, 12-1/2 in. deep.

Rack Mount:



Weight:
Cabinet Mount: Net 13 lbs, shipping 19 lbs
Rack Mount: Net 17 lbs, shipping 29 lbs

Accessories Furnished:
41A-16E Cable Assembly

Accessories Available:
Plug-In Filter 415B-42B 315 to 700 cps (specify frequency), $60.00
Plug-In Filter 415B-42C 700 to 2020 cps (specify frequency), $50.00
AC-16D Cable Assembly, 44 in. of 50-ohm coaxial cable terminated at one end only with a BNC male connector, $3.50.
AC-16K Video Cable Assembly, 4 ft of 50-ohm coaxial cable terminated at each end with BNC male connectors, $6.50

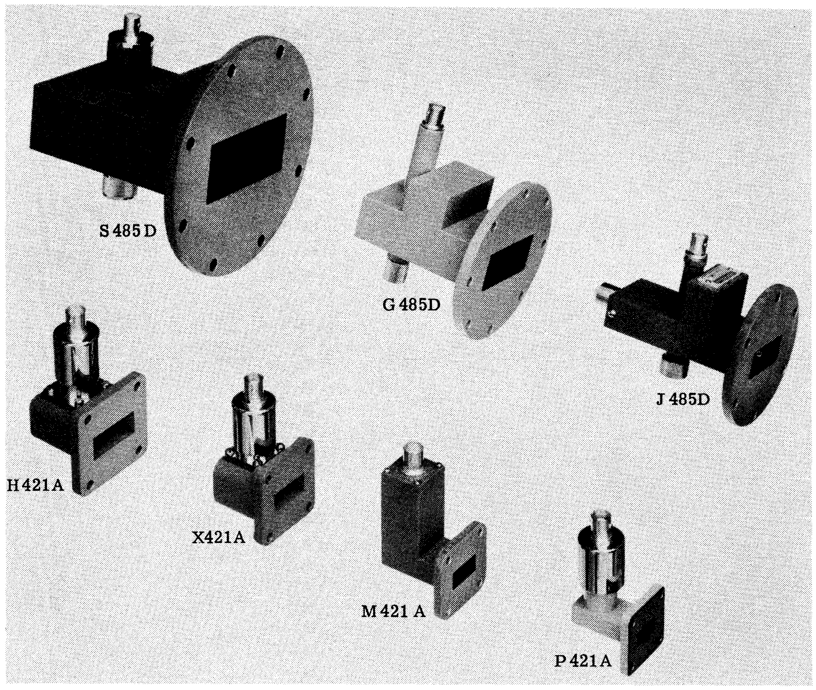
Price:
ⓗⓟ Model 415B Standing Wave Indicator, Cabinet Mount $225.00
ⓗⓟ Model 415BR Standing Wave Indicator, Rack Mount $230.00

Prices f.o.b. factory 4/1/61
DATA SUBJECT TO CHANGE WITHOUT NOTICE 9/30/61

TECHNICAL DATA

HEWLETT-PACKARD COMPANY · 1501 PAGE MILL ROAD · PALO ALTO, CALIFORNIA, U.S.A.
CABLE "HEWPACK" TELEPHONE DAVENPORT 6-7000

WAVEGUIDE REFLECTOMETER DETECTOR MOUNTS
hp MODEL S485D, G485D, AND J485D BARRETTER MOUNT
hp MODEL H421A, X421A, M421A AND P421A CRYSTAL DETECTOR MOUNT



ACCURATE, SQUARE-LAW DETECTORS FOR DETERMINING SYSTEM FLATNESS

Reflectometer Detector Mounts are available for the seven most commonly used waveguide sizes. They are supplied as barretter mounts for the three lower frequency ranges and as silicon diode mounts for the higher frequency ranges. All the mounts are characterized by a low SWR, a flat frequency response, a true square-law characteristic and a high sensitivity to provide maximum accuracy in a reflectometer installation.

All other elements required for a waveguide reflectometer setup in any of the seven frequency ranges are available from Hewlett-Packard Company. For details and application information, see J. K. Hunton and Elmer Lorence, "Improved Sweep Frequency Techniques for Broadband Microwave Testing", Hewlett-Packard Journal, Vol. 12, No. 4, December, 1960.

485D WAVEGUIDE BARRETTER MOUNTS

These instruments are available with barretters installed and tested for SWR, frequency response, and square-law characteristics. They are supplied complete with factory selected barretters to insure maximum accuracy. The hp AC-60K Barretter Matching Transformer is required to interconnect the 485D Detector Mounts and the hp 416A Ratio Meter.

421A WAVEGUIDE CRYSTAL DETECTOR MOUNTS

At higher waveguide frequencies, better standing wave characteristics can be obtained with crystals than with barretters. Instruments for the H, X, and P bands therefore use 1N26 silicon diodes, which are furnished installed in the mounts. These crystals exhibit an accurate square law characteristic over a 40 db range when operated into a selected value video load resistor. Such a resistor is factory selected and included in each mount. It is possible to select pairs of 421A Crystal Detector Mounts which exhibit similar frequency response and square-law characteristics for reflectometer applications. The units are available as "matched pairs" for a small additional charge.

SPECIFICATIONS

	S485D	G485D	J485D	H421A	X421A	M421A	P421A
Frequency Range, GC:	2.6 - 3.95	3.95 - 5.85	5.2 - 8.2	7.05 - 10	8.2 - 12.4	10.0 - 15.0	12.4 - 18.0
► Sensitivity (Typical Value):	← 2 volts rms/1 mw* →			← 0.05 volts/1 mw →			
SWR (maximum):	1.5	1.5	1.5 †	1.5	1.5	1.5	1.5
Frequency Response. Maximum variation over full range, less than:	±1 db	±1 db	±1 db	±2 db √	±2 db √	±2 db √	±2 db √
Square-Law Characteristic (maximum variation over 40 db range, maximum input power less than 1 mw:	±0.5 db	±0.5 db	±0.5 db	±1 db √	±1 db √	±1 db √	±1 db √
Detector Element:	Barretter	Barretter	Barretter	Crystal	Crystal	Crystal	Crystal
Video Load:	Not required	Not required	Not required	Selected & Installed	Selected & Installed	Selected & Installed	Selected & Installed
Waveguide Size:	3" x 1-1/2"	2" x 1"	1-1/2" x 3/4"	1-1/4" x 5/8"	1" x 1/2"	.850 x .475	.702" x .391"
Shipping Weight, approx.:	5 lbs	3 lbs	3 lbs	2 lbs	1 lb	1 lb	1 lb
Accessories Available:	ⓗⓟAC-60K Barretter Matching Transformer for interconnection with ⓗⓟ416A Ratio Meter $80.00			Matching Transformer not Required.			
► Price, f.o.b. factory	$200.00	$170.00	$170.00	$ 95.00 $105.00#	$ 75.00 $ 85.00#	$175.00 $185.00#	$150.00 $160.00 #

† Over frequency range 5.2 to 7.5 gc. Increases to approximately 2.0 at 8.2 gc.

√ When ordered as "matched pairs", tolerance on frequency response and square-law characteristics combined, but excluding basic crystal sensitivity, is held to within ±2 db for the pair.

Unit price for detectors of a matched pair.

* Fundamental component 100% square wave modulated at 1 kc; measured on secondary of AC-60K Barretter Matching Transformer.

00354-3

DATA SUBJECT TO CHANGE WITHOUT NOTICE

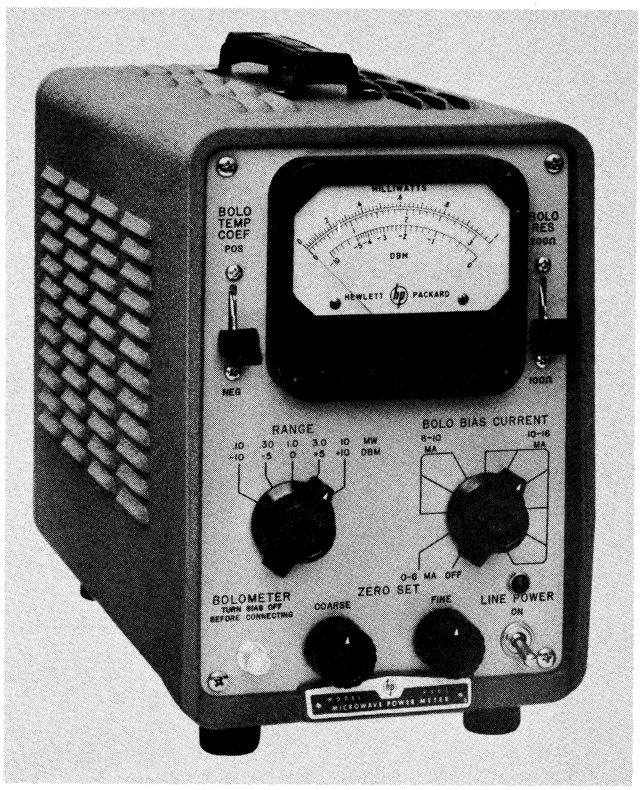
5/15/61
12/15/61

TECHNICAL DATA

HEWLETT-PACKARD COMPANY • 1501 PAGE MILL ROAD • PALO ALTO, CALIFORNIA, U. S. A.
CABLE "HEWPACK" TELEPHONE DAVENPORT 6-7000

hp MODEL 430C

MICROWAVE POWER METER



DIRECT, AUTOMATIC READINGS OF PULSED OR CW POWER

This hp Microwave Power Meter gives instantaneous rf power readings in dbm or mw - and completely eliminates tedious computation and troublesome adjustments during operation. The instrument may be used at any frequency for which there are bolometer mounts - and measurements are entirely automatic.

In measuring cw power, hp 430C can use an instrument fuse, barretter, or thermistor as the bolometer element. Pulsed or cw power may be measured using either a negative or positive temperature coefficient element at 100 - or 200-ohms. Power is read directly in milliwatts, from 0.01 to 10 mw, or in dbm from -20 to +10 dbm. Higher powers may be measured by inserting attenuators or by using directional couplers to sample energy.

When used in an appropriate bolometer mount, instrument fuses are generally satisfactory for measuring cw, and modulated power at frequencies up to 4 kmc. Barretters and thermistors can be used for these measurements at much higher frequencies, up to 12.4 kmc for barretters (in hp mounts) and up to 40.0 kmc for certain thermistors.

hp waveguide bolometer mounts are available covering frequencies from 2.6 kmc to 40 kmc. Each waveguide bolometer mount covers a complete waveguide band. In addition, coaxial bolometer mounts cover the frequency spectrum from 10 mc to 10 kmc. Model 430C Microwave Power Meter furnishes dc bias current for bolometer mounts which require up to 16-ma bias current. Fine as well as coarse control of the bias current permits exact balancing of the bolometer element in the bridge over wide-range ambient temperature variations.

ADVANTAGES

- Fast readings
- Read power directly in mw and dbm
- Use 100 or 200 ohm, positive or negative temperature coefficient bolometers
- Read cw and modulated power from 0.01 mw to 10 mw
- Extend range with attenuators, directional couplers
- Bolometer mounts available from 10 mc to 40 kmc
- Simple operation

CIRCUIT DESCRIPTION

hp Model 430C consists of an audio bridge, one arm of which is a power-sensitive element. Initially, the bridge is balanced with no rf power in the element. When rf power is applied, an equivalent audio power is automatically removed, so the bridge remains balanced. The change in audio power level is indicated directly on a vacuum tube voltmeter which is calibrated to show rf power.*

*See "Power Measurements from 10 to 12,400 Megacycles" and "More Conveniences in the Microwave Power Meter", Hewlett-Packard Journal V2 #7-8 and V6 #7.

hp 430C EEM2900

00342-1

SPECIFICATIONS

POWER RANGE:
5 ranges. Full scale readings of 0.1, 0.3, 1, 3, and 10 milliwatts. Also calibrated in dbm from -20 dbm to +10 dbm (0 dbm = 1 mw).

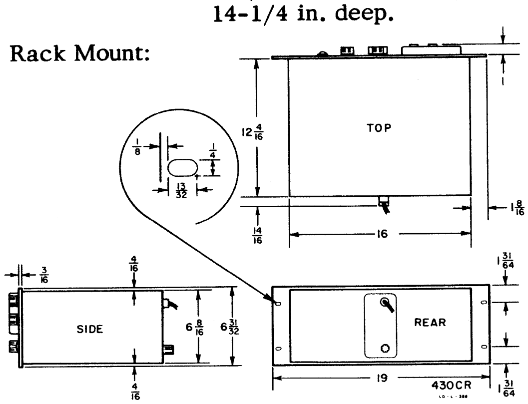
EXTERNAL BOLOMETER:
Frequency range depends on bolometer mount. Bolometers operate at resistances of 100 or 200 ohms and can have positive or negative temperature coefficients. Any dc bias current up to 16 ma is available for biasing positive or negative temperature coefficient bolometers. DC bias current is continuously adjustable and is independent of bolometer resistance and power level range.

ACCURACY:
Within ±5% of full scale.

POWER SUPPLY:
115/230 volts ±10%, 50-1000 cycles, approximately 90 watts.

SIZE:
Cabinet Mount: 7-3/8 in. wide, 11-1/2 in. high, 14-1/4 in. deep.

Rack Mount:



WEIGHT:

Cabinet Mount:	Net 14 lbs.,	Shipping 19 lbs.
Rack Mount:	Net 18 lbs.,	Shipping 30 lbs.

►ACCESSORIES AVAILABLE:
AC-16D Cable Assembly, consisting of 44 in. RG-58/U cable terminated on one end with a UG-88/U BNC connector. $3.50.
► AC-16K Cable Assembly, BNC to BNC, 48 in. long $6.50.

COAXIAL BOLOMETER MOUNTS:

hp Model 476A, 10 to 1000 mc $85.00

hp Model 477B, 10 mc to 10 kmc, $75.00

WAVEGUIDE BOLOMETER MOUNTS:
► hp Model 485A (less detector), S band, $185.00

hp Model 487B, G through R band, $75.00 to $275.00 each

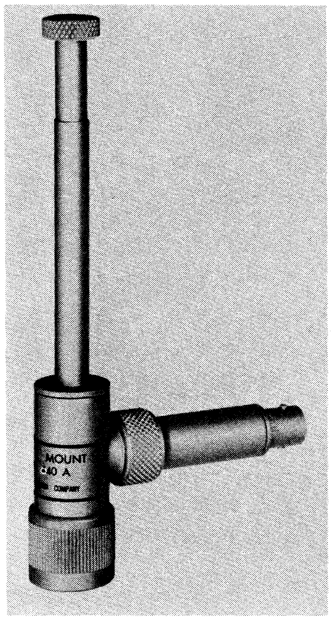
hp Model 485B (less detector), G through X band, $75.00 to $100.00

PRICE:
hp Model 430C Microwave Power Meter, Cabinet Mount, $250.00

hp Model 430CR Microwave Power Meter, Rack Mount, $255.00.

► Prices f.o.b. factory
DATA SUBJECT TO CHANGE WITHOUT NOTICE

11/15/60
2/1/61
00342-2



ⓗⓟ MODELS 440A, 442B, and 444A

MODEL 440A DETECTOR MOUNT - Simplifies Detection of RF Energy in Coaxial or Waveguide Systems.

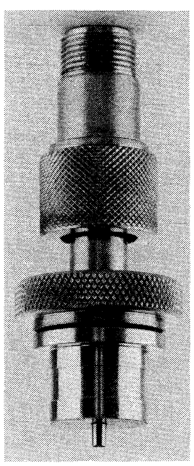
Model 440A is a simple, easy-to-use instrument for detecting rf energy in either coaxial or slotted waveguide systems.

In coaxial use, the equipment operates at any frequency from 2.4 to 12.4 kmc. Just one adjustment is required for tuning. Silicon crystals or bolometers may be used interchangeably in the same holder. A built in rf bypass is provided. The coaxial connector is equivalent to a UG21B/U Type N plug. Detector output appears at a BNC jack.

In conjunction with ⓗⓟ 442B Broadband Probe, ⓗⓟ 440A becomes a sensitive and easy-to-tune detector for use with slotted waveguide sections. The ⓗⓟ 809B carriage is designed to accept this combination.

SPECIFICATIONS -

Frequency Range:	2.4 to 12.4 kmc.
Detector (not supplied):	1N21 or 1N23 silicon crystals or Sperry 821 barretter.
Tuning:	Single stub.
Connectors:	UG21B/U (rf input); BNC jack (Detector output).
Price:	$85.00 f.o.b. Palo Alto, California.



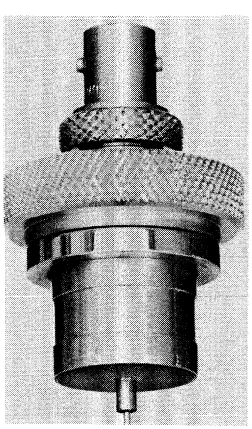
MODEL 442B BROADBAND PROBE -

Model 442B is a probe whose depth of penetration into a waveguide section is variable. It is held in position by friction, and may be fixed in place by a locking ring. Sampled rf appears at a Type N jack, permitting direct connection to a receiver, spectrum analyzer or other instrument.

This broadband Probe may be connected to a Model 440A Detector Mount to form a sensitive and convenient rf detector for slotted waveguide sections.

The probe is shielded and polyiron inserts are provided to prevent spurious resonances. Model 442B fits ⓗⓟ 809B Universal Probe Carriage or other carriages with a 3/4" diameter mounting hole.

► Price: $50.00 f.o.b. Palo Alto, California.



MODEL 444A UNTUNED PROBE -

This probe consists of a crystal plus a small antenna in a convenient housing that permits probe penetration to be varied quickly and easily. The probe is held in position by friction, or may be fixed in place by a locking ring. No tuning is required, and sensitivity is equivalent to or excels many elaborate single-and double-tuned probes, particularly over the 8.0 to 18.0 kmc range. Polyiron inserts damp out spurious resonances.

Model 444A fits ⓗⓟ 809B Universal Probe Carriage or other carriages with a 3/4" mounting hole.

SPECIFICATIONS -

Frequency Range:	3.0 to 18.0 kmc.
Output Connector:	BNC
Detector:	Supplied
Replacement Crystal:	ⓗⓟ Stock Number 444A-25E, $15.00
► Price:	$55.00 f.o.b. Palo Alto, California.

ⓗⓟ 440 EEM 3400

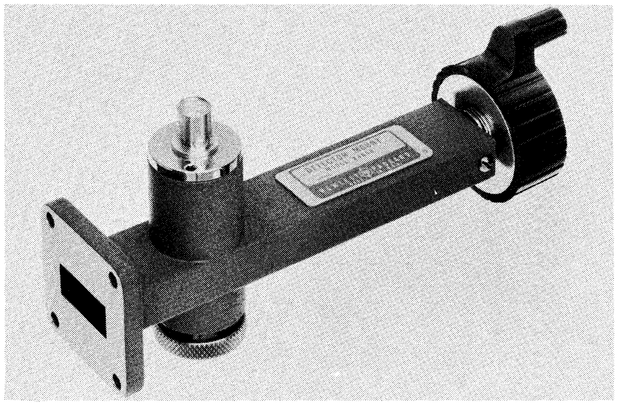
DATA SUBJECT TO CHANGE WITHOUT NOTICE

8/15/58
1/1/59

TECHNICAL DATA

HEWLETT-PACKARD COMPANY • 1501 PAGE MILL ROAD • PALO ALTO, CALIFORNIA, U. S. A.
CABLE "HEWPACK" TELEPHONE DAVENPORT 6-7000

hp MODEL 485A/B DETECTOR MOUNT



USES

- Measure microwave power
- Sensitive RF detector

SIMPLE DEVICE FOR MEASURING OR DETECTING RF POWER

Offered in two basic models, these detector mounts are designed so that a single tuning control is sufficient to match accurately waveguide sections to a bolometer element for measuring power. For maximum sensitivity in the detection of rf energy where swr is not critical the bolometer element of the 485B series can be replaced by a crystal.

Model S485A (2.6 to 3.95 gc) uses a Sperry 821 Barretter or a Narda N821 and requires no tuning. It has swr of less than 1.35 over entire waveguide band.

The 485B series, for G, J, H, and X bands (3.95 to 12.4 gc) is tuned by a variable short. When a Sperry 821 or a Narda N821 barretter is used, these mounts can be adjusted to a swr of less than 1.25 over the respective waveguide bands. For power measurements this results in a reflection loss of less than 0.1 db. For maximum sensitivity in the detection of rf energy in applications where the swr is not critical, the barretter element can be replaced with a 1N21 or a 1N23 silicon crystal. If a low swr is desired, precede the 485B with an appropriate slide screw tuner such as an hp 870A.

In all models detected output appears at a BNC jack mating with a UG88/U plug. Detector elements can be quickly interchanged. For measuring maximum power from a mismatched source, these detectors may be preceded by a slide screw tuner such as hp 870A. The detectors are ideal for use with hp 430C Microwave Power Meters, or hp 415 Standing Wave Indicators.

SPECIFICATIONS

Model	Maximum swr[1]	Frequency Range (gc)	Fits Waveguide Size (in.)	Length (in.)	► Shipping Weight	► Price
S485A[2]	1.35	2.60 - 3.95	3 x 1.5	4-11/16	5 lbs	$185.00
G485B[3]	1.25	3.95 - 5.85	2 x 1	9-5/16	5 lbs	95.00
	1.25	5.85 - 8.20				
J485B[3]	1.35	5.50 - 5.85	1.5 x 0.75	7-3/8	4 lbs	90.00
	1.50	5.20 - 5.50				
H485B[3]	1.25	7.05 - 10.0	1.25 x 0.625	6-3/8	3 lbs	85.00
X485B[3]	1.25	8.20 - 12.40	1 x 0.5	6	2 lbs	75.00

Accessories Available:

hp AC-16D Cable Assembly, 44 inches of RG-58/U 50 ohm coaxial cable terminated at one end only with a UG-88/U Type BNC male connector, $3.50

hp AC-16K Cable Assembly, 4 feet of RG-58/U 50 ohm coaxial cable terminated at each end with UG-88/U Type BNC male connectors, $6.50.

[1] With Sperry 821 or Narda N821 barretter

[2] Sperry 821 or Narda N821 only

[3] May use 1N21 or 1N23 for maximum detection sensitivity where swr is not critical.

(Detector elements are not supplied)

hp 485A/B EEM 3400

Prices f.o.b. factory
DATA SUBJECT TO CHANGE WITHOUT NOTICE

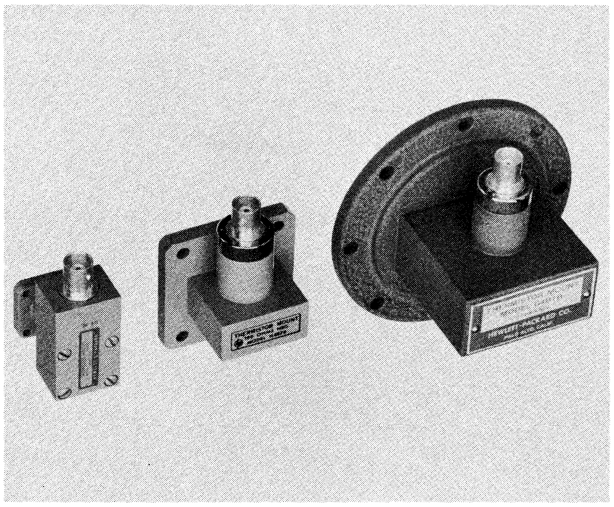
7/30/60
4/1/61

00353-2

TECHNICAL DATA

HEWLETT-PACKARD COMPANY • 1501 PAGE MILL ROAD • PALO ALTO, CALIFORNIA, U.S.A.
CABLE "HEWPACK" TELEPHONE DAVENPORT 6-7000

hp MODEL 487 BROAD BAND WAVEGUIDE THERMISTOR MOUNTS
2.6 to 40,000 mc



WAVEGUIDE THERMISTOR MOUNTS PROVIDE SWIFT WAVEGUIDE POWER MEASUREMENTS, 2.6GC (KMC) to 40.0GC

These thermistor mounts each cover full frequency range of the waveguide size. They have a low swr at all frequencies and require no tuning. Their use simplifies setups, saves operator time and provides maximum accuracy in measurement of microwave power. They employ permanently installed thermistors; thus are ideal for measuring average power of low duty cycle pulses. The units are rugged, and because the thermistors have high temperature coefficients and large overload factors, time consuming replacement operations due to detector element burnout are virtually eliminated.

The thermistor mounts are designed for use with a microwave power meter such as the hp Model 430C, which is responsive to negative temperature coefficient bolometers which operate at a 100 or 200 ohm level. They have BNC-type connectors for output connection to the microwave power meter. K and R band models are available having either precision cover flanges or circular flanges; all other bands have precision cover flanges. Models with circular flanges are designated by a "C" following the standard model number.

ADVANTAGES

- Cover full waveguide frequency range
- No tuning required
- Low swr
- Not susceptible to burn-out
- Long time-constant for accurate modulated power measurements

SPECIFICATIONS

Model	Frequency (kmc)	Maximum swr	Detector (VECO)	Oper. Resis.	Fits Waveguide Size (OD-in.)	Equiv. Flange	► Approx. Length (in.)	Ship. Weight (lb.)	Price
► S487B	2.6 - 3.95	1.35	32A5	100	3 x 1-1/2	UG-53/U	2-3/8	5	$105.00
► G487B	3.95 - 5.85	1.5	32A5	100	2 x 1	UG-149A/U	2-1/8	2	95.00
► J487B	5.3 - 8.2	1.5	32A5	100	1-1/2 x 3/4	UG-344/U	1-3/4	2	90.00
H487B	7.05 - 10.0	1.5	32A5	100	1-1/4 x 5/8	UG-51/U	1-5/16	2	80.00
X487B	8.2 - 12.4	1.5	32A5	100	1 x 1/2	UG-39/U	1-3/16	1	75.00
► M487B	10.0 - 15.0	1.5	33A9	100	0.850 x 0.475	CMR 159	15/16		110.00
P487B	12.4 - 18.0	1.5	33A9	100	0.702 x 0.391	UG-419/U	13/16	3/4	110.00
K487C	18.0 - 26.5	2	33A9	200	1/2 x 1/4	UG-595/U	1-5/8	1/2	225.00
K487BC	18.0 - 26.5	2	33A9	200	1/2 x 1/4	UG-425/U	1-5/8	1/2	260.00
R487B	26.5 - 40.0	2	33A9	200	0.360 x 0.220	UG-599/U	1-3/8	1/2	275.00
R487BC	26.5 - 40.0	2	33A9	200	0.360 x 0.220	UG-381/U	1-3/8	1/2	315.00

ALL MODELS

Maximum Power Level: 10 mw when used with an hp Model 430B or 430C Microwave Power Meter.

Thermistor Time Constant: Approximately one second when cooling on an open circuit.

Accessories Available: hp AC-16D Cable Assembly, 44 in. of RG-58/U 50 ohm coaxial cable terminated at one end with a UG-88/U Type BNC male connector, $3.50.

hp AC-16K Cable Assembly, 4 ft. of RG-58/U 50 ohm coaxial cable terminated at each end with UG-88/U Type BNC male connectors, $6.50.

Prices f.o.b. factory
DATA SUBJECT TO CHANGE WITHOUT NOTICE

00357-2

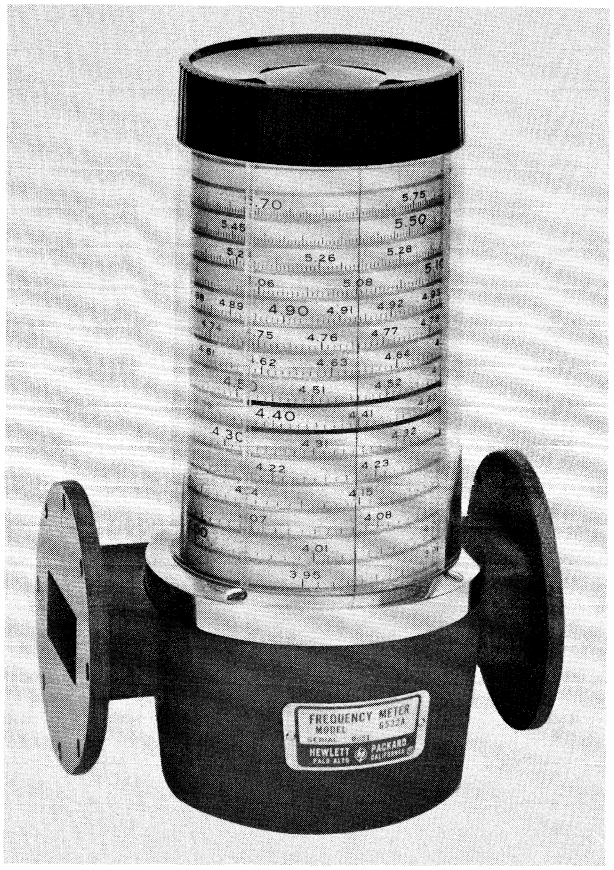
5/31/60
2/1/61

hp 487 EEM 2100

TECHNICAL DATA

HEWLETT-PACKARD COMPANY • 1501 PAGE MILL ROAD • PALO ALTO, CALIFORNIA, U.S.A.
CABLE "HEWPACK" TELEPHONE DAVENPORT 6-7000

MODEL 532 DIRECT READING FREQUENCY METERS
3,950 to 40,000 Megacycles



Model G532A

ADVANTAGES

- High resolution, easy to read dial
- Direct reading
- Broadband
- Accuracy specified over 20°C and 0 to 100% relative humidity

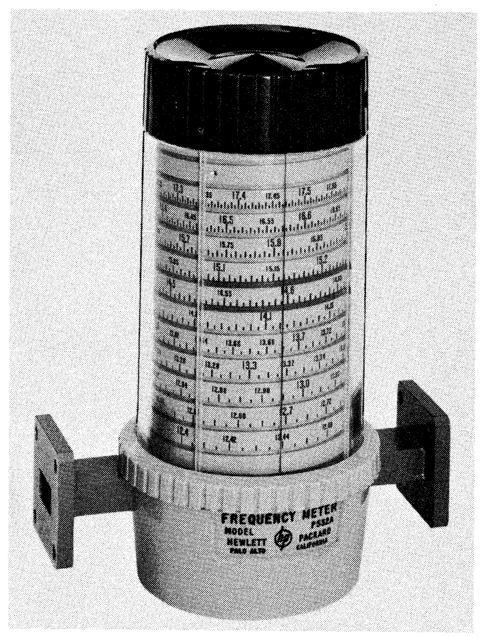
DESCRIPTION

These direct reading frequency meters allow you to measure frequencies from 3.95 to 40 GC (KMC) quickly and accurately. Their long scale length and numerous calibration marks provide a high resolution which is particularly useful when measuring frequency differences or small frequency changes. Because resolution and resetability is high, individual scale correction charts may be made and readings repeated with even higher accuracy. The sliding cursor leaves all frequency calibrations visible so that you can tell at a glance the specific portion of the band you are measuring.

Overall accuracy of the 532 Frequency Meters includes such variables as dial calibration, temperature variation over a 20° Centigrade range and relative humidity effects. The tuning plungers are spring loaded to eliminate backlash and provide good resetability. Because the calibration increments are small and the marks well separated small frequency differences may be easily resolved. Even at the high frequency end of the bands, minimum spacing of the calibration marks is 1/32 of an inch.

Except for the J532A, each frequency meter responds only to the TE_{111} mode for frequencies in their ranges; there are no spurious modes or resonances. Model J532A essentially combines two wavemeters in one unit so that the complete J-band can be covered with one rather than two instruments. Consequently, greater usefulness, convenience and savings are obtained. Because of the inherent behavior of a half-wave cavity, frequencies in the upper portion of J-band (7.6 to 8.2 GC) excite the TE_{112} mode when the dial is set between 5.3 and 5.6 GC; settings which are often excluded from instruments operating in the extended range 1-1/2 x 3/4 inch waveguide.

Resonance of the frequency meters is indicated by a small dip in transmitted power and insertion loss off resonance is approximately that of the same length of regular waveguide. Smooth tuning and long life is the result of using non-contacting plungers. K and R-band instruments are available with circular flanges (UG-425/U and UG-381/U respectively) in place of the precision cover flanges.



Model P532A

532 EEM 2900

SPECIFICATIONS

For all Models: Dip at Resonance: 1 db or more
Minimum Calibration Spacing: 1/32"

Model	G532A	J532A[1]	H532A	X532B	M532A	P532A	K532A K532AC	R532A R532AC
Frequency Range	3.95-5.85	5.3-8.2	7.05-10	8.2-12.4	10-15	12.4-18	18-26.5	26.5-40
Overall Accuracy[2]	0.065%	0.065%	0.075%	0.08%	0.085%	0.10%	0.11%	0.12%
Calib. Increments	1 MC	2 MC	2 MC	5 MC	5 MC	5 MC	10 MC	10 MC
Scale Length (in.)	155	140	125	77	74	75	72	75
Dial Accuracy[3]	0.033%	0.033%	0.040%	0.050%	0.053%	0.068%	0.077%	0.083%
Fits Waveguide (in.)	2 x 1	1-1/2 x 3/4	1-1/4 x 5/8	1 x 1/2	.850 x .475	.702 x .391	1/2 x 1/4	.360 x .220
Equiv. Flange	UG-149A/U	UG-344/U	UG-51/U	UG-39/U	- - - -	UG-419/U	K532A:UG-595/U K532AC:UG-425/U	R532A:UG-599/U R532AC:UG-381/U
Max. Temp. Coef. %/°C	0.0012	0.0012	0.0015	0.0010	0.0012	0.0012	0.0013	0.0017
Size(in.) Length[4] Height Depth	 6-1/4 9-1/2 5	 6-1/4 9-1/8 4-1/2	 6-1/4 8 4-3/8	 4-1/2 6-1/4 2-3/4	 4-1/2 6-1/4 2-3/4	 4-1/2 6-1/4 2-3/4	 4-1/2 5-1/2 2-3/4	 4-1/2 5-1/2 2-3/4
Weight (lb.) Net Shipping	 8-1/4 9	 7-1/4 8	 3-1/2 5-1/2	 3-1/2 5-1/2	 3-1/2 5-1/2	 3 5	 1-1/2 4	 1-1/2 4
Price	$375.00	$350.00	$300.00	$200.00	$300.00	$275.00	$350.00 $420.00	$400.00 $480.00

1 Because of the wide frequency range of the J532A, frequencies from 7.6 to 8.2 GC can excite the TE_{112} mode when the dial is set between 5.3 and 5.6 GC.

2 Includes dial accuracy, 20°C temperature variation (23°±10°C) and 0.02% for 0 to 100% relative humidity.

3 Includes mechanical tolerances and backlash.

4 K532AC, 5-1/8". R532AC, 5-1/8".

00379-2

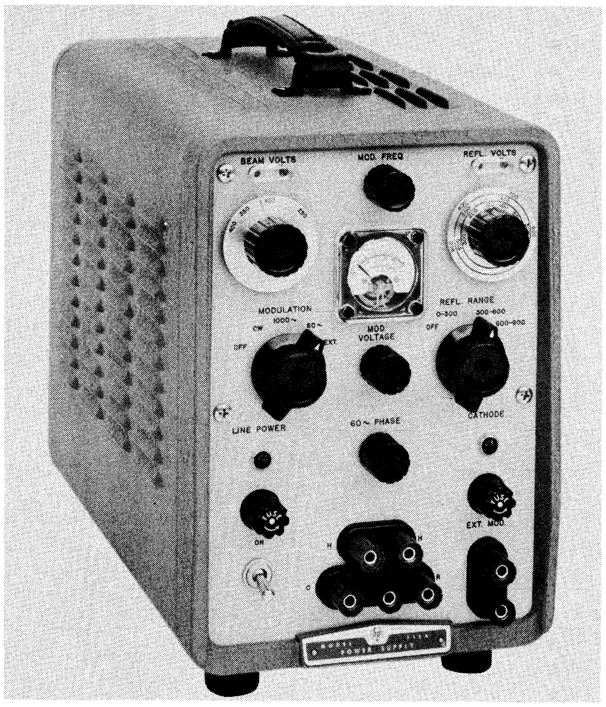
Prices f.o.b. factory
DATA SUBJECT TO CHANGE WITHOUT NOTICE

12/31/60
2/16/61

TECHNICAL DATA

HEWLETT-PACKARD COMPANY • 1501 PAGE MILL ROAD • PALO ALTO, CALIFORNIA, U.S.A.
CABLE "HEWPACK" TELEPHONE DAVENPORT 6-7000

MODEL 715A KLYSTRON POWER SUPPLY



VERSATILE POWER SOURCE FOR LOW-POWER KLYSTRONS

The Model 715A Power Supply was designed to meet the need for a compact, portable bench supply capable of operating many different types of low-power klystrons.

The supply offers a regulated -250 to -400 volt beam voltage (continuously variable), a 0 to -900 volt regulated and continuously variable reflector supply, and a 6.3-volt ac filament supply. The reflector supply, which is stacked on the beam voltage, can be squarewave modulated internally at the nominal frequency of 1000 cps or sinewave modulated at the power line frequency.

To minimize the chance of damage to a klystron, the instrument's reflector supply is arranged with a protective circuit preventing the reflector from becoming appreciably more positive than the resonator.

DESCRIPTION

The beam current is obtained from a conventional regulator circuit providing approximately 1% regulation under various combinations of load and line voltage. The reflector voltage is obtained from a regulated rf supply that provides high dc voltages. This circuit provides extreme economy of weight and size.

The reflector 1000-cps square-wave modulation is adjustable in frequency ±100 cps and in amplitude 0 to 110 volts peak-to-peak. To keep incidental fm of klystrons low, the time of rise and fall is less than 10 microseconds.

The unit is provided with a plug-in output cable that is shielded to minimize hum pick-up. Direct reading controls set the regulated voltages and a meter monitors the beam current.

SPECIFICATIONS

Supply No. 1 (Beam Supply):
-250 at 30 ma to -400 v at 50 ma; regulation less than 1% change no load to full load or for nominal line voltage variations of ±10%; ripple, less than 7 mv; calibrated voltage controls provided.

Supply No. 2 (Reflector Supply):
0 to -900 v (10 μa max.) with respect to supply no. 1; regulation, within 1% for line voltage variations of ±10% (constant current); ripple, less than 10 mv; calibrated voltage controls provided.

Filament Supply:
Provides 1.5 amperes maximum at 6.3 volts ac.

► Modulation:
Square-wave modulation provided on supply no. 2; adjustable from 0 to 110 v p-p; Rise and decay times less than 10 μs each; frequency, 1000 ± 100 cps, adjustable. Sinusoidal modulation on supply no. 2 at the power line frequency; adjustable 0 to 350 volts p-p.

External Modulation:
Terminals and circuit provided for modulation from external source. Input impedance at external modulation terminals is approximately 100,000 ohms.

Power:
115/230 volts ±10%, 50/60 cps, 200 watts.

Dimensions:
Cabinet Mount: 7-3/8 in. wide, 11-1/2 in. high, 13-3/4 in. deep.

Weight:
Net 19 lbs, shipping 24 lbs

Accessories Furnished:
715A-16C Shielded Output Cable (for connection to klystron).

Price:
Model 715A Klystron Power Supply $325.00

715A EEM 3400

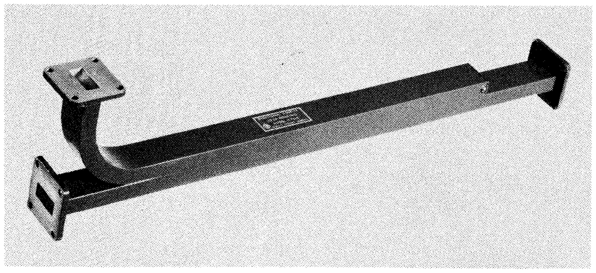
Prices f.o.b. factory
DATA SUBJECT TO CHANGE WITHOUT NOTICE

9/30/58
5/1/61

00443-2

HEWLETT-PACKARD COMPANY · 1501 PAGE MILL ROAD · PALO ALTO, CALIFORNIA, U.S.A.
CABLE "HEWPACK" TELEPHONE DAVENPORT 6-7000

MODELS 750/752 DIRECTIONAL COUPLERS



EASY-TO-USE, PRECISION COUPLERS
SIMPLIFY WAVEGUIDE MEASUREMENTS

Directional couplers[1] such as hp 752 and hp 750 are important tools in waveguide measurements. They may be used to monitor power, measure reflections, mix signals or isolate signal sources or wavemeters. They have the property of inducing into an auxiliary guide a power proportional to that flowing in the main guide. In addition, power flowing in one direction in the main guide induces uni-directional power in the auxiliary.

The ratio between power applied to the main guide and power delivered from the auxiliary is known as the "coupling factor" and is generally expressed in decibels (db).

hp 752 has an overall directivity of better than 40 db (including reflection from built-in termination and flange) over the entire range of the guide.

hp 752 MULTI-HOLE COUPLERS

Coupling is obtained from a series of graduated holes. These holes are accurately machined in two rows along the broad faces of the waveguides. Power flowing down the primary guide couples through the holes, exciting waves which propagate in both directions in the auxiliary. The coupling holes are spaced 1/4 wavelength apart, and thus waves traveling in the reverse direction are out of phase and cancel each other. Waves traveling in a forward direction re-inforce each other.

The auxiliary guide of Model 752 is terminated in a low reflection load at one end. Detection of power in the auxiliary arm can be achieved readily by connecting a crystal detector or bolometer mount to the open end.

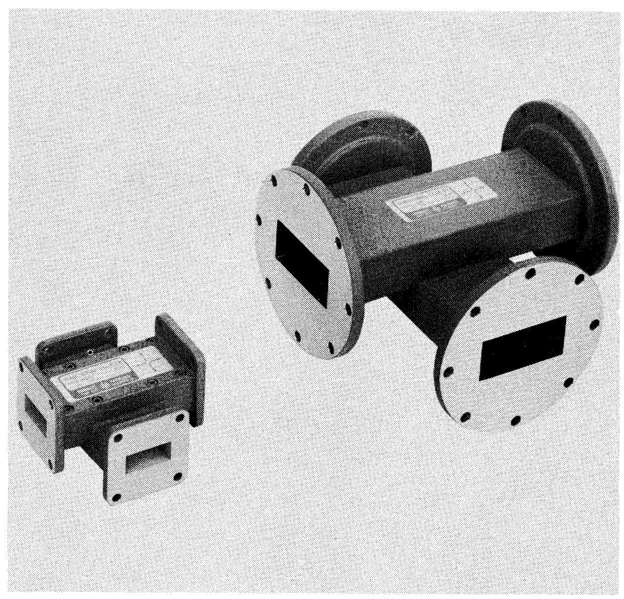
K and R band models are available having either precision cover flanges or circular flanges; all other bands have precision cover flanges which mate with other precision flanges or with choke flanges. Models with circular flanges are designated by a "C" following the standard model number.

USES AND ADVANTAGES

Because of its high directivity, this equipment is particularly suited for measurement of very small reflections, for rapidly adjusting transmission line flatness over the entire frequency range of the guide or for broadband reflectometer applications. With Model 752, a single oscilloscope presentation of swr vs. frequency is easily made. In this operation, output of the auxiliary arm of the coupler is detected, amplified and applied to the vertical plates of the oscilloscope tube. The frequency applied to the system is swept and a voltage proportional to this frequency is applied to the horizontal plates of the oscilloscope. The resulting trace is a plot of reflection vs. frequency.

REFLECTOMETER SETUP

For reflectometer measurements[2] two directional couplers are connected back-to-back between a swept frequency source, such as an hp Electronic Sweep Oscillator, and the device being measured. One directional coupler samples power traveling to the load, while the other samples power reflected by the load. The output signals of the directional couplers may be measured separately and reflection coefficient



X750D G750E

[1] E. F. Barnett and J. K. Hunton, "A Precision Directional Coupler Using Multi-Hole Coupling", Hewlett-Packard Journal, Vol. 3 No. 7-8, Mar., April, 1952.

[2] J. K. Hunton and N. L. Pappas, "The hp Microwave Reflectometers", Hewlett-Packard Journal, Vol. 6, No. 1-2, Sept.-Oct., 1954.

computed, or the output signals may be connected to the ⓗⓟ 416A Ratio Meter and reflection coefficient read directly.

ⓗⓟ 752A 3 DB COUPLER

Since the 3 db coupler has the high directivity of the 10 and 20 db couplers, it can usually be used in place of hybrid Tees. The 3 db multi-hole coupler, unlike the hybrid Tee, is a matched device having a low swr of 1.10 or less over a waveguide frequency range.

ⓗⓟ 750 CROSS-GUIDE COUPLERS

ⓗⓟ 750 Cross-Guide Coupler is an inexpensive and compact instrument suited to numerous laboratory tests. Model X750 Cross-Guide couplers are cast then broached to tight tolerances. Compact X750 is only 3 inches flange-to-flange.

Model 750 consists of two waveguide sections joined at right angles across their broad faces. It is available in coupling factors of 20 or 30 db, and connections may be made to both ends of the main and auxiliary guides. This provides a "four-terminal" network of maximum usefulness and versatility. The unit is well suited for power monitoring, for isolation and for mixing powers.

SPECIFICATIONS ⓗⓟ 750 Cross-Guide Couplers

Model	Coupling (db)	Waveguide Size (inches)	Frequency Range (GC)	Physical Size (in.)	Net Weight (lbs.)	Shipping Weight (lbs.)	Price
S750D S750E	20 30	3 x 1-1/2	2.6 - 3.95	9 x 9	11	18	$150.00
G750D G750E	20 30	2 x 1	3.95 - 5.85	6 x 6	5	7	120.00
J750D* J750E*	20 30	1-1/2 x 3/4	*5.85 - 8.20	5 x 5	3	4	100.00
H750D H750E	20 30	1-1/4 x 5/8	7.05 - 10.0	4 x 4	1-1/2	3	75.00
X750D X750E	20 30	1 x 1/2	8.2 - 12.4	3 x 3	1	2	60.00

* J750 couplers usable to 5.2 GC. Directivity same as above. Coupling within 3 db of nominal value.

DIRECTIVITY: Approximately 20 db or more.

COUPLING ACCURACY: Less than ±1.7 db variation from nominal value over entire frequency range of guide.

FLANGES: Precision cover flanges.

SPECIFICATIONS

ⓗⓟ MODEL 752 MULTI-HOLE DIRECTIONAL COUPLERS

Band[1,2] (Prefix)	Freq. GC	Fits Waveguide	Mean Coupling Accuracy[3,4]	SWR Main Guide 752A	SWR Main Guide 752C/D	Average Power Aux. Guide Load(w)	Length (in.) 752A	Length (in.) 752C	Length (in.) 752D	Weight(lbs.) Net	Weight(lbs.) Ship.	Price
S	2.6 - 3.95	3 x 1-1/2	±0.4	1.1	1.05	2	50-1/4	48	48	25	40	$400.00
G	3.95 - 5.85	2 x 1	±0.4			2	34-5/8	33	33	10-1/4	19	$300.00
J*	5.85 - 8.2	1-1/2 x 3/4	±0.4			1	26-1/2	25-9/16	25-9/16	5-1/4	16	$190.00
H	7.05 - 10	1-1/4 x 5/8	±0.4			1	18-5/8	17-1/2	17-1/2	2-3/4	5	$135.00
X	8.2 - 12.4	1 x 1/2	±0.4			1	16-11/16	15-11/16	15-11/16	1-3/4	4	$110.00
M	10.0 - 15.0	.850 x .475	±0.4			1	16-5/16	15-11/16	15-11/16	1-1/4	4	$175.00
P	12.4 - 18.0	.702 x .391	±0.4			1	13-3/4	12-1/4	12-1/4	3/4	3	$125.00
K†	18.0 - 26.5	.500 x .250	±0.7			1/2	10-3/8	9-15/16	9-15/16	1/2	3	$200.00
R†	26.4 - 40.0	.360 x .220	±0.7			1/2	11-5/8	8-5/8	7-5/8	1/4	2	$250.00

1. First letter suffix indicates nominal coupling. A for 3 db, C for 10 db, D for 20 db (Example: S-band, 3 db coupling, Model S752A). Second suffix "C" indicates circular flange model, UG-425 for K-band, UG-381 for R-band (Example: K-band, 20 db coupling, circular flanges, Model K752DC).
2. Directivity is at least 40 db.
3. Mean coupling is the average of the maximum and minimum coupling values in the rated frequency range.
4. Coupling variation over rated frequency range is not more than ±0.5 db about mean coupling.

* J752 Couplers operate to 5.2 GC with reduced performance. DIRECTIVITY: >40 db, 5.85 to 5.5 GC >36 db, 5.5 to 5.2 GC. Variation of Coupling from nominal value: not more than -1.2 db at 5.5 GC, not more than -2 db at 5.2 GC.

† Circular flange model available see note 1.

Prices f. o. b. factory

DATA SUBJECT TO CHANGE WITHOUT NOTICE

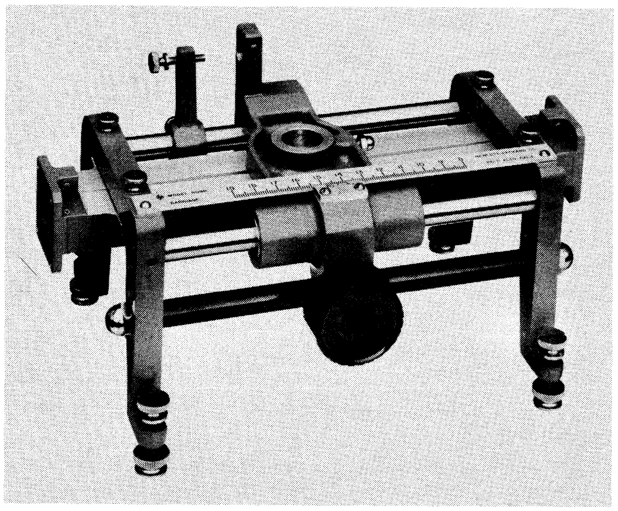
12/1/60
6/1/61

00450-2

TECHNICAL DATA

HEWLETT-PACKARD COMPANY · 1501 PAGE MILL ROAD · PALO ALTO, CALIFORNIA, U. S. A.
CABLE "HEWPACK" TELEPHONE DAVENPORT 6-7000

hp MODEL 809B UNIVERSAL PROBE CARRIAGE
hp MODEL 810 WAVEGUIDE SLOTTED SECTION
hp MODEL 806B COAXIAL SLOTTED SECTION



hp Model 809B shown with hp Model 810B

LOW-COST, PRECISION TOOLS FOR MICROWAVE READINGS

Model 809B Universal Probe Carriage is a precision-built mechanical assembly designed to operate with five hp 810B Waveguide Slotted Sections covering frequency ranges from 3.95 kmc to 18.0 kmc and with hp 806B Coaxial Slotted Section, 3.0 to 12.0 kmc.

Model 809B provides, in one compact instrument, equipment that greatly simplifies waveguide measurements over a number of frequency bands and eliminates the cost of a probe carriage for each waveguide band. It saves appreciably on engineering time since waveguide sections can be interchanged in 30 seconds or less. It is lightweight and easily portable and is designed for use with either hp 444A Untuned Probe or hp 440A Detector and hp 442B Broadband Probe in combination. The unit has a centimeter scale with a vernier reading to 0.1 mm. Provision is also made for mounting a dial gauge where more accurate readings are required.

The instrument is simple in mechanical design and is carefully manufactured to assure trouble-free operation. The probe carriage moves on ground stainless steel rods, and its 3-point suspension system includes two linear-motion ball bearings with dust seals and permanent lubrication. A conventional ball bearing forms the third point of suspension. Accuracy is superior or equal to the most expensive custom-made slotted lines.

00454-2

ADVANTAGES

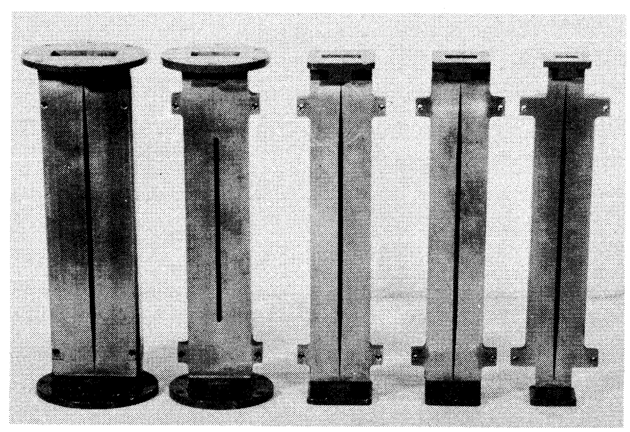
- Universal Carriage mounts 6 different slotted sections
- Broad Usefulness, 3,000 to 18,000 mc
- Carriage operates with hp waveguide or coaxial slotted sections
- Precision accuracy, high stability
- Sections interchange in 30 seconds
- Mounts dial gauge for high accuracy
- Simple operation, compact, low cost

USE IT TO MEASURE

- Characteristics of rf waveguide systems or coaxial transmission lines
- Standing wave magnitude and phase
- Impedance
- System flatness, connector reflection
- Degree of antenna-match
- Percent of transmitted or reflected power

hp 810B WAVEGUIDE SLOTTED SECTIONS

Waveguide slotted sections are fundamental tools for the measurement of magnitude and phase of standing waves in a waveguide system. Such data may be transformed readily into impedance of terminal load of the system or components. Slotted sections may also be used to measure reflection, percent of transmitted power, degree of antenna match and other waveguide characteristics.



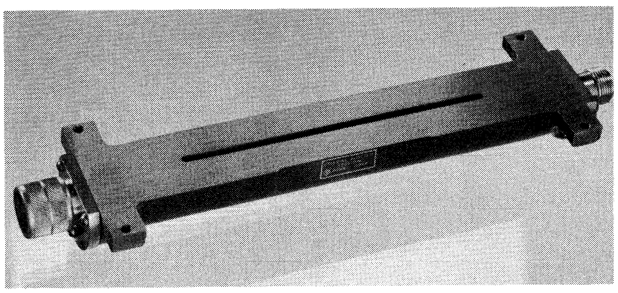
hp 810B Waveguide Slotted Sections comprise an accurately-machined section of waveguide in which a small longitudinal slot is cut. They are designed and finished to fit hp 809B Carriage in a precisely indexed

hp 806B EEM 3400

position. A traveling probe mounted on the 809B Carriage samples the waveguide's electric fields along the slot and permits precise plotting of variations along the entire length of probe travel. The slotted sections are carefully machined from normalized aluminum castings to insure a uniform cross-section.

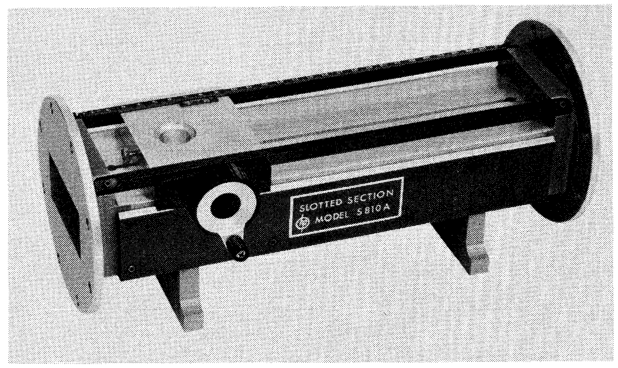
Ends of the slots are tapered to reduce slot reflection to less than 1.01 swr. A high order of accuracy and stability is maintained. ⓗⓟ 810B sections are available for seven common waveguide frequency-sizes.

ⓗⓟ 806B COAXIAL SLOTTED SECTION



This instrument provides continuous coverage from 3.0 to 12.0 kmc and is designed for use with ⓗⓟ 809B Universal Probe Carriage. Impedance is 50 ohms to match flexible coaxial cables. This broadband coaxial slotted section has special fittings mating with Type N connectors to assure a minimum swr.

ⓗⓟ 810A WAVEGUIDE SLOTTED SECTION



This instrument is a conventional type of slotted waveguide complete with probe carriage mounted directly on the section. Model S810A is available in the 3 x 1-1/2 inch (2.6 to 3.95 kmc) frequency only. The carriage accepts ⓗⓟ Model 442B Broadband Probe for sampling rf or Model 444A Untuned Probe which supplies a detected output.

SPECIFICATIONS

ⓗⓟ 809B Universal Probe Carriage

Carriage:
Mounts all ⓗⓟ 810B Waveguide Slotted Sections and ⓗⓟ 806B Coaxial Slotted Sections.

Probe Required:
ⓗⓟ 442B Broadband Probe in combination with ⓗⓟ 440A Detector or ⓗⓟ 444A Untuned Probe.

Probe Travel:
10 Centimeters

Calibration:
Metric. Vernier permits readings to 0.1 mm. Provision for dial gauge installation.

Leveling Screws:
Knurled thumb screws provided on all 4 carriage legs.

►Accuracy:
When used with waveguide sections, standing wave ratios to 1.02 can be easily read. Slope error of slotted sections may be eliminated by adjustment.

Dimensions:
8 in. long, 6-1/4 in. wide, 5 in. high.

►Price:
$175.00

ⓗⓟ S810A Waveguide Slotted Section

Conventional waveguide slotted section with probe carriage mounted directly on waveguide. Will accept ⓗⓟ 442B or 444A Probes.

Frequency Range:
2.6 to 3.95 kmc

► Slope and Irregularities:
1.01 swr

Residual SWR:
Less than 1.01

Waveguide Size:
3 x 1-1/2 inches

Length:
12-3/4 inches

Price:
$450.00

SPECIFICATIONS (CONT'D)

ⓗⓟ 806B Coaxial Slotted Section

Carriage:
Fits ⓗⓟ 809B Universal Probe Carriage

Frequency Range:
3.0 to 12.0 kmc

► Impedance:
50 ohms

Connections:
Type N, one male, one female. Special fittings provide minimum swr. Either end may be connected to load. Includes shorting connectors, male and female, for phase measurements.

Residual SWR:
Less than 1.04, 3.0 to 8.0 kmc. Approx. 1.06, 8.0 to 10.0 kmc. Approx. 1.1, 10.0 to 12.0 kmc.

Pick-up Error:
Probe pick-up variation along line is less than 0.1 db except at extreme ends where it is less than 0.4 db.

Length:
10 inches

Price:
$200.00

ⓗⓟ 810B Waveguide Slotted Sections

Model	Frequency Range kmc	Fits Waveguide Size (inches)	Equiv. Flange	Length (inches)	Price
► G810B	3.95 - 5.85	2 x 1	UG-149A/U	↑	$125.00
► C810B	4.9 - 7.05	1.718 x .923	--------		150.00
J 810B	5.20 - 8.20	1-1/2 x 3/4	UG-344/U		110.00
H810B	7.05 - 10.00	1-1/4 x 5/8	UG-51/U	10-1/4	110.00
M810B	10.0 - 15.0	.850 x .475	--------		175.00
X810B	8.20 - 12.40	1 x 1/2	UG-39/U		90.00
P810B	12.40 - 18.00	.702 x .391	UG-419/U	↓	110.00

► Slope and irregularities: 1.01 swr

Disontinuity due to slot results in swr of less than 1.01

Prices f.o.b. factory

DATA SUBJECT TO CHANGE WITHOUT NOTICE

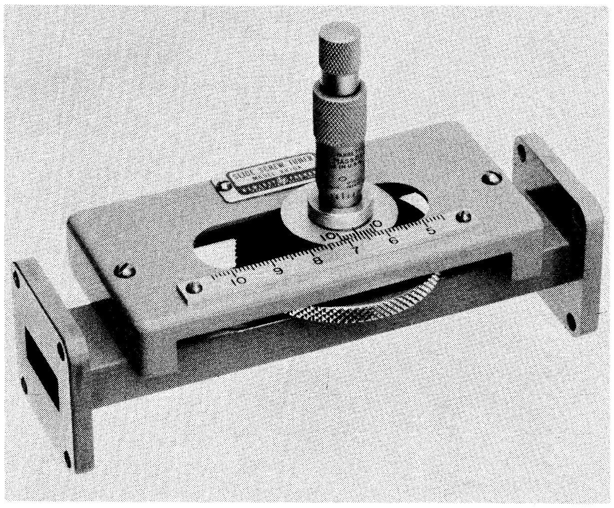
8/30/60
2/1/61

00454-2

TECHNICAL DATA

HEWLETT-PACKARD COMPANY · 1501 PAGE MILL ROAD · PALO ALTO, CALIFORNIA, U. S. A.
CABLE "HEWPACK" TELEPHONE DAVENPORT 6-7000

ⓗⓟ MODEL 870A SLIDE SCREW TUNERS



ⓗⓟ Model X870A

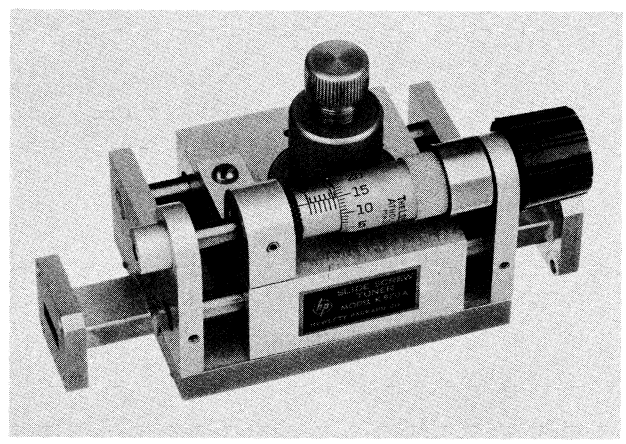
USES

Waveguide slide-screw tuners are used primarily for correcting discontinuities or for "flattening" waveguide systems. They are also used to match loads, terminations, bolometer mounts or antennas to the characteristic admittance of the waveguide. They are particularly valuable in determining experimentally the position and magnitude of matching structures required in waveguide systems.

DESCRIPTION

The tuners consist of a waveguide slotted section with a precision-built carriage on which is mounted an adjustable probe. The position and penetration of the probe is adjusted to set up a reflection which is used to cancel out an existing reflection in a system.

Probe penetration into the guide is varied with a threaded adjustment on the S, K and R band units and by a micrometer drive on the others. Position of the probe along the guide of the G, J, H, X and P band units is adjusted by a thumb-operated wheel and position is read to 0.1 mm on a vernier scale. K and R band instruments have a micrometer drive to move the probe along the guide and its graduations can be read to 0.01 mm. The maximum swr values (see specifications) can be corrected with an accuracy of 1.02, and small swr's may be exactly corrected. K and R band models are available having either precision cover flanges or circular flanges; all other bands have precision cover flanges.



ⓗⓟ Model K870A

SPECIFICATIONS

Model	Freq. Range gc	Fits Waveguide Size (in.)	Equivalent Flange Type	Length (in.)	Ship Wt. (lbs.)	Price
S870A	2.6 - 3.95	3 x 1-1/2	UG-53/U	11	15	$250.00
G870A	3.95 - 5.85	2 x 1	UG-149A/U	8-1/4	7	200.00
►C870A	4.9 - 7.05	1.718 x .923	CMR-159	9-1/4	7	225.00
J 870A	5.3 - 8.20	1-1/2 x 3/4	UG-344/U	7-5/8	7	165.00
H870A	7.05 - 10.00	1-1/4 x 5/8	UG-51/U	6	4	140.00
X870A	8.20 - 12.40	1 x 1/2	UG-39/U	5-1/2	3	130.00
M870A	10.0 - 15.0	0.850 x 0.475	----	5-7/8	3	170.00
P870A	12.40 - 18.00	0.702 x 0.391	UG-419/U	5	3	140.00
K870A	18.00 - 26.50	1/2 x 1/4	UG-595/U	4-1/4	3	250.00
►K870AC			UG-425/U	4-7/8		320.00
R870A	26.50 - 40.00	0.360 x 0.220	UG-599/U	4-3/8	3	300.00
►R870AC			UG-381/U	5		380.00

Correctable swr on all models: 20

Insertion loss db at corrected swr of 20: 2 db max.; 3 db max., for K and R bands

Prices f.o.b. factory 8/30/60

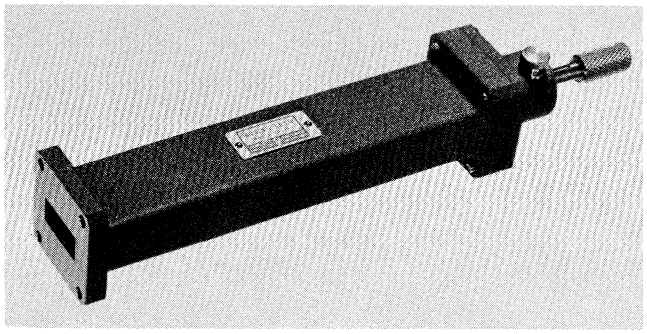
00459-2 DATA SUBJECT TO CHANGE WITHOUT NOTICE 4/1/61

ⓗⓟ 870A EEM 3400

TECHNICAL DATA

HEWLETT-PACKARD COMPANY · 1501 PAGE MILL ROAD · PALO ALTO, CALIFORNIA, U. S. A.
CABLE "HEWPACK" TELEPHONE DAVENPORT 6-7000

MODEL 914 A/B MOVING LOADS



DESCRIPTION

Model 914 Moving Load consists of a section of waveguide in which is mounted a sliding, tapered, low-reflection load. A plunger controls the position of the load, moving it at least 1/2 wavelength at the lowest waveguide frequency. Thus the phase of the residual load reflection may be reversed so that this reflection can be separated from the other small reflections in the waveguide system. K and R band models are available having either precision cover flanges or circular flanges; all other bands have precision cover flanges which mate with other precision flanges or with choke flanges.

Models with circular flanges are designated by a "C" following the standard model number. Models 914A differ from Models 914B in construction. The "A" models are made of standard waveguide stock, whereas the X914B is an aluminum casting broached to very close tolerances to keep the waveguide swr low and Models K914B and R914B are precision machined from brass stock.

In addition, all "B" models have a locking nut which prevents accidental movement of the load.

SPECIFICATIONS

	Model	Freq. Range (gc)	Fits Waveguide Size (in.)	Equivalent Flange	Max. SWR Load	Max. SWR Wavg.	Average Power Rating (Watts)	Length (in.)	Price
	S914A	2.6 - 3.95	3 x 1-1/2	UG53/U	↑	1.01	2W	25-3/4	$125.00
	G914A	3.95 - 5.85	2 x 1	UG149A/U		1.01	2W	17	95.00
►	J914A	5.2 - 8.2	1-1/2 x 3/4	UG344/U		1.01	2W	13-1/2	85.00
	H914A	7.05 - 10.0	1-1/4 x 5/8	UG51/U		1.015	1W	11-1/4	70.00
	X914B	8.2 - 12.4	1 x 1/2	UG39/U		1.005	1W	10	60.00
	M914A	10.0 - 15.0	0.850 x 0.475	-----	1.01	1.02	1W	8-1/16	85.00
►	P914A	12.4 - 18.0	0.702 x 0.391	UG419/U		1.02	1/2W	9-3/4	70.00
	K914B	18.0 - 26.5	1/2 x 1/4	UG595/U		1.01	1/2W	8-1/4	250.00
	K914BC	18.0 - 26.5	1/2 x 1/4	UG425/U		1.01	1/2W	8-1/4	285.00
	R914B	26.5 - 40.0	0.360 x 0.220	UG599/U		1.01	1/2W	7	250.00
	R914BC	26.5 - 40.0	0.360 x 0.220	UG381/U	↓	1.01	1/2W	7	290.00

914 EEM 3400

Prices f.o.b. factory
DATA SUBJECT TO CHANGE WITHOUT NOTICE

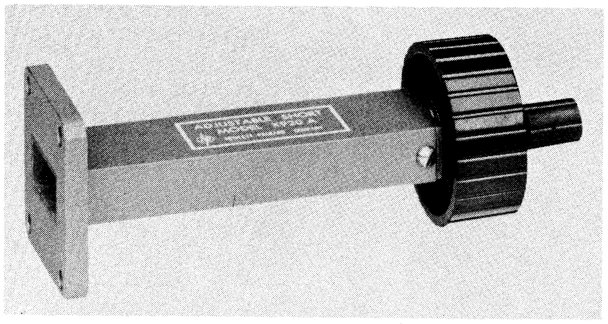
8/30/60
4/1/61

00465-2

TECHNICAL DATA

HEWLETT-PACKARD COMPANY · 1501 PAGE MILL ROAD · PALO ALTO, CALIFORNIA, U.S.A.
CABLE "HEWPACK" TELEPHONE DAVENPORT 6-7000

hp MODEL 920
ADJUSTABLE SHORT



USES

Adjustable shorts are convenient instruments for introducing a variable element in waveguide systems. They can be used to provide a variable short-circuit reference point. With a waveguide tee section, they can form a stub-transformer or tuner providing variable reactance. They may also be used as a convenient tuner for crystal or bolometer mounts.

DESCRIPTION

Mechanically, hp 920 Shorts are a waveguide section in which a movable low-loss contacting finger wiper* is mounted. The position of the short is varied by a fine tuning control. K and R band models are available having either precision cover flanges or circular flanges; all other bands have precision cover flanges. Models with circular flanges are designated by a "C" following the standard model number.

SPECIFICATIONS

Model	Fits Waveguide Size (in.)	Equiv. Flange Type	Freq. Range kmc	Length (in.)	Ship. Wt. (lbs.)	Price
S920A	3 x 1-1/2	UG-53/U	2.60 - 3.95	10-7/16	10	$150.00
G920A	2 x 1	UG-149A/U	3.95 - 5.85	7-13/16	4	125.00
J920A	1-1/2 x 3/4	UG-344/U	5.20 - 8.20	6-1/4	3	100.00
H920A	1-1/4 x 5/8	UG-51/U	7.05 - 10.00	4-7/8	2	85.00
X920A	1 x 1/2	UG-39/U	8.20 - 12.40	4-7/8	2	75.00
M920A	.850 x .475	------	10.00 - 15.00	4-13/16	2	125.00
P920B	.702 x .391	UG-419/U	12.40 - 18.00	5-3/4	2	125.00
► K920B	.500 x .250	UG-595/U	18.00 - 26.50	5-1/2	2	155.00
K920BC	.500 x .250	UG-425/U	18.00 - 26.50	5-13/16	2	190.00
R920B	.360 x .220	UG-599/U	26.50 - 40.00	4-1/2	2	155.00
R920BC	.360 x .220	UG-381/U	26.50 - 40.00	4-13/16	2	195.00

* In the P, K, and R bands a choke-type short is employed. Position of the choke is varied by a micrometer adjustment.

Prices f.o.b. factory

DATA SUBJECT TO CHANGE WITHOUT NOTICE

5/15/61
10/30/61

hp 920A EEM 3400

00467-2

APPENDIX C

Bibliography

The following reference books may be consulted for more detailed (and often more theoretical) discussions on microwave measurements.

Ginzton, Edward L., *Microwave Measurements.* New York: McGraw-Hill Book Company, Inc., 1957.

King, Donald D., *Measurements at Centimeter Wavelength.* New York: D. Van Nostrand Company, Inc., 1952.

Reich, Herbert J., Philip F. Ordung, Herbert L. Krauss, and John G. Skalnik, *Microwave Theory and Techniques.* New York: D. Van Nostrand Company, Inc., 1953.

Wind, Moe and Harold Rapaport, eds., *Handbook of Microwave Measurements.* Polytechnic Institute of Brooklyn, Microwave Research Institute, 1955.