Errata

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HP References in this Application Note

This application note may contain references to HP or Hewlett-Packard. Please note that Hewlett-Packard's former test and measurement, semiconductor products and chemical analysis businesses are now part of Agilent Technologies. We have made no changes to this application note copy. The HP XXXX referred to in this document is now the Agilent XXXX. For example, model number HP8648A is now model number Agilent 8648A.

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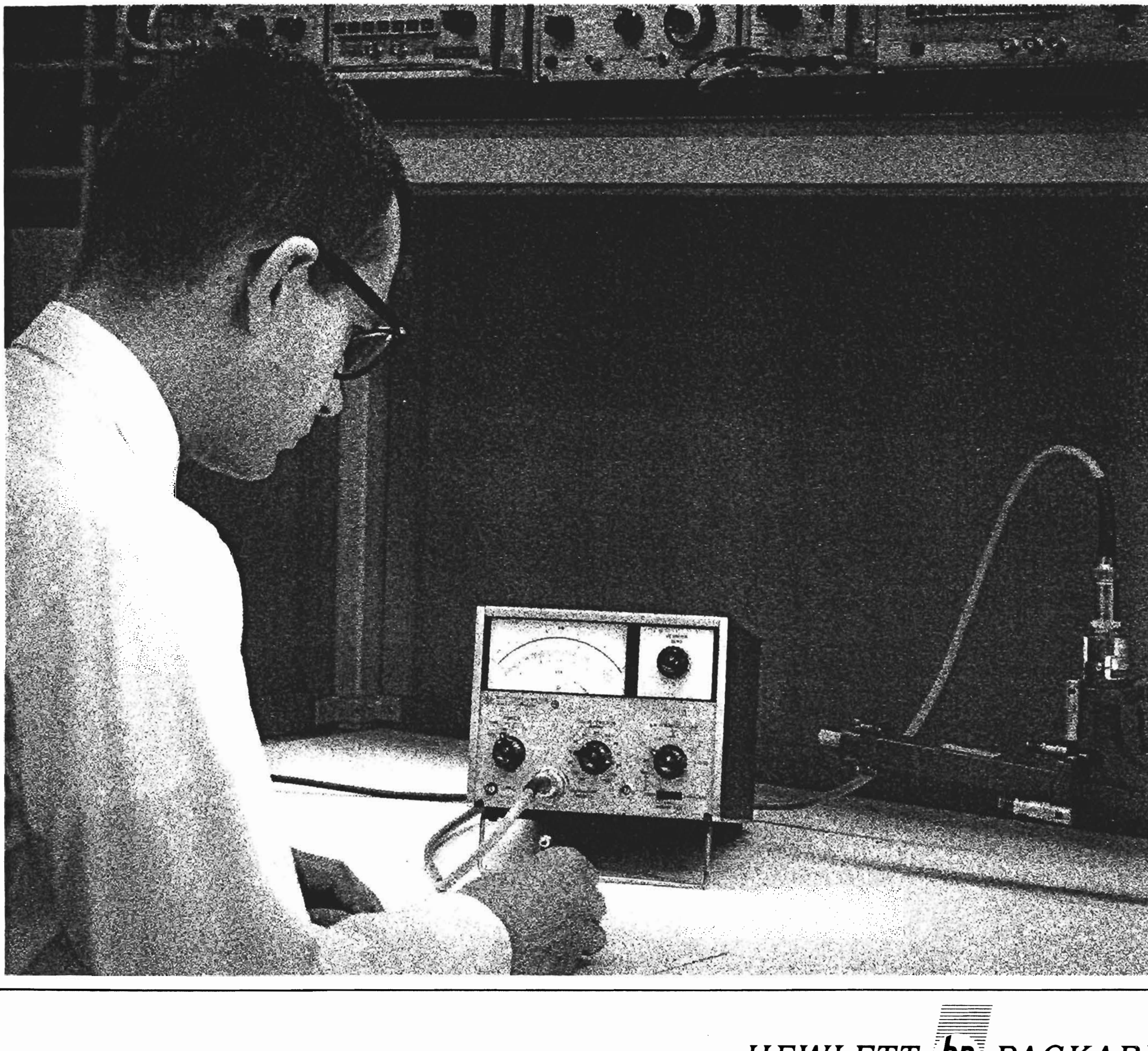
APPLICATION NOTE 64

Microwave Power Measurement



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APPLICATION NOTE 64

MICROWAVE POWER MEASUREMENT

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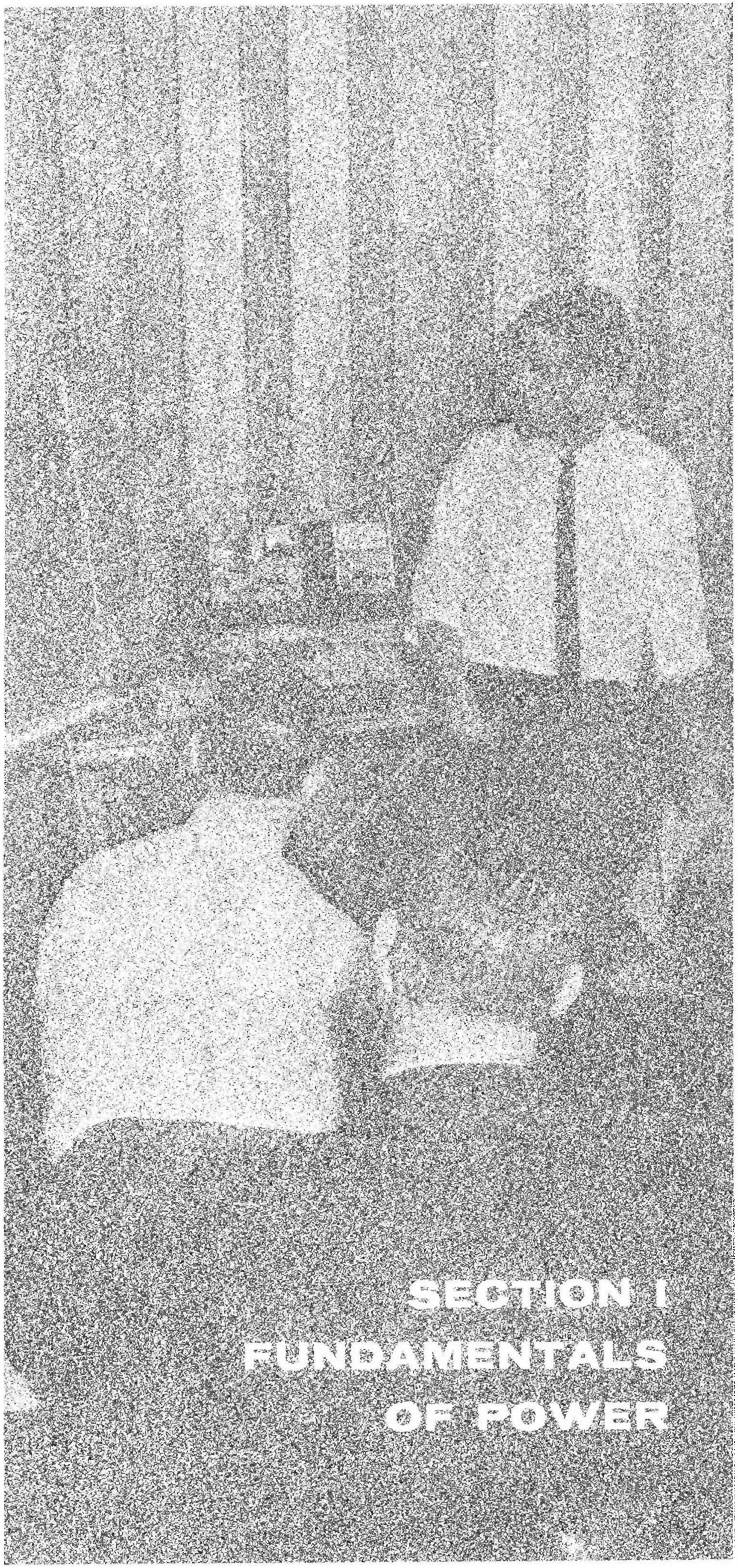


SECTION I
FUNDAMENTALS OF POWER

INTRODUCTION

The term "power" is used to describe the rate at which energy is made available to do work. In the field of microwave, this "work" is usually the transmission of aural, visual, or coded (radar) intelligence over a given distance. Other examples of work accomplished by microwave power include the excitation of molecules in a medium to produce heat, or the acceleration of particles for nuclear studies.

From an economic standpoint we are interested in measuring power to determine the cost of work performed in terms of energy expended. In applied science, power measurement establishes a means for evaluating the work capability and efficiency of a given device. The continuous effort to design more efficient systems and increase the work capability of microwave devices leads to higher accuracy requirements in the measurement art. Along with this, greater measuring speed is desirable so advancements are not limited by impractical test methods. This Application Note will define power, discuss instruments available for microwave power measurements and show typical equipment setups with emphasis on modern techniques, accuracy considerations and sources of error.

Considerable information regarding power has been published over the years, some of which discusses the subject in great detail. The purpose of this Note is to present a composite reference of techniques and Hewlett-Packard instruments for those interested in measuring microwave power, with a bibliography for those wanting greater detail in a specific area.

WHY POWER IS BASIC AT MICROWAVE FREQUENCIES

Voltage and current measurements are basic at low frequencies and DC because they can be made conveniently with high accuracy. Voltage, current, and resistance are fundamental electrical standards whereas power is a derived term of necessarily lower accuracy. At low frequencies, power may be calculated from voltage and impedance measurements using Ohm's Law,

$$P = \frac{E^2}{Z} \cos \phi.$$

At microwave frequencies, we are faced with distributed, circuit constants and transmission line lengths that are appreciable fractions of a wavelength or more. Impedance at any given measurement point is not easily determined and any impedance mismatch between source and load sets up standing waves along the transmission line such that voltage measurement becomes arbitrary.(1) Figure 1-1 illustrates how voltage and current vary on a mismatched line. Since power remains invariant with position in a lossless line, the practical method of determining energy available in a microwave circuit becomes one of power measurement.

(1) See "Microwave Theory and Measurements", Eng. Staff, Hewlett-Packard, Prentice-Hall, 1962, pp 7-21.

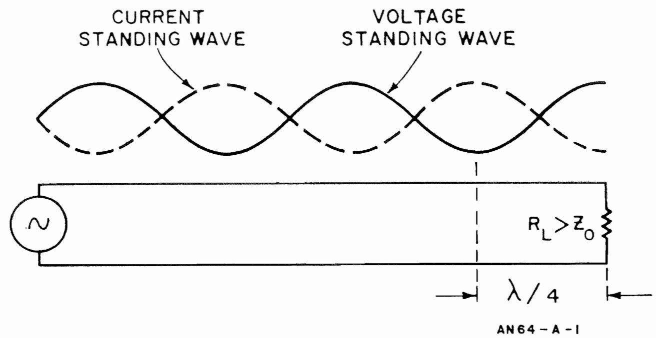


Figure 1-1. Mismatched transmission line exhibits standing waves, making voltage or current measurements arbitrary. At microwave frequencies, energy levels are more readily determined by power measurment which is independent of position (lossless line).

DEFINITIONS

Two types of power measurement are common in microwave work. Many applications require a knowledge of average power in a device while others require peak power of a periodic pulse. Before describing the instruments and methods for measuring these, it might be well to briefly review the definitions of each.

Average Power:

Figure 1-2 shows a load resistor, connected to a source producing a sinusoidal output voltage. By definition, the power in R_L at any instant in time is the product of e and i. From this we see that instantaneous power varies at a rate twice the frequency of the generator. Instantaneous power does not relate the amount of work done over any interval in time, which is of primary importance in most cases. We therefore define power with a time dimension in the following equation:

$$P = \frac{1}{T} \int_0^T ei\, dt \tag{1.1}$$

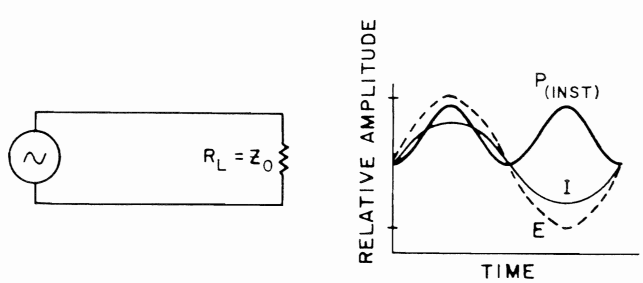


Figure 1-2. Instantaneous power in load resistor R_L varies at a rate twice the applied generator frequency. Average power is the integral of instantaneous power averaged over one or more cycles.

where

$$\begin{aligned} P &= \text{average power} \\ T &= \text{time} \\ e \text{ and } i &= \text{instantaneous values of voltage and current} \end{aligned} \qquad (1.1)$$

This simply states that the average power over a given time period is the integral or summation of all instantaneous values of power divided by the period.

With a CW source, the shortest time of interest is generally the period of one cycle of output signal. The average power in this period is the same as for any integral number of cycles, provided steady-state conditions exist. Assigning the period of one cycle to the general equation (1.1), we may write:

$$P = \frac{1}{2\pi} \int_0^{2\pi} E_p \sin\theta \cdot I_p \sin(\theta - \phi)\, d\theta \qquad (1.2)$$

where

E_p and I_p = peak value of voltage and current
θ = phase angle of I with respect to E

Performing the indicated operation results in the expression:

$$P = \frac{E_p I_p}{2} \cos\phi \qquad (1.3)$$

For sinusoidal waveforms,

$$E_p = \sqrt{2}\, E_{rms} \text{ and } I_p = \sqrt{2}\, I_{rms}$$

which may be substituted into equation (1.3) giving us the familiar expression for average power in an AC circuit;

$$P = E_{rms} \cdot I_{rms} \cos\phi . \qquad (1.4)$$

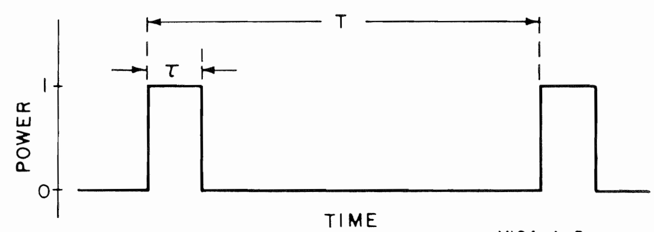


Figure 1-3. Peak power of periodic rectangular pulse is readily defined by average power-duty cycle relationship. Pulse width (τ) and repetition rate ($\frac{1}{T}$) is customarily measured with reference to 50 percent amplitude points.

Peak Pulse Power:

Peak pulse power may be defined in one way as the average power during the time the pulse is on. Figure 1-3 shows the envelope of a periodic rectangular pulse of rf energy of width τ and period T. The general integral (1.1) for average power may be used again to define peak power in this case by assigning pulse width as the limit and constant in the integral as follows:

$$P_{pk} = \frac{1}{\tau} \int_0^{\tau} ei\, dt \qquad (1.5)$$

where

τ = pulse width at 50% peak amplitude.

It should be noted that any aberrations in the pulse, such as overshoot or ringing, are averaged out by this definition and their instantaneous peak values cannot be considered as peak pulse power. Peak power has commonly been defined by an extension of equation 1.5 since the early days of microwave. Early radar applications of pulsed microwave power involved only the use of rectangular pulses which led to the expression

$$P_{pk} = \frac{P_{ave}}{\text{Duty Cycle}} \qquad (1.6)$$

where

Duty Cycle is $\frac{\tau}{T}$.

Continued development of microwave applications and accuracy improvement has required better definitions since many times the pulse is intentionally not rectangular or, because of aberrations, does not allow accurate determination of τ. For example, consider the case of a TACAN (Tactical Air Navigation) pulse which is gaussian as shown in Figure 1-4. Considerable difference exists between the "peak" power at the pulse's maximum amplitude and peak power as defined in equation 1.6. A constant of correction can be calculated or determined with a planimeter for many non-rectangular pulses, however this is a time consuming task. In applications involving non-rectangular pulse shapes, peak pulse power may be defined as

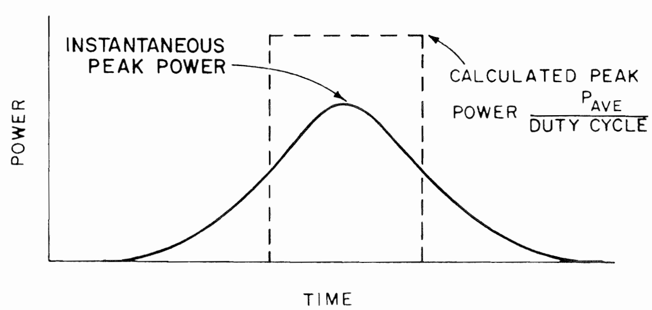


Figure 1-4. Gaussian pulse such as used in TACAN and DME applications illustrates different meanings of "peak power". Complex pulse shapes require better definition of peak power than rectangular pulses.

$$P_{pk} = \frac{e_p^2}{R} \tag{1.7}$$

where

R = a constant load resistance absorbing the power

e_p = Peak voltage across a constant resistance, R.

This expression does not rely on pulse shape for an accurate determination of the pulses' maximum amplitude point. This is especially useful for voltage breakdown checks. Since peak pulse power may have different meanings depending upon the application, the method of measurement should be flexible enough to allow a choice as to what is considered the peak power of a pulse. Such methods are discussed at the end of Section II.

SECTION II

INSTRUMENTS FOR MICROWAVE POWER MEASUREMENT

The fundamental standards of microwave power lie in DC or low frequency -AC Power which may be used for comparison or substitution, then accurately measured. The instrumentation used for such comparison or substitution must be efficient in design and properly employed for minimum loss of time and accuracy in the transfer. As might be expected, these criteria often impose divergent requirements so the approach must be one of compromise. This section describes various instruments for measuring microwave power and summarizes their advantages and limitations. Accuracy considerations for bolometric and calorimetric power measurements are also included in this section because of their direct relation to instrument design.

BOLOMETRIC METHOD

Microwave power up to about 10 mw average is usually measured by bolometric means. The basis for bolometric power measurement is that when power is dissipated in a resistive element, a corresponding change occurs in the element's resistance. By proper construction of the element, the heating effect of DC and microwave power will be nearly the same, and the resistance change may be sensed by associated circuitry to indicate power. The resistive elements are termed "bolometers" and fall into two classifications: 1) Barretters, which consist of a short length of fine wire suitably encapsulated or a strip of thin metallic film deposited on a base of glass or mica, and 2) Thermistors, which are made of a compound of metallic oxides. Barretters exhibit a positive temperature coefficient, i. e., as more power is dissipated in the element its resistance becomes greater and vice versa. Thermistors have a negative temperature coefficient and are physically and electrically more rugged than barretters. Figure 2-1 shows the resistance versus power characteristics of a typical wire barretter and bead thermistor. More detail is given later on bolometer characteristics but for now we need only the description given so far to explain the basic operation of a bolometric power meter.

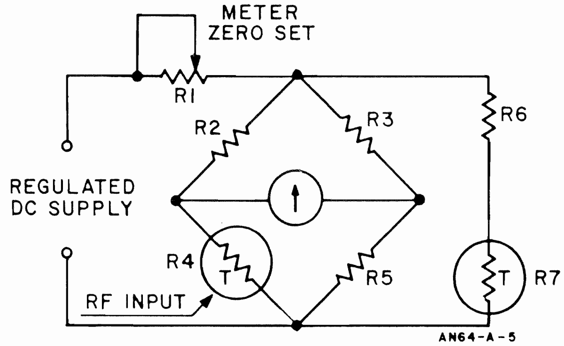


Figure 2-2. Unbalanced thermistor bridge provides simple power monitor for limited range and accuracy requirements.

UNBALANCED BRIDGE.

One of the simplest methods for bolometric power measurement is to place a bolometer (usually a thermistor) in one leg of a wheatstone bridge as in Figure 2-2. The bridge is excited by a regulated DC supply whose amplitude may be adjusted with R-1. Since R-4 is a thermistor, its resistance may be controlled by the amount of current allowed to pass through it. In operation, R-1 is adjusted until just enough current passes through the bridge to make the thermistor's resistance equal to R-5 bringing the bridge into balance and causing the meter to read zero. Microwave power is then applied to the thermistor and the heating effect causes the thermistor's resistance to decrease, unbalancing the bridge in proportion to the power applied. The unbalance current is indicated on the meter which is calibrated directly in milliwatts.

The bolometer resistance curves shown in Figure 2-1 vary with temperature, requiring some form of temperature compensation to minimize meter drift. Therefore, a disc-shaped thermistor R-7, is placed in close thermal proximity with R-4 but isolated from the microwave power. This is usually done by placing the disc against the outside wall of the thermistor mount for maximum heat transfer. The temperature charac-

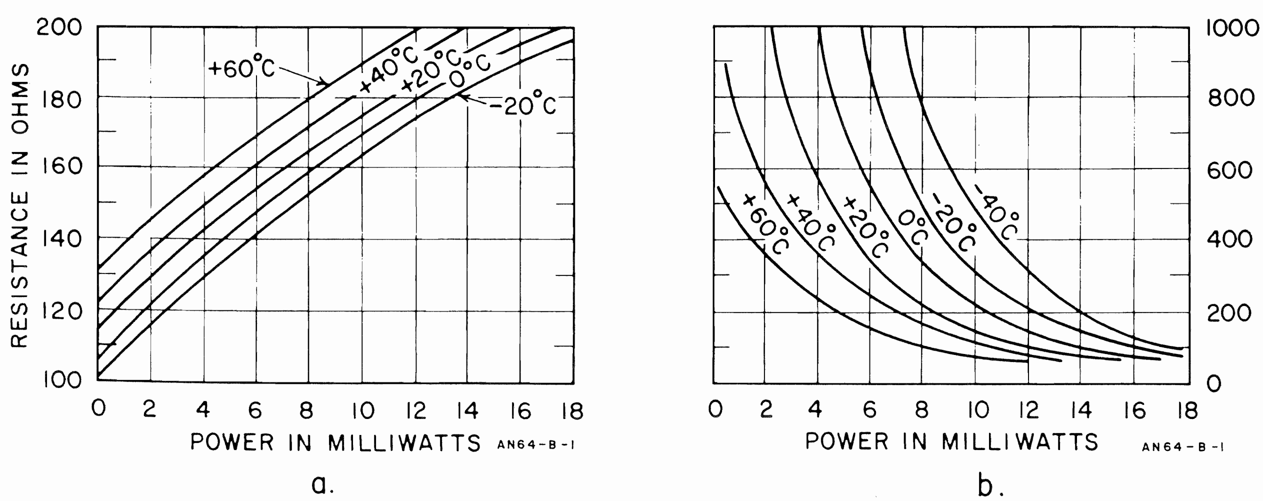


Figure 2-1. Resistance vs. dissipated power in (a) typical wire barretter (b) typical bead thermistor, at various ambient temperatures.

teristics of R-7 are chosen to be as nearly equal to R-4's as possible so that as ambient temperature increases, for example, resistance decreases proportionately in both thermistors. The reduction in R-4's resistance is seen as an unbalance in the bridge as if more power were being applied to R-4. The reduction in R-7's resistance results in more current shunted away from the bridge thus lowering the dc power in R-4. This action causes R-4's resistance to increase, compensating for the temperature change and restoring bridge balance.

This scheme works well over limited temperature variations but depends upon the R-P curves of both thermistors being the same which is seldom the case. Also, the system has slow response because of the time required for the thermistor mount and compensating disc to reach equilibrium with the measuring thermistor's temperature. With a full scale range of 2 mw, meter drift over an extended period of time might be as high as half scale in a typical unbalanced bridge with this form of temperature compensation.

Note in Figure 2-1(b) that the thermistor's resistance is only linear with applied power over a very small range. This range extends for about a 2 mw increment of power in the knee of the R-P curves, thus limiting dynamic range. At an operating temperature of 40°C a "bias" power of approximately 6 mw is required to place the thermistor in the linear portion of its R-P curve. Thermistor resistance at this point is about 200 ohms so this value is chosen for the other resistors in the bridge. As power in excess of 8 mw (6 mw dc bias + 2mw microwave) is applied, thermistor resistance change for a given power change becomes less and resolution is drastically reduced.

One might ask the question, "why not bias the thermistor at a much lower power and take advantage of the very steep portion of the R-P curves for higher sensitivity"? The meter could be scaled accordingly and very good resolution would be possible. There are several reasons why this is not practical: 1) As the bias power is reduced, bridge driving power is also reduced, lowering bridge sensitivity even though detector sensitivity is increased; 2) Ambient temperature variations represent a larger proportion of the thermistor's operating temperature resulting in greater effects on bridge balance; 3) Impedance match of the thermistor to microwave transmission lines is difficult at high bolometer operating resistances; 4) Bridge balance may become impossible at elevated temperatures.

One consideration in the accuracy of the unbalanced bridge is the fact that bolometer resistance varies with the level of microwave power applied. This changes the impedance of the bolometer mount resulting in mismatch error(2) since part of the microwave power is reflected, rather than absorbed in the bolometer element. Mismatch error due to a 2:1 resistance change in the bolometer can be as high as 0.5 db.

(2)Mismatch error is covered in more detail under "Accuracy Considerations".

In summary, then, the general characteristics of the unbalanced thermistor bridge are:

a. Simple design and operation.
b. Dynamic range limited to about 2 mw.
c. Subject to drift with temperature variations.
d. Limited accuracy.

For these reasons, the unbalanced bridge is generally used where simplicity and convenience is desired for setting a reference power level in the 1 to 10 mw range. This method is used for example as the output power monitor for signal generators and radar test sets.

MANUALLY BALANCED BRIDGE.

The manually balanced bridge provides a means for direct substitution of DC or low-frequency AC power (which may be accurately measured) for an unknown microwave power. This method offers high accuracy and wide dynamic range but is much slower and less convenient than the unbalanced bridge. Manually balanced bridges have long been used in standards laboratories where high accuracy is of prime importance. Figure 2-3 is a simplified circuit diagram of the hp K04-999D manually balanced bolometer bridge. There are three modes of operation possible with this bridge, all involving substituted power: 1) DC voltage, 2) DC current, and 3) AC voltage. Bridge operation is as follows for each mode:

1. DC Voltage: S-1 is switched to the "on" position connecting an external voltmeter across the bolometer. Nothing is connected to the AF INPUT jack in this mode of operation. S-2 is closed, bypassing the milliammeter jack and connecting the bridge to the regulated DC power supply. R-2 is now adjusted for exact bridge balance as indicated by a sensitive galvanometer. As balance is approached, R-1 is adjusted for maximum galvanometer sensitivity. When balance is achieved, the DC voltage across the bolometer is noted as E_1. Microwave power is then applied to the bolometer, unbalancing the bridge. By removing an equivalent amount of DC power from the bridge with R-2, balance is restored and at this point the bolometer's DC voltage drop is again measured and noted as E_2. Assuming the DC and microwave power to have equal heating effects on the bolometer(3), the unknown microwave power may be calculated from the equation

$$P = \frac{E_1^2 - E_2^2}{R} \qquad (2.1)$$

where

R = bolometer resistance at bridge balance.

(3)DC-microwave substitution error is covered in this Section under "Accuracy Considerations".

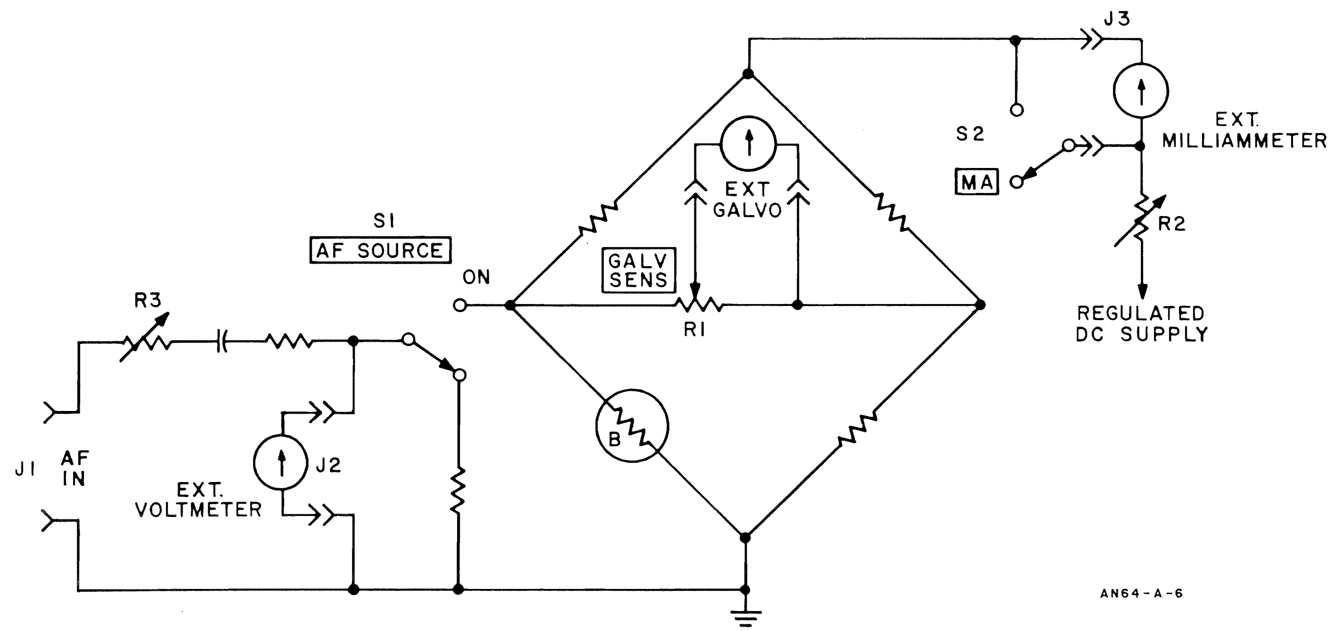


Figure 2-3. HP K04-999D manually balanced Bolometer Bridge operates with common 100 or 200 ohm bolometers to measure power up to 20-30 mw. Substituted power to bolometer may be furnished from internal DC supply or obtained externally from an audio oscillator. Accuracy of substitution is typically 0.5% to 5% depending primarily on external meters used.

2. DC Current: The procedure of balancing the bridge in the DC current mode is the same as used in the DC voltage mode. Switch S-2 is opened and bridge current noted on an external milliammeter at J-3. Microwave power is then calculated from the difference of the current readings as follows:

$$P = R(I_1^2 - I_2^2) \qquad (2.2)$$

where

I_1 and I_2 = one half the total bridge current without and with microwave power applied, respectively.
R = bolometer resistance at bridge balance.

3. AC Voltage: In this mode S-2 is closed, providing DC bias to the bridge for initial balance and by-passing the milliammeter jack. An accurate AC voltmeter is connected at J-2 and S-1 turned on. Microwave power is applied to the bolometer and the bridge is balanced with R-2. Now the microwave power is removed and sufficient audio power substituted from an audio oscillator at J-1 to restore bridge balance. At balance, the AF voltage at J-2 is read and power calculated:

$$P = \frac{E^2_{af}}{R} \qquad (2.3)$$

where

E_{af} = Substituted audio voltage
R = Bolometer resistance at bridge balance

Again assuming the substitution to be valid, the calculated audio power equals the microwave power dissipated in the bolometer. The AC mode is generally preferred when the microwave power to be measured is small compared to the required bias power. If the DC mode was used, measurement accuracy would depend upon taking a small difference between the squares of two relatively large voltages; viz E^2_1 and E^2_2 in equation (2.1). Even with a 0.1% accurate DC meter at J-2 the calculated difference in voltages could be in error by 7% for a 0.5 mw power measurement. With AC substitution and DC bias, the difference in substituted power is determined from the AC voltage only, which is first zero, then some small value, say 0.5 volts. This may be measured to an accuracy of typically 0.25% with laboratory AC voltmeters, resulting in 0.5% accuracy in substituted power.

Characteristics of the manually balanced bolometer bridge may be summarized as follows:

a. The bolometer always operates at one resistance, allowing about 20 db of dynamic range and a good impedance match to microwave.

b. Substituted power may be measured very accurately, providing direct reference of microwave power to known voltage and resistance standards.

c. The measurement is slow requiring at least four balance and read operations plus a calculation for each microwave power measurement.

d. The system is not temperature-compensated, requiring frequent zero adjustments and limiting measurement of very small amounts of power.

e. Several auxiliary instruments are required, including a galvanometer, high accuracy voltmeter or current meter, and an audio oscillator if AC substitution is desired.

AUTOMATICALLY BALANCED BRIDGE

In the automatically balanced bridge, AC substitution and readout is done automatically, eliminating all operations required in the manually balanced bridge except the zero adjustment for initial balance. The technique combines much higher accuracy and wider range than the unbalanced bridge with far greater convenience and speed than the manual bridge. These features have led to the automatic bolometer bridge long being the workhorse of the industry. The hp 430C is a typical example of this type of power meter. The 430C is entirely self-contained, with no auxiliary instruments required to measure microwave power other than the bolometer mount. The hp 430C will operate with thermistors or barretters requiring up to 16 ma of bias current to reach an operating resistance of 100 or 200 ohms. Full scale power ranges are from 0.1 mw to 10 mw in a 1, 3, 10 sequence. A simplified schematic of the hp 430C is shown in Figure 2-4.

To illustrate the 430C operation, we will assume that a thermistor mount is being used which requires a total of 28 milliwatts* at its ambient temperature to reach the operating resistance set by the circuit. This power may be DC, AF, microwave, or any combination thereof. In the 430C, a combination of audio and DC is initially applied to the mount. Whatever DC power is applied, the circuit automatically supplies the balance in audio power up to a total of, in this example, 28 milliwatts. Any microwave power applied then automatically reduces the audio power by a like amount to maintain the total constant. A VTVM then measures this reduction and indicates it as applied power.

Referring to Figure 2-4, it may be seen that feedback from the amplifier to the bridge is positive for one side of the bridge and negative for the other. The positive feedback is temperature-sensitive, depending on the resistance of the thermistor. Any increase in feedback power, for example, decreases the feedback factor. The negative feedback, on the other side of the bridge, is frequency-sensitive, and reaches a minimum value at the resonant frequency of the tuned circuit L-1, C-1. When the circuit is first turned on, the thermistor will be cold, its resistance high, and the positive feedback very large. Oscillations will immediately start up, putting power into the thermistor and reducing its resistance until the positive feedback exceeds the negative by a very small margin. This condition occurs at the resonant frequency, where the negative feedback is a minimum. The degree by which the positive feedback exceeds the negative is set by the gain of the circuit and is made very small by making the gain high.

*Typical of dual element thermistor mounts.

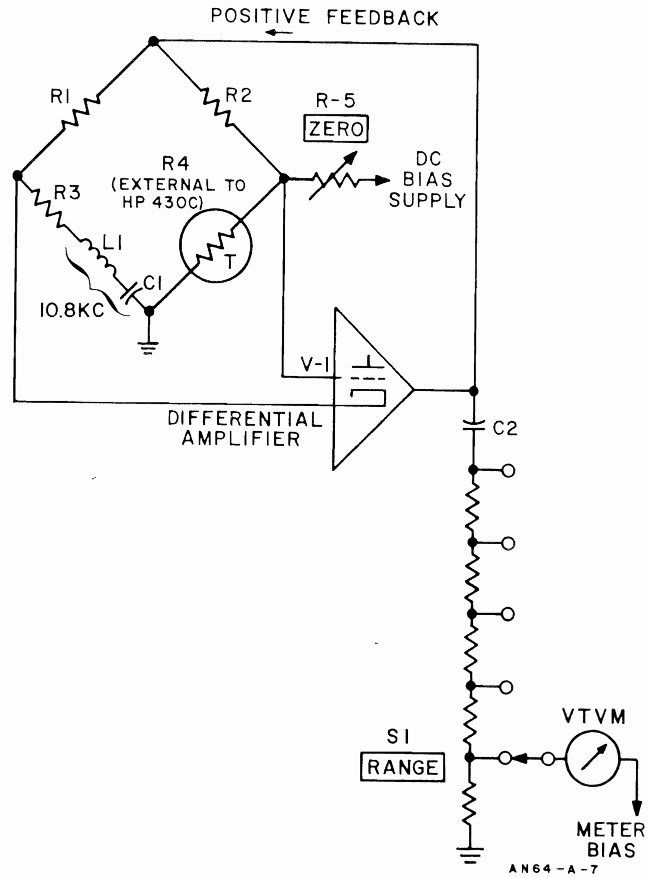


Figure 2-4. Simplified schematic of hp 430C Power Meter shows bridge, feedback, and metering system for automatic power measurements of 0.1 mw to 10 mw. Accuracy of substituted power is ±5%.

Now if DC or microwave power is added to the thermistor, the thermistor resistance tends to decrease. This tendency automatically decreases the net positive feedback, however, causing the oscillations to decrease by an amount exactly equal to the applied DC or microwave power. The small degree of bridge unbalance is always preserved and the thermistor resistance does not change.

In order to indicate power on the VTVM meter scale, it is necessary to start from a known reference condition. If there were, say 20 milliwatts of audio power in the bolometer and 5 milliwatts of RF were added, a certain scale would be required to operate between audio power of 20 and 15 milliwatts. If the audio decreased instead from 10 to 5 mw, a different scale would be required. A simple gain adjustment in the VTVM would not suffice, because of the square-law relation between the power and the rectified voltage actually applied to the meter. Hence, a DC bias supply is incorporated into the instrument. Enough DC power is added to reduce the audio power to a convenient reference value. In the 430C this is arranged to be 1.2 times the full-scale value on each range. Thus, on the 10-milliwatt range, the DC is adjusted so there is exactly 12 milliwatts audio in the bolometer, and the meter circuit is calibrated to read zero. When 10 milliwatts RF is applied, the audio drops to 2 milliwatts and the circuit is calibrated to read full-scale.

This 1.2 to 0.2 relation holds on all ranges and the meter scale is accurate at all points.

Table 1 illustrates how combinations of power are used to balance a typical bolometer in the hp 430C bridge for zero and full scale indications on the 10 mw range.

Table 1

Power to Bolometer	hp 430C at ZERO	hp 430C at FULL SCALE
Microwave	0 mw	10 mw
DC	16 mw	16 mw
Audio	12 mw	2 mw
Total	28 mw	28 mw

Since it is desirable to have the meter read upscale as RF power is applied, a residual DC current is applied to the meter movement in the forward direction and the rectified audio in the reverse direction.

Range-switching is accomplished by changing the sensitivity of the VTVM to the audio level. As lower ranges are selected, VTVM sensitivity is increased and the DC bias increased to reduce the audio level. Eventually, drift becomes a problem since instability in the DC supply becomes magnified. In addition, any change in the ambient temperature of the bolometer changes its total power requirement, causing the audio level to change. If this occurs during the course of a measurement, an error results, since the temperature change is indicated along with the RF power. Dynamic range is normally limited to about 20 db.

Characteristics of the hp 430C automatically balanced bolometer bridge may be summarized as follows:

a. Measurement is rapid and no auxiliary instruments are required other than the bolometer mount.

b. The bolometer always operates at one resistance, making possible a good impedance match at all power levels within the dynamic range of the system.

c. Accuracy of <u>substituted</u> power is better than 5% of full scale which is primarily determined by the VTVM circuit. (Other sources of error external to the 430C will be covered under "Accuracy Considerations".)

d. Dynamic range is about 20 db, limited by bolometer burnout at the high end and zero drift on the low power end. On the lowest range of the 430C, which is 0.1 mw full-scale, slight ambient temperature variations at the bolometer cause large variations in meter reading. This is due to the inability of the system to distinguish between applied power changes and environmental temperature variations at the bolometer element. This is true of <u>all</u> non-temperature compensated bolometer bridges.

BOLOMETER MOUNTS FOR MANUAL AND AUTOMATIC BRIDGES

So far, we have only discussed the instrumentation portion of the bolometer bridge in detail. Before introducing temperature compensated automatic bridges it would be well to complete the discussion on uncompensated systems with more detail on bolometers and how they are mounted for practical use.

Bolometer elements are mounted in either coaxial or waveguide structures so they are compatible with common transmission line systems used at microwave and RF frequencies. The bolometer-mount combination must be designed to satisfy four important requirements so the bolometer element absorbs as much of the power incident to the mount as possible. In this regard the mount must: 1) present a good impedance match to the transmission line over the frequencies of interest, 2) keep I^2R and dielectric losses within the structure minimized so that power is not dissipated in electrical contacts, waveguide walls or insulators, 3) provide isolation from thermal and physical shock and 4) keep leakage small so microwave power does not escape from the mount in a shunt path around the bolometer. Shielding is also important so that extraneous power does not enter the mount. One other consideration should not be overlooked, although it is usually accounted for by other mount construction requirements. The mount should have enough mass to minimize the effects of ambient temperature variation.

Bolometer mounts may be sub-divided into tunable, fixed tuned, and broadband untuned types. Some mounts allow convenient field replacement of the bolometer elements while others have their elements soldered or spot welded in place, often requiring factory replacement should they be damaged. Figure 2-5 shows a variety of HP bolometer mounts which may be used with manual or automatically balanced bridges such as the K04-999D and 430C.

BROADBAND COAXIAL THERMISTOR MOUNT

The hp 477B is a broadband, untuned thermistor mount designed to operate from 10 Mc to 10 Gc with a maximum SWR of 1.5:1, (less than 1.3:1 from 50 Mc to 7 Gc). Two thermistor beads are connected in the coaxial mount as shown in Figure 2-6. Connection to a bolometer bridge such as the 430C is made through J-2. The bridge provides approximately 30 mw of bias power to the two thermistors in series reducing their resistance to 100 ohms each. Total resistance presented to the bridge is thus 200 ohms since reactance of C1 and C2 is high at the DC and 10.8 kc bias frequency of the 430C. When microwave power is applied at J-1, it passes unattenuated through the DC blocking capacitor C-1 to the junction of the two thermistors. Since C-2 is a low impedance at microwave frequencies, the two 100 ohm thermistors appear to be in parallel, presenting a combined resistance of 50 ohms necessary for impedance matching to the coaxial transmission line. This design was chosen primarily because DC return to the bridge would be difficult over broadbands using a single thermistor. All components are shielded by the coaxial

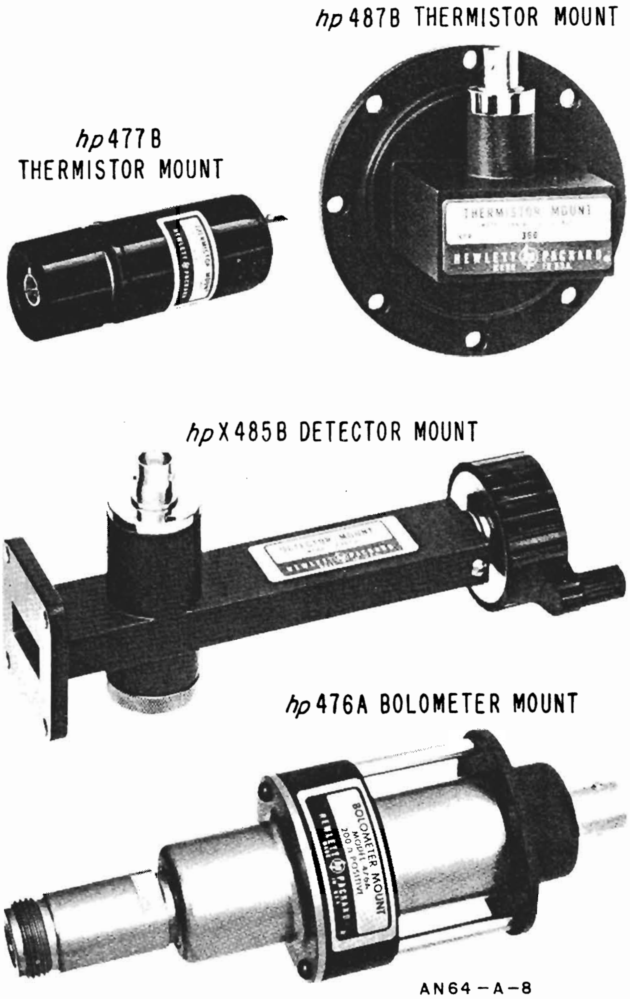


Figure 2-5. Variety of coaxial and waveguide bolometer mounts which may be used with hp K04-999D or 430C Power Meter.

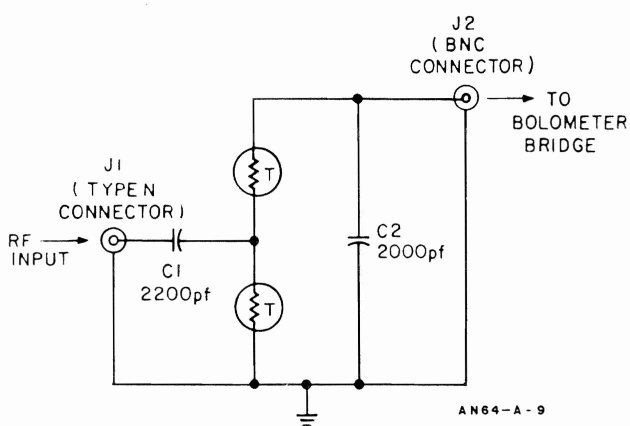


Figure 2-6. HP 477B Dual-element coax thermistor mount operates over 10 Mc - 10 Gc frequency range. SWR is 1.3 or less over most of the band. Circuit arrangement results in 50 ohm impedance match to RF input, and 200 ohm resistance to bolometer bridge.

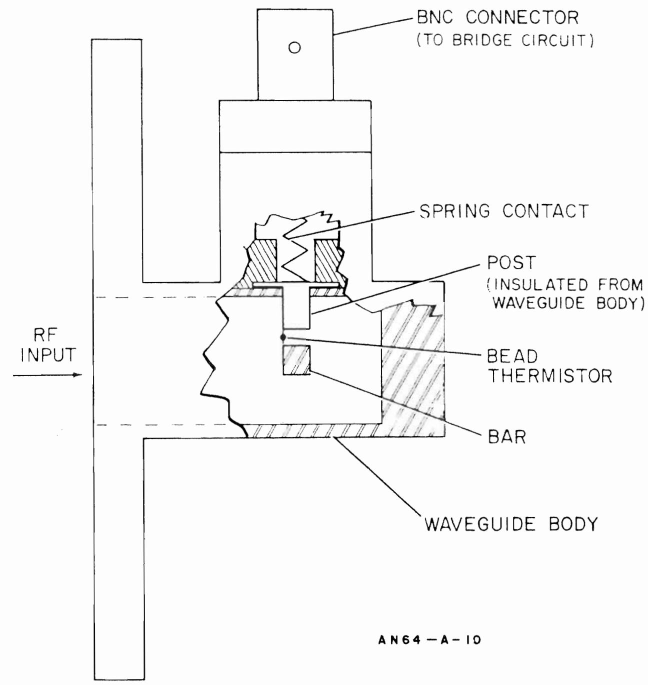


Figure 2-7. Cutaway view of hp 487B broadband waveguide thermistor mount shows post and bar arrangement for mounting thermistor. Configuration enables good impedance match over waveguide band without tuning.

housing of the mount to prevent extraneous pickup and radiation. The thermistor elements are field replaceable using kits available through Hewlett-Packard Service Centers.

The hp 476A is a similar coaxial mount using four instrument fuses as barretters instead of thermistors. The 476A has long been used for measurements in the 10 Mc to 1 Gc range. This mount as well as the 477B operate with the 430C at power inputs up to 10 mw.*

UNTUNED WAVEGUIDE BOLOMETER MOUNTS

Waveguide bolometer mounts are used for power measurements in waveguide systems to avoid the losses and mismatch errors caused by waveguide-to-coax adapters. Figure 2-7 shows a cutaway side view of the hp X487B untuned broadband waveguide thermistor mount. It consists of a shorted section of fabricated brass waveguide with a thermistor bead mounted on a post and bar arrangement in the center of, and parallel to, the E field. The thermistor is biased to an operating resistance of 100 ohms by applying DC and/or AF bias current through the BNC jack shown in the figure. The size and location of the post and bar are designed to present a conjugate impedance match to the shorted waveguide section over a wide range of frequencies. This arrangement provides broadband impedance matching of the mount for low SWR.

The hp 487 waveguide series of broadband thermistor mounts cover the frequency range of 2.6 Gc to 40 Gc with a maximum SWR of 1.35:1 and 1.5:1 in the 2.6 to

*The 477B is capable of measuring up to 20 mw with suitable manual bridges such as the K04-999D.

18 Gc range, and 2:1 in the 18 to 40 Gc bands. Average power up to 10 mw may be measured directly with these mounts connected to a suitable bridge. Although it is not recommended, these thermistors will withstand considerable overloads without burnout. As excess power is applied, the bridge can no longer remove enough bias power to maintain bridge balance. The thermistor therefore decreases in resistance rapidly, causing a large impedance mismatch to reflect a portion of applied power which prevents its being dissipated in the bead. Accidental overloads of 25 to 30 mw will not harm the element.

TUNABLE WAVEGUIDE BOLOMETER MOUNTS

The hp 485B-series of tunable detector mounts cover the frequency range of 3.95 Gc to 12.4 Gc and are designed for use with Sperry type 821 or Narda N 821 encapsulated barretters. The mounts may also be used with suitable crystal diodes for detection of modulated RF signals. Figure 2-8 is a cutaway side view of the hp X485B with a barretter installed. The barretter is positioned in the center of, and parallel to the E field in the guide the same as the untuned mount previously described. The barretter's lower contact is made to ground through a screw-in plug which allows easy removal of the barretter element for replacement. The upper contact is insulated from the guide and passes through a shunt capacitance to minimize RF leakage and allow connection to the bias and bridge circuit. The movable shorting block is highly conductive and has silver contact fingers which ensure a low loss short across the guide. The block is positioned at any odd multiple of a quarter wavelength behind the barretter element at the operating frequency. It is best to use the position closest to the element which still allows peaking of the indicated power since this minimizes I^2R losses in the waveguide. When the plunger is properly tuned, maximum power is absorbed by the barretter and the mount presents a good impedance match to the line. SWR is specified to be better

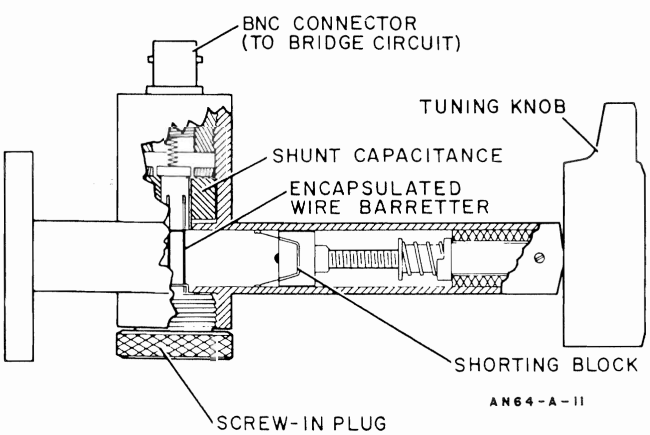


Figure 2-8. HP 485B Detector Mount uses encapsulated wire barretter for power measurements in waveguide bands from 2.6 to 12.4 Gc. Moveable shorting block is adjusted for best impedance match at test frequency. Screw-in plug provides easy access for changing barretter or installing crystal detector for other applications.

than 1.25:1 in most waveguide bands when the unit is properly tuned. The small reduction in SWR offered by tunable mounts over broadband mounts is generally not significant enough to warrant the tuning operation required for each frequency change or measurement location change. When the best possible impedance match is required for maximum accuracy, a slide screw or E-H tuner is used ahead of the bolometer mount. The chief advantage of using the 485B is that it employs a barretter rather than a thermistor. From the curves in Figure 2-1 we see that ambient temperature variations have much less effect on a barretter than on a thermistor in non-temperature compensated systems. For this reason bridge zeroing is less frequent using the 485B instead of the 487B. The disadvantages of using a barretter are 1) easy burnout 2) mechanically delicate 3) fast thermal time constant can cause error in average power measurements of AM signals.

TEMPERATURE COMPENSATED POWER METERS

In the preceding balanced bridge power measuring systems, there has been no way to distinguish between ambient temperature variations and applied power changes at the bolometer. The temperature compensation described for the unbalanced bridge is satisfactory for slow temperature changes in relatively insensitive bridges. With increased sensitivity, the problem becomes more serious until ultimately "zero drift" prevents making a measurement. If power is to be measured in the microwatt region, some effective means of temperature compensation must be employed to make highly sensitive bridges practical.

HP 431C Power Meter:

The hp431C is an automatically balanced, dual bridge power meter designed to operate with hp 478A and 486A-series temperature compensated thermistor mounts. Power may be measured with these mounts, in 50 ohm coaxial systems from 10 Mc to 10 Gc, and in waveguide systems from 2.6 Gc to 40 Gc. Zero drift is typically less than 1/100th of that in non-compensated systems allowing measurements as small as 10 microwatts full scale. Many other advantages result from temperature compensation which will be covered in Section III. First let's see how the system solves temperature variation problems.

Figure 2-9 shows the 431C block diagram in two halves to facilitate description of circuit operation. On the left is the RF Substitution bridge with the RF detection thermistor R_d connected. On the right is the compensation and metering bridge with the compensating thermistor R_c connected. Both R_d and R_c are contained in the same thermistor mount with R_d being mounted to intercept applied microwave power. R_c is electrically isolated from R_d and the microwave power, however the two elements are physically and thermally in very close proximity. Thus any change in mount temperature affects both thermistors identically, but only R_d is exposed to RF.

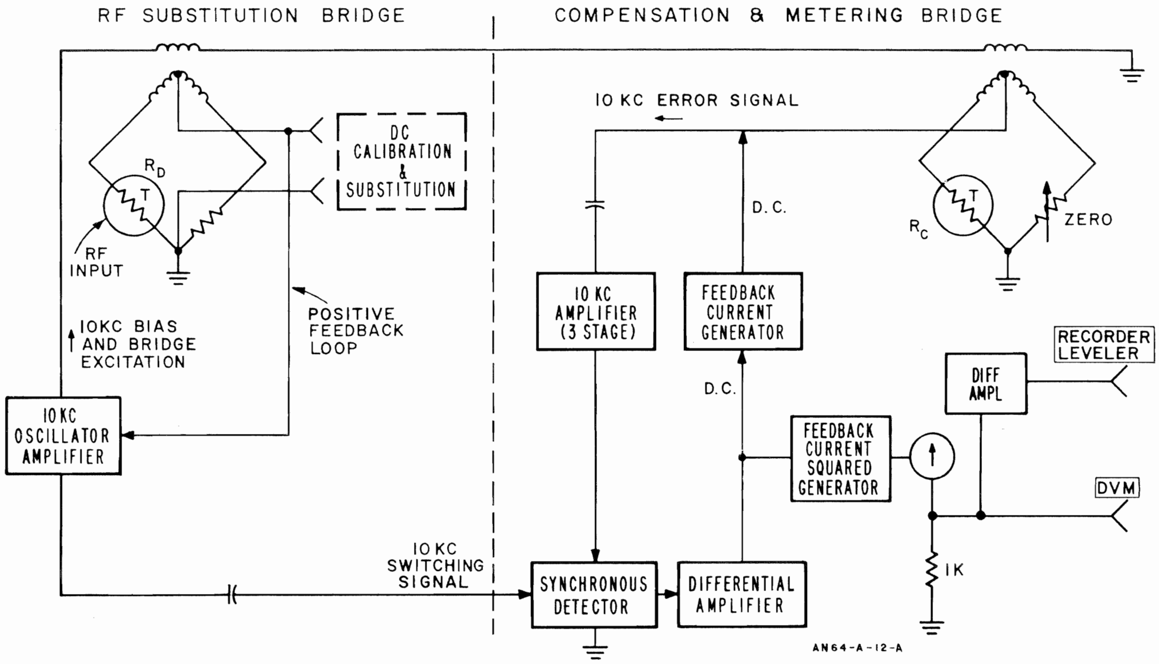


Figure 2-9. Block diagram of hp 431C Temperature Compensated Power Meter. Dual bridge design differentiates between changes in applied power and unwanted temperature variations at the thermistor mount. Used with hp 478A or 486A compensated thermistor mounts, this automatic power meter provides useful ranges from 10 μwatts to 10 mw with a single zero and null adjustment.

Each thermistor forms one leg of two separate bridges which are excited by a common 10 kc oscillator-amplifier through transformers T101 and T102. The action of the RF bridge is such that it is self-balancing, similar to the hp 430C bridge described earlier, except that in normal operation only AC bias is applied to R_d. Initially, the ZERO control is adjusted until balance is also reached in the metering bridge. At balance, minimum 10 kc error signal is applied to the 3-stage amplifier so the synchronous detector and differential amplifier outputs are minimum and the meter indicates zero. The circuit is arranged to avoid perfect balance since this would result in zero DC feedback to the metering bridge. With zero feedback, the metering loop would appear open. With an open meterint loop, the circuit would be unstable and a zero reference could not be accurately established. The bridge is therefore operated at a point very near balance and compensation made in the current-squared generator for zero indication on the meter

When microwave power is applied to R_d, the RF Substitution bridge tends to unbalance and the 10-kc oscillator backs off an equal amount of bias power to maintain bridge balance. Since the primaries of T101 and T102 are in series, the metering bridge feels an equal reduction in bias power. R_c is no longer biased for bridge balance and an error signal is fed to the 3-stage amplifier and synchronous detector. The pulsating DC from the detector is filtered and amplified by the differential amplifier and fed back to the metering bridge to rebalance it. Part of the DC feedback current is squared and applied to the meter to indicate the equivalent microwave power applied.

The squaring circuit allows a linear calibration of power on the meter scale since the power is proportional to bridge current squared. The 0 to 1 ma DC current to the meter movement is made available at a rear panel jack on the instrument for operation of a recorder or other auxiliary circuit where a current proportional to input power is useful. Automatic power leveling of sweep oscillators, such as the hp 690-series, is an important application utilizing the recorder output current provided by the 431C.*

Now that we have described how microwave power is sensed and measured by the 431C system, let's change the ambient temperature around the thermistor mount and see how it is compensated. After the meter is zeroed, suppose the temperature increases. Since both thermistors are in equal thermal environments, and of identical types, their resistance will decrease by an equal amount. The RF bridge attempts to unbalance, causing a reduction in 10-kc bias power applied to T101 and T102 which brings the RF bridge and R_d back into balance. The reduction in bias necessary for R_d to rebalance in the RF bridge is sensed by R_c in the compensating bridge as being just enough reduction to compensate for the heating effect of the ambient temperature increase. Thus the metering bridge remains in balance and no error signal is fed to the 3-stage amplifier. Consequently no change in DC feedback occurs and the meter remains at zero. The same compensation would take place if the meter were measuring power and an ambient temperature change oc-

*See HP Application Note #65, "Swept-Frequency Techniques".

curred, since the system is able to distinguish between RF and mount temperature variation.

The synchronous detector is a solid-state circuit used to minimize common-mode signals and noise entering the metering circuit. Substituted power equivalent to RF power sensed by the thermistors, is automatically measured to an accuracy of 3% of full scale and read on a mirror-backed taut-band meter calibrated in mw and dbm. Power ranges are in 5-db steps from 10 μw to 10 mw full scale. Range change is accomplished by changing the gain in the 3-stage amplifier and the feedback current generator. With this arrangement the meter only needs zeroing on the most sensitive range to be used and it will remain zero within 0.5% when switched to any of the other ranges. This capability is called "zero-carryover" and is a valuable convenience when successively measuring several power levels which differ greatly. Zero carryover also allows use of the power meter as an attenuator in power leveling setups such as described in Section III and IV.

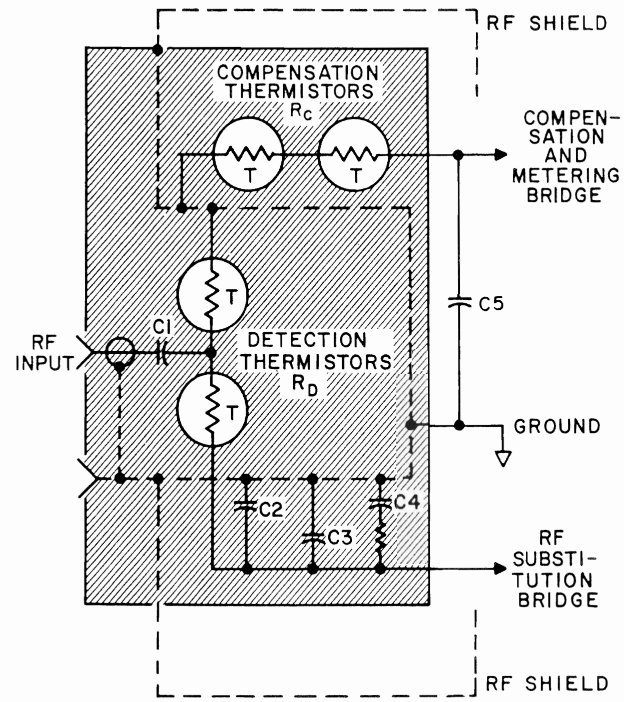
For greater accuracy in measuring substituted power, the DC Substitution and calibration feature of the 431C may be used. This circuit provides access to the RF thermistor bridge for applying DC from an external supply. With suitable instrumentation, the substituted DC power may be measured to an accuracy of 0.5%. This feature is highly important for critical applications or standards laboratory type work.

HP Temperature-Compensated Thermistor Mounts:

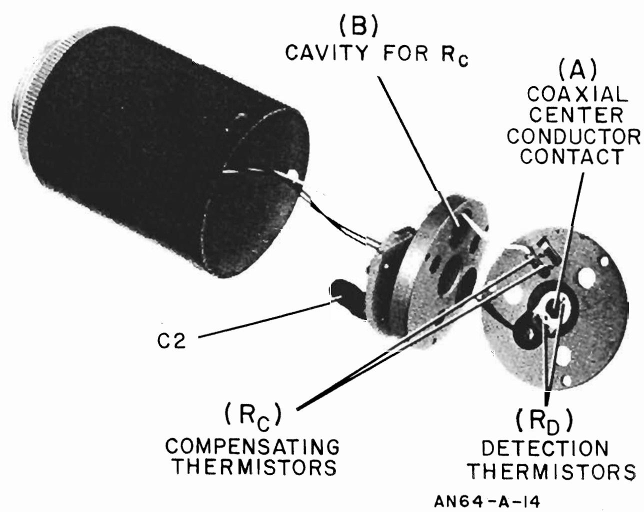
Temperature-compensated power meters depend heavily on thermistor mount design and construction for drift-free, accurate measurements. In addition to the general requirements listed for uncompensated bolometer mounts, the compensated mount must: 1) provide as closely as possible, identical temperature-resistance characteristics of R_d and R_c, 2) maintain electrical isolation between R_d and R_c, and 3) keep both elements in thermal proximity.

The hp 478A Coaxial Thermistor Mount contains four identical thermistors electrically connected as in Figure 2-10a. The mount is similar to the non-compensated 477B described earlier in the respect that two 100 ohm thermistors appear in series to the 10-kc bias and bridge excitation signal from the 431C, and in parallel to applied RF. These two thermistors are the detection thermistors R_d which are mounted within a comparatively massive thermal block as shown in Figure 2-10b. The coaxial center conductor from the RF input jack connects to the junction of the two thermistors at point A through a DC and audio blocking capacitor (not shown).

The compensating thermistors R_c are mounted on small blocks which are completely enclosed in the cavity B when the unit is assembled. With this arrangement, electrical isolation and thermal proximity of R_d and R_c is accomplished with the thermal block massive enough to prevent sudden temperature gradients across the thermistors. Capacitors C2 through C5 are mounted on a circuit board on the back of the thermal block which is shielded by the metal case. Capacitors



a.



b.

Figure 2-10. Temperature variations around hp 478A Coaxial Thermistor Mount are compensated by circuit arrangement (a) and construction (b) shown above. Operating with hp 431C Power Meter, RF power is sensed by dual-element arrangement of detection thermistors R_d. Unwanted temperature variations are sensed by identical compensation thermistors R_c, to reduce meter zero-drift by 100:1

C2 through C4 provide the RF bypass required for a parallel thermistor configuration for RF. Phase shift of the 10 kc signal in the 431C metering bridge must be cancelled in order to achieve bridge balance. Capacitor C5 is used to swamp the bridge-to-mount connecting cable capacitance so variations in cable capacity will have little effect on 10 kc phase shift.

The 478A thermistor assembly is field replaceable should damage occur from excessive RF power or serious mechanical shock. Some RF overloads are not sufficient to burn out the thermistor element but cause permanent change in the thermal characteristic of R_d so it no longer matches R_c. This change degrades the temperature compensation effectiveness causing excessive zero drift. A small screw near R_d may be adjusted in the field to restore compensation in many such cases.

Requirements 1, 2, and 4 at the beginning of the bolometer mount discussion on Pages 19 and 20 deal with mount SWR and efficiency and apply to all bolometer mounts whether they are compensated or not. Temperature compensation improves the overall accuracy possible and therefore adds emphasis to the importance of such considerations. Because of this importance, mount SWR and efficiency of coax and waveguide mounts are treated in detail under "Accuracy Considerations" which will show typical test results from several hundred production mounts manufactured to date by Hewlett-Packard.

The hp 486A-series of waveguide thermistor mounts cover frequencies from 2.6 Gc to 40 Gc. Those mounts in the 2.6 Gc to 18 Gc bands utilize a post-and-bar mounting arrangement for the detection thermistor similar to that described for the 487B non-compensated mounts. The detection thermistor R_d is a single bead thermistor mounted with its axis parallel to, and centered in the E field within a shorted section of waveguide. Figure 2-11a shows the electrical configuration of R_d and the compensation thermistor R_c. Figure 2-11b shows the post and bar which are thermally isolated from the waveguide structure by a circular section of glass epoxy. Electrical continuity across the epoxy is obtained by a thin gold plating on the epoxy surface inside the guide. This thin plate exchanges a minimum of heat from the waveguide to the bar and thermistor element, isolating them from ambient temperature changes while providing a good electrical path. The detection thermistor is further isolated from ambient thermal changes by a block of polystyrene foam inserted into the waveguide opening. This foam prevents convective temperature changes at R_d and also serves to physically protect the tiny thermistor bead from foreign objects which could enter the waveguide input. The foam has very little effect on the mount SWR and does not change mount efficiency. The compensating thermistor R_c is also a single bead which is thermally strapped to the bar as shown in Figure 2-11b. Thermistor R_c and the swamping capacitor C1 are shielded by a metal case which also serves to prevent convective thermal changes from reaching the compensating thermistor.

The 486A-series mounts covering K and R band (18-26.5 Gc and 26.5 - 40 Gc) utilize thermistors which are biased to an operating resistance of 200 ohms rather than the 100 ohms used in lower frequency units. These mounts differ somewhat from the post and bar arrangement but are schematically the same as that shown in Figure 2-11a.

ACCURACY CONSIDERATIONS

A number of factors are responsible for the overall accuracy attained in a microwave power measurement. By knowing the cause and quantitative effects of errors, we can correct or account for them thereby improving the measurement accuracy. Temperature compensation establishes a drift-free starting or reference point and we may now turn our attention from constant re-zeroing of the bolometer bridge to other sources of error.

MISMATCH ERROR.

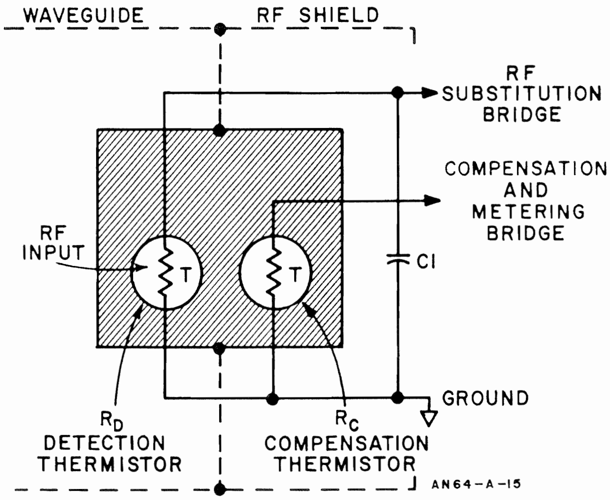
A. Basic Principles:

Consider a DC source with an internal resistance R_G and an external load R_L. Obviously, no power can be delivered to R_L when it is either zero or infinite, so there must be some in-between value at which the power delivered is maximum. It is easily shown that this occurs when R_L equals R_G. Any other value of R_L results in the delivery of less than the maximum available power, or a "mismatch loss".

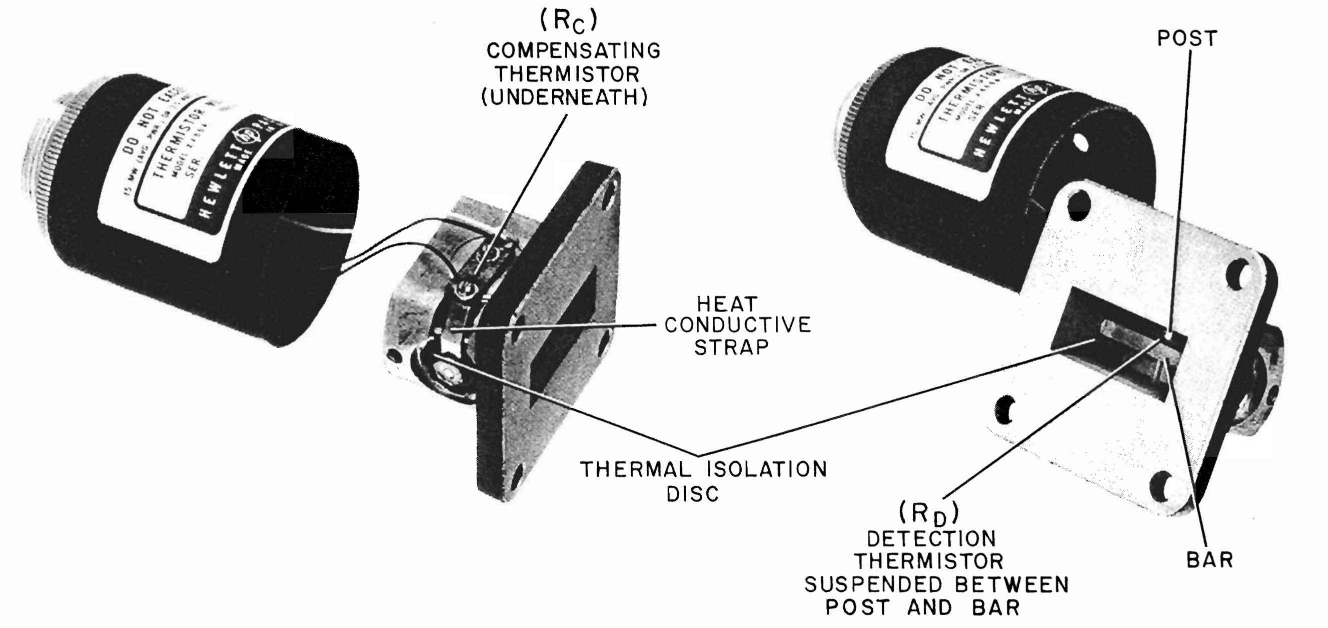
Consider next the extension to the general case, where the source has an internal impedance which at any frequency can be represented by a resistance and a reactance. Whether series or parallel equivalent circuits are used is immaterial, but the reactances must be of equal magnitude and opposite sign so that they will be in resonance and therefore have no effect on the power delivered. Again, the two resistances should be equal. Hence, to get maximum available power from the source, the load impedance should be the complex conjugate of the source impedance. When the two actual impedances are known, the power delivered can be calculated and compared with the maximum available power to determine the mismatch loss.

B. Mismatch at Microwave Frequencies:

At microwave frequencies, a complication arises. The length of transmission line used to connect the load and source can be long enough electrically to transform the load impedance to some other value at the source terminals. What the source "sees" is determined by the actual load impedance, the electrical length of the line, and the characteristic impedance, (Z_o) of the line. In the optimum situation, all elements in a system have the characteristic impedance of the line and there is a maximum transfer of power. In general, however, neither source nor load has Z_o impedance. Furthermore, the actual impedances are almost never known completely. They are given only in the form of SWR's, which lack phase information. As a result, the power delivered to the load, and hence the mismatch loss, can be described only as lying somewhere between two limits. This uncertainty increases with SWR, which is one of the fundamental reasons why manufacturers strive to reduce the SWR's of microwave components.



a.



b.

Figure 2-11. HP 486A waveguide thermistor mount circuit (a) and construction (b) show details of detection and temperature compensation design. Single thermistor R_d is mounted in a post-and-bar arrangement to sense RF power. Post and bar are thermally isolated from the waveguide body by epoxy discs with thin gold plating to maintain electrical continuity. Compensating thermistor (R_c) senses equal temperature variations as R_d because of large heat conductive strap connected to bar.

In the special case where either source or load has unity SWR, the mismatch loss is unique and calculable from the other SWR. The accuracy specification on some commercial power measurement systems is based on the assumption of a Z_o source. Practically speaking, however, almost no sources have exactly Z_o impedance and such an accuracy specification is unrealistic.

C. Basis of Analysis:

To analyze a particular case of mismatch, a convenient basis of calculation must be chosen. If the power actually delivered to the load is to be compared with the maximum available from the source, this is on a conjugate basis. If comparison is made with the power the source will deliver to a Z_0 load, this is on a Z_0 basis. These are two of a number of possibilities.

Considerable confusion has arisen over the use of terms such as "match" and "mismatch" since it is not always clear just what basis is intended. R. W. Beatty(4, 5) of NBS has proposed a complete set of specific terms and definitions which, if generally adopted, should eliminate this confusion. The terms and definitions pertaining to mismatch analysis in power measurement are as follows:

Conjugate Match: The condition for maximum power absorption by a load, in which the impedance seen looking toward the load, at a point in a transmission line, is the complex conjugate of that seen looking toward the source.

Conjugate Mismatch: The condition in the situation above in which the load impedance is not the conjugate of the source impedance.

Conjugate Mismatch Loss: The loss resulting from conjugate mismatch.

Z_0 Match: The condition in which the impedance seen looking into a transmission line is equal to the characteristic impedance of the line.

Z_0 Mismatch: The condition in which the impedance seen looking into a transmission line is not equal to the transmission line characteristic impedance Z_0. (In general, conjugate match is a case of Z_0 mismatch.)

Z_0 Mismatch Loss: The loss resulting from a Z_0 mismatch.

Conjugate Available Power: Maximum available power.

Z_0 Available Power: The power a source will deliver to a Z_0 load.

All mismatch terms used in this Note will follow these definitions where any possibility of ambiguity might otherwise exist. Thus, "conjugate mismatch" and "Z_0 mismatch" are used to describe the actual figures obtained in the examples.

The general expression for power transfer between a source and a load of reflection coefficients Γ_G and Γ_L is*:

$$\frac{(1-|\Gamma_G|^2)\,(1-|\Gamma_L|^2)}{|1-\Gamma_G\Gamma_L|^2}, \qquad (2.4)$$

where

$|\Gamma_G|$ and $|\Gamma_L|$ can be obtained from the SWR's σ_G and σ_L by the simple relation:

$$|\Gamma| = \frac{\sigma-1}{\sigma+1}. \qquad (2.5)$$

The expression in 2.4 is the fraction of the maximum available power actually absorbed by the load. The Z_0 mismatch loss associated with the source is expressed by the term $(1-|\Gamma_G|^2)$. The Z_0 mismatch loss resulting from the load is expressed by the term $(1-|\Gamma_L|^2)$. The uncertainty in the power transfer is expressed by $(1-\Gamma_G\Gamma_L)^2$ since Γ_G and Γ_L are complex quantities. The limits of uncertainty are obtained by evaluating $(1\pm|\Gamma_G||\Gamma_L|)^2$. (It can be seen that the rather vague term "mismatch error" applies in general to a combination of calculable mismatch losses and uncertainties.)

If a conjugate basis is to be used, the effects of all three terms in the expression of 2.4 are included and the entire expression lies between two limits never exceeding unity. Now consider the case if a Z_0 basis is to be used; since the first term in the numerator gives the fractional power delivered to a Z_0 load by the source, only the remaining two terms in equation 2.4 need be evaluated to determine the Z_0 mismatch loss. Note, however, that Γ_G must still be known in order to determine the uncertainty in the loss calculation. This fact must be recognized somehow in any statement of power measurement accuracy. The expression for Z_0 mismatch can have limits above and below unity, as well as both below unity.

When the conjugate basis is used, mismatch is always expressed as a loss. Figures 2-12 and 2-13 are charts giving conjugate mismatch loss limits for different ranges of SWR. The diagonal lines running upward to the right give the minimum possible loss for any combination of source and load SWR's, while the lines running upward to the left give the maximum possible loss.

When the Z_0 basis is used, the Z_0 mismatch loss is obtained from the bottom scale on Figure 2-14 and the uncertainty from the chart above. In Figure 2-14 there is only one set of diagonal lines, but note that the upper left half of the chart gives the upper limit of uncertainty, the lower right half the lower limit, and that these begin to differ appreciably in the upper right corner.

D. Mismatch Uncertainties in Power Measurements:

An example may help clarify the use of these charts in power measurement analysis. Suppose that the power output of a signal generator having a SWR not greater than 1.80 is measured with a power meter and a bolometer having a SWR not greater than 1.35. If it is inconvenient to measure the actual SWR's these

(4) Beatty, R. W., "Intrinsic Attenuation", IEEE Transactions on Microwave Theory and Techniques, Vol. MTT-11, No. 3, May 1963, p. 179.

(5) Beatty, R. W., "Insertion Loss Concepts", Proceedings IEEE, Vol. 52, No. 6, June 1964, p. 663.

*The symbol Γ is used in this section to denote reflection coefficient magnitude and phase. Later the note the symbol ρ is used to denote reflection coefficient magnitude only. (Thus $\Gamma = \rho e^{-j\theta}$)

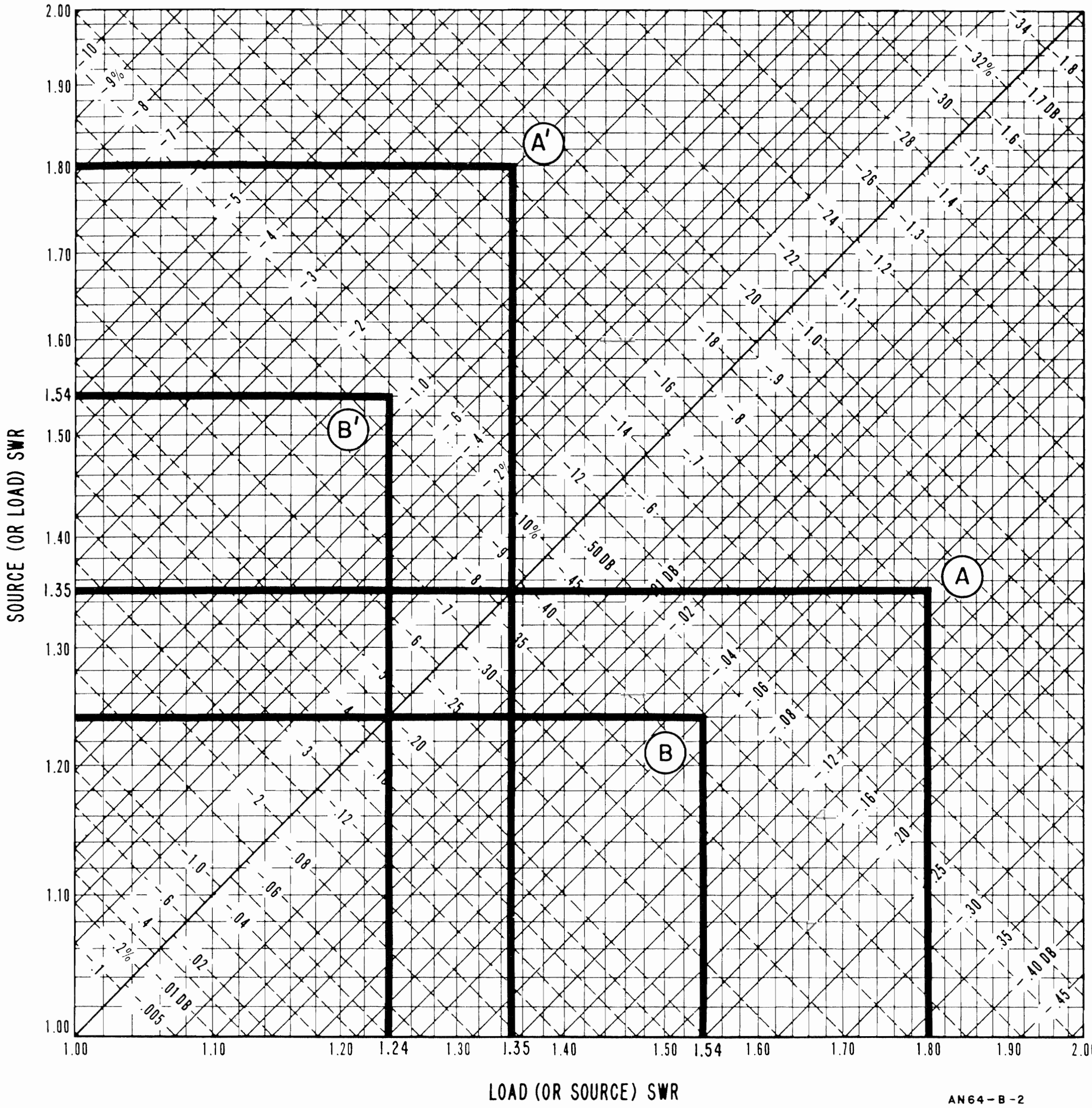


Figure 2-12. Conjugate mismatch-loss chart. Read losses at either intersection of SWR's in question (load and source SWR scales are interchangeable). Diagonal lines running up to right indicate minimum loss; diagonal lines running up to left show maximum loss. Upper left half of chart shows losses in percentages, while lower right half shows losses in db.

values may be taken as a worst possible case. Figure 2-12 (point A) shows a lower conjugate mismatch loss limit of -.090 db, and an upper limit of -.83 db, an uncertainty range of .74 db. The loss limits may also be found in percentage of the conjugate available power by interchanging source and SWR data to enter Figure 2-12 as shown by Point A'. Now suppose the actual SWR's are measured and found to be 1.54 and 1.24, respectively. Points B (and B') on Figure 2-12 now give limits of -.050 db (1.2%) and -.445 db (-9.8%), an uncertainty range of .395 db.

Signal generators are customarily rated in terms of the power they will deliver to a Z_0 load ("Z_0 available power"). When testing such generators, use the Z_0 mismatch loss chart to determine if the generator meets output power specifications. For example, suppose the generator SWR is 1.54 and bolometer mount SWR is 1.24; using the bolometer mount SWR of 1.24, we enter the lower scale in Figure 2-14 (point A) and find the Z_0 mismatch loss of the mount is -.050 db. On this loss is an uncertainty of +.200, -.195 db (point B). Adding the loss and uncertainty data results in a total Z_0 mismatch "loss" of +.15 db to -.245 db. Thus the power available to a Z_0 load would be between .150 db below and .245 db above the power absorbed by the bolometer mount.

E. Maximum and Minimum Loss Calculations:

In case a mismatch loss chart is not available, the maximum and minimum conjugate mismatch losses corresponding to two SWR's, σ_1 (the larger) and σ_2, may readily be determined as follows: The maximum loss corresponds to that which would occur if one SWR were equal to the product $\sigma_1\sigma_2$ and the other to unity, while the minimum loss corresponds to that which would occur if one SWR were equal to the quotient σ_1/σ_2 and the other to unity. Using the relation between SWR and reflection coefficient, the fractional expression for minimum power transfer (maximum loss) is

$$\frac{4\sigma_1\sigma_2}{(\sigma_1\sigma_2+1)^2},$$

while the fractional expression for maximum power transfer (minimum loss) is

$$\frac{4\sigma_1\sigma_2}{(\sigma_1+\sigma_2)2}.$$

The loss in db is ten times the log of either expression (or, conventionally, its reciprocal, for a positive number of db).

F. SWR Characteristics of Thermistor Mounts:

Figure 2-15a shows what is typically being achieved in SWR of the hp 478A temperature compensated coaxial thermistor mount. Note that over a major portion of the frequency range of the instrument, SWR is less than 1.2:1 and, indeed, over a very large range it is less than 1.1:1. It is only at the high and low frequency ends of the band that a given mount may deviate from this typical curve. For this reason it is unnecessary in many cases to consider the use of a tuner with this thermistor mount. A tuner or other effective means of reducing mismatch error is recommended, however, where source SWR is high or high

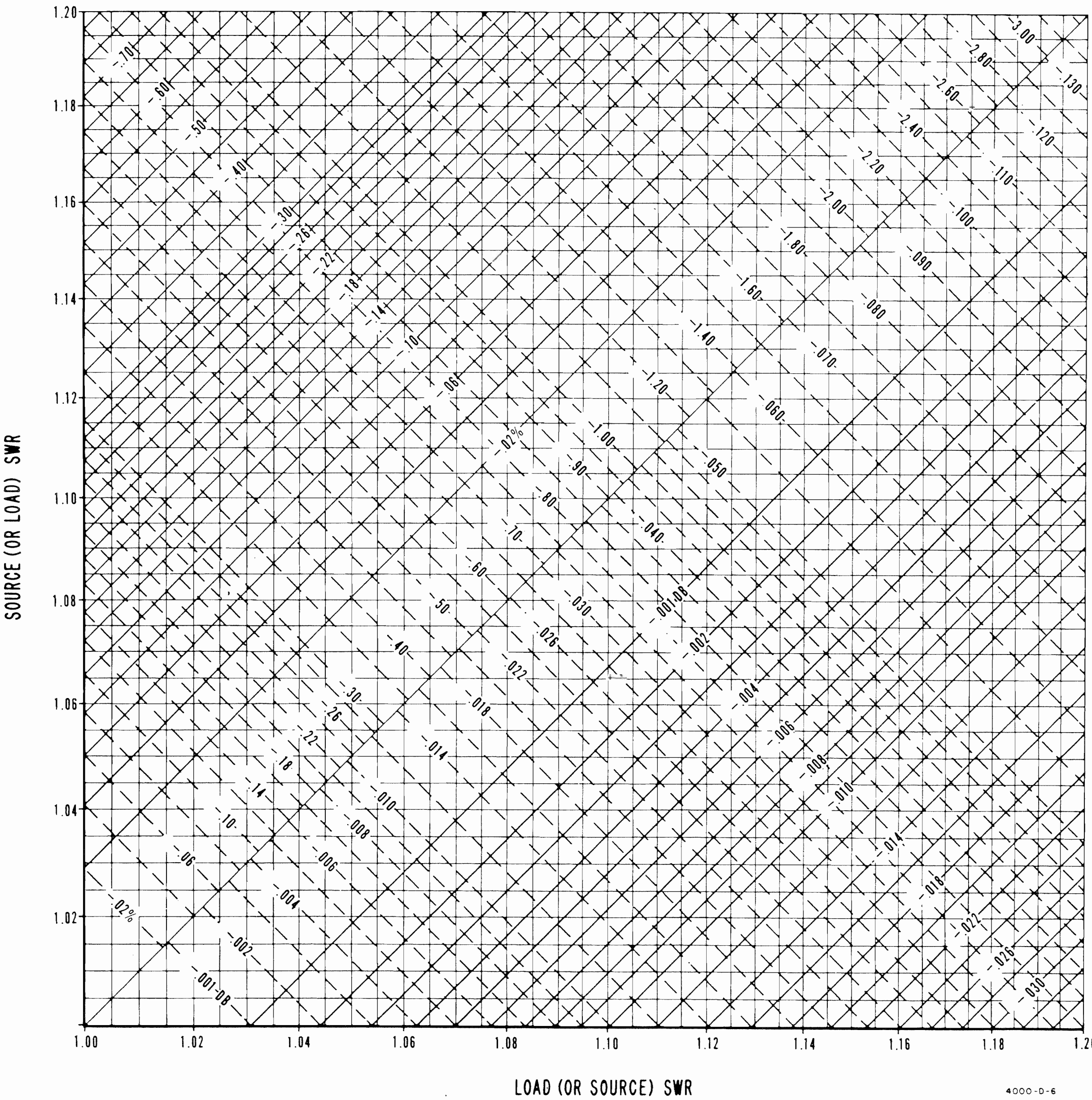


Figure 2-13. Expanded conjugate mismatch-loss chart covering SWR's from 1.00 to 1.20. Chart is read in same manner as Figure 2-12.

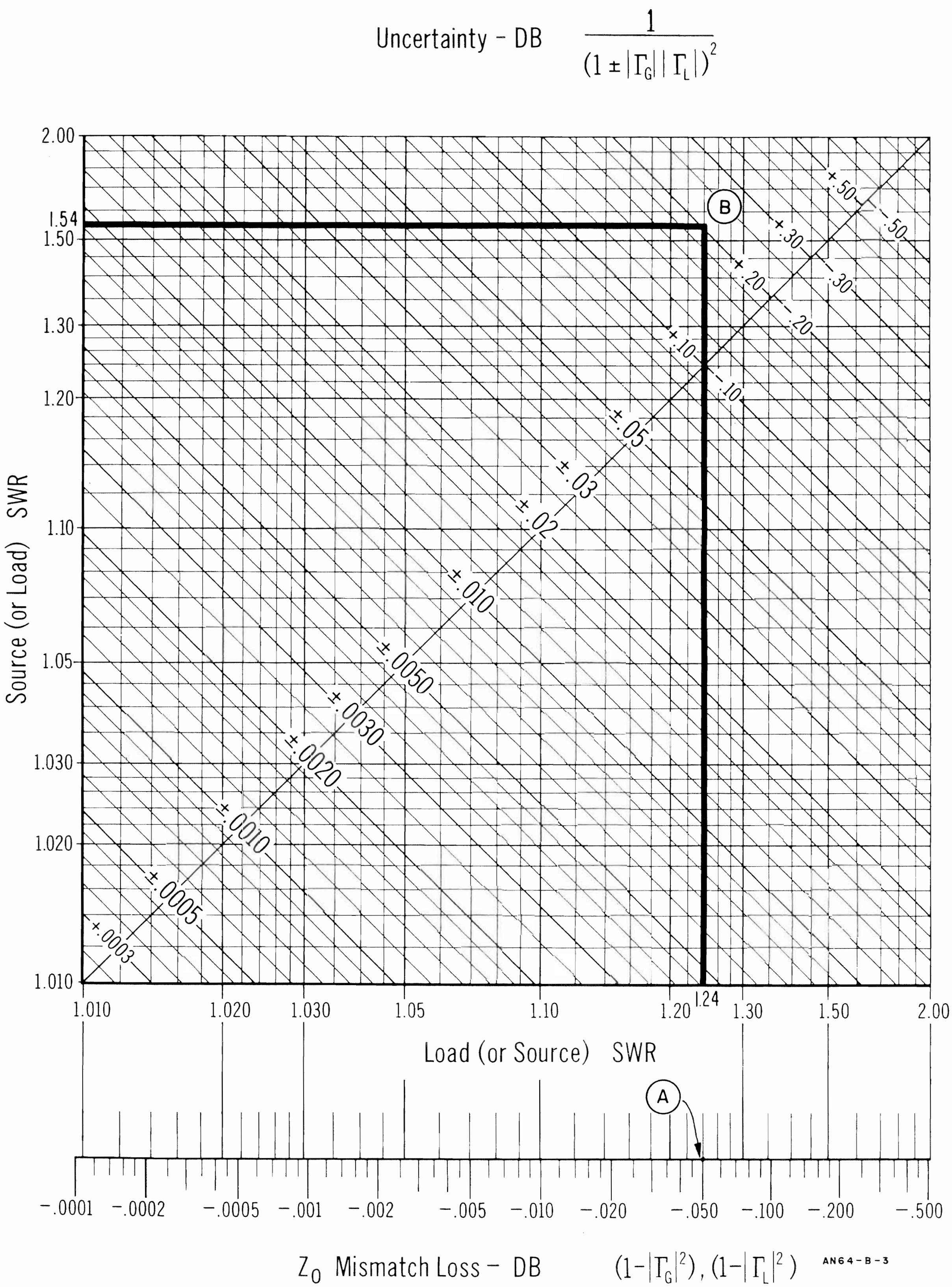


Figure 2-14a. Z_0 mismatch-loss and uncertainty chart. Z_0 mismatch loss (in DB) is found on bottom scale by entering the appropriate load SWR above the scale, and reading the loss immediately below. Z_0 mismatch uncertainty (in DB) is found at the intersection of Source and Load SWR's in the chart.

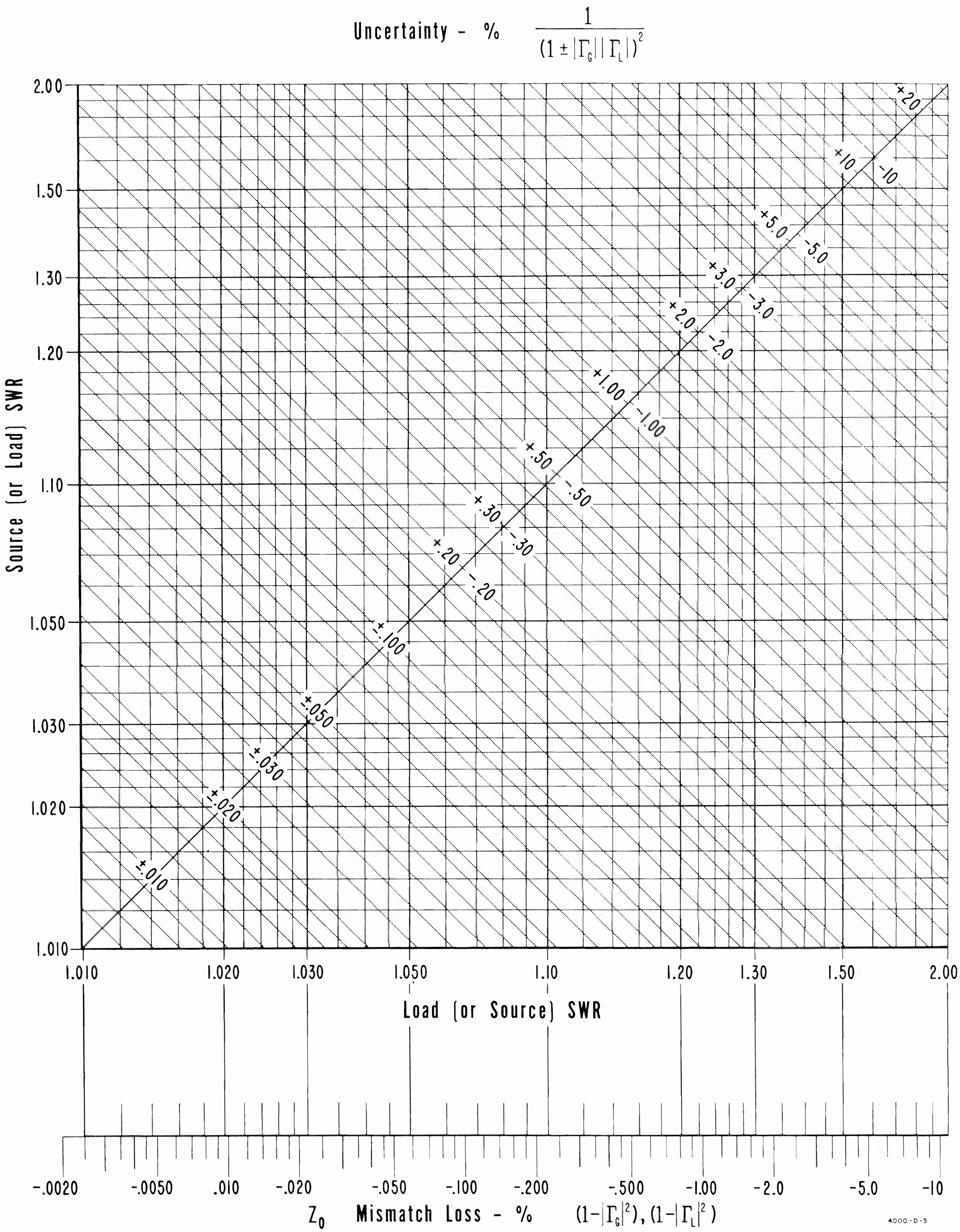


Figure 2-14b. Z_O mismatch loss and uncertainty chart calibrated in percent of measured power.

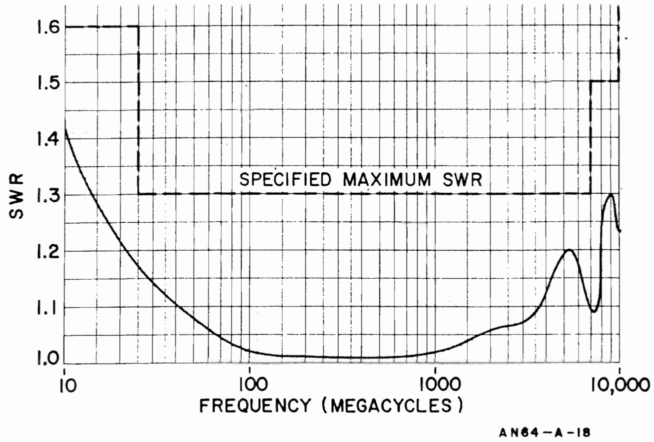


Figure 2-15a. Typical SWR of the hp 478A Coaxial Thermistor Mount. Z_O mismatch resulting from such low reflections is less than 0.2% from 40 Mc to 6 Gc, and less than 1.5% to 10 Gc.

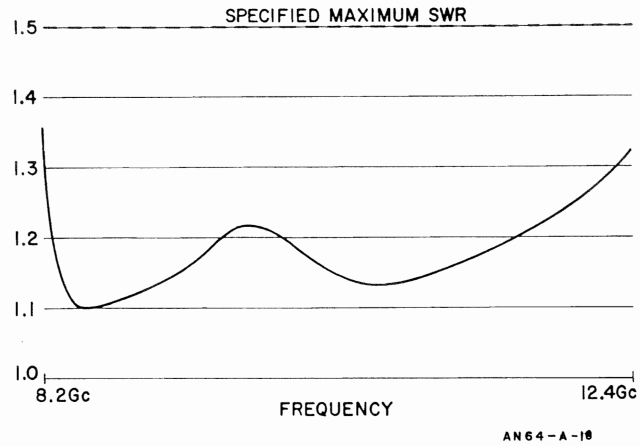


Figure 2-15b. Typical SWR of the hp X486A Waveguide Thermistor Mount. Z_O mismatch due to this mount's reflections would be less than 0.8% over most of the X-band.

accuracy is required. The typical SWR of the hp X486A waveguide thermistor mount is shown in Figure 2-15b. Deviations from this curve by individual mounts may be rather great even though they remain well below the 1.5:1 specified limit. When SWR of a waveguide mount is being considered in a power measurement error analysis, it is recommended that the actual SWR of the specific mount be measured rather than assuming that it will be close to the value indicated by this curve. Each Hewlett-Packard waveguide thermistor mount is swept-frequency tested for SWR specifications as part of its final test prior to shipment. This testing ensures satisfactory performance of each mount throughout its frequency range in the field.

RF LOSSES AND DC-TO-MICROWAVE SUBSTITUTION ERROR

RF losses account for power entering the thermistor mount but not dissipated in the detection thermistor and not reflected by mount mismatch. Such losses may be in the walls of a waveguide mount or the center conductor of a coaxial mount, losses in capacitor dielectric or in poor connections within the mount, radiation, etc. From the previous discussion on the operation of bolometer power meters, we know that the power meter can sense only the power dissipated in the detector (thermistor(s) or barretter).

DC-to-microwave substitution error is caused by the difference in heating effects of the DC or audio bias power substituted in a bolometer and the microwave power being measured. This difference results from the fact that the spatial distributions of current, power, and resistance within the bolometer element are different for DC and RF power.

DC-to-microwave substitution error and RF losses in thermistor mounts are virtually impossible to separate in a quantitative measurement. For this reason, the total effect of these two errors is measured by the National Bureau of Standards (NBS) and some commercial standards laboratories and presented as a figure of merit called the Effective Efficiency (η_e) of a mount.

Effective Efficiency is defined as the ratio of substituted DC power in the bolometer element to the microwave power dissipated within the bolometer mount. This may be stated symbolically as

$$\eta_e = \frac{P_{(\text{dc substituted})}}{P_{(\mu\text{wave dissipated})}} \qquad (2.6)$$

a ratio less than unity in actual practice*, generally expressed in percentage. Note that this expression does not include the effects of mismatch error. Effective efficiency is largely independent of the level of input power; therefore, efficiency test results at say 10 mw are valid at 10 μw in well designed mounts.

All hp thermistor mounts are swept frequency tested in production for consistently high efficiency throughout each operating band. This testing, performed with a reflectometer-type system, assures good broadband performance of each mount and prevents efficiency "holes" that could cause large power measurement errors. After each swept test, absolute calibrations are made at several discrete frequencies within the band. This data is now recorded on all temperature compensated mounts and available on a special handling basis on uncompensated thermistor mounts. These figures are traceable to NBS wherever services are available, and provide actual correction data for

*DC-microwave substitution error could be positive or negative so in theory η_e could exceed unity. In actual practice; however, RF losses are always large enough to cause the combined effect to be less than unity.

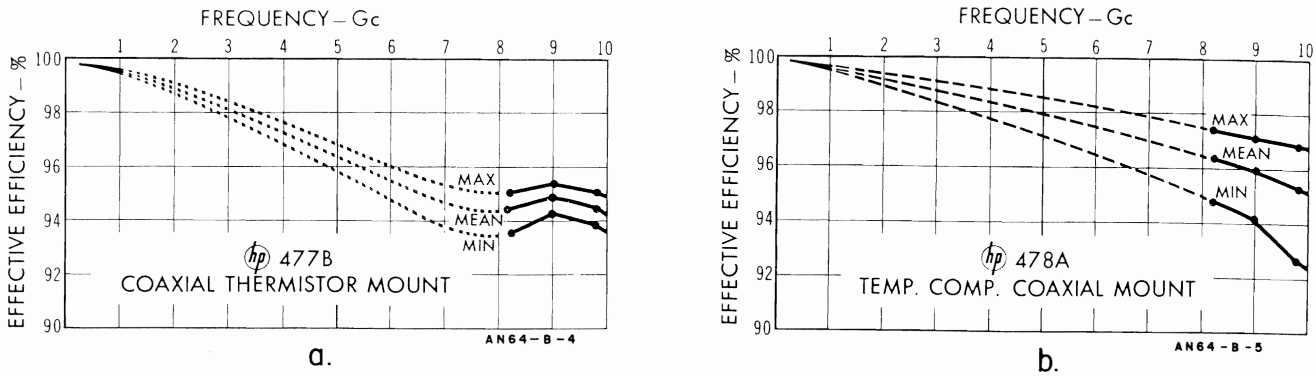


Figure 2-16. Typical efficiency data on (a) HP 477B coaxial thermistor mounts and (b) HP 478A temperature compensated coax mounts. Solid lines indicate data with NBS traceability. Recent NBS calibrations indicate dashed line interpolation is valid at 3 Gc also. Actual data is now stamped on each hp 478A compensated mount for highest accuracy.

high accuracy power measurements. Section III contains information on mount efficiency calibration services and the system used for production testing at HP.

Figure 2-16 shows the efficiency of several hp 478A temperature-compensated coax mounts and hp 477B non-compensated coax mounts. Note the improved mean efficiency of the compensated mounts over the older 477B's, especially at the higher end of the band. The curves show the maximum efficiency of the 478A's tested to be about 97% and the minimum from 92 to 95% in the 8. 0 to 10. 0 Gc band. In the VHF region up to 1 Gc mount efficiency is above 99%. Comparisons against waveguide mounts and hp 423A flat crystal detectors have given consistent indication that 478A efficiency curves are smooth between 1Gc and 8 Gc. Therefore, the dashed lines of Figure 2-16b show the interpolated values of efficiency that can be expected. Interpolation in Figure 2-16a is less certain because less data has been accumulated on the uncompensated mounts in the 1 to 8 Gc region. Figure 2-17 shows the efficiency plots of several hp X486A and X487B waveguide thermistor mounts over the X-band. Uniformly high efficiency is typical in these mounts with the temperature compensated X486A running slightly higher than the non-compensated mounts at the higher frequencies.

One H-band (7. 05 to 10. 0 gc) compensated mount was included with the results of seven X-band mounts so a comparison of different waveguide size units could be made. The particular H486A tested compared very favorably with the X-band units as can be seen from the curves.

Another calibration service available from NBS and some commercial standards laboratories is the Calibration Factor (K_b) of a bolometer mount. Calibration Factor is defined as the ratio of the substituted DC power in the bolometer element to the microwave power incident upon the bolometer mount. This may be stated symbolically as

$$K_b = \frac{P_{(\text{dc substituted})}}{P_{(\mu\text{wave incident})}}$$

Note that this expression includes the effects of mismatch loss since $P_{dc\ substituted}$ decreases when RF power is reflected. If a tuner is not used to cancel mismatch effects in a power measurement, Calibration Factor can be directly applied as a correction to the meter reading to improve overall accuracy. Since

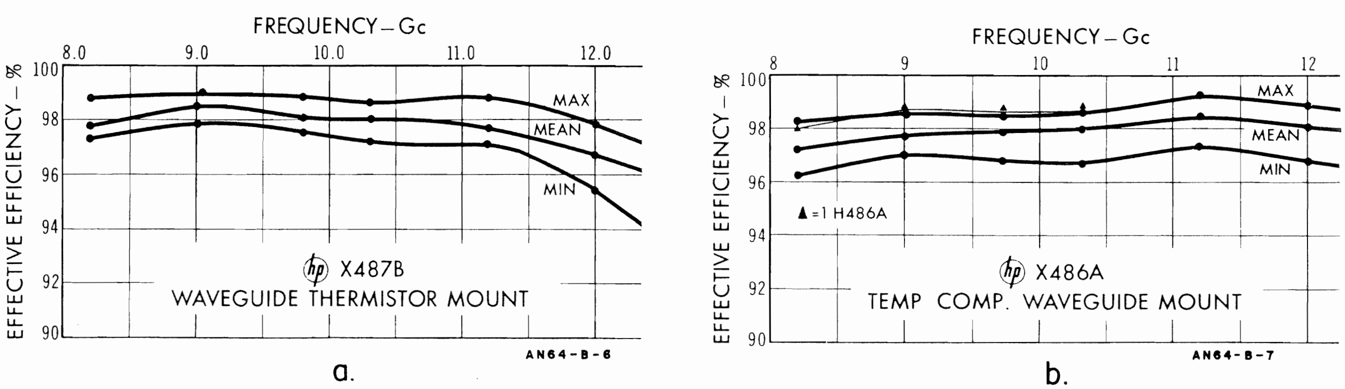


Figure 2-17. Typical efficiency data on (a) HP X487B waveguide thermistor mounts and (b) HP X486A temperature compensated waveguide mounts. Actual data is now recorded on each 486A compensated mount.

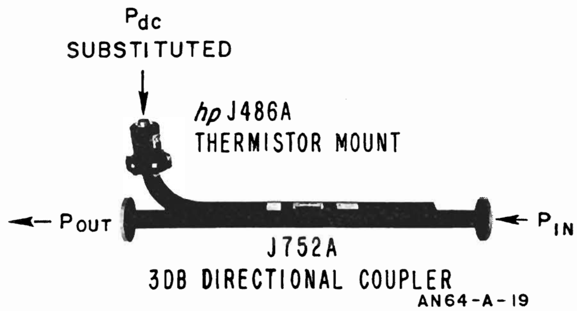


Figure 2-18. Typical J-band Power Standard comprises thermistor mount permanently attached to 3 db directional coupler auxiliary arm. Calibration factor of Standard relates DC (or audio) substitution power in thermistor element to main RF power out.

a tuner is often not used, calibration factor is a highly practical term and is therefore measured and included with the efficiency data on each HP temperature compensated thermistor mount.

Another common and very useful calibration is the Calibration Factor of a bolometer mount-directional coupler combination, or "Power Standard", as shown in Figure 2-18. The Power Standard Calibration Factor is defined as follows:

$$K_c = \frac{P_{dc\ substituted}}{P_{\mu wave\ incident}},$$

where $P_{dc\ substituted}$ = DC power substituted in the bolometer element, and $P_{incident}$=microwave power incident to a non-reflecting load terminating the coupler main arm output. The bolometer mount is of the fixed-tuned or untuned broadband variety such as the hp 486A shown, and is permanently attached to the auxiliary arm of a high-directivity coupler of 3 to 20 db coupling factor. The Calibration Factor is measured at several specific frequencies within the band and includes the effects of mismatch between bolometer mount and auxiliary arm, and the bolometer mount efficiency. The Power Standard provides a means for accurately monitoring power in a circuit rather than terminating all of the power in the measuring device. This configuration is useful as a standard of comparison for calibrating other power measuring devices connected to the coupler main arm output or for any case of making accurate in-line measurements of incident power. Examples of how to use bolometer mount and mount-coupler calibrations are shown in Section III.

DUAL ELEMENT BOLOMETER MOUNT ERROR

A detailed analysis of dual element error has been made by G. F. Engen(1) of NBS and by L.A.Harris(2). This error does not occur in the waveguide thermistor mounts but must be considered in nearly all coaxial mount designs.

Figure 2-19a shows the basic circuit of a coaxial mount. R_{T1} and R_{T2} are the detection thermistors. Figure 2-19b shows the equivalent circuit for DC and 10 kc. The thermistors are in series and if the resistance division is unequal, the greatest power will be dissipated in the thermistor with the largest resistance. However, referring to the equivalent circuit for RF (Figure 2-19c) the RF power sees the two thermistors in parallel. In this case, if the resistance division is unequal, the greatest power will be dissipated in the thermistor with the least resistance. Unfortunately, the situation at microwave frequencies is not quite this simple inasmuch as a different power split may result due to circuit reactance. In any case, Mr. Engen has shown that the resulting error is equal to

$$E = \left(\frac{1}{\gamma 2} - \frac{1}{\gamma 1}\right)\Delta r$$

where γ_1 and γ_2 are the "ohms per milliwatt" coefficients of the thermistors and Δr is the shift in resistance division.

Dual element error may be as large as 1% for conventional thermistor mounts at 10 mw RF levels increasing sharply as power increases. Conversely, as the measured RF power decreases, this error decreases rather rapidly. This, of course, represents one of the problems involved with dual element error, inasmuch as it is not a constant error, but rather is a function of the RF power being measured. Since automatic bridges such as the 430C and 431C have 10 mw upper range limits, this error is generally quite small when using mounts of good quality.

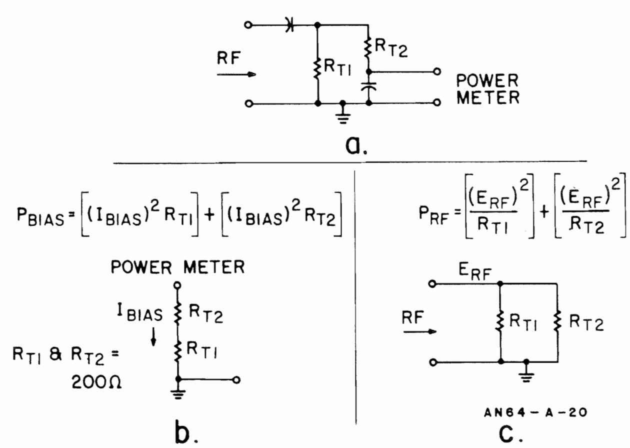


Figure 2-19. Actual circuit of dual-element coaxial thermistor mount is shown in (a). DC and low frequency equivalent circuit (b) appears as two series thermistors to power meter bridge. Equivalent circuit for RF input power (c) appears as two parallel thermistors. If resistance of thermistors is unequal, DC and RF power will have unequal effects on circuit, causing "dual element error."

(1)Engen, G. F., "A DC-RF Substitution Error in Dual Element Bolometer Mounts", NBS Report 7934.

(2)Harris, I. A., "A Coaxial Film Bolometer for the Measurement of Power in the UHF Band", Proc. IEE (British), Vol. 107, Part B, No. 31 (Jan. 1960).

By virtue of the techniques necessary to get thermal balance in the 478A coaxial thermistor mount, the parameters which cause dual-element error are controlled to a great degree. Initial tests on fifteen 478A thermistor elements at Hewlett-Packard revealed a maximum error of about 0.1% at 10 mw RF power. Most of the elements were somewhat below this value. The measurement technique used, however, did not include reactive effects encountered at microwave frequencies. These effects have been tested more recently with tentative results showing virtually no further error above 100 Mc.

THERMOELECTRIC EFFECT ERROR

Thermoelectric error is peculiar to the 431-type of power meter. This error is found in power meters where nearly the total bias power is supplied by AC. While this thermoelectric phenomena has not been fully investigated, it is felt that it can be explained on the basis of the two thermocouples that are formed by the contacts of the thermistor leads to the thermistor oxide (see Figure 2-20). The bead thermistor used in thermistor mounts operates more than 100°C above room temperature. It is likely that both contacts to the bead will not be at the same temperature; thus, a thermocouple will result. The AC bias power heats the thermistor and if there is not perfect cancellation of these thermocouples, the result will be a DC current flow. This DC current flow will either add to or partially cancel any DC currents that are being applied by the power meter.

The magnitude of this thermoelectric effect is measured in all of the 478A and 486A thermistor mounts built at Hewlett-Packard. In production mounts, thermoelectric effect measurement error should not exceed 0.3 μw. Obviously, this error is significant only on the more sensitive ranges of the power meter. However, this effect also becomes significant in the RF thermistor when a DC calibration or DC substitution measurement is being made. Again, this can amount to a 0.3 μw error. Where accurate substitution or calibration is required at low power levels, this thermoelectric effect can be eliminated by using a dc supply whose polarity is easily reversed to supply the substitution or calibration power to the RF thermistor. By reversing polarity, the sum and difference of substitution power and thermoelectric power can be determined. Section III shows how to apply these data for correcting meter readings of very low power level.

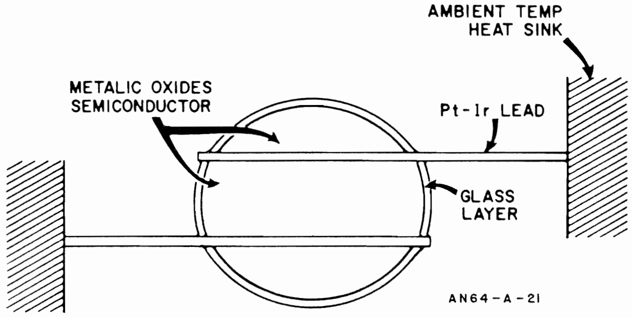


Figure 2-20. Cross section of a bead thermistor connected to a heat sink for thermal stability.

INSTRUMENTATION ERROR

Instrumentation error is the inability of the power meter to exactly measure the DC or AC Substituted power in the bolometer element. This type of error is somewhat analogous to the error found in VTVM's. Examples of the causes of instrumentation error are: 1) Range resistor tolerance, 2) Non-linearity in feedback amplifiers, 3) Matching of bridge transformers (431C), and 4) Meter tracking error. In specifying the accuracy of a power meter, instrumentation error is the figure generally used. As mentioned earlier, this error does not exceed 3% of full scale in the hp 431C power meter.

The ability to make a DC calibration or substitution measurement is a very significant feature provided by the circuitry in the 431C. This technique, described under "Methods for Improving Accuracy" in Section III, can reduce instrumentation error to less than 0.5%. Some manually balanced bridges allow this error to be reduced to 0.1% using precision potentiometric methods.

CALORIMETRIC POWER METERS

The most fundamental method of measuring microwave power is to dissipate it as heat and measure the resulting temperature rise. True calorimetry involves dissipating a certain amount of energy, containing the heat resulting, and measuring the temperature rise. This principle has been used for years in chemical and physical measurements. The calorimetric power meter dissipates power (a continuous flow of energy), controls the resulting flow of heat through some thermal path, and measures the temperature rise set up by this flow.

Calorimetric power meters used for microwave work may be divided into two categories: 1) Flow (liquid) and 2) Static (dry). While there are several designs of both types, an example of each will acquaint the reader with the principles involved. The dry calorimetric power meter was an outgrowth of the basic calorimeter, and is still used in some applications. The Hewlett-Packard Calorimetric Power Meter is a more recent development which uses oil flow as a means of carrying away heat continuously in a controlled manner. This gives a convenient range of dissipation capability... 10 mw to 10 w in a structure small enough to provide a good impedance match to a 50 ohm transmission line from DC through X-band.

FLOW CALORIMETERS

Figure 2-21a shows the basic circuitry of the hp 434A 10-watt calorimetric power meter. The circuit consists of a self-balancing bridge which has identical temperature-sensitive resistor gauges (one in each leg), a high-gain amplifier system, an indicating meter, and two load resistors, one for unknown input power and one for comparison power. The input load resistor and one gauge are in thermal proximity so that heat generated in the input load resistor is carried to its gauge by the oil-stream, tending to unbalance the bridge. The unbalance signal is amplified and applied to the comparison load resistor which is

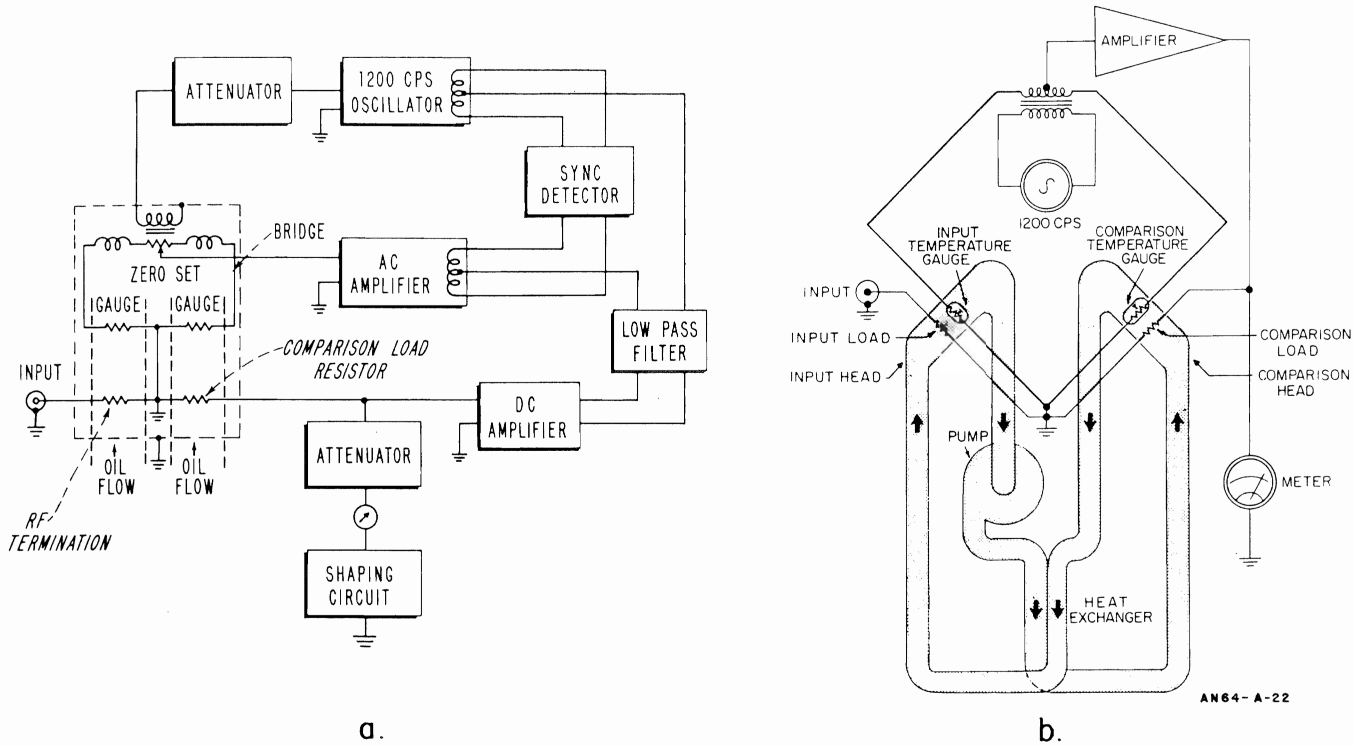


Figure 2-21. Block diagram (a) and oil flow diagram (b) of hp 434A Calorimetric Power Meter. Moving oil stream and bridge-feedback arrangement results in fast response time to input power changes enabling dynamic tuning of units in test. Series oil flow arrangement eliminates flow rate errors usually encountered in flow calorimeters.

in thermal proximity to the other gauge. The heat generated in the comparison load resistor is carried to its gauge by the oil-stream which automatically rebalances the bridge. The meter measures power supplied to the comparison load to rebalance the bridge which is equivalent to the RF power applied to the input load. Because of this feedback arrangement total response time of the 434A is less than 5 seconds for full-scale deflection of the meter.

Resistance characteristics of the gauges are the same, as are the heat transfer characteristics of the loads, so the meter may be calibrated directly in input power. The calorimetric-power-meter measurement is accurate because volumetric oil flow through the two heads is made identical by placing all elements of the oil system in series, as shown in Figure 2-21b. Differential temperature between oil entering the two heads is also zero because the oil to each head is brought to the same temperature by passing it through a parallel-flow heat exchanger.

Microwave power dissipated in the RF head resistor is matched by DC power in the comparison head to give a meter reading. The 434A is capable of high accuracy because the RF head can be DC-calibrated like the 431C. A 1% accurate DC power calibrating source of 100 mw is built into the 434A for standardizing the calorimeter before making power measurements.

CALORIMETRIC POWER METER ACCURACY

The calorimetric power meter is inherently accurate due to its circuit stability, controlled thermal flow, and rigid mechanical design. Overall measurement accuracy of the 434A is better than 5% of full scale excluding mismatch loss. Accuracy can be even higher by being aware of, and correcting for, various sources of error such as mismatch, head efficiency, and instrumentation error.

Mismatch Error.

As in the case of the bolometric power meter, mismatch error is a very important factor in making accurate calorimetric power meter measurements. The same laws of impedance and phase apply to the calorimetric head as to bolometer mounts. When neither source nor load is Z_0, there is an ambiguity as to how much power is being delivered. On a conjugate basis, there is a mismatch loss when the calorimetric input impedance is not the conjugate of the source.

Table 2 lists maximum SWR specifications for the hp 434A and Figure 2-22 shows a typical SWR curve plotted from 500 Mc to 12.4 Gc. In cases where high accuracy is required, the hp 872A coaxial tuner may be used for correcting mismatch in the 500 Mc to 4 Gc region. In higher bands, coaxial stub-tuners are available for impedance matching or, if waveguide

Table 2. SWR Specifications for HP 434A Calorimetric Power Meter

Frequency	Max Input SWR
DC to 5 Gc	1.3:1
5 Gc to 11 Gc	1.5:1
11 Gc to 12.4 Gc	1.7:1

systems are being used, the hp 870-series of slide screw tuners are recommended. Hewlett-Packard sweep tests each 434A in production to insure quality and adherence to SWR specification.

Head Efficiency.

The calorimetric power meter does not depend upon the validity of the substitution process in the comparison of microwave power and DC or low-frequency power. This is because all power delivered to the terminating resistor raises the temperature of the oil stream, and skin effect or thermal distribution within the sensor is not of primary importance as it is in the bolometer mount.

Efficiency in a calorimeter head is defined as the microwave power dissipated in the head resistor divided by the microwave power dissipated in the entire RF circuit. Calibration factor is the microwave power in the head resistor divided by the microwave power incident upon the RF input circuit.

It may be seen that these definitions are essentially equivalent to those for bolometers except that the sensors are different. Loss in the RF circuit occurs in the transmission line between the front panel connector and the head load resistor which heats the oil stream. Figure 2-23 shows an X-band efficiency plot of three 434A Calorimetric Power Meters, each with 1 watt of microwave power input. Typical efficiencies are between 97 and 99% in the X-band region, and approach 100% at lower frequencies down to DC.

Instrumentation Error.

Causes of instrumentation error in the 434A-type calorimeter is largely the same as in bolometer power meters in that measurement of the feedback power cannot be exact. Instrumentation error accounts for 1-2% in the overall 434A accuracy specification of 5%. The calibration accuracy (which includes head efficiency) is usually better than 3% to 10 Gc, and within 4% through 12.4 Gc. On the 10 and 100 mw power ranges however, errors to about 5% can occur

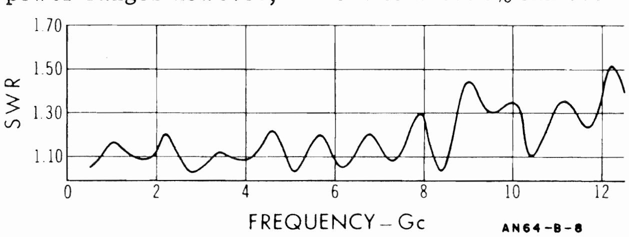


Figure 2-22. Input SWR of a typical hp 434A Calorimetric Power Meter. Broadband impedance is carefully controlled by terminating input power in a coaxial film resistor in a tapered outer conductor.

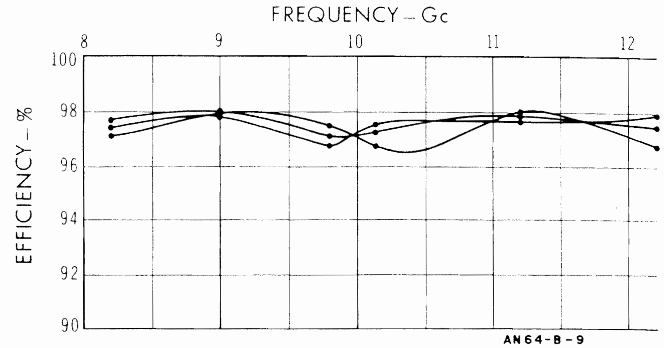


Figure 2-23. Typical efficiency data from hp 434A Calorimetric Power Meters in X-band. Consistency of data taken on these and other 434A's over several years results in a high degree of confidence in predicting instrument accuracy.

due to noise and instability. The internal calibrator of the 434A provides a DC standard for checking and adjusting amplifier gain for full scale on the 100 mw range. The calibrator is accurate to within 1%.

The 434A is factory-tested and adjusted with an hp K02-434A Test Set which provides accurate and stable DC reference power levels in 1/3 scale increments throughout the 434A power range of 10 mw to 10 watts. DC calibrations to 0.5% are possible with this test set which is available to 434A users wishing to make such calibrations conveniently.

DRY CALORIMETERS

The basic dry calorimeter measures the thermal EMF generated in a thermopile placed between a reference load and an active load dissipating input power. There are a number of dry calorimeter designs, one of which is of particular interest. This is the microcalorimeter described by G. F. Engen* of NBS which is the U.S. national reference standard for Effective Efficiency measurements at microwave frequencies. NBS Working Standard bolometer mounts are calibrated in the microcalorimeter and then used to calibrate suitable bolometer mounts submitted for efficiency and calibration factor measurements.

Microcalorimeter Design.

Figure 2-24 is a cutaway view of the microcalorimeter showing a double walled thermal shield housing two identical bolometer mounts of special design. RF entry is made through a length of waveguide thermally isolated from the bolometer mounts to prevent conduction of heat away from the mounts. Applied RF power passes through a waveguide bend and is dissipated in the lower (active) bolometer mount. A physical junction is made at the waveguide bend with an isometric bend connecting to the upper (reference) bolometer mount. RF is prevented from entering the upper bend by a partition at the guide junction. Thus the reference mount remains constant in temperature with power applied to the entry point. This arrange-

*Engen, G.F., "A Refined X-Band Microwave Microcalorimeter", NBS J. of Res. 63C 77 (1959).

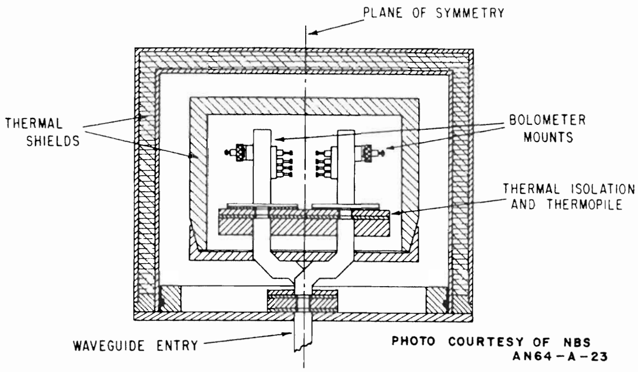


Figure 2-24. Cross section of NBS microcalorimeter with special bolometer mounts installed. Efficiency Calibrations in the 8.2 to 18.0 Gc region by NBS are based upon instruments of this design.

ment provides equal mass in the two arms so they are isothermal with no power applied and remain so with ambient temperature variation. A number of series-connected thermo-junctions are made alternately between the waveguide flanges facing each bolometer mount. This thermopile senses the temperature gradient between the bolometer mounts that results from the application of power to the system. The thermopile output can be accurately calibrated by applying known amounts of DC power to the active bolometer mount.

Operation of the Microcalorimeter.

Recall that Effective Efficiency of a bolometer mount is the ratio of substituted power in the bolometer element, to the total microwave power dissipated in the mount. In the microcalorimeter, substituted (or retracted) power in the bolometer element is measured by an external self-balancing DC bridge with the active bolometer forming one leg. Total power dissipated in the mount is measured by the thermopile output.

In operation, the bolometer mount to be tested is placed in the active arm of the microcalorimeter and DC bias power applied to balance the external bridge.

When equilibrium in the mount temperature is reached, the thermopile output is noted as e_1 and the total bridge current as I_1. RF is then applied and enough DC bias removed from the bolometer to maintain bridge balance. When equilibrium is again reached, the thermopile output is noted as e_2 and the total bridge current as I_2. Power retracted from the bolometer element is calculated from the equation

$$P_{dc} = \frac{R}{4}\left(I_1^{\,2} - I_2^{\,2}\right) \qquad (2.9)$$

where R = Bolometer resistance in an equal arm bridge.

The power dissipated in the mount during the RF "off" and RF "on" periods is ideally related by the thermopile response e_1 and e_2 respectively. The power resulting in thermopile output e_1 is from the DC bias $I_1^2\ R/4$. The total power for e_2 is the sum of the applied RF and the DC bias remaining, viz

$$P_t = P_{rf} + I_2^{\,2}\ R/4 \qquad (2.10)$$

We may now write the following proportionality:

$$\frac{e_2}{e_1} = \frac{P_t}{I_1^{\,2}\ R/4}$$

$$\text{transposing } P_t = \frac{e_2 I_1^{\,2} R}{4e_1} \qquad (2.11)$$

The RF power applied is the difference in P_t and the DC bias remaining, thus

$$P_{rf} = \frac{e_2 I_1^{\,2} R}{4e_1} - \frac{I_2^{\,2} R}{4}$$

$$\text{simplifying } P_{rf} = \frac{R}{4}\left(\frac{e_2}{e_1} I_1^{\,2} - I_2^{\,2}\right) \qquad (2.13)$$

We are now able to express Effective Efficiency as a ratio of DC bias power retracted (eq. 2.9) to RF power dissipated (eq. 2.13) as follows:

$$\eta_e = \frac{R/4\ (I_1^{\,2} - I_2^{\,2})}{R/4\left(\frac{e_2}{e_1} I_1^{\,2} - I_2^{\,2}\right)}$$

simplified, this becomes:

$$\eta_e = \frac{1 - (I_2/I_1)^2}{e_2/e_1 - (I_2/I_1)^2} \qquad (2.14)$$

Figure 2-25a illustrates the cycle of power applied to the microcalorimeter. Figure 2-25b shows the thermopile output resulting from heating the mount with applied power, first with only DC then with DC and RF, then only DC again. The small amount of RF dissipated in the mount, but not in the bolometer element, accounts for the overall increase in thermal output to e_2. In practice, two cycles of RF "off" are used and the average thermopile output taken as e_1. This tends to reduce the effects of thermal background

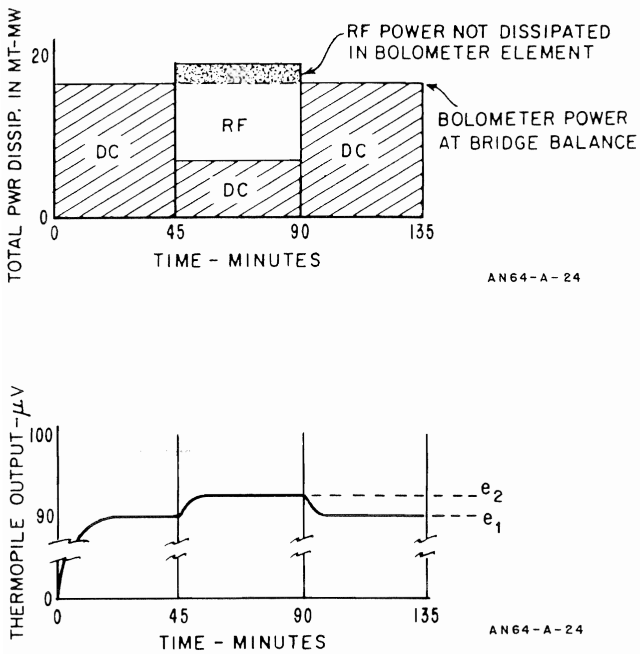


Figure 2-25(a). Cycle of power applied to active bolometer mount in NBS microcalorimeter; (b) Thermopile output of NBS microcalorimeter resulting from power cycle shown in (a)

drift in the system. As Figure 2-25 shows, 45 minutes are required for the mount to reach thermal equilibrium after each power change so the task is very time consuming, especially when one considers that this procedure is required for every discrete frequency where efficiency is desired. The primary consideration in this application, however, is accuracy rather than speed. Efficiency calibrations in the microcalorimeter described are estimated to achieve the remarkable accuracy of 0.1%. Using faster methods for working calibrations and allowing for transfer error, NBS efficiency calibrations are stated to be within 1%.

The National Bureau of Standards presently provides calibration factor and efficiency calibrations at specific frequencies in the 8.2 to 18.0 Gc bands* based on the microcalorimeter just described. This service is being extended to cover the entire microwave region, wherever activity warrants, as quickly as facilities can be implemented. Using standard mounts thus calibrated, it becomes practical to test thermistor mounts in quantity using the technique described in Section III under "Methods for Improving Accuracy".

*NBS Calibrations in the 5.85 to 8.2 Gc band are based upon interim standards until the microcalorimeter for that band is operational.

OTHER METHODS OF AVERAGE POWER MEASUREMENT

Bolometric and calorimetric techniques are universally the most common in microwave power measurements because they are absolute. However, a number of simple, indirect techniques based on bolometer or calorimeter calibrations have been devised which are of interest. A few of these methods are subsequently described, to complete the discussion on instruments for measuring average power at RF.

RECTIFIER METERS.

The fundamental requirement for measuring electrical phenomena is to accurately translate the electrical quantity to some physical quantity which may be observed. The diode offers a simple conversion of RF power to a direct current which may be used to deflect a microammeter. The meter scale may be calibrated in terms of power by applying known input levels and noting the meter deflection.

Figure 2-26 shows the circuit of a typical "termination wattmeter" which employs a semiconductor diode. Operation is simple and easily explained. The majority of input power is dissipated in a coaxial load resistor R_O which ideally is equal to the characteristic impedance of the power source, e.g., 50 ohms. In some wattmeters a small sample of input power is tapped from the load resistor and fed to the diode detector as shown in the Figure. Others use a simple pickup loop to sample the input power. After detection, the RF component is bypassed to ground through C1 and the DC component deflects the microammeter which is calibrated in watts. Switch selected resistors R1 and R2, are meter shunts that allow a choice of two full-scale power ranges, typically 0 - 150 and 0 - 500 watts. The load resistor R_O is a large film type immersed in an oil bath to dissipate the large input power. The impedance of R_O is controlled by its design and manufacture for a maximum SWR of about 1.2:1 over the instrument's operating frequency range of typically 20 Mc to 1000 Mc.

Figure 2-27 shows a bi-directional power monitor also using a diode detector. Here, RF may be applied to either of two coaxial connectors and pass unattenuated through the primary line to some external load. A sample of the input power is loosely coupled from the

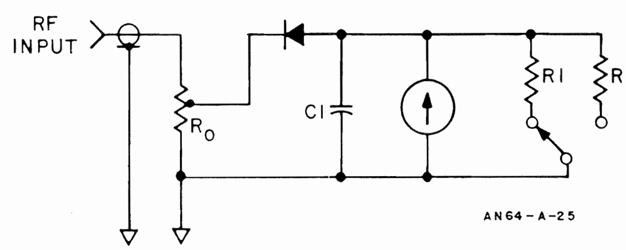


Figure 2-26. Basic circuit of a dual range termination wattmeter employing a semi-conductor diode. Typical wattmeters of this type operate up to 1 Gc at power levels of 150 and 500 watts full scale.

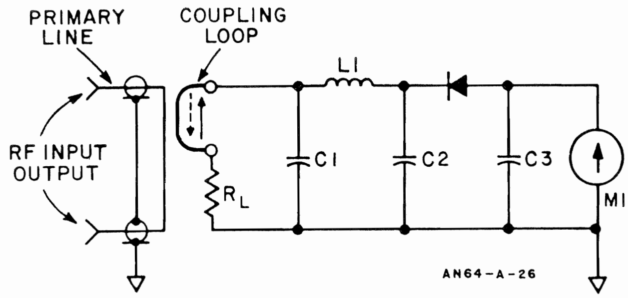


Figure 2-27. Circuit diagram of a bi-directional power monitor. Resistive-loop coupler rotates in 180° steps to enable forward and reverse power monitoring in the primary line.

primary line by a resistive-loop coupler and fed to the diode through the low-pass filter formed by L1, C1 and C2. In order to maintain the physical size within convenient limits, the coupling loop is much shorter than a quarter wavelength at the operating frequencies involved. For this reason coupling varies greatly with frequency. The low-pass filter components are chosen to present a frequency versus output curve approximately equal and opposite the coupling loop curve thereby cancelling the frequency sensitivity over several octaves. The primary line and meter are usually supplied as a basic unit with a number of plug-in elements (containing the loop, filter, and diode) available in various frequency and power ranges from about 30 to 1000 Mc at 1 to 1000 watts average. Because of the use of a resistive loop coupler, the system is ideally sensitive to power flowing in one direction only in the primary line. By rotating the loop axis 180° with respect to the primary line, power flowing in the reverse direction only may be observed which gives an indication of SWR at the load. Bi-directional and termination wattmeters generally have a stated accuracy of 5% of full scale (excluding mismatch).

THERMOCOUPLE POWER METERS

The use of high frequency thermocouples for measuring power has the same attractive feature as diodes in that only a simple DC millivolt meter is needed to indicate the detector output. Most high frequency thermocouples are of the direct heating variety which have a resistive wire heating element situated between two thermocouple wires forming a series circuit. As RF current is passed through the circuit, power is dissipated in the heating element causing a thermal voltage to be developed by the thermocouple wire. The millivolt output may be calibrated by applying known low frequency power levels to the heater element.

There are a number of disadvantages, however, which have precluded significant use of this technique at microwave frequencies. One is the problem of impedance matching the thermocouple to the transmission line. As in the unbalanced bolometer bridge, element resistance varies with applied power resulting in a mismatch error partially dependent on power level. This makes accurate measurements difficult over any appreciable dynamic range. Another problem is that the thermocouple heater must have very small diameter to minimize skin effects so the output is not frequency sensitive. With small heater diameter, the element operates very near burnout and will not withstand even short duration or transient overloads of more than about 50%.

The use of thin film techniques is a recent approach to microwave power measurement with thermocouples. This has led to a power meter comparable in range to bolometric meters that uses a number of thin film thermoelectric heads for various power ranges in certain frequency bands. Each head uses a thin film metallic load which absorbs the input power. A number of thermo-junctions are formed by thin films of bismuth and antimony which are heated by the metallic load causing a thermal EMF to be developed. The output is then applied to a DC amplifier and meter circuit to indicate power. Unlike a bolometer bridge, the system is open loop and subject to undetected calibration changes due to aging or usage of the thermoelectric heads.

Maximum SWR in commercial coaxial heads is stated to be 1.5:1 and overloads of 2:1 may be absorbed without damage. Accuracy of the DC meter is rated at 1% of full scale.

PEAK POWER METERS

Measurement of peak pulse power has been a frequent requirement in microwave work since the early development of pulse radar. Various approaches to peak pulse power measurement include the following techniques: 1) Average power - Duty cycle, 2) Notch Wattmeter, 3) Direct Pulse, 4) DC-Pulse Power Comparison, 5) Barretter Integration-Differentiation.

The first three methods simply involve the interconnection and use of a number of standard instruments such as crystal detectors, bolometers, and oscilloscopes. For this reason we prefer to describe them as "techniques" of peak power measurement in Section III rather than include them in this discussion as peak power meters.

MODEL 8900B PEAK POWER CALIBRATOR.

The HP Model 8900B is a DC-Pulse Power comparator meter which accurately measures pulsed power up to 200 mw peak. Unlike some systems, the 8900B does not rely on pulse width nor repetition rate measurements for its accuracy. From Equation 1.6 in Section I we see that if the duty cycle of a pulsed CW source were made equal to 1, its average CW power and peak power would be equal. Moreover, there would be no indecision as to τ (pulse width) because of pulse rise and decay time. Equation 1.7 is also satisfied if we can accurately detect the peak value of the pulse and dissipate its power in a constant resistance R. The principle of operation in the 8900B provides the opportunity to use either definition of pulsed power.

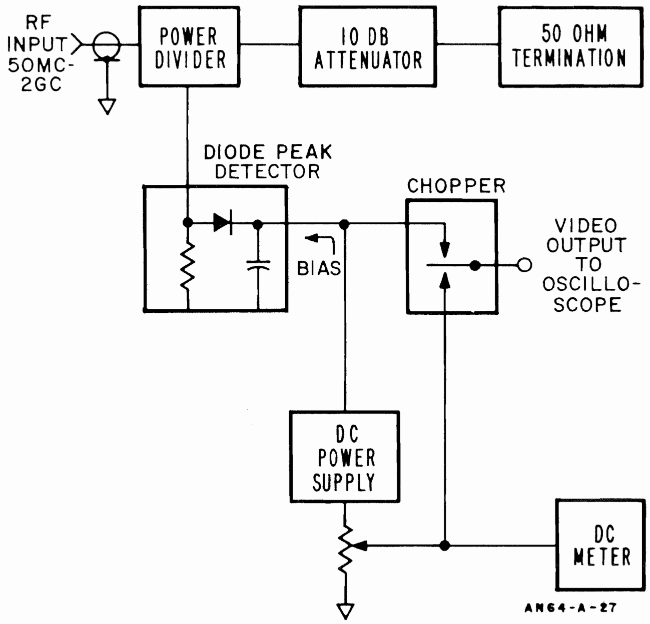


Figure 2-28. Block diagram of HP Model 8900B Peak Power Calibrator. Sample of input power is peak detected and compared with a known DC voltage on an auxiliary oscilloscope. DC meter is calibrated directly in milliwatts of peak power.

Figure 2-28 shows a block diagram of the HP Model 8900B. The power divider splits the input pulse power and feeds part of it to a 10-db attenuator and 50 ohm termination. The remainder of input power is fed to a diode peak detector which develops a DC level proportional to the peak voltage of the input RF pulse. The diode is forward-biased to bring its operating point to maximum stability and away from the square-law region for a more linear relationship in output voltage with applied RF. The peak voltage developed by the diode is connected to one contact of a mechanical chopper. The regulated DC supply that biases the diode also provides a variable comparison voltage which is fed to the other contact of the chopper and a DC meter accurately calibrated in peak power. The center arm of the chopper alternately switches between the detector output and the DC supply allowing an oscilloscope comparison of the two outputs arriving at the video output jack. The static DC bias on the diode is effectively erased from the video output by a front panel null control which permits compensation for long term aging effects in the diode.

In operation, unknown pulsed power is applied to the 8900B input, the envelope peak detected and displayed on an oscilloscope. A DC voltage supplied by the 8900B is adjusted for coincidence with the detected peak pulse amplitude. At this point the DC voltage is equal to the peak pulse voltage input, and the meter indicates the equivalent peak power. An example is given in Section III of how the 8900B is used in a typical application.

The 10-db attenuator and 50 ohm termination built into the 8900B provides an arrangement for convenient calibration of the instrument. If the 50 ohm termination is replaced by an accurate CW power standard such as a bolometer or calorimeter, and a CW power applied to the 8900B input connector, the effect of applied power can be monitored on the average reading CW standard and the peak reading diode detector simultaneously. Therefore, one only needs to know accurately the attenuation between the input connector and the CW standard to determine what effect a known power level has on the peak detector. The 10-db pad reduces input power so bolometric power meters such as the hp 431C and 478A Thermistor Mount may be used for calibration. The hp 434A Calorimeter may also be used for a calibration standard in the same manner.

The HP Model 8900B enables convenient and rapid peak power measurements of pulses greater than 0.25 μsec in duration at RF frequencies of 50 Mc to 2 Gc. PRF's up to 1.5 Mc may be measured because of the wideband detector. The 0.25μsec specification gives the peak detector time to charge to the true peak value of an input pulse and is typically 0.1μsec with normal cable lengths and oscilloscopes connected to the video output. Overall accuracy is ±1.5 db, however, an optional correction chart for frequency effects reduces this error to ±0.6 db maximum. This accuracy is based on an absolute worst-case error analysis which includes the following sources of error: 1) Attenuation measurement between input connector and 10-db pad output, 2) CW power standard, 3) Mismatch, 4) Meter tracking and repeatability, 5) Aging effects on detector diode, and 6) Operator readout.

BARRETTER INTEGRATION-DIFFERENTIATION.

Peak power meters are available which operate in conjunction with barretter mounts and special barretter elements that integrate the pulsed power input due to the barretter's comparatively long thermal time constant (greater than 120μsec).* The barretter forms one leg of a wheatstone bridge that receives excitation and bias current from a constant current supply. Input pulses to the barretter change the element resistance resulting in a bridge output signal which is the integral of the pulses. By amplifying and differentiating the integrated bridge signal, the input pulse shape is reconstructed. The reconstructed pulses are peak detected in a vacuum tube voltmeter circuit calibrated in peak power. Pulse sensitivity characteristics of the barretter must be known in order to translate the relative pulse amplitude to an absolute power level. Calibration of typical peak power meters using the Barretter Integration technique is based on: 1) previous measurements of representative barretter output-per-milliwatt input-per-microsecond (pulse sensitivity) and 2) a calibrating signal generator used to adjust amplifier gain to a predetermined level on the meter before measuring pulse power.

While the characteristics of the amplifier and differentiator are important considerations, the barretter's thermal time constant is the basic limitation to pulse width and repetition rate as well as maximum input power capability. If the pulse input is too narrow, the barretter cannot heat to the true peak value of the input waveform and the bridge signal is reduced into the

*Rudolph E. Henning, Sperry Engineering Review, May-June 1955 noise level of the amplifier. If the pulse is too long with respect to repetition rate (long duty cycle), the barretter builds up a residual temperature which prevents a true on-off ratio in the element resistance. Maximum input power to the barretter is, of course, limited by the physical characteristics of the element.

Typically the pulse width is specified at 0.25μsec minimum to 10μsec maximum with a PRF of 50 to 5000 pulses per second. Pulses up to about 300 mw peak may be handled within the duty cycle specified for the system.

Frequency range is determined by the barretter mounts which are available in common waveguide bands from 2.6 Gc to 12.4 Gc and in coax. Overall accuracy of this system is estimated to be about 17% or 0.8 db in power. Included in the error analysis are: 1) Barretter tolerance, 2) Barretter linearity, 3) Mount efficiency, 4) Bias current, 5) Amplifier linearity, and 6) Calibration signal.

SECTION III
POWER MEASURING TECHNIQUES

INTRODUCTION

Solving a given power measurement problem requires the answers to three basic questions: 1) How much power is to be measured and its classification - average or pulse 2) Over what frequency range, and 3) To what accuracy and precision. The answers to these questions will suggest the instrumentation required. This section of the Application Note is devoted to practical measuring techniques and error reduction using instruments described in the previous section. Even though the basic limitation to measuring power remains in the instrumentation available, much depends upon the care and technique employed in its application.

GENERAL PRECAUTIONS

The following precautions, as applicable, should be observed when making microwave power measurements.

1. When calibrating an instrument, the accuracy of the calibrating system must be greater then the instrument in test. In microwave measurements, the calibrating system generally should have an accuracy of at least three times that of the test instrument.
2. Carefully align waveguide flanges to minimize mismatch loss, and use clamps or bolts to hold mating flanges together.
3. Type N coaxial connectors are commonly encountered in microwave power measurements. Inspect coax connectors often for wear and improper fitting, replacing them when necessary.
4. Avoid the use of coaxial adapters between the measurement point and measuring device whenever possible. If adapters must be used, account for their loss*. The same precaution is especially true for coaxial cables, since the loss can vary with cable flexing.
5. Operate all equipment within manufacturer's specifications. Instruments such as barretters burn out easily when subjected to even momentary overloads. Do not exceed the maximum peak power rating of thermistors or calorimeters when measuring average power in pulsed sources, even though the average power may be within specifications.

NBS AND HP STANDARDS LABORATORY CALIBRATION

The National Bureau of Standards provides a number of services in the microwave region for scientific research and industrial laboratories. NBS does not approve standards, but rather, issues reports of calibration. These reports contain the appropriate calibration data on the instrument submitted, test conditions at the time of measurement, and uncertainty of the data. Table 3-1 shows the services currently available from NBS for microwave power measuring instruments. A complete schedule of NBS services may be obtained for a nominal fee from the Superintendent of Documents, U.S. Government Printing Office, Washington, D.C., 20402.

The HP Standards Laboratory frequently consults with NBS and submits various instruments for calibration. Table 3-2 shows the calibration services available from the HP Standards Laboratory for microwave power measurements. These services are available for special calibrations of new HP instruments before shipment or for suitable instruments already in the field. All calibrations by the HP Standards Laboratory are traceable to NBS to the extent allowed by the Bureau's capabilities. Special calibrations on new HP equipment should be arranged through the local HP Sales Offices when placing an order. Calibrations on existing equipment are handled through the HP Western Service Center at Palo Alto, California, and may be coordinated through any HP Sales Office.

Figure 3-1 shows the NBS traceability of HP production instruments designed for measuring microwave power. Where direct traceability is not yet available, HP maintains interim standards whose characteristics are indicative of high quality through indirect measurements, experimental data, and theoretical considerations.

MEASURING AVERAGE POWER — 10 μWATTS TO 1 WATT

Average power measurements in the 10 μwatt to 10 mw range are nearly always accomplished by bolometric means. Bolometers offer the advantages of high sensitivity, speed, and accuracy. Higher power, sometimes up to 1 kw, may be measured using bolometers if the power is reduced to the bolometer's range with a calibrated directional coupler or attenuator. Couplers and attenuators will reduce accuracy somewhat and their use with bolometers should be considered carefully against the accuracies available using direct calorimetric techniques. This section shows examples of systems measuring power from 10 μw full scale to about 1 watt.

BASIC COAXIAL SYSTEM - 10 MC TO 10 GC.

Figure 3-2a illustrates the simplest setup for measuring power up to 10 mw in a coax system using the hp 431C power meter and 478A thermistor mount. The procedure is as follows:

1. Select the 200Ω position on the MOUNT RES switch of the 431C and energize the power meter.
2. With the source power turned off, connect the thermistor mount to the source.
3. Null and zero the power meter according to the 431C Operating and Service Manual.
4. Switch the power meter to the appropriate range for the maximum power expected and energize the source, reading the average power on the meter.

*A method for measuring the loss in adapters is included later under "Methods for Improving Accuracy".

Table 3-1. NBS Calibration Services Available for RF and Microwave Power

Item	Quantity Measured	Transmission Line	Frequency	Calibration Range	Calibration Accuracy
Bolometer Mount	Effective Efficiency	WR 137	5.85 - 8.2 Gc		2%
		WR 90	8.2 - 12.4 Gc		1%
		WR 62	12.4 - 18.0 Gc	1 - 10 mw	1%
	Calib. Factor	Coax	0.1 & 1.0 Gc		1%
		WR 137	5.85 - 8.2 Gc		2%
		WR 90	8.2 - 12.4 Gc		1%
		WR 62	12.4 - 18.0 Gc	1 - 10 mw	
Bolometer Mount-Directional Coupler Combination	Calib. Factor	WR 137	5.85 - 8.2 Gc	1 - 10 mw in Auxiliary Arm Bolometer Mount for 10 - 3 db directional coupler *	2%
		WR 90	8.2 - 12.4 Gc		1%
		WR 62	12.4 - 18.0 Gc		
Calorimetric Power Meters, Power Measuring Bridges, Absorption Power Meters Feed-through Power Meters	Power	Coax	100, 300 kc 1, 3, 10, 30 Mc	1 mw 200 w	2%
			100, 200, 300, 400 Mc	1 mw - 100 w	
Dry Calorimeter	Output voltage vs. input power	WR 90	8.2 - 12.4 Gc	1 - 100 mw	1%

* Add 1% error for 0.1 mw in auxiliary arm bolometer mount for 20 db directional coupler.

The chart in Figure 3-2a lists sources of error and examples of error limits for both "conjugate available" and "Z_0 available" power measurements. Note that before "correction" there are definite limits of error possible, i.e., the meter reading can be in error up to a certain percentage, depending on the magnitude of each error source. After "correction" the measured power is known with an uncertainty depending on how accurately each source of error is known. Examples of how to correct readings is given later under "Methods for Improving Accuracy."

IMPROVED COAXIAL SYSTEM - 500 MC TO 10 GC.

Figure 3-2b shows an improved setup for coaxial power measurements using a slide screw (or double stub) tuner for minimizing mismatch loss above 500 mc. When using a tuner, follow the same procedure given for Figure 3-2a to Step 4 adjusting the tuner for maximum power reading if conjugate available power is desired. If a Z_0 available power measurement is desired, the tuner must first be adjusted for minimum reflection using a reflectometer or slotted line. Without disturbing the tuner setting, the mount and tuner combination is then connected to the power source and the meter reading noted. The chart in Figure 3-2b shows the tuners' effectiveness in reducing limits of error.

EXTENDED POWER RANGE - 10 MC TO 10 GC.

Figure 3-2c shows two methods of measuring power up to 1 watt in 50 ohm coaxial systems. Either a coaxial attenuator or directional coupler (215 Mc - 4 Gc) is used for reducing power to the thermistor by a known amount. The indication on the 431B power meter is then corrected by accounting for the power loss in the attenuator or coupler-termination. Sources of error in this measuring technique are:

1. Instrumentation error
2. Thermistor mount efficiency
3. Attenuator or coupling factor calibration
4. Tuner loss
5. Mismatch loss

Table 3-2. HP Standards Lab Services Available for Microwave Power

Item	Quantity Measured	Transmission Line	Frequency	Calibration Range	Calibration Accuracy
Power source	Power	50Ω coax	DC - 1 Gc	10 mw - 100 mw	±3%
	Power	50Ω coax	DC - 1 Gc	100 mw - 10 w	±1%
	Power	50Ω coax	10 Mc - 4 Gc	10 μw - 20 mw	±2% (approx)
	Power	50Ω coax	1 Gc - 4 Gc	10 mw - 100 mw	±4% (approx)
	Power	50Ω coax	1 Gc - 4 Gc	100 mw - 10 w	±2% (approx)
	Power	Waveguide	4 Gc - 8 Gc	10 μw - 10 mw	±3% (approx)
	Power	50Ω coax	4 Gc - 8 Gc	10 mw - 100 mw	±5% (approx)
	Power	50Ω coax	4 Gc - 8 Gc	100 mw - 10 w	±3% (approx)
	Power	Waveguide	8 Gc - 12 Gc	10 μw - 10 mw	±2%
	Power	50Ω coax	8 Gc - 12 Gc	10 mw - 100 mw	±4%
	Power	50Ω coax	8 Gc - 12 Gc	100 mw - 10 w	±2%
	Power	Waveguide	12 Gc - 18 Gc	10 μw - 10 mw	±2%
Self balancing bolometer bridge	Full scale accuracy			0.1 - 10 mw 100 & 200Ω Neg, 200Ω Pos.	±2%
Self balancing temp. comp. thermistor bridge				10 μw - 10 mw 100 & 200Ω Neg.	±1/2%
Bolometer Mount (100 or 200Ω Pos. or Neg.)	Effective Efficiency or Calib. Factor	50Ω coax (type N connector)	0.1, 1.0, 3.0, 6.45, 7.00, 7.40, 8.2, 8.5, 9.0, 9.8, 10.0, 10.3, 11.0, 11.2, 12.0, 12.4 Gc		±2% (±2.5% @ 3 Gc)
Bolometer Mount (100 or 200Ω Pos. or Neg.)	Effective Efficiency or Calib. Factor	Waveguide; WR 137, WR 112, WR 90, WR 75, WR 62, WR 51	6.45, 7.00, 7.40, 8.2, 8.5, 9.0, 9.8, 10.0, 10.3, 11.0, 11.2, 12.0, 12.4, 13.0, 13.5, 14.0, 15.0, 16.0, 16.5, 17.0, 17.5, 18.0 Gc	1 or 10 mw	±1.5%
3 db Directional coupler/thermistor mount combination	Calibration Factor				±2%
Calorimetric Power Meter	Effective Efficiency or Calibration Factor	50Ω coax (Type N Connector)	8.2, 9.0, 9.8, 10.125, 11.2, 12.2 Gc	1 Watt	2%

The chart in Figure 3-2c relates these sources of error and typical values which may be expected.

POWER IN WAVEGUIDE SYSTEMS - 2.6 GC TO 40 GC.

Figure 3-3 breaks power measurements in waveguide systems into the same three categories shown for coaxial systems. These categories are illustrated as follows:

Figure 3-3a: A basic setup for measurements to 10 mw in waveguide bands from 2.6 Gc to 40 Gc.

Figure 3-3b: An improved setup over that shown in Figure 3-3a using a tuner for eliminating mismatch loss.

Figure 3-3c: Measuring power up to 1 watt using a directional coupler to reduce power to the thermistor mount.

The measurement procedure is as follows:

1. Select the proper thermistor operating resistance on the 431C MOUNT RES switch, and energize the meter. (The thermistor operating resistance is marked on each HP thermistor mount.)

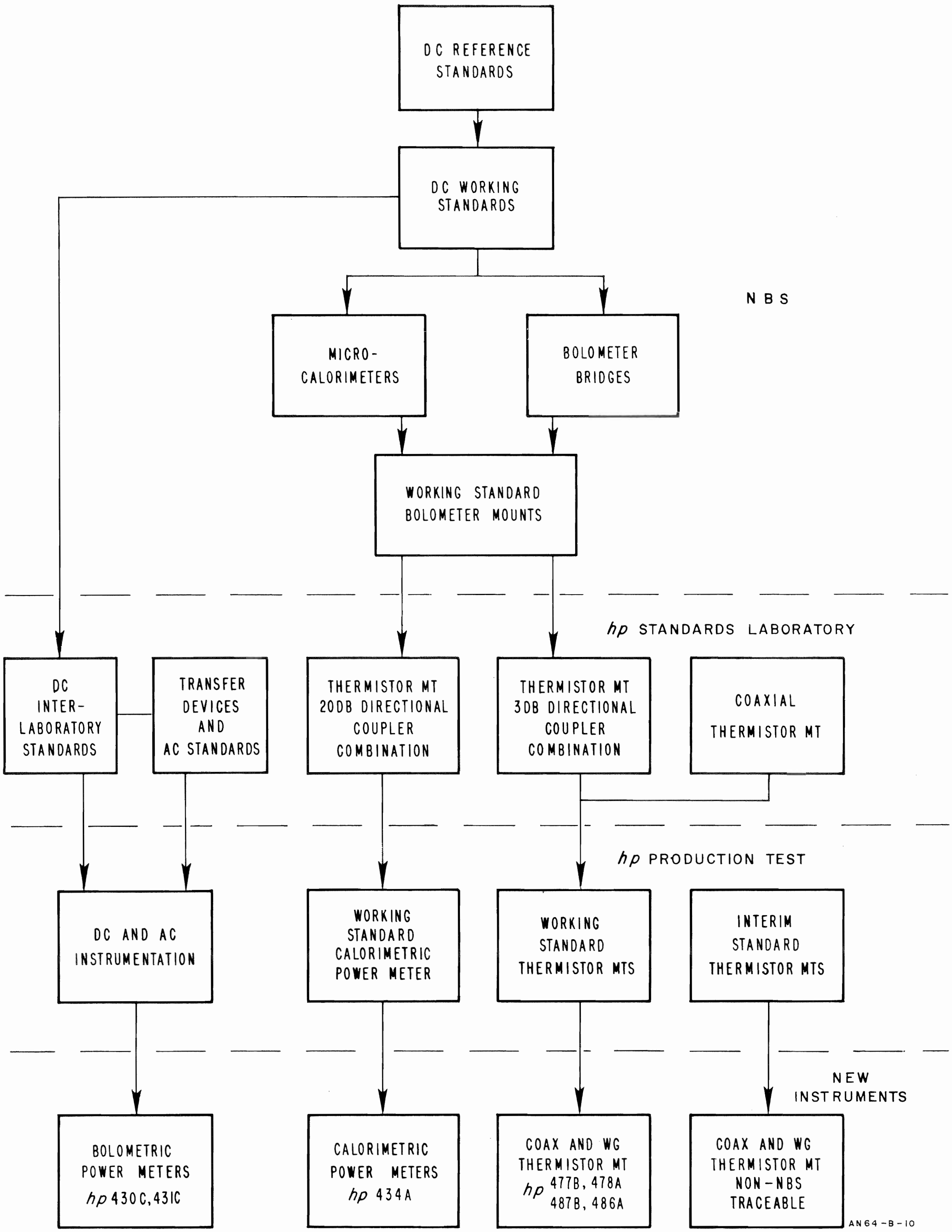
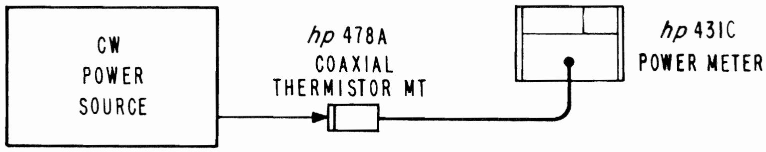


Figure 3-1. Calibration traceability of HP power measuring instruments from production line to NBS. Regular calibration of HP standards assures valid measurements on the production line and adherence to published specifications.



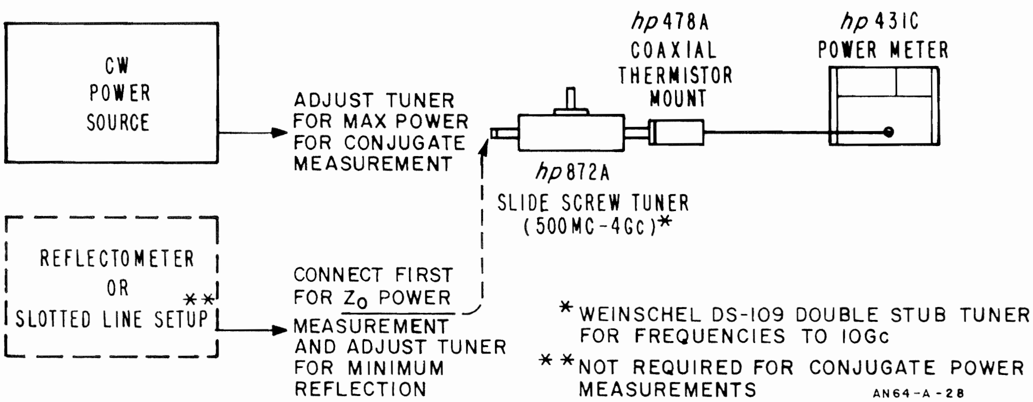
$$P_c = \frac{P_{indicated}\,(1 \pm \rho_m \rho_g)^2}{K_b\,(1 - \rho_g^2)} \qquad (3.4)$$

$$P_o = \frac{P_{indicated}\,(1 \pm \rho_m \rho_g)^2}{K_b} \qquad (3.7)$$

Error Source	Conjugate Available Power Measurement		Z_o Available Power Measurement	
	Value Measured	Uncertainty	Value Measured	Uncertainty
Instrumentation (431C)	1 mw	±3%*	1 mw	±3%*
Calib. Factor (K_b) of Thermistor Mount (478A)	.944 (±2%)	-3.6 to -7.6%	.944 (±2%)	-3.6 to -7.6%
Mismatch Loss ρ_g = .26 (SWR 1.7) ρ_m = .13 (SWR 1.3)		-1.7 to -14%		+8 to -7%
Limits of Error Before Correction	-2.3 to -24.6%		+7.6 to -17.4%	
Total Uncertainty After Correction	+3.3 to -19%		+13 to -12%	

*May be reduced to ±0.5% with D.C. substitution in 431C.

Figure 3-2a. Basic system for coaxial power measurements of 10 μw to 10 mw full scale, 10 Mc - 10 Gc. Chart lists sources of error and examples of their effects on the measurement. Equations show relationship of indicated power to conjugate available power (P_c) and Z_o available power (P_o).



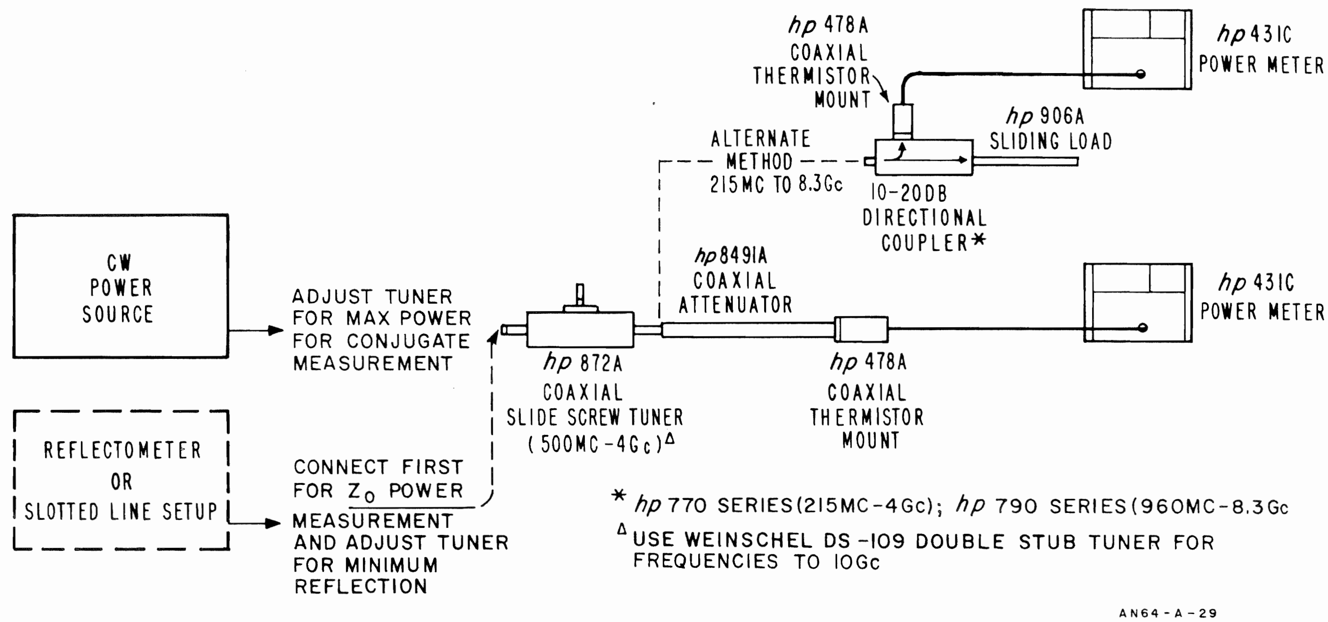
$$P_c = \frac{P_{indicated}}{T_L(\eta_e)} \qquad (3.6)$$

$$P_o = \frac{P_{indicated}}{T_L(\eta_e)} \qquad (3.9)$$

Error Source	Conjugate Available Power Measurement		Z_o Available Power Measurement	
	Value Measured	Uncertainty	Value Measured	Uncertainty
Instrumentation (431C)	1 mw	±3%*	1 mw	±3%*
Effective Efficiency (η_e) of Thermistor Mount (478A)	.96 (±2%)	-2 to -6%	.96 (±2%)	-2 to -6%
Tuner Loss Ratio (T_L)	.99	-1%	.99	-1%
Limits of Error Before Correction	0 to -10%		0 to -10%	
Total Uncertainty After Correction	±5%		±5%	

*May be reduced to ±0.5% with D.C. substitution in 431C.

Figure 3-2b. Improved system for coaxial power measurements 500 Mc to 10 Gc. Tuner eliminates mismatch between source and thermistor mount, reducing uncertainty of measurement. Equations are simplified by removal of reflection coefficient terms.



$$P_c = \frac{P_{\text{indicated}}}{T_L(\eta_e)\left[\log_{10}^{-1}\left(\frac{-DB}{10}\right)\right]}$$

$$P_o = \frac{P_{\text{indicated}}}{T_L(\eta_e)\left[\log_{10}^{-1}\left(\frac{-DB}{10}\right)\right]}$$

Error Source	Conjugate Available Power Measurement		Zo Available Power Measurement	
	Value Measured	Uncertainty	Value Measured	Uncertainty
Instrumentation (431C)		±3%*		±3%*
Effective Efficiency of Thermistor Mount (478A)	.96 (±2%)	-2 to -6%	.96 (±2%)	-2 to -6%
Attenuation or Coupling Factor of Power Reducer (DB)	20 db	±0.2 db ±(4.5%)	20 db	±0.2 db ±(4.5%)
Tuner Loss Ratio	.99	-1%	.99	-1%
Mismatch Loss Coupler or Pad $\rho = .09$ Thermistor Mount $\rho = .13$		-.15 to -4.8%		-.15 to -4.8%
Limits of Error Before Correction	+4.35 to -19.3%		+4.35 to -19.3%	
Total Uncertainty After Correction	+9.35 to -14.3%		+9.35 to -14.3%	

*May be reduced to 0.5% with D.C. substitution in 431C.

Figure 3-2c. Extended range coaxial system uses fixed attenuator or directional coupler to reduce power at thermistor. System above will handle up to 1 watt average. Equations include coupling factor or attenuation between source and thermistor mount.

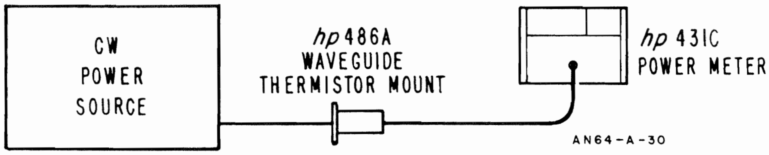
2. Null and zero the 431C power meter.

3. Select the appropriate power range on the 431C for the maximum power expected.

4. Connect the equipment as shown in Figure 3-3 appropriate to the particular measurement situation, and energize the power source.

5. Read average power on the 431C for setup 3-3a. For setup 3-3b and 3-3c, adjust the tuner for a maximum power indication on the 431C if conjugate available power is desired. If Z_0 available power is desired, adjust the tuner-mount combination for minimum reflection on a reflectometer or slotted line before connecting to the source in test. The charts in each of the Figures 3-3a through 3-3c show examples of errors and their overall effect on the measurement uncertainty.

Hewlett-Packard waveguide equipment is identified with its frequency range and waveguide size by a letter prefix to each model number. For example, suppose



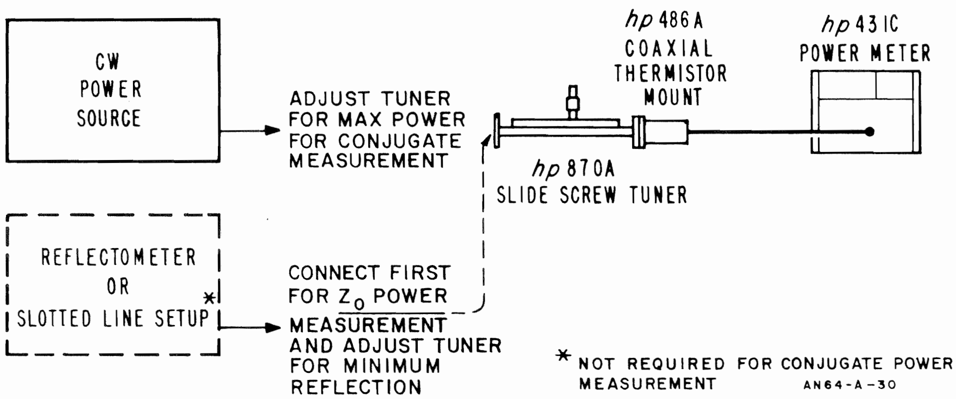
$$P_c = \frac{P_{indicated}\,(1 \pm \rho_m \rho_g)^2}{K_b(1-\rho_g^2)} \qquad (3.4)$$

$$P_o = \frac{P_{indicated}\,(1 \pm \rho_m \rho_g)^2}{K_b} \qquad (3.7)$$

Error Source	Conjugate Available Power Measurement		Z_0 Available Power Measurement	
	Value Measured	Uncertainty	Value Measured	Uncertainty
Instrumentation (431C)	-	±3%* Max	-	±3%* Max
Calib. Factor (K_b) of Thermistor Mount (486A)	.975 (±2%)	-0.5 to -4.5%	.975 (±2%)	-0.5 to -4.5%
Mismatch Loss ρ_g = .05 (SWR = 1.10) ρ_m = .07 (SWR = 1.15)	-	-.05 to 1.38%	-0.7%	0 to -1.4%
Limits of Error Before Correction	+2.45 to -6.38%		+2.5 to -8.9%	
Total Uncertainty After Correction	+4.95 to -6.38%		±5.7%	

*May be reduced to ±.15% with D.C. substitution in 431C.

Figure 3-3a. Basic system for waveguide power measurements of 10 μw to 10 mw full scale, 2.6 Gc to 40 Gc.



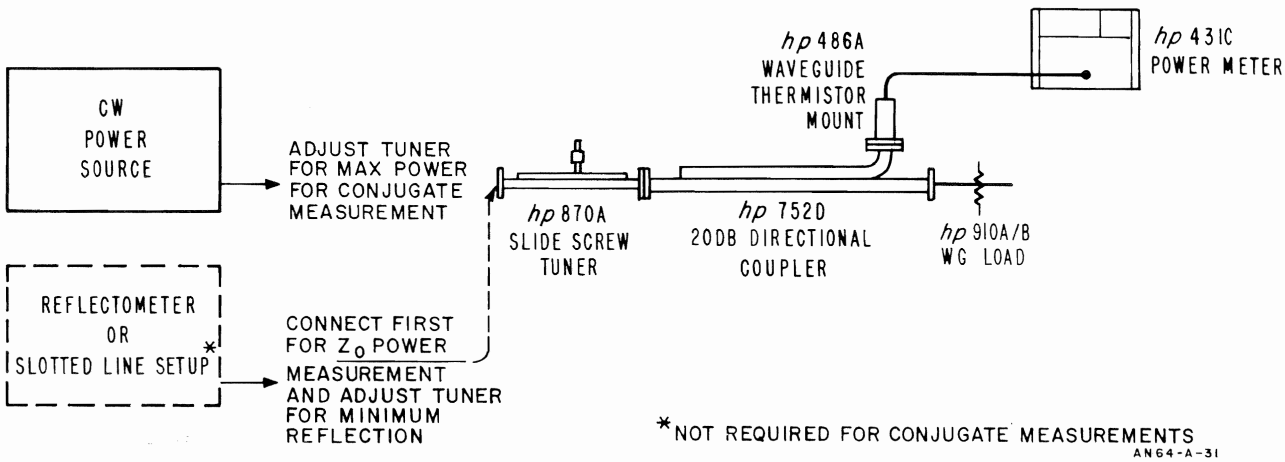
$$P_c = \frac{P_{indicated}}{T_L(\eta_e)} \qquad (3.6)$$

$$P_o = \frac{P_{indicated}}{T_L(\eta_e)} \qquad (3.9)$$

Error Source	Conjugate Available Power Measurement		Z_0 Available Power Measurement	
	Value Measured	Uncertainty	Value Measured	Uncertainty
Instrumentation (431C)		±3%*Max		±3%*Max
Effective Efficiency (η_e) of Thermistor Mount (486A)	.98 (±2%)	0 to -4%	.98 (±2%)	0 to -4%
Tuner Loss Ratio (T_L)	.993	-0.7%	.993	-0.7%
Limits of Error Before Correction	+2.3 to -7.7%		+2.3 to -7.7%	
Total Uncertainty After Correction	±5%		±5%	

*May be reduced to ±0.15% with D.C. substitution in 431C.

Figure 3-3b. Improved waveguide system using tuner to minimize mismatch loss.



$$P_c = \frac{P_{indicated}}{T_L(\eta_e)\left(\log_{10}^{-1}\frac{-DB}{10}\right)}$$

$$P_o = \frac{P_{indicated}}{T_L(\eta_e)\left(\log_{10}^{-1}\frac{-DB}{10}\right)}$$

Error Source	Conjugate Available Power Measurement		Z_0 Available Power Measurement	
	Value Measured	Uncertainty	Value Measured	Uncertainty
Instrumentation (431C)	-	±3%* Max	-	±3%* Max
Effective Efficiency (η_e) of Thermistor Mount (486A)	.98 (±2%)	0 to -4%	.98 (±2%)	0 to -4%
Coupling Factor (DB)	20 db	±0.2 db ±(4.5%)	20 db	±0.2 db ±(4.5%)
Tuner Loss Ratio (T_L)	.997	-0.3%	.997	-0.3%
Mismatch Loss Coupler Aux. Arm ρ = .07 Thermistor Mount ρ = .07	-	-0 to -1.9%	-	-0 to -1.9%
Limits of Error Before Correction	+7.2 to -13.7%		+7.2 to -13.7%	
Total Uncertainty After Correction	+9.5 to -11.4%		+9.5 to -11.4%	

*May be reduced to ±.15% with D.C. substitution in 431C.

Figure 3-3c. Waveguide system for extending 431C power range to 1 watt, 2.6 Gc to 40 Gc. Greater power may be measured by substituting directional couplers of higher coupling factor (>20 db) and waveguide loads of higher power rating than shown. HP 750E-series 30 db cross-guide couplers are suitable in this application, for frequencies of 2.6 to 12.4 Gc.

the power source in Figure 3-3b has an X-band waveguide output system operating over a frequency range of 8.2 to 12.4 Gc. The appropriate slide screw tuner and waveguide thermistor mount in Figure 3-3b would have model numbers of X870A and X486A respectively. Table 3-3 lists HP, EIA and JAN waveguide designations for the various frequency ranges.

METHODS FOR IMPROVING ACCURACY

It should be particularly noted that the types of error listed in Figures 3-2 and 3-3 are common to all bolometric power measurements. The values depend entirely on the equipment being used and the care exercised in its use. The compatibility of HP coaxial and waveguide equipment, plus instrumentation versatility, allows greater accuracy through special techniques. In the following discussion we will show how to systematically eliminate or reduce the effects of each source of error listed using this compatibility and versatility.

DC SUBSTITUTION TO REDUCE INSTRUMENTATION ERROR.

The error in measuring substituted power in a thermistor element can be greatly reduced by using DC substitution techniques. The provision for DC substitution is a significant feature of the hp 431C power meter, whereby instrumentation error may be reduced from ±3% of full scale to ±0.5% of reading. Briefly, the technique involves: 1) applying the unknown RF power to the thermistor mount and noting the 431B power meter reading 2) removing the RF power from the thermistor and substituting DC current from an external source for a 431C reading equal to that noted in Step 1, then 3) computing substituted DC power from the DC current and thermistor operating resistance.

Table 3-3. Standard Waveguide Specifications

DESIGNATIONS			DIMENSIONS	TE_{10} OPERATING RANGE			FREESPACE WAVELENGTH
hp	EIA	JAN	ID (Inches)	Frequency (Gc)	Wavelength (cm)	Cutoff Freq. (Gc)	(cm)
S	WR 284	RG-48/U	2.840 x 1.340	2.60 - 3.95	19.18 - 8.92	2.078	11.53 - 7.59
G	WR 187	RG-49/U	1.872 x 0.872	3.95 - 5.85	12.59 - 6.08	3.152	7.59 - 5.12
C	WR 159	———	1.590 x 0.795	4.90 - 7.05	9.37 - 5.01	3.711	6.12 - 4.25
J	WR 137	RG-50/U	1.372 x 0.622	5.30 - 8.20	9.68 - 4.29	4.301	5.66 - 3.66
H	WR 112	RG-51/U	1.122 x 0.497	7.05 - 10.0	6.39 - 3.52	5.259	4.25 - 3.00
X	WR 90	RG-52/U	0.900 x 0.400	8.20 - 12.4	6.09 - 2.85	6.557	3.66 - 2.42
M	WR 75	———	0.750 x 0.375	10.0 - 15.0	4.86 - 2.35	7.868	3.00 - 2.00
P	WR 62	RG-91/U	0.622 x 0.311	12.4 - 18.0	3.75 - 1.96	9.487	2.42 - 1.67
K	WR 42	RG-66/U	0.420 x 0.170	18.0 - 26.5	2.66 - 1.33	14.050	1.67 - 1.13
R	WR 28	RG-96/U	0.280 x 0.140	26.5 - 40.0	1.87 - 0.88	21.075	1.13 - 0.749

Figure 3-4a shows the 431C set up for a DC substitution measurement. The hp 8402B Power Meter Calibrator provides DC power and switching logic to perform substitution measurements with the 431C conveniently. However, any well regulated DC supply capable of providing 20 ma into a 10K ohm load may be used in place of the hp 8402B, if desired. Figure 3-4b illustrates how power supplies may be connected for DC substitution. The 10K ohm series resistor prevents the low output impedance of the power supply from shunting the 431C bridge circuit.

<u>DC Substitution Procedure Using HP 8402B Calibrator.</u>

1. Connect the equipment as shown in Figure 3-4a with the digital voltmeter (DVM) input connected across the 1K ohm resistor (1K ohm ±10% @ 1W). This arrangement allows the 431C readings to be duplicated with greater precision than possible with direct meter indication.

2. With the RF source off, set the 8402B OUTPUT CURRENT to OFF and energize the power meter, calibrator, and digital voltmeter.

3. Null and zero the 431C power meter, and select the desired power range.

4. Set the 8402B FUNCTION switch to SUB and RANGE (mw) to the 431C power range in use.

5. Apply unknown RF power to the thermistor mount and <u>note the DVM reading</u> across the 1K ohm resistor.

6. Remove the RF power from the thermistor mount.

7. Set the 8402B CURRENT CONTROL fully counterclockwise and turn OUTPUT CURRENT on.

8. Increase the 8402B CURRENT CONTROL and VERNIER until the DVM indicates the same reading as noted in Step 5.

9. Move the DVM input leads from the 1K ohm resistor to the VOLTMETER (ISOLATED) jack of the 8402B and note the reading. The substituted DC current in milliamps is equal to the DVM reading in volts when the 8402B is on the .01 to 1 mw power range, and 10 times the DVM reading in volts for the 3 and 10 mw ranges.

10. Calculate the substituted DC power to the thermistor (which is now equal to the RF power previously in the thermistor) by the formula:

$$P_{dc} = \frac{I_{ma}^2 R_m}{4 \times 10^3} \qquad (3.1)$$

where

I = current measured in Step 9.
R_m = actual thermistor operating resistance in ohms.*

*Note: The calculation in Step 10 may be carried out using the nominal thermistor operating resistance marked on the thermistor mount which may cause approximately 0.2% error in the calculation. For best accuracy, use the actual measured value of mount resistance. The 8402B Operating and Service Manual includes directions for measuring mount resistance.

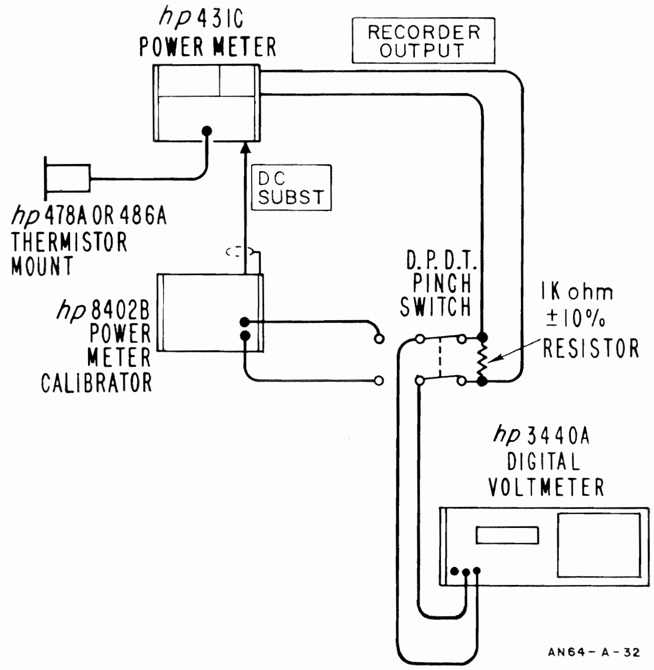


Figure 3-4a. DC Substitution reduces instrument error to ±0.5% using hp 431C Power Meter and 8402B Power Meter Calibrator. DVM monitors recorder output of 431C during application of RF power to thermistor mount. Reading is duplicated by removing RF and applying equivalent DC Substitution power. DVM is then switched to read DC substituted.

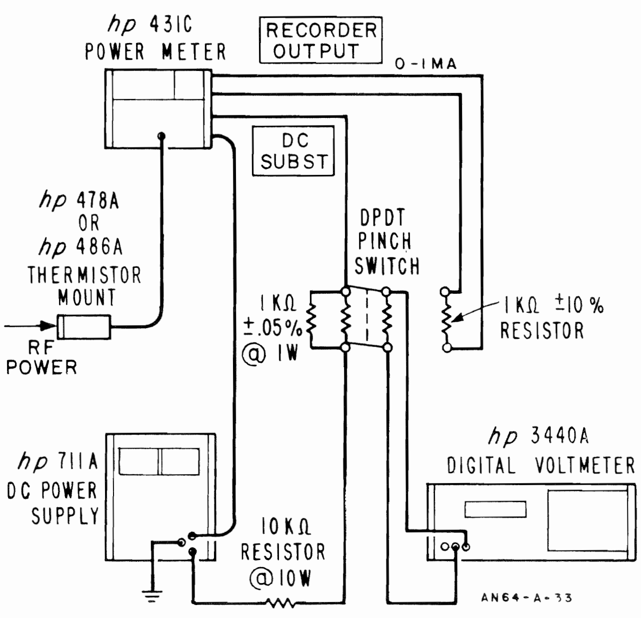


Figure 3-4b. DC Substitution up to 10 mw is possible using hp 431C Power Meter and regulated DC supply as shown. Accuracy of DC substitution power is ±0.5%.

DC Substitution Procedure Using a DC Power Supply.

1. Connect the equipment as shown in Figure 3-4b with the DVM connected across the RECORDER resistor (1K ohm ±10% @ 1W).

2. With the RF source off and the power supply DC VOLTAGE turned off, energize the test equipment.

3. Null and zero the 431C power meter and select the desired power range.

4. Apply unknown RF power to the thermistor mount and note the DVM reading.

5. Remove RF power from the thermistor mount.

6. With the power supply DC VOLTS ADJ at minimum (CCW), turn DC VOLTAGE on.

7. Increase the power supply DC VOLTS ADJ coarse and fine controls until the DVM indicates the reading noted in Step 4.

8. Move the DVM input leads from the 431C recorder load resistor to the 1K ohm monitor resistor shown in Figure 3-4b.

9. Note the DVM reading. Substituted DC current, in milliamps, is equal to the DVM reading in volts.

10. Calculate the substituted DC power to the thermistor (which is now equal to the RF power previously in the thermistor) using equation 3.1.

The thermoelectric error discussed in Section II under "Accuracy Considerations" can be significant on the two most sensitive ranges of the 431C when making DC substitution measurements. Typically the error is 0.1 μw and never exceeds 0.3 μw in HP production thermistor mounts. When operating on the 431C 10 μw or 30 μw range, correction for thermoelectric error may be made as follows:

1. Apply 10 μw from the 8402B to the 431C DC SUBST jack and record the power meter reading as P_1.

2. Reverse the connection at the DC SUBST jack and note the power meter reading as P_2.

3. Calculate the thermoelectric correction factor -

$$\text{correction } (\mu w) = \frac{P_1 - P_2}{2} \qquad (3.2)$$

If DC Substitution is used, add the thermoelectric correction to all power meter readings when operating on the 10 and 30 μw ranges of the 431C. Be sure to leave the polarity of the DC Substitution connection the same as in Step 2 so the sign of the correction factor will be correct.

BOLOMETER MOUNT EFFICIENCY AND MISMATCH CORRECTION

Corrections for bolometer mount error are by far the most important in achieving accurate power measurements. As may be noted from the systems shown in Figure 3-2 and 3-3, a power measurement may fall into one of four general classifications: These include measurement of:

1. Conjugate available (maximum) power, without using a tuner to eliminate mismatch loss between source and bolometer mount.
2. Conjugate available power, using a tuner to eliminate mismatch loss.
3. Z_0 available power, without using a tuner to eliminate bolometer mount reflection.
4. Z_0 available power, using a tuner to eliminate mount reflection.

Each situation requires the use of specific bolometer mount and source data in order to correct readings observed on the power meter. Each temperature compensated HP thermistor mount has both its effective efficiency and its calibration factor recorded on a special label affixed to the mount. The following four examples illustrate the simplest usage of these data in the four measurement situations just described. Conversion from effective efficiency to calibration factor may be performed if only one set of data is known using the equation

$$K_b = \eta_e(1 - \rho_m^2) \qquad (3.3)$$

where

ρ_m = Reflection coefficient magnitude of the bolometer mount.

Example 1: Conjugate Power Measurement Without Tuner

When a tuner is not used, the indicated power may be corrected using reflection coefficient data and the thermistor mount Calibration Factor (K_b). The corrected reading will be the actual power available to a conjugate load, with an uncertainty determined by the source and thermistor mount reflection coefficient magnitudes. The equation for determining conjugate available power when not using a tuner is:

$$P_c = \frac{P_{indicated}\left(1 \pm \rho_m \rho_g\right)^2}{K_b\left(1 - \rho_g^2\right)} \qquad (3.4)^*$$

where

P_c = Conjugate available power.

K_b = Calibration factor of thermistor mount

ρ_g and ρ_m = Reflection coefficient magnitude of source and thermistor mount respectively.

*Derivation of Equation 3.4 is given in Appendix I.

Under conditions shown in Figure 3-2a for conjugate available power, we may apply equation (3.4) for correcting a power meter reading of 1 milliwatt for this example.

$$P_c = \frac{1\text{ mw}\left[1 \pm (.13)(.26)\right]^2}{.944\left[1 - (.26)^2\right]} = \begin{matrix}1.06\text{ mw to}\\ 1.21\text{ mw}\end{matrix}$$

Note the uncertainty in the corrected value due to the unknown phase relationship of ρ_g and ρ_m. Total uncertainty of the corrected power meter indication depends upon the uncertainty of each term substituted in the equation. Typical values of each uncertainty are shown in the chart of Figure 3-2a along with the resulting limits of error before correction, and overall uncertainty after correction.

The conjugate mismatch loss chart in Figure 2-12 (or 2-13) simplifies this correction by eliminating all terms in equation 3.4 containing ρ. The chart also accounts for the $(1-\rho_m^2)$ portion of K_b* leaving only the effective efficiency term. Thus, the correction becomes

$$P_c = \frac{P_{indicated}}{\eta_e(1 - \text{chart loss})} \qquad (3.5)$$

Continuing the example using the loss chart,

$$P_c = \frac{1\text{ mw}}{.96\left(1 - \begin{matrix}.017\\ .14\end{matrix}\right)} = \begin{matrix}1.06\text{ mw to}\\ 1.21\text{ mw}\end{matrix}$$

Example 2: Conjugate Power Measurement Using Tuner

When a slide screw or multiple stub tuner is used to maximize power transfer from source to thermistor mount, the effective efficiency (η_e) of the mount is used to correct the meter reading for conjugate available power. Tuner loss will be small in most cases. However, its effects on the measurement can be included in the correction if the loss is known.

The equation for determining conjugate available power when using a tuner is:

$$P_c = \frac{P_{indicated}}{T_L \eta_e}, \qquad (3.6)$$

where

P_c = Conjugate available power
T_L = Tuner loss ratio, P_{out}/P_{in}
η_e = Thermistor mount effective efficiency

*See equation 3.3

Under conditions shown in Figure 3-2b for conjugate available power, equation 3.6 may be applied to a power meter reading of say 1 mw as follows:

$$P_c = \frac{1 \text{ mw}}{(.99)(.96)} = 1.05 \text{ mw}$$

Note there is no mismatch uncertainty in the corrected value because mismatch has been tuned out. Uncertainties of each term substituted into equation 3.6 must be evaluated, however, to determine total measurement uncertainty. Typical figures are shown in the chart of Figure 3-2b for conjugate available power. Note the marked improvement when using a tuner. If DC substitution was used to reduce instrumentation error to 0.5% the overall uncertainty would decrease to essentially the thermistor mount η_e calibration uncertainty or 2%. Going a step further, one could obtain NBS calibration on the mount decreasing η_e calibration uncertainty to 1%. Overall measurement uncertainty in this example could therefore approach 1% using appropriate techniques with the equipment shown.

Example 3: <u>Z_o Available Power Measurement Without Tuner</u>

When a tuner is not used to cancel the thermistor mount reflection, the Z_o available power may be determined using the following equation:

$$P_o = \frac{P_{indicated}\,(1 \pm \rho_m \rho_g)^2}{K_b}, \qquad (3.7)^*$$

where

P_o = Z_o available power
K_b = Calibration factor of thermistor mount
ρ_m and ρ_g = Reflection coefficient magnitude of mount and source respectively

Assuming the conditions shown in Figure 3-2a for Z_o available power with an indicated power of 1 mw, the equation becomes:

$$P_o = \frac{1 \text{ mw}\,[1 \pm (.13)(.26)]^2}{.944}$$

$$= .99 \text{ mw to } 1.13 \text{ mw.}$$

Note the mismatch uncertainty resulting from the unknown phase relationship of ρ_g and ρ_m. <u>Total</u> uncertainty in this example is given by the table of Figure 3-2a for Z_o available power.

By using the "uncertainty" portion of the Z_o mismatch chart of Figure 2-14 the correction can be simplified by eliminating all terms of equation 3.7 containing ρ. The correction for this example then becomes:

$$P_o = \frac{P_{indicated}}{K_b\,(1 \pm \text{chart uncertainty})} \qquad (3.8)$$

Continuing the example for corrections with the Z_o mismatch chart,

$$P_o = \frac{1 \text{ mw}}{.944\left(1 \pm {}^{.072}_{.065}\right)}$$

$$= .99 \text{ mw to } 1.13 \text{ mw}$$

The "Loss" scale is not used in this case because the term K_b includes the mismatch loss associated with the thermistor mount.

Example 4: <u>Z_o Available Power Measurement Using Tuner</u>

When a tuner is used to cancel thermistor mount reflection, the Z_o available power is calculated from the equation:

$$P_o = \frac{P_{indicated}}{T_L \eta_e}, \qquad (3.9)$$

where

P_o = Z_o available power
T_L = Tuner loss ratio, P_{out}/P_{in}
η_e = Effective efficiency of bolometer mount

Using the values shown in Figure 3-2b for Z_o available power and an indicated power of 1 mw,

$$P_o = \frac{1 \text{ mw}}{.99\,(.96)} = 1.05 \text{ mw}$$

As in the previous examples, total uncertainty of the 1.05 mw value is determined by the uncertainty of each term substituted in the equation. As shown in Example 2, total uncertainty could be reduced to 1%, using appropriate techniques.

It is interesting to note how much error could have been encountered if corrections had not been applied to the power meter readings shown in the 4 examples just given. In Example 1, overall measurement error could have been as high as 24.6%. The reading in Example 2 could have been in error by 10%. It should be emphasized that the figures given in these examples are <u>typical</u> using the equipment shown in Figure 3-2 and 3-3. Larger errors could be encountered with

*Derivation of Equation 3.7 is given in Appendix I.

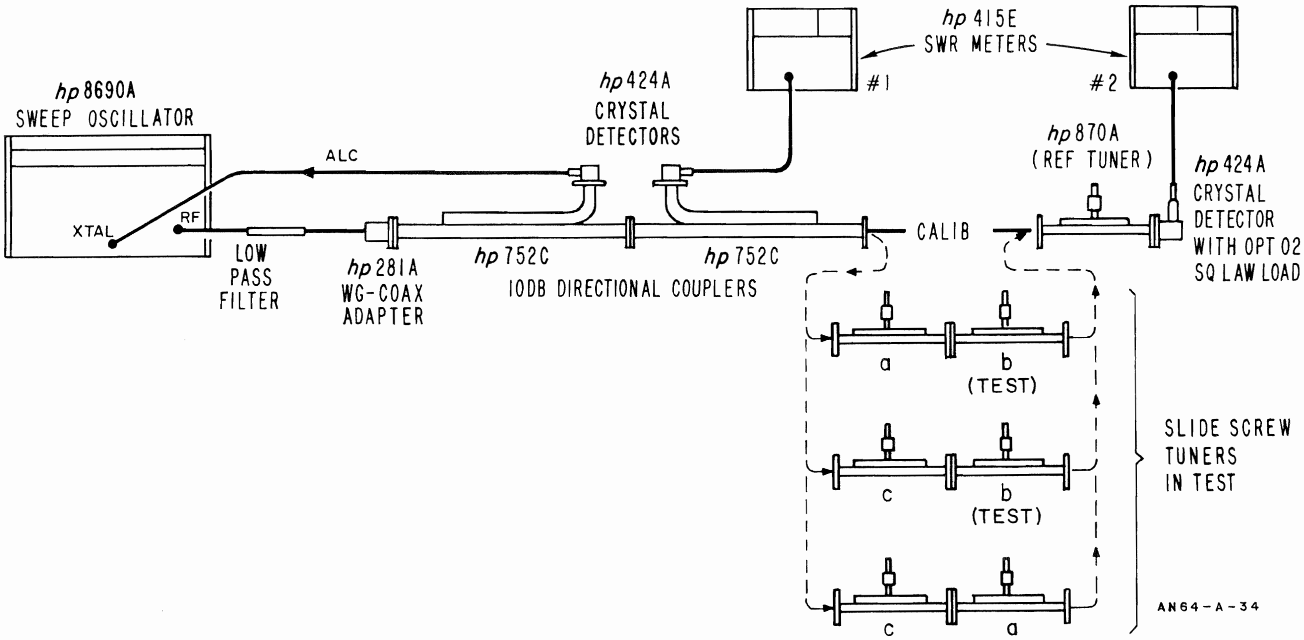


Figure 3-5. Slide screw tuner loss measurement system. After minimizing detector reflections with reference tuner, a reference reading is established on 415E #2. Tuner in test (b) is set to specific depth and carriage position where loss calibration is desired. Mismatch is tuned out with a second tuner (a or c) and DB loss noted on 415E #2. Evaluation of simultaneous equations reveals each individual tuner loss.

defective thermistor mounts or uncalibrated mounts of some other designs.

SLIDE-SCREW TUNER LOSS MEASUREMENT

The following procedure may be used to measure slide-screw tuner loss so its effects on power measurements may be corrected. Tuner loss must be measured at the same settings of probe depth, carriage position and frequency as those used in the actual power measurement being corrected. All HP slide-screw tuners have micrometer depth calibration and precise carriage position scales for accurate duplication of settings when making loss measurements. The calibration of a waveguide tuner is illustrated in this procedure. However, coaxial tuner loss may be measured in a similar manner using the appropriate HP coaxial counterparts to the waveguide equipment shown. A leveled source is used to maintain good source impedance match and maintain constant output power for a stable reference. Leveling is extremely important in this application since the measured loss will be very small and mismatch uncertainties or reference power drift cannot be tolerated. To measure tuner loss set up the equipment as shown in Figure 3-5 and adjust the source for 1 kc AM and leveled operation at the test frequency. Power output from the system should be maintained at -6 dbm to remain in square law on the 424A detector and #2 415E. Proceed as follows:

- With the equipment connected as shown for CALIBRATE, adjust the reference tuner for minimum reflection as indicated by 415E #1.

- Set a convenient reference level on 415E #2 using the GAIN control and expanded DB scale.

- Insert tuners (a and b) as shown. Set tuner (b) to the same settings used for the actual power measurement in which the tuner was involved. Adjust tuner (a) for minimum reflection as noted on 415E #1 using as nearly as possible the same settings of probe depth and carriage position as on tuner (b).

- Note the DB loss on 415E #2.

- Replace tuner (a) with tuner (c) and tune (c) for minimum reflection. Again note DB loss on 415E #2.

- Repeat the same procedure with tuner (a) and (c), adjusting tuner (a) to the same settings as tuner (b).

This test will produce three DB loss values which represent all combinations of tuner pairs. The three losses may be set into a system of three linear equations with three unknowns (Eq. 3.10). Evaluating these equations will reveal the individual loss of each tuner (a, b, and c). Figure 3-6 shows the typical loss of an hp X870A Slide-screw tuner at specific SWR settings and frequencies throughout its operating band.

Simultaneous Equations for Determining Tuner Loss.

$$\begin{aligned} a + b + 0 &= x \\ b + c &= y \\ a + 0 + c &= z \end{aligned} \tag{3.10}$$

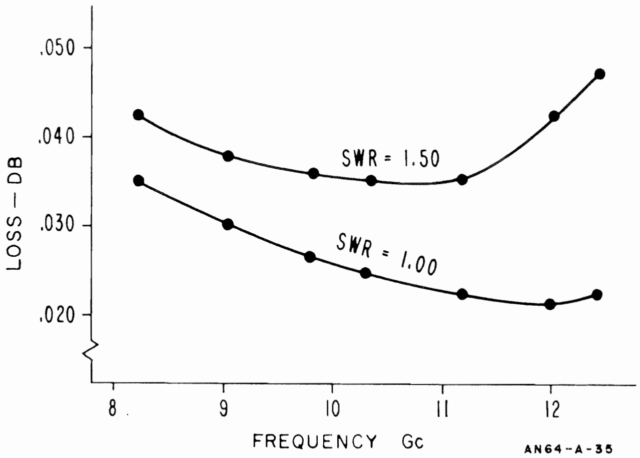


Figure 3-6. Typical loss calibration of an hp X870A slide screw tuner.

where

a, b, c = tuner loss in DB for each individual tuner.
x, y, z = measured loss in DB for each combination of tuners.

Example: Tuner Loss Calculation

measured values:

$x = 0.1$ db

$y = 0.11$ db

$z = 0.09$ db

then,

$a + b + 0 = 0.1$

$b + c = 0.11$

$a + 0 + c = 0.09$

Solution of the resulting third order determinant reveals the following individual tuner losses:

$a = .04$ db

$b = .06$ db

$c = .05$ db

The db loss values can then be converted to power ratios using the standard equation db = 10 $\log_{10}$ (power ratio).

WAVEGUIDE TO COAX ADAPTER LOSS CALIBRATION

The use of adapters in the measuring circuit should be avoided wherever possible. However, there are occasions when their use is necessary. This is particularly true when using broadband calorimetric power meters such as the hp 434A which have type N coaxial input systems. The frequency range of the 434A extends from DC through X-band traversing several waveguide bands for maximum versatility of the instrument. Power in the various waveguide bands may be measured with the 434A using appropriate waveguide-to-coax adapters and accounting for their loss.

A modified hp 281A waveguide-to-coax adapter* is recommended for greatest accuracy in such measurements. The adapter is modified to have a male type N coax connector in place of the normal female connector, thus it mates directly with the 434A input. To make connection to the calorimeter more convenient, a 90° E-plane waveguide bend is connected to the waveguide flange of the adapter. Finally a slide-screw tuner is connected to the bend to match the calorimeter input system to the transmission line or power source. The loss of the tuner and bend must be included, therefore, in calibration of the adapter for use with the calorimeter. The overall loss may be determined as follows:

1. Connect the equipment as shown in Figure 3-7 for "CALIBRATE".
2. Turn on the test equipment and adjust the sweep oscillator CW frequency to that used in the power measurement involving the adapter.
3. Adjust the Sweep Oscillator for a leveled RF output and apply 1 kc amplitude modulation. (RF power output not to exceed -6 dbm)
4. Set #1 415E Standing Wave Indicator gain and range for an upscale reading and minimize indication by adjusting the reference tuner (#1). With 415E #2 on EXPAND-DB set up a convenient reference reading using the meter GAIN control.
5. Insert test assembly and an identical adapter/tuner combination to form the symmetrical configuration shown in Figure 3-7 for "TEST".
6. Set test assembly tuner (#2) to probe depth and carriage position used in the actual power measurement.
7. Adjust tuner #3 for minimum reflection from the entire assembly as indicated by #1 415E. As nearly as possible use the same probe depth and carriage position away from the plane of symmetry on tuner #3 as used on tuner #2.
8. Note loss in db on 415E #2. The test adapter assembly loss is about one half the db loss noted. If there is sufficient test equipment available, two additional test assemblies may be measured and the individual loss of each found by simultaneous equations. The system of equations is identical to those used in the slide-screw tuner loss measurement given previously. Overall loss in the test adapter assembly will typically be about 3 or 4 percent at X-band frequencies. Figure 3-8 shows the loss calibration on an hp X281A waveguide to coax adapter and tuner assembly.

*Available from Hewlett-Packard on special order.

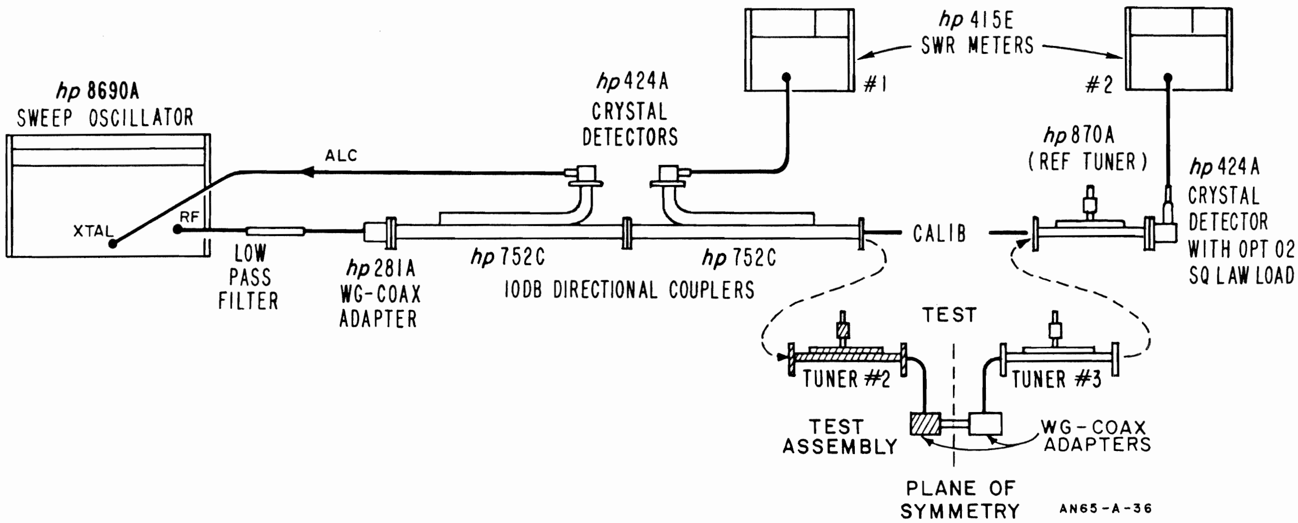


Figure 3-7. Setup for loss measurement on waveguide-to-coax adapter assembly. Technique is similar to tuner loss test shown in Figure 3-5.

A METHOD FOR MEASURING THERMISTOR MOUNT EFFICIENCY

The importance of thermistor mount efficiency and calibration factor has been illustrated by the examples shown in this section, and by the accuracy discussion in Section II. Efficiency and calibration factor of HP thermistor mounts may be obtained in one of three ways: 1) NBS Calibration 2) HP Production Test (new mounts only) and 3) standards lab calibration (HP and other commercial facilities). Since it is often impractical to send all mounts from the field to NBS or some distant Standards Lab, it is desirable to have some local method of measuring effective efficiency and calibration factor. The calibration system subsequently described* is the same system used on HP production lines to test new thermistor mounts prior to shipment.

The system is based on a comparison of test mounts with a "working standard" thermistor mount whose efficiency calibration is traceable to NBS wherever possible. Until NBS efficiency calibrations are available throughout the microwave region, interim standards of efficiency are used. The system accomodates thermistor mounts of 100 or 200 ohm operating resistance, temperature compensated or not, and is capable of either fixed-frequency or swept-frequency operation. Figure 3-9 shows the Efficiency Calibrator connected for operation. RF power is furnished in the frequency range of interest (X-band in this example) by an hp 690-series sweep oscillator which allows either swept-frequency operation with readout on the X-Y recorder, or fixed frequency programming by a DC voltage supplied to the EXTERNAL FM jack from the Efficiency Calibrator. The sweep oscillator's EXTERNAL AM input is used for closed-loop level control from a differential amplifier in the Calibrator. Two directional couplers are connected to the oscillator output similar to a conventional reflectometer setup, with reflected power sampled by power meter #1 and incident power by meter #2. P_{dc} substituted in the working standard and test mounts connected to the reference plane, is monitored by power meter #3. The 0 to 1 ma recorder output from each power meter is fed to a switching arrangement and load resistors in the Calibrator allowing digital voltmeter readout of each meter.

When measuring efficiency, #1 and #2 power meter outputs are connected differentially to the input of an amplifier in the calibrator. This amplifier develops a feedback voltage which is applied to the sweeper's EXTERNAL AM input, maintaining a constant difference between incident and reflected power at the reference plane. This arrangement assures a constant power dissipation in the total mount under test regardless of its SWR.

Initially a cavity wavemeter is used to calibrate six preset frequency programming potentiometers on the Calibrator front panel for fixed frequency operation. This is done so the sweeper later may be instantly tuned to predetermined test frequencies by a front

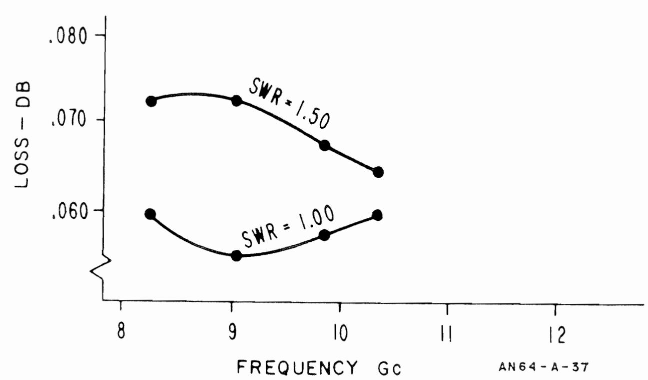


Figure 3-8. Typical loss calibration of hp X281A waveguide-to-coax adapter assembly.

*Macrie, James J., "Efficiency: The Missing Link in Bolometer Power Measurements", Microwave Journal, Vol. 8, No. 6, June 1965

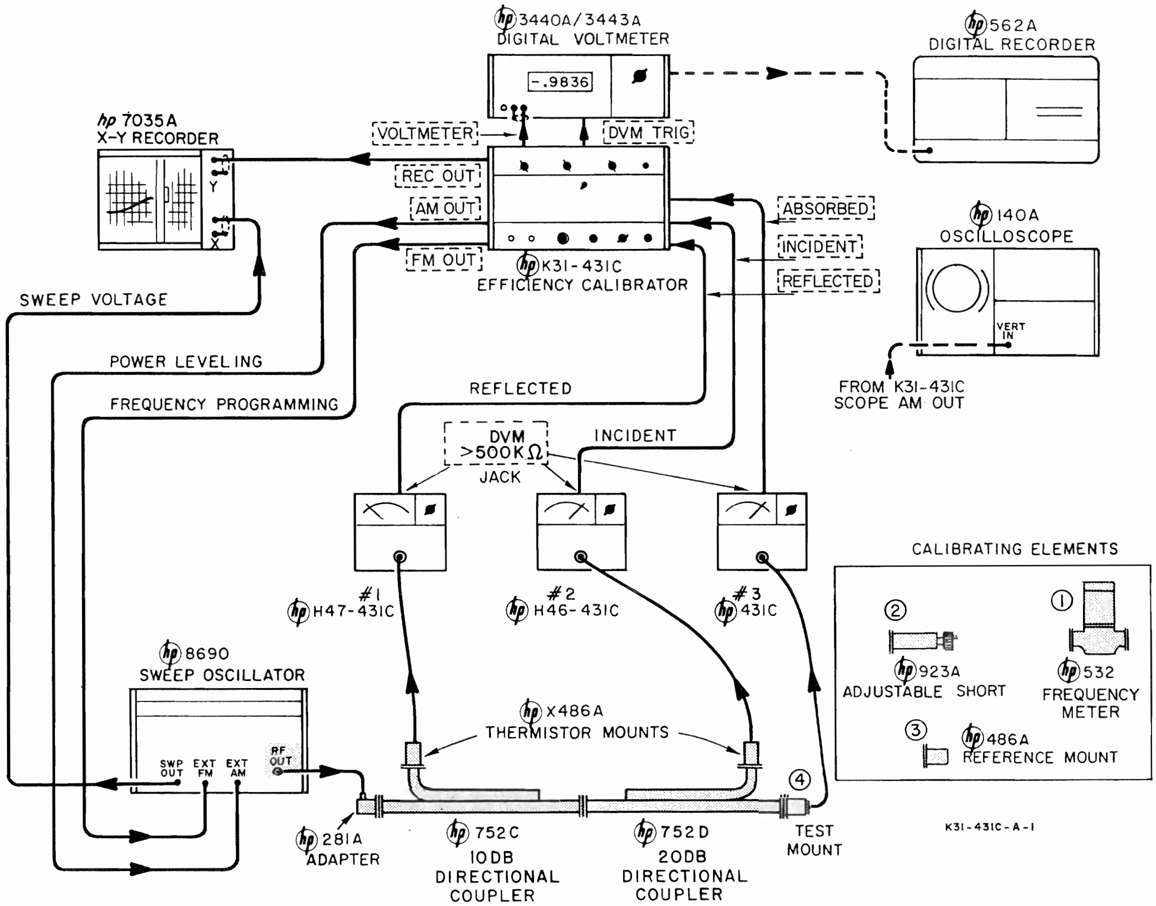


Figure 3-9. Thermistor mount efficiency, Calibration Factor, and reflection coefficient is quickly evaluated on a swept and fixed frequency basis by HP using system shown. Calibration data thus obtained is used for quality control and can be applied for greater power measuring accuracy.

panel switch on the Calibrator. After frequency adjustment is complete, the wavemeter is removed. An adjustable short reflecting all of the incident power is then connected to the reference plane. As the short is phased through a half wavelength, the mean output voltages of meter 1 and 2 are made equal. This adjustment compensates for differences in coupler and power meter output characteristics and calibrates the system for reflection coefficient measurements.

The only operation remaining in the system's overall calibration is to connect the working standard thermistor mount to meter #3, as shown in Figure 3-9, and set up reference efficiency readings on the digital voltmeter at each fixed calibration frequency. This is done by presetting six individual efficiency adjustments on the calibrator corresponding to the fixed test frequencies already set up. Adjusting the efficiency potentiometers on the calibrator at each test frequency adjusts the combination of #3 meter's output and the digital voltmeter to indicate the standard mount efficiency figures. For example, suppose a working standard thermistor mount is calibrated by a standards laboratory resulting in the efficiencies shown in Table 1.

The Calibrator would be set up to program the sweeper to each test frequency in the Table, and the sweeper's

Table 1. Typical efficiency calibration figures of an HP X486A Waveguide Thermistor Mount

TEST FREQUENCY (Gc)	EFFECTIVE FREQUENCY (%)
8.2	97.2
9.0	97.9
9.8	98.0
10.3	97.9
11.2	98.6
12.0	98.2

output level programmed for the corresponding efficiency figures to read out directly on the digital voltmeter.

An efficiency-adjust potentiometer is also included for the swept-frequency mode of operation. In the swept mode, the calibrator connects an X-Y recorder to the output of #3 power meter. The X axis of the recorder is driven by the sweep output voltage from the 690-series sweeper. Arbitrary calibration grid lines are plotted on the X-Y recorder by sweeping the frequency band of the working standard thermistor mount at different settings of the swept "efficiency-adjust" pot. For example, suppose it is anticipated that the mounts to be tested are within 4.5% of the working standard efficiency. The swept efficiency adjust would be set for a DVM reading of .955 volts with the 690 in the "TRIGGER" sweep mode (i.e. not sweeping.). The manual trigger is then depressed on the 690 and a reference line swept on the X-Y recorder. After completing the sweep, the swept efficiency adjust is set for a DVM reading of 1.000 volts and another reference line swept. Now the mounts to be calibrated are connected to the system in place of the standard mount and swept-efficiency tested with relation to the calibration grid previously drawn. Figure 3-10 shows swept-frequency tests on both coax and waveguide mounts. If the swept check looks good, the sweeper is set for EXTERNAL FM operation in the CW sweep mode and programmed to each of the preset calibration frequencies by stepping the Calibrator "FREQUENCY" switch through its range. The digital voltmeter automatically reads out the effective efficiency of the test mount at each test frequency.

By setting the "FUNCTION" switch of the Calibrator to "Calibration Factor" the system automatically reads out K_b of the test mount at each test frequency without need for additional calibration.

How does the technique just described and the definition for Effective Efficiency given in Section II correlate? To show this we recall the equation for effective efficiency is:

$$\eta_e = \frac{P_{dc\ substituted}}{P_{\mu\ wave\ dissipated}} \qquad (2.6)$$

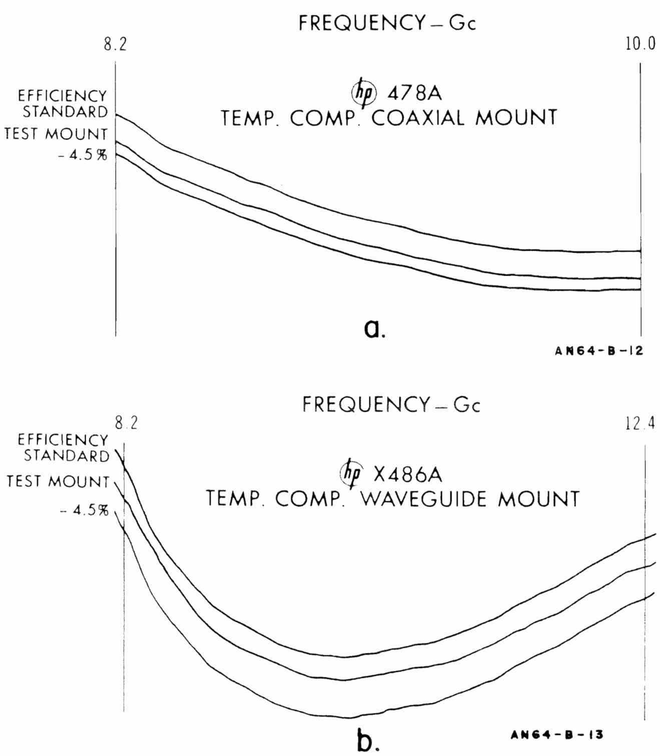


Figure 3-10. Swept frequency efficiency plots of (a) HP 478A coax thermistor mount (b) HP X486A waveguide mount, compared to a standard mount with system in Figure 3-9. Note absence of efficiency "holes" sometimes found in defective mounts.

Power at the reference plane in Figure 3-9 is controlled by the difference between incident and reflected power. Thus, any reflection from the mount increases the sweeper output just enough to maintain a constant dissipated power in the test mount. Under this condition, the term $P_{\mu wave\ dissipated}$ in equation 2.6 may be considered a constant at any fixed frequency. Next, with a working standard mount of known efficiency (η_e) we may adjust the digital voltmeter reading of #3 power meter, which reads $P_{dc\ substituted}$, to satisfy equation 2.6. By proper adjustment of #3 meter's recorder output load resistor, the DVM reading is normalized to read efficiency directly; e.g., the actual power meter reading may be 9.3 mw and the load resistor adjusted for a DVM reading of .9800 volts representing 98% efficiency. This eliminates the need for setting absolute power levels out of the sweeper in the initial calibration. Replacing the working standard mount with a test mount, we may now effectively read $P_{dc\ substituted}$ on meter #3 with $P_{\mu\ wave\ dissipated}$ in the mount held constant. Since meter #3's output has been calibrated to read efficiency, the test mount efficiency is read directly on the DVM.

Correlation of system operation with the definition for calibration factor is similar. By leveling forward power only (meter #2), the $P_{incident}$ term in the Calibration Factor equation can be made constant, and the readings of meter #3 normalized to read K_b directly.

The primary sources of error present in this system are: 1) the directivity of the 10-db directional coupler sampling reflected power 2) error in the standard thermistor mount efficiency calibration, and 3) discontinuities in the waveguide system. The worst case error, due to coupler directivity and waveguide discontinuities, has been determined to be less than 0.7% up to 12.4 gc, and less than 0.9% to 18 gc when the vector difference between the standard mount and test mount reflection coefficients is .20.* This error arises from the fact that any reflected signal from the standard mount adds vectorially in different phase and magnitude with the fixed directivity signal in coupler #1 than does the reflected signal from the test mount. This small variation in the resultant power at meter #1 prevents $P_{\mu wave\ dissipated}$ from being held exactly constant by the leveler feedback to the sweep oscillator, resulting in a comparison error between standard and test mounts.

Standard mount calibrations obtained from NBS are accurate to 1% in X and P band, and temporarily 2% in J-band. This error is transferred to the initial calibration of the system so adding the two sources of error directly, we obtain an overall efficiency calibration accurate to 1.9% or better from 8.2 to 18.0 gc. This is essentially the accuracy figure used in the previous examples of correcting power meter readings.

Experience has shown that thermistor mounts of good design and construction tend to have smooth curves of efficiency between calibration points thus lending themselves well to swept-frequency efficiency measurements.

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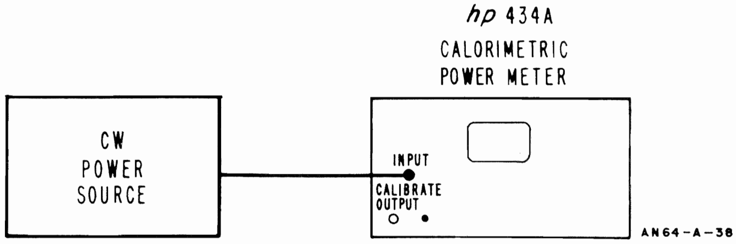
*Pramann, Robt. F., "A Quick, Accurate Method for Measuring Thermistor Mount Efficiency and Calibration Factor", paper presented to the 20th annual ISA Conf., Los Angeles, Calif., Oct. 4-7, 1965.



Error Source	Conjugate Available Power Measurement		Z_o Available Power Measurement	
	Value Measured	Uncertainty	Value Measured	Uncertainty
Instrumentation (434A)	-	±1%	-	±1%*
Effective Efficiency (η_e) of Head Box. (434A)	.99 (±2%)	0 to -3%	.99 (±2%)	0 to -3%
Mismatch Loss ρ_g = .43 (SWR = 2.5) ρ_L = .05 (SWR = 1.1)	-	-15 to -21.5%	-18.6%	±4%
Limits of Error Before Correction	-14 to -25.5%		-13.6 to -26.6%	
Total Uncertainty After Correction	-13 to -24.5%		+6 to -7%	

*May be reduced to 0.5% by calibrating 434A with -hp- K02 434A.

Figure 3-11a. Basic coaxial power measuring setup using hp 434A Calorimetric Power Meter. Power range is 10 mw to 10 watts full scale from DC to 12.4 Gc.

It is conceivable that a particular thermistor mount design could have efficiency "holes" between the efficiency calibration frequencies chosen. In this case, merely comparing two mounts of identical design on a swept-frequency basis would not necessarily yield valid efficiency information between calibration points, since both the reference mount and test mount would have holes at the same frequencies. To preclude such false data, Hewlett-Packard has cross-checked a number of thermistor mounts from production runs against mounts of different design and found that no holes exist within the frequency range of each standard mount and the efficiency curves are indeed smooth.

MEASURING AVERAGE POWER — 10 MW TO 10 KW

Calorimetric power meters are used for best accuracy in power measurements beyond the range of bolometers. Calorimetric power meters are available to measure high power levels directly with greater accuracy than possible using directional couplers, attenuators, and bolometers. The hp 434A Power Meter has seven overlapping power ranges from 10 mw to 10 watts full scale, providing smooth transition between low level bolometric and higher power calorimetric measurements. As in bolometric measurements, varying degrees of accuracy are possible with the calorimeter depending upon the technique of measurement and care exercised. The 10 watt range of the 434A may be extended through the use of directional couplers and high power terminations to about 10 kw with some sacrifice in accuracy.

BASIC COAXIAL SYSTEM - DC to 12.4 GC

Figure 3-11a shows a basic setup for measuring power up to 10 watts in coaxial systems using the hp 434A. The procedure is as follows:

1. Initial 434A Adjustments:
   a. Make initial checks of oil level and flow rate according to 434A Operating and Service Manual and zero the meter.
   b. Set the RANGE switch to 0.1 watt.
   c. Connect the 434A CALIBRATE OUTPUT jack to the INPUT with a short length of coax and set ADJUST TO 0.1 WATT F.S. front panel control for full-scale meter reading of 0.1 watt.
   d. Disconnect the CALIBRATE OUTPUT from the input and recheck meter zero.
2. Set the RANGE switch for the maximum power expected from the source to be measured.
3. Connect the source to the 434A INPUT and read average power.

Measurement uncertainty using the technique just described, depends upon mismatch loss, 434A head-box efficiency and instrumentation error which is determined primarily by the accuracy of the 0.1 watt calibration power. A typical error analysis for a measurement at 3 Gc is given by the chart in Figure 3-11a.

As may be seen from the chart, mismatch loss causes the greatest uncertainty in this example. Further investigation reveals that high source SWR is primarily responsible for the large mismatch loss. Note that the greatest total uncertainty occurs in the conjugate available power measurement, or the power that would otherwise be delivered if the source and load were a conjugate match. The measured power with respect to P_0, or the power that would be dissipated in a nonreflecting load, is reasonably accurate since the 434A input SWR is only 1.1 in this example.

IMPROVED COAXIAL SYSTEM - 500 MC TO 12.4 GC

Figure 3-11b illustrates a method for measuring power up to 10 watts with improved accuracy. In this setup, a slide screw (or double stub) tuner is used either to match the 434A input impedance to the transmission line for minimum SWR, or to present a conjugate match to the source for maximum transfer of power. To measure power with the 434A and a tuner, proceed as follows:

1. Follow the same turn-on, zeroing, and calibration procedure outlined in Step 1 for Figure 3-11a. Instrumentation error may be reduced from 1% to less than 0.5% by performing the initial calibration of the 434A with the hp K02-434A DC Test Set instead of using the 434A CALIBRATE OUTPUT.

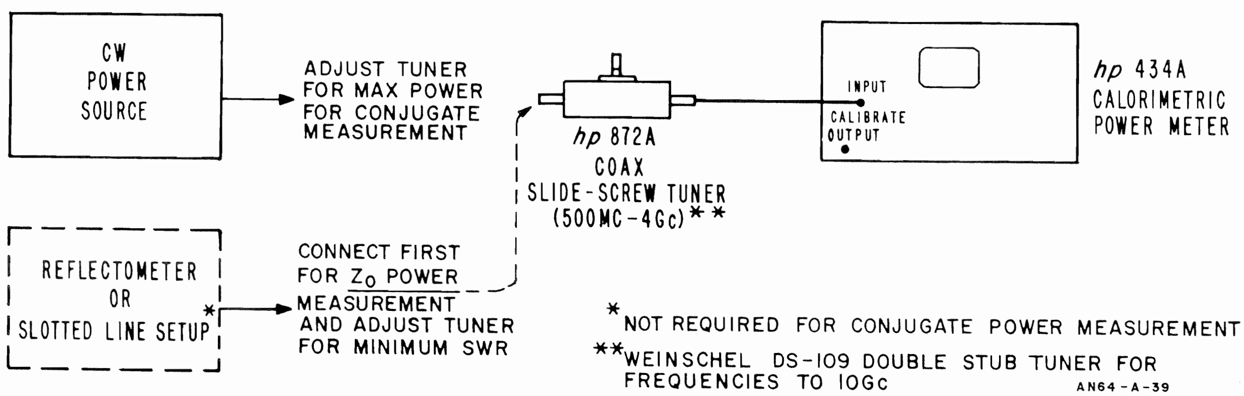
2. Set 434A RANGE to accommodate highest power anticipated.

3. If conjugate available power is desired, connect source directly to slide-screw tuner and 434A as shown in Figure 3-11b. Adjust tuner for maximum power indication on 434A, and read conjugate available power. If Z_0 available power is desired, connect tuner and 434A to appropriate reflectometer or slotted line setup and adjust tuner for minimum SWR. Then connect tuner and 434A to power source and read Z_0 available power. Equations 3.6 and 3.9, given earlier, may be applied to the 434A reading to correct for the sources of error present in this system.

Overall uncertainty is reduced considerably by the use of the tuner as can be seen in the error analysis chart of Figure 3-11b. The most notable improvement is in the measurement of power to a conjugate load since the comparatively large conjugate mismatch loss has been eliminated.

POWER IN WAVEGUIDE SYSTEMS - 2.6 GC TO 12.4 GC

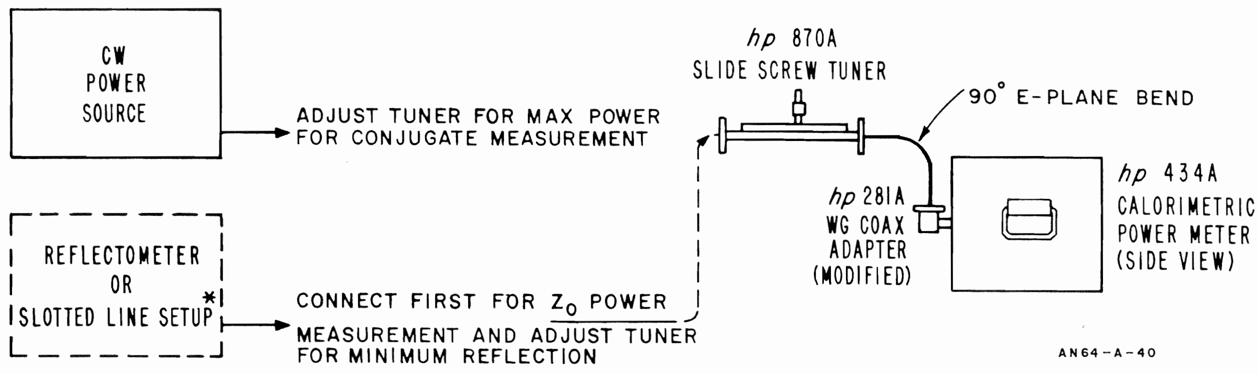
The coaxial input system of the 434A Calorimetric Power Meter may be adapted to various waveguide systems in the frequency range of 2.6 to 12.4 Gc using the appropriate waveguide-to-coax adapter assembly. For best accuracy and convenience a slide-screw tuner and waveguide bend are assembled with a modified hp 281A adapter to make the connection. The loss introduced by the adapter assembly may be measured using



Error Source	Conjugate Available Power Measurement: Value Measured	Conjugate Available Power Measurement: Uncertainty	Z_0 Available Power Measurement: Value Measured	Z_0 Available Power Measurement: Uncertainty
Instrumentation (434A)	-	±1%*	-	±1%*
Effective Efficiency (η_e) of Head Box (434A)	.99 (±2%)	0 to -3%	.99 (±2%)	0 to -3%
Tuner Loss Ratio (T_L)	.995	-0.5%	.995	-0.5%
Limits of Error Before Correction	+0.5 to -4.5%		+0.5 to -4.5%	
Total Uncertainty After Correction	+2 to -3%		+2 to -3%	

*May be reduced to 0.5% by calibrating 434A with K02-434A test set.

Figure 3-11b. Improved coaxial power measuring setup with hp 434A includes tuner for eliminating mismatch loss.



*NOT REQUIRED FOR CONJUGATE POWER MEASUREMENT

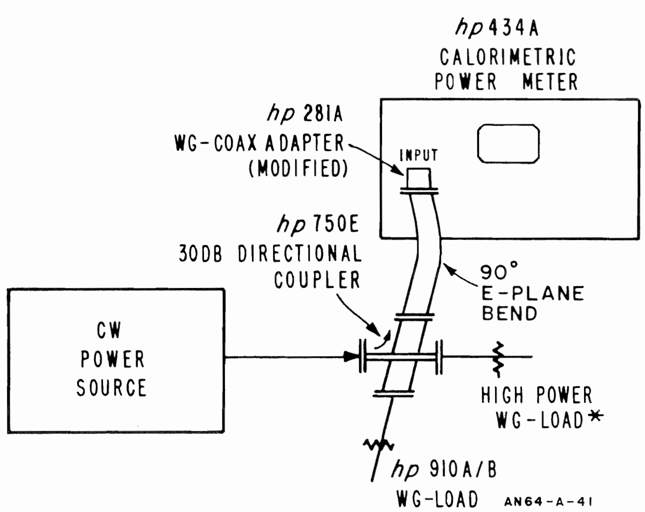
Figure 3-12a. Power up to 10 watts in waveguide systems can be measured with the hp 434A using the adapter-tuner-bend assembly shown above. Loss caused by assembly should be evaluated using system shown in Figure 3-7 for highest accuracy.

the procedure shown earlier under "Methods for Improving Accuracy". To measure power up to 10 watts in waveguide systems, use the setup shown in Figure 3-12a, and proceed as follows:

1. Make the initial 434A adjustments as outlined in Step 1 for Figure 3-11a.

2. If conjugate available power is desired, connect the source to the tuner-calorimeter input and adjust the tuner for maximum power indication on the 434A meter. If Z_0 available power is desired, connect the tuner and 434A to a reflectometer or slotted line setup and adjust the tuner for minimum reflection. Then connect the power source to the tuner input.

3. Read average power on the 434A meter. Apply equation 3.6 or 3.9 to the reading to correct for the sources of error present.



*hp X913A UP TO 500 WATTS AVERAGE 8.2-12.4Gc

Figure 3-12b. Extended power range in waveguide enables measurements to 10 kw average using suitable loads and coupler. Some uncertainty is added to the measurement because of coupling factor calibration uncertainty.

Refer to the chart in Figure 3-11b again for a breakdown of error sources also present in the system of Figure 3-12a. Power in waveguide systems below 2.6 gc may be measured with a similar setup if appropriate adapters, tuners, and bends are available.

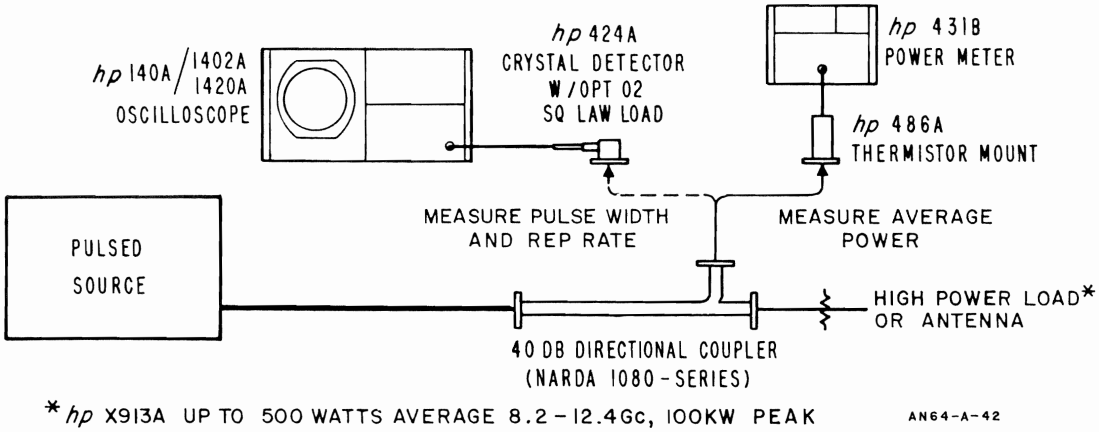
EXTENDED RANGE WAVEGUIDE SYSTEMS - 2.6 GC TO 12.4 GC.

Figure 3-12b illustrates how the 434A may be used to measure average power in excess of 10 watts. As in the case of extended range bolometer measurements, the addition of directional couplers or attenuators adds uncertainty. The measurement procedure for this setup is the same as previously described under "Basic Coaxial System - DC to 12.4 Gc."

MEASURING PEAK PULSE POWER

AVERAGE POWER-DUTY CYCLE METHOD #1

Peak power of a rectangular RF pulse envelope may be determined by measuring average power, pulse width and repetition frequency. The product of pulse width and repetition frequency is defined as the duty cycle of the pulsed source and relates the time the pulse is on to the period of the pulse rate. Figure 3-13 shows a setup for determining peak power in a waveguide system using the average power-duty cycle method. It is important that the maximum peak power and maximum energy-per-pulse ratings of the hp 486A thermistor



$$P_{PK} = \frac{P_{average}^{**}}{(PRF \times \tau)\left(\log_{10}^{-1} \frac{-DB}{10}\right)} \quad (3.11)$$

Error Source	Peak Pulse Power Measurement (Average Power - Duty Cycle) Method #1	
	Value Measured	Uncertainty
Average Power Measurement after Correction**	7.5 mw	±6%
Coupling Factor (DB)	40 db	±0.4 db ≈ ±9%
Pulse Repetition Frequency (PRF)	1×10^3 PPS	±2%
Pulse Width (τ)	1.5×10^{-6} sec.	±2%
Total Uncertainty	±19%	

**431C indication corrected using equation 3.5.

Figure 3-13. On-line test of peak power sources may be conveniently made using system shown. Power meter reads average power which is then divided by system duty cycle and power ratio of coupler.

mount not be exceeded. At pulse repetition frequencies less than 1kc, the maximum energy per pulse is 2.5 watt-microseconds, i.e., any combination of peak power and pulse width which when multiplied together produces 2.5 watt-microseconds. For pulse repetition frequencies above 1kc, the maximum is 5 watt-microseconds. The maximum peak power in any case should not exceed 100 watts at the thermistor mount. Another highly important factor in peak power systems is that all waveguide flanges be very carefully aligned and that the waveguide is clean and dry to avoid arc-over.

The procedure for Figure 3-13 is as follows:

1. Connect the hp 424A Crystal Detector and its square-law load* to the auxiliary arm of the directional coupler as shown.

2. Measure and note pulse repetition frequency (PRF) and pulse width (τ) using the oscilloscope's calibrated sweep.

3. Replace the 424A detector with the 486A thermistor mount and 431C power meter.

4. Note the average power reading on the 431C, correcting for coupling factor, thermistor mount calibration factor and mismatch uncertainty between mount and the coupler's auxiliary arm.

5. Calculate peak pulse power using the following equation:

$$P_{pk} = \frac{P_{average}}{(PRF \times \tau)\left[\log_{10}^{-1}\left(-\frac{DB}{10}\right)\right]} \quad (3.11)$$

where

P_{pk} = Peak power of source

P_{ave} = Average power indicated on 431B

DB = Coupling factor of directional coupler

Example:

Suppose the following readings are made from a pulsed source with the setup shown in Figure 3-13:

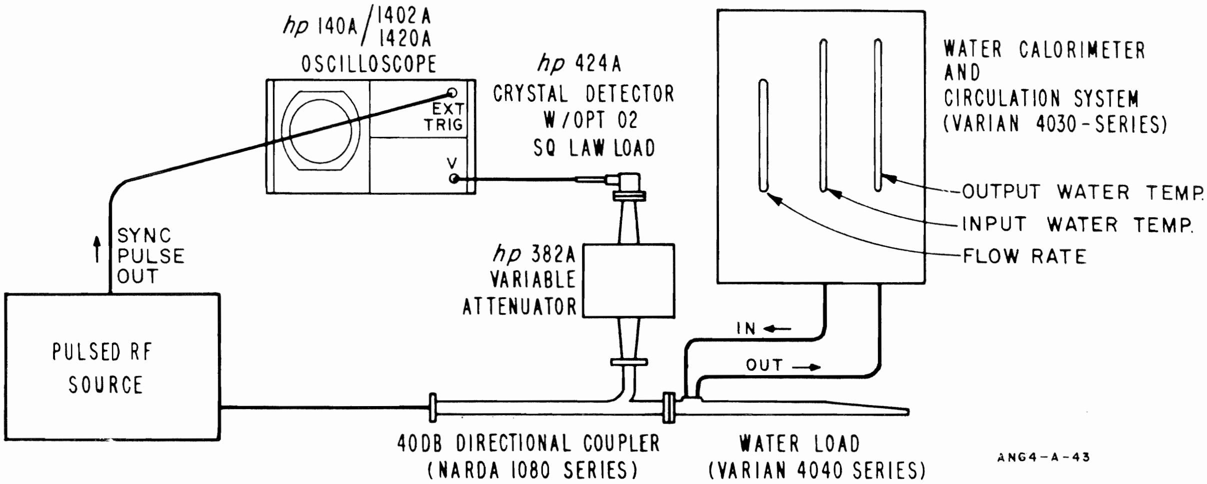
PRF = 1kc

$\tau = 1.5 \times 10^{-6}$ seconds

$P_{average} = 7.5 \times 10^{-3}$ watts

Coupling Factor = 40 db

*In addition to good square law performance up to 50mv output, square law load improves rise (and decay) time characteristics of 424A detectors. Both factors are essential for accurate pulse shape determination.



$$P_{PK} = \frac{P_{ave}}{(PRF \times \tau)}$$

$$P_{ave} = 4.18\, mCp\Delta T \text{ watts}$$

where

m = Water flow rate through LOAD (gallons/sec).

Cp = Specific heat of water (gallons/°C).

ΔT = Temperature rise due to power absorbed (°C)

Error Source	Peak Pulse Power Measurement (Average Power - Duty Cycle) Method #2	
	Value Measured	Uncertainty
Average Power Measurement	500 watts	±3%
Pulse Repetition Frequency (PRF)	1×10^3 PPS	±2%
Pulse Width (τ)	1.5×10^{-6} sec	±2%
Total Uncertainty	±7%	

Figure 3-14. Water load and calorimeter offer best accuracy in high power measurements (over 10 watts average) largely because directional coupler uncertainty is not involved. Coupler shown merely samples pulsed power for determination of PRF and pulse width on oscilloscope. System is useful when all the source power can be terminated during the measurement and water circulation equipment can be readily set up.

Using equation 3.11 peak pulse power is calculated:

$$P_{pk} = \frac{7.5 \times 10^{-3} \text{ watts}}{(1 \times 10^{3})\,(1.5 \times 10^{-6})\left[\log_{10}^{-1}\left(\frac{-40}{10}\right)\right]} = 50\,\text{kw}$$

Sources of error and examples of the resulting uncertainties are summarized in the table of Figure 3-13. Since mismatch cannot be corrected with a tuner in pulse power measurements, it is important to calculate the possible limits of mismatch loss using the coupler auxiliary arm and thermistor mount SWR* as indicated by Step 4 of the foregoing procedure. The accuracy of this method of pulse power measurement depends upon the pulse being rectangular. As the RF pulse envelope departs from a true rectangle, error increases rapidly. A better method for peak power measurements of non-rectangular pulses is described under "Direct Pulse Power Technique" which does not depend on pulse shape. Peak Power range of the system shown is about 100kw maximum at X-band, limited by the maximum power ratings of the termination, thermistor mount, detector and coupler. This power range decreases at waveguide sizes smaller than WR-90. Average power should not exceed 500 watts using the X913A termination unless it is liquid cooled.

*Thermistor mount ρ may be calculated from efficiency and calibration factor data included with new HP thermistor mounts using the relation

$$\rho = \sqrt{1 - K_{b}/\eta_{e}} \qquad SWR = \frac{1+\rho}{1-\rho}.$$

AVERAGE POWER - DUTY CYCLE METHOD #2

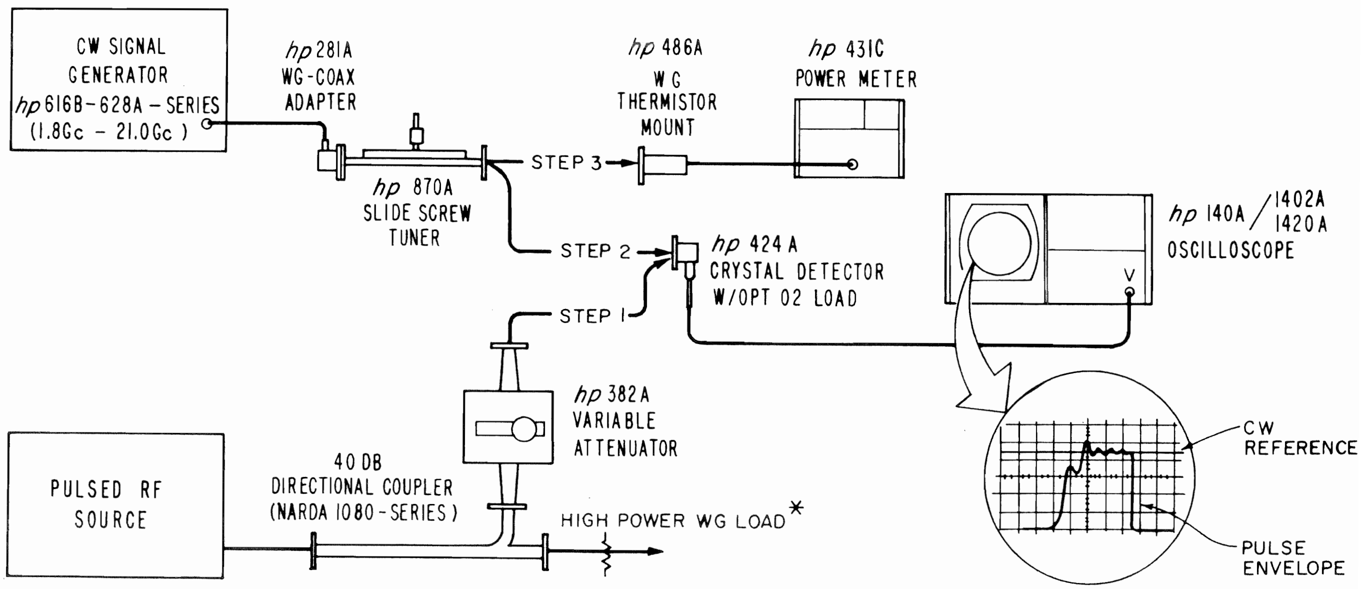
One method of improving overall accuracy is to improve the method of average power measurement and again use the duty cycle calculation for peak power. A more accurate average power measurement is possible in situations where all of the power may be terminated and measured in a water load, thus avoiding directional coupler calibration error. The system shown in Figure 3-14 is suitable for permanent laboratory installations that will accommodate the required calorimeter and water circulation equipment. Average power is calculated from the temperature difference in water entering and leaving the load, flow rate, and specific heat of the water using the equation shown in Figure 3-14. Duty cycle is determined from the oscilloscope measurements of pulse width and repetition rate.

The chart in Figure 3-14 indicates various sources of error for measuring rectangular RF pulse envelopes and probable uncertainties. Common water load systems handle average power levels from 5KW to 30KW in X-band, with capabilities increasing at the lower microwave bands to about 100KW. Larger systems utilizing power splitters handle average power in the megawatt region.

DIRECT PULSE POWER TECHNIQUE.

An improved waveguide setup for peak power measurement of non-rectangular pulses or multipulse systems is shown in Figure 3-15. In this arrangement, the pulse envelope is first detected by a fast-response crystal detector, hp 424A with Option 02 load. The detected pulse amplitude is noted on a DC-20MC oscilloscope then accurately duplicated with a known CW power. The directional coupler and attenuator should limit the power into the 424A detector to 1 to 2 mw peak since this is about the maximum CW power available for comparison. The procedure is as follows:

1. Connect the equipment as shown in Figure 3-15, step 1. Set attenuator to a convenient db reference that will limit the expected peak power to the detector to about 1 mw.



*hp X913A UP TO 500 WATTS AVERAGE 8.2-12-4Gc, 100 KW PEAK

AN64-A-44

$$P_{PK} = \frac{P_{average}^{*}}{\left[\log_{10}^{-1}\left(\frac{-DB}{10}\right)\right]}$$

*CW Generator Power from Step 3.

Error Source	Peak Pulse Power Measurement (Direct Pulse Method) Value Measured	Uncertainty
Average Power Measurement	0.5 mw	±5%
Coupling Factor and Attenuation (DB)	80 db	±0.8 db ≈ ±18%
Comparison of Pulse Amplitude to CW reference including Mismatch of Detector and Attenuator		±2%
Total Uncertainty	±25%	

Figure 3-15. Direct pulse method for peak power in waveguide system. Technique is useful for measuring peak powers of non-rectangular pulse envelopes.

2. Adjust the oscilloscope trigger and vertical sensitivity controls for a stable pulse display. Use DC coupling at the oscilloscope vertical input, and make no further adjustments to the oscilloscope sensitivity once the reference pulse amplitude is set.

3. Move the detector/square-law load combination to the output of the slide screw tuner and signal generator as shown for step 2.

4. Adjust the signal generator for a CW output frequency equal to the pulsed source's RF frequency.

5. Adjust the slide screw tuner for maximum deflection of the oscilloscope trace. DO NOT ADJUST THE OSCILLOSCOPE VERTICAL SENSITIVITY. Use the generator's output attenuator to keep the trace on the screen during the tuner adjustment.

6. Carefully set the generator output attenuator for a trace deflection on the oscilloscope equal to the pulse amplitude noted in Step 1. Make no further adjustments to the output attenuator.

7. Connect the 486A thermistor mount and 431C power meter to the generator output as shown for Step 3.

8. Re-adjust the slide screw tuner for a maximum reading on the 431C and note the average power, correcting for mount efficiency and tuner loss (Eq. 3.6).

9. Calculate peak power at the source output from the equation

$$P_{pk} \text{ source} = \frac{P_{average}}{\left[\log_{10}^{-1}\left(-\frac{DB}{10}\right)\right]}$$

where

DB = Coupling Factor of directional coupler in db.

$P_{average}$ = Average Power of generator measured on 431B.

Example:

Suppose $P_{average}$ is 0.5 mw and the combined attenuation of the directional coupler and attenuator used is 80 db at the test frequency. Peak power is then:

$$P_{pk} \text{ source} = \frac{5 \times 10^{-4} \text{ watts}}{\left[\log_{10}^{-1}\left(-\frac{80}{10}\right)\right]} = 50 \text{ kw}$$

The chart in Figure 3-15 shows the sources of error and examples of the resulting uncertainties using the direct pulse technique. Comparing the uncertainties of Figure 3-13 and 3-15 for equal values of peak power indicates the direct pulse technique to be less desirable. Greater ambiguity exists because of the greater decoupling required between source and detector. However, advantages of the system in Figure 3-13 are quickly lost if the pulse envelope departs considerably from a rectangular shape or if multiple pulses are being measured.

In addition to the usual limitations of peak and average power ratings of the components, this system must provide a CW power level equal to the peak power detected at the variable attenuator output. Using the generators shown, this power level is on the order of 1 to 2 mw. At X-band, the system shown would be limited to about 100 kw peak power and 500 watts average with convection cooling of the load.

NOTCH WATTMETER.

The Notch Wattmeter gets its name from a technique that involves gating a reference CW power source off during the interval that the pulsed source in test is on. Thus, the reference output is "notched" at the same rate, duration, and amplitude as the pulse output of the device in test. The resulting CW power is then measured by conventional means. This system can be useful in situations where the average to peak power ratio of a pulse is too great to be measured directly by a power meter. An example of this would be a radar of very low duty cycle.

Figure 3-16 shows a notch wattmeter setup. The technique requires that the pulsed source in test has a sync pulse output of sufficient amplitude and polarity to operate an external pulse generator. The hp 214A Pulse Generator is recommended in this application partly because of its versatility in accepting sync pulses of either polarity at 5 to 15 volts peak amplitude. Other advantages are continuously variable pulse widths of 50 nsec to 10 msec with repetition rates of 10 cps to 1 Mc, positive or negative output pulse polarity and excellent pulse shape characteristics.

The measurement procedure is as follows:

1. Connect the equipment as shown, pre-setting the 214A PULSE AMPLITUDE to about 50 volts positive when operating with the hp 618B or 620B Signal Generators.

2. Null and zero the 431C and turn on the pulsed source in test.

3. Adjust the Signal Generator to the pulsed source's RF frequency.

4. Set the 214A Pulse Generator as follows:

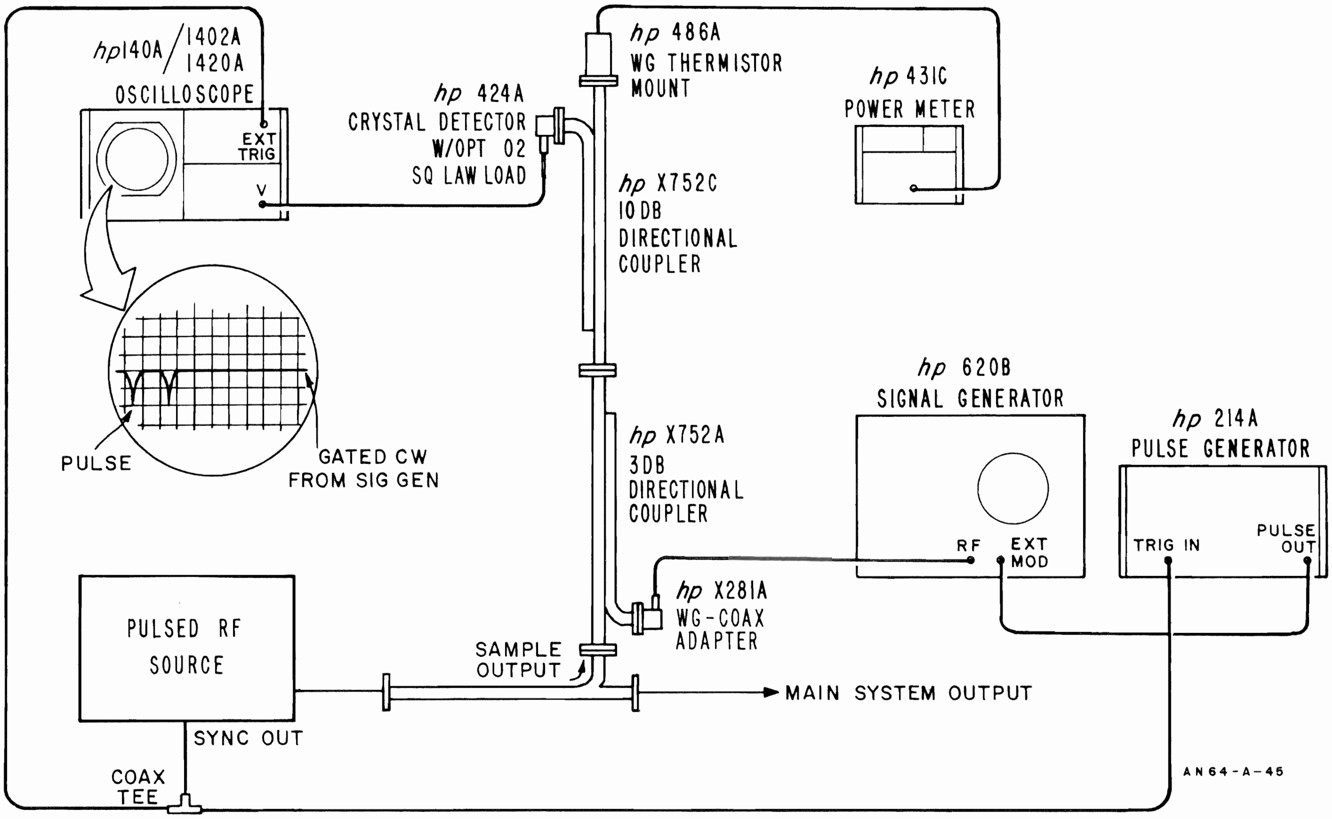
   TRIGGER MODE to EXT.

   SLOPE for same polarity as sync pulse out of source in test.

   INT. REP. RATE to approximately the same frequency as the source PRF.

   PULSE WIDTH to approximately that of the pulsed source.

   PULSE POSITION to 0-1 μsec, PULSE DELAY.



Error Source	Peak Pulse Power Measurement (Notch Wattmeter Method)	
	Value Measured	Uncertainty
CW Power	1 mw	±10%
Total Coupling Factor	73 db	±0.8 db ≈ 18%
Total Uncertainty	±28%	

Figure 3-16. Notch wattmeter requires more elaborate setup than other methods but is useful with very low duty cycles and non-rectangular pulse shapes.

5. With the generator output attenuator set for minimum output power, adjust the oscilloscope vertical sensitivity, sweep time, and trigger controls for a stable display of the source's detected output pulse. Use DC coupling at the oscilloscope vertical input.

6. Now increase the generator output until the oscilloscope baseline rises to an amplitude equal to the detected pulse top.

7. Adjust the 214A INT. REP. RATE, PULSE WIDTH, and PULSE POSITION as required to obtain a straight and nearly unbroken trace on the oscilloscope. The signal generator output is now being pulsed off by the 214A for the precise interval that the source is pulsed on. The generator's output amplitude has also been made equal to the peak pulse power at the output of the 70 db directional coupler, resulting in a CW power which is easily measured with the 431C power meter.

8. Read the power level indicated by the 431C and calculate peak power at the pulsed source output by the following equation:

$$P_{pk} = \frac{P_{indicated}}{\left|\log_{10}^{-1}\left(-\frac{DB}{10}\right)\right|}$$

where

P_1 = Power read on 431C.

DB = Total attenuation of directional couplers between pulsed source and thermistor mount.

Example:

Suppose P_1 is 1 mw and the total attenuation between source and thermistor mount is 73 db at the test frequency. Peak power out of the source may then be calculated as follows:

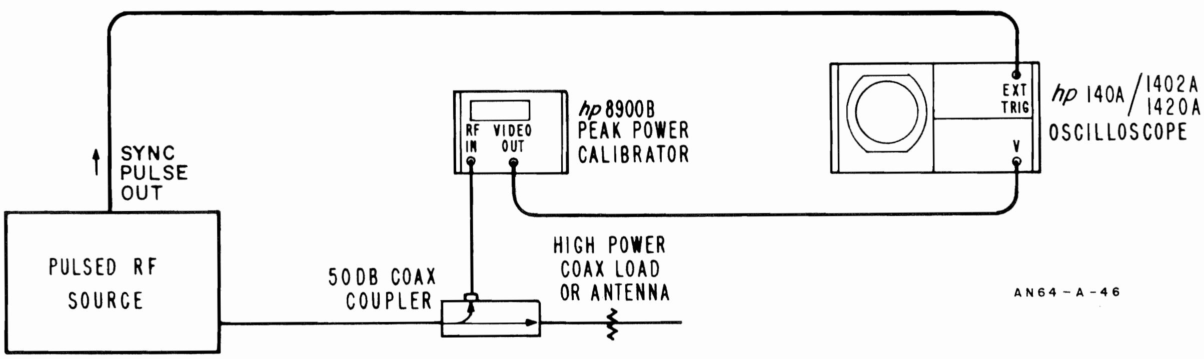
$$P_{pk} = \frac{1 \times 10^{-3} \text{ watts}}{\left[\log_{10}^{-1}\left(\frac{-73}{10}\right)\right]} = 20 \text{ kw}$$

Overall uncertainty of the notch wattmeter depends on factors listed in the error analysis of Figure 3-16. The values shown are examples of X-band measurements which result in an overall uncertainty of about 28%. The notch wattmeter does not depend upon pulse shape for its accuracy provided the duty cycle is low. The waveform shown in Figure 3-16 illustrates how a pulse of poor rise and decay time might appear in the notch display. At low duty cycles the "dead" time between the generator output and pulsed source power is so small that the thermistor averages out the error. Conventional barretters are not satisfactory for this application because of their comparatively fast thermal time constant which would tend to follow the instantaneous power variation, introducing pulse shape error into the measurement.

PEAK POWER IN COAX SYSTEMS - 50 MC TO 2 GC.

Convenient peak power measurements are possible in coaxial systems using the HP Model 8900B Peak Power Calibrator. Operation of the 8900B is based on a DC-to-Pulse-Power Comparison which was described in Section II. Figure 3-17 shows the 8900B in a typical setup. A 50 db directional coupler, in this example, reduces source power entering the 8900B input to 200 mw peak or less. The procedure is as follows:

1. Connect the equipment as shown and check to see that the 50 ohm termination furnished with the 8900B is properly connected at the BOL. OUT connector on the Calibrator's rear panel.
2. Turn on the test equipment, leaving the source power off.
3. Set the 8900B function switch to CAL and check that the meter indication is within +0. 2 and -0. 4 db of the calibration mark.
4. Set the oscilloscope VERTICAL SENSITIVITY to approximately 50 mv/cm and select DC input and amplifier coupling.
5. Set the oscilloscope sweep time and trigger controls for a free running trace. The display should appear as two horizontal lines on the CRT.



Error Source	Peak Pulse Power in Coax Boonton 8900B	
	Value Measured	Uncertainty
8900B (Includes Mismatch)	135 mw	±0.6 db* (≈±14%)
Directional Coupler - Filter db	50 db	±0.5 db ≈ (±11%)
Pulse - DC Comparison		Negligible
Total Uncertainty	$^{+1.0}_{-1.2}$ db (≈±25%)	

*With custom calibration chart.

Figure 3-17. Convenient peak power measurements at frequencies from 50 Mc - 2 Gc are possible using HP Model 8900B and auxiliary oscilloscope. Direct reading 8900B does not rely on pulse shape for accuracy.

6. Adjust the 8900B VIDEO NULL until the two horizontal lines are superimposed. (This balances out the detector diode bias from the video output waveform which might otherwise be read as RF power.)

7. Set the 8900B function switch to MEAS and turn on the pulsed source.

8. Synchronize the oscilloscope sweep for a stable display of the 60 cps chopper output of the 8900B. Use the external sync mode if the pulsed source provides a suitable sync pulse for the oscilloscope. The oscilloscope may also be synchronized by connecting the MON OUT jack of the 8900B to the oscilloscope's EXT SYNC jack.

9. Using the COARSE and FINE controls of the 8900B, adjust the DC reference voltage (which now appears as a straight line on the CRT) until it coincides with the peak of the displayed pulse. The oscilloscope display should now appear as illustrated in Figure 3-18.

   NOTE: For optimum resolution, use the highest practical vertical sensitivity of the oscilloscope keeping the trace in view with the vertical position control.

10. Read the peak power indicated by the 8900B and calculate peak power from the source as follows:

$$P_{pk\ source} = \frac{P_{pk\ 8900B}}{\left[\log_{10}^{-1}\left(\frac{-DB}{10}\right)\right]}$$

where

$P_{pk\ 8900B}$ = Peak power measured in Step 10.

DB = Actual coupling factor of the directional coupler in db, at the test frequency.

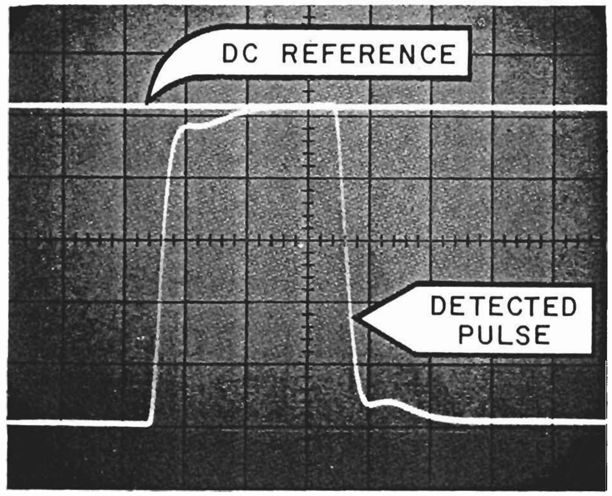


Figure 3-18. Oscilloscope display of HP Model 8900B video output enables selective positioning of DC reference on RF pulse envelope. Meter on 8900B then reads reference directly in peak pulse power.

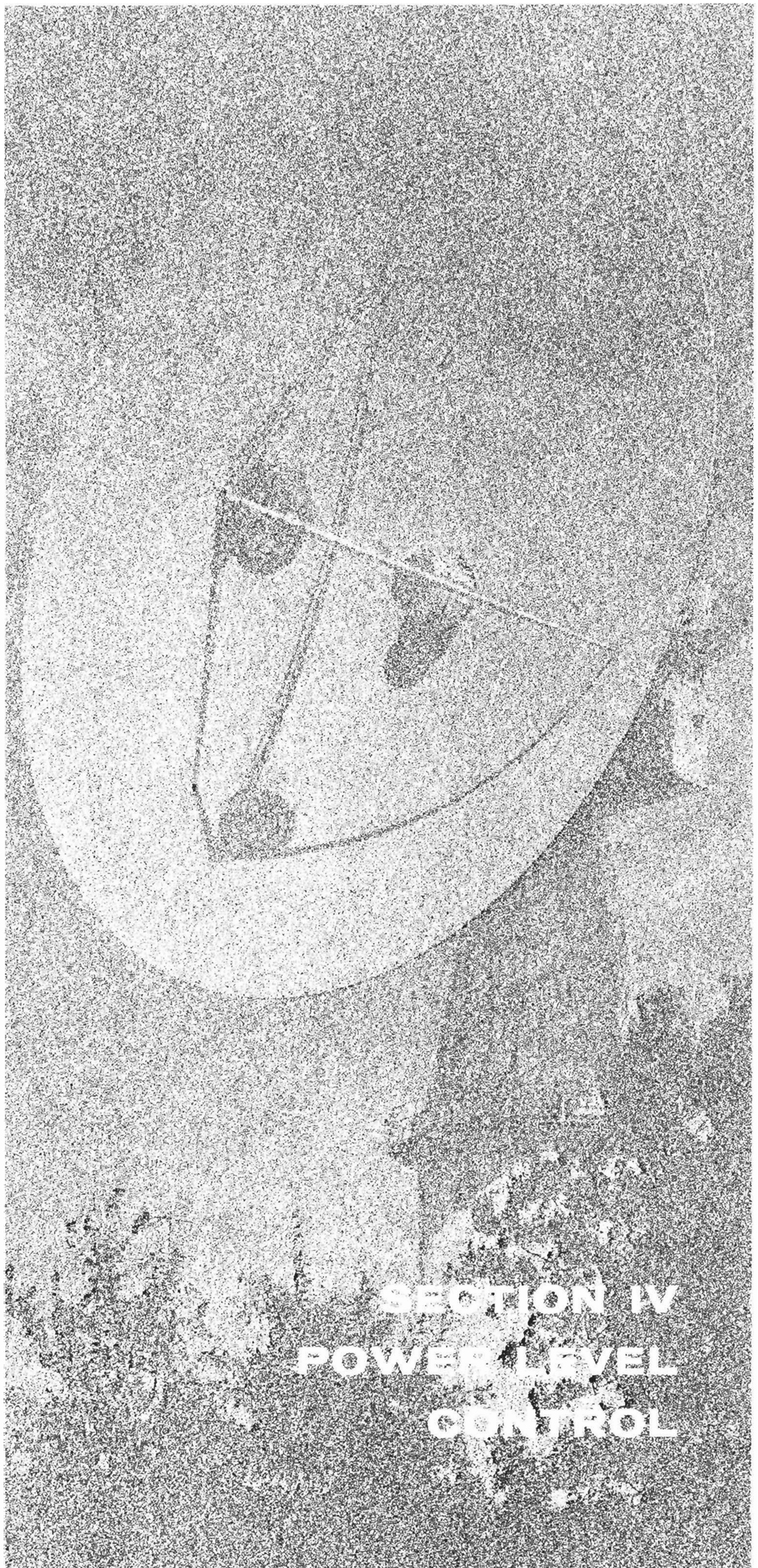
Example:

Suppose the 8900B measures 135 mw peak and the coupling factor is 50 db. Peak power from the source then is:

$$P_{pk\ source} = \frac{135 \times 10^{-3} \text{ watts}}{\left[\log_{10}^{-1}\left(\frac{-50}{10}\right)\right]}$$

$$= \frac{(135 \times 10^{-3} \text{ watts})}{1 \times 10^{-5}} = 13.5 \text{ kw}$$

The basic accuracy of the HP Model 8900B is ±1.5 db. An optional calibration chart for the 8900B may be used to correct readings to an accuracy of ±0.6 db including mismatch due to input SWR. The error analysis in Figure 3-17 relates the various error sources for this example.



SECTION IV
POWER LEVEL CONTROL

OUTPUT POWER LEVELING

BASIC LEVELING

Output leveling of microwave sources is a valuable application of temperature-compensated power meters. Figure 4-1 shows a simple leveler setup. The thermistor mount, connected to the auxiliary arm of the directional coupler, senses the incident power entering the directional coupler. The power meter produces a DC current at its recorder jack proportional to the microwave power sensed. The recorder current can be applied directly to the power meter leveling input of the hp8690-series of sweep oscillators. Here it is amplified and applied as negative feedback to maintain power at the thermistor mount constant. (In the older hp 680-series of sweepers the hp H01-8401A Leveler Amplifier can be connected between the Power Meter and Sweeper External Modulation Input.)

Typically the amount of negative feedback is so great that the quality of the leveled output appearing at the output of the primary arm of the directional coupler is a function of the coupling factor of the directional coupler and to a lesser extent, variations in the effective mount efficiency of the mount used to sense the power.

A second feature of leveling is that it creates a composite generator having a very low reflection coefficient. With high leveler loop gain it can be shown that this apparent generator reflection coefficient is essentially equal to the directivity of the coupler. In other words, a 40-db directivity directional coupler gives an apparent reflection coefficient of .01 which is a generator SWR of 1.02.

A third feature of using a power meter leveler is that one knows the output power of the primary arm of the directional coupler. The swept plot of Figure 4-2 was obtained very easily and very quickly by successively decreasing the power at the secondary arm of the directional coupler by 1-db steps using the Sweep Oscillator Power Level control. One can go in steps of 5 db by changing the range switch on the monitoring power meter. This is how the -5 and -10 db lines in Figure 4-2 were plotted. In concept, this is a very flat, accurate, low SWR microwave attenuator.

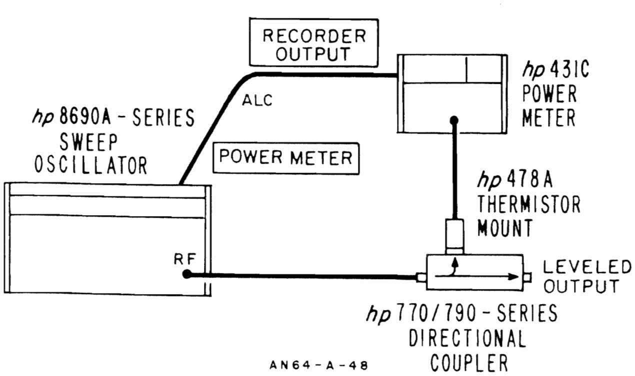


Figure 4-1. Conventional power meter leveled system in coax.

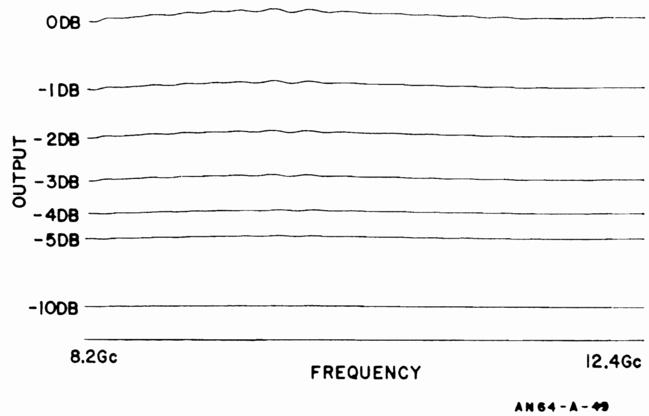


Figure 4-2. Power output of a super-leveled waveguide system using power meter control.

SUPER LEVELING

One of the characteristics of the multi-hole waveguide directional coupler is that it has a 1-db variation across a waveguide band. This means that when using a waveguide coupler for leveling, the output will vary because the coupling is not flat. A way of compensating for this is to use the setup of Figure 4-3, namely adding a 3-db waveguide coupler with a load as indicated in the block diagram. The variation in coupling factor of a waveguide coupler such as the hp 752-series, is the same for a 20-db, a 10-db, or a 3-db coupler. With a 3-db coupler, the power arriving at the output of the primary arm of the coupler is the inverse of the variation of coupling appearing at the secondary arm. Since this is the case we now have, in effect, a flat 13 or 23-db directional coupler using this connection. The flat output achieved in Figure 4-2 was a result of using this super leveling combination. Figure 4-4 shows the variation in output power of the same super leveling greatly expanded. In situations where two couplers are used, such as in a reflectometer, one

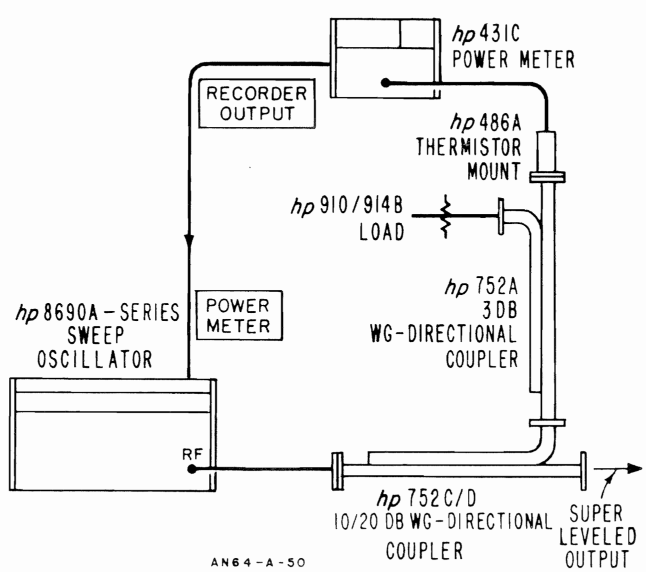


Figure 4-3. Super-leveled waveguide system.

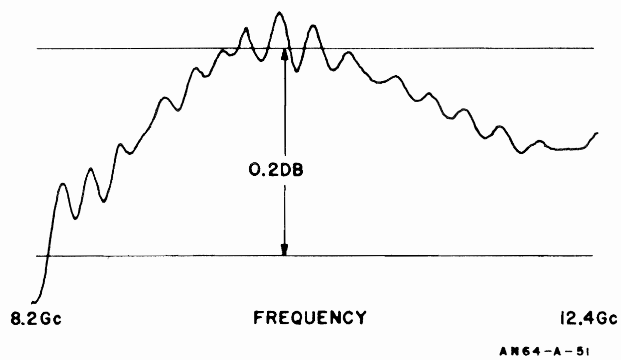


Figure 4-4. High resolution X-Y Plot shows output flatness of a super leveled system.

compensates the other and use of the super leveling is unnecessary.

LEVELED 1-WATT SWEEPER

Placing a traveling wave tube amplifier between the output of the sweeper and the directional coupler of Figure 4-1 yields a 1-watt leveled sweeper. With this setup it is normally best to level the sweeper rather than the TWT amplifier for two reasons: First, the sweeper usually has better control characteristics than the traveling wave tube amplifier and second, one is much less likely to overload the traveling wave tube amplifier. Overloading the TWT can cause spurious oscillation and high level harmonics in the output.

PRECISE OUTPUT MICROWAVE GENERATOR

A generator with an accurate power output can be created by using the system shown in Figure 4-5 if one has a thermistor mount in combination with a directional coupler which has been calibrated by NBS or other standards lab for calibration factor*. Dividing the power desired out of the main arm of the directional coupler by the calibration factor yields the substituted DC power required into the thermistor mount on the secondary arm. With the signal generator turned off, one applies this required amount of DC power from the calibrator to the power meter and notes the indication of the voltmeter across the 1K ohm resistor. Then one turns off the DC current from the calibrator, turns on the generator and varies the output to get the same indication on the voltmeter across the resistor. If the X382 Variable Attenuator in the Figure is set to approximately 1/2 db, it acts as a fine power control and the signal generator attenuation is used for the coarse power adjustments. If one does not have a thermistor mount in combination with a directional coupler which has been calibrated, but rather a mount which has been calibrated for effective efficiency or calibration factor, this mount and the power meter can be used on the output of the primary arm of the directional coupler, and the attenuator adjusted for exactly the desired microwave power output.

A DC TO 100 CPS POWER METER

The DC Calibration and Substitution Jack of the hp 431 Power Meter will not only accurately measure DC power but it will measure any power from DC through 100 cps. Sometimes in measuring noise or other phenomena, difficulties are encountered because most RMS voltmeters cut off at 10 cycles. Figure 4-6 shows the setup and relationships between the power meter reading and input power for making measurements of power from DC through 100 cps.

*See last paragraph "RF Losses and DC-Microwave Substitution Error" in Section II.

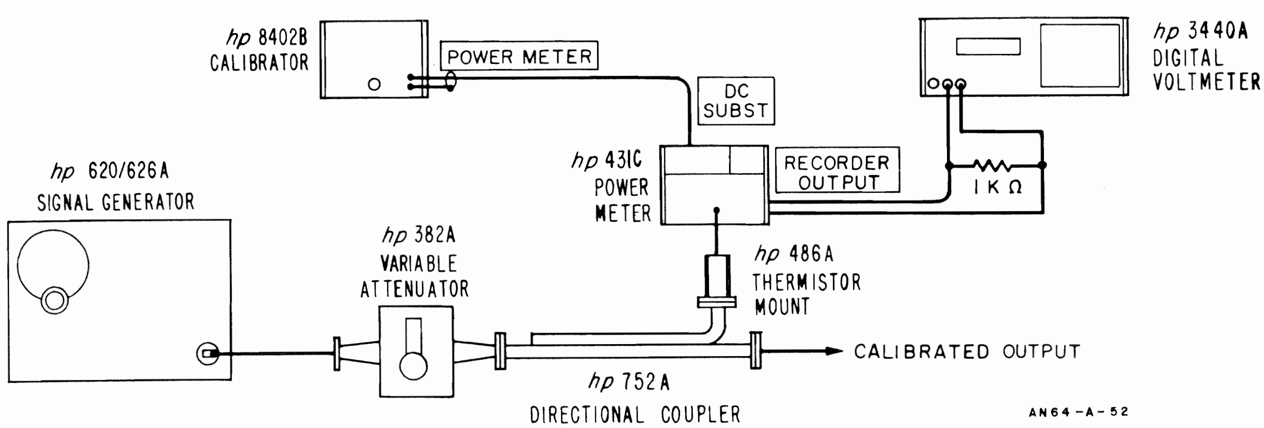


Figure 4-5. A precise output generator uses versatility of hp 431B Power Meter and a calibrated power standard (3 db coupler-thermistor mount combination).

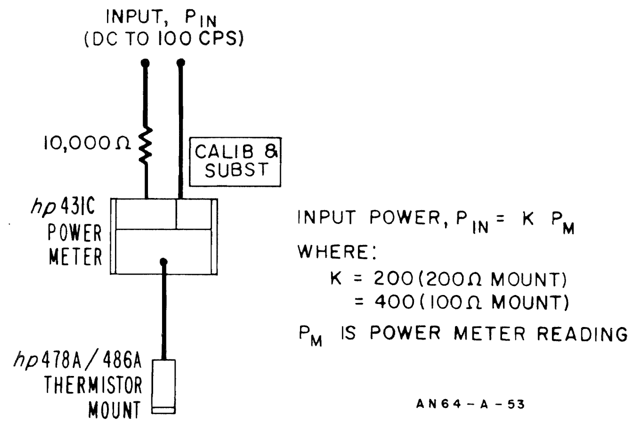
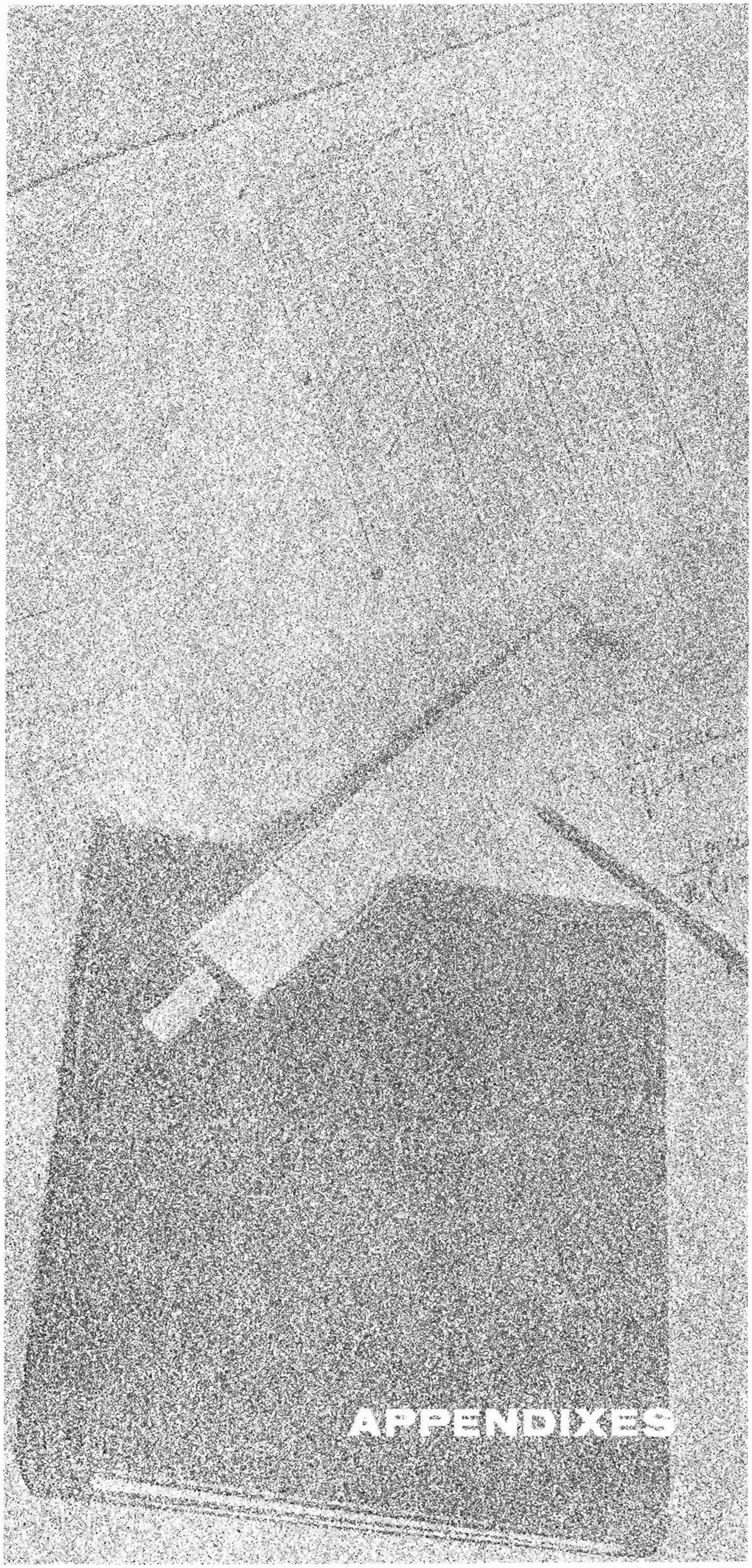


Figure 4-6. Setup for reading power from DC to 100 CPS. Technique is particularly useful in measuring power below the frequency range of RMS voltmeters.

MISCELLANEOUS TECHNIQUES

The hp 431C temperature compensated power meter lends itself well to many applications. Some of the most significant applications have been described here in detail. Other possibilities include use of the DC Substitution input circuitry on the 431C for remote zeroing of the meter, use of the recorder output for readout of transmitter power at remote locations or actuation of automatic alarm circuits. The ratio of two power meter recorder outputs may be used for precise measurements of small attenuations.

A special version of the 431 provides input switching and individual bridge balance circuits for up to five compensated thermistor inputs. This instrument (hp H20-431B) is especially useful for systems power monitoring at several points without the need for connecting and disconnecting the thermistor mount for each measurement.


APPENDIXES

APPENDIX I
POWER TRANSFER EQUATIONS

Equations 3.4 and 3.7, used in Section III for correcting power meter readings are derived from the flowgraph shown in Figure A-1. From the flowgraph we may write the following relationships:

$$E_i = \frac{E_g e^{-j\theta}}{1 - \Gamma_m \Gamma_g e^{-2j\theta}} \text{ and,} \tag{1}$$

$$E_r = E_i \Gamma_m = \frac{E_g \Gamma_m e^{-j\theta}}{1 - \Gamma_m \Gamma_g e^{-2j\theta}} \tag{2}$$

Power dissipated in the thermistor mount is given by the expression:

$$P = \left|\frac{E_i^2}{Z_o}\right| - \left|\frac{E_r^2}{Z_o}\right| \tag{3}$$

Taking mount efficiency into account, the power indicated by the meter is:

$$P_{\text{indicated}} = \eta_e \left[\left|\frac{E_i^2}{Z_o}\right| - \left|\frac{E_r^2}{Z_o}\right|\right] \tag{4}$$

Substituting for E_r,

$$P_{\text{indicated}} = \eta_e \left[\left|\frac{E_i^2}{Z_o}\right| - \left|\frac{\Gamma_m^2 E_i^2}{Z_o}\right|\right.$$

$$= \eta_e \left|\frac{E_i^2}{Z_o}\right| \left(1 - \left|\Gamma_m^2\right|\right) \tag{5}$$

Substituting for E_i with the expression given in (1),

$$P_{\text{indicated}} = \eta_e \left|\frac{E_g^2 e^{-2j\theta}}{Z_o \left(1 - \Gamma_m \Gamma_g e^{-2j\theta}\right)^2}\right| \left(1 - \Gamma_m^2\right) \tag{6}$$

Recall from the mismatch discussion in Section II that by definition

$$\Gamma_m = \rho_m e^{-j\theta}$$

thus

$$\left|\Gamma_m\right| = \rho_m$$

Also by definition, e^{-2j} is a phasor of unit magnitude therefore,

$$\left|e^{-2j\theta}\right| = 1$$

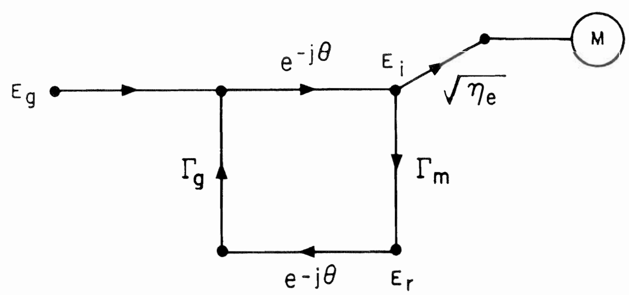
Thus the general equation for indicated power becomes,

$$P_{\text{indicated}} = \eta_e \frac{E_g^2}{Z_o} \frac{\left(1 - \rho_m^2\right)}{\left|1 - \Gamma_m \Gamma_g e^{-2j\theta}\right|^2} \tag{7}$$

Z_o Available Power

Now consider the case of power delivered to a Z_o load. In this case $\rho_m = 0$, so the general equation (7) becomes,

$$P_{\text{indicated}} = \eta_e \frac{E_g^2}{Z_o}$$



AN64-A-54

E_g = Generator Voltage
E_i = Voltage Incident to Load (Thermistor Mount)
E_r = Voltage Reflected by Load (Thermistor Mount)
η_e = Effective Efficiency of Thermistor Mount
M = Power Meter
Γ_g = Generator Reflection Coefficient
Γ_m = Load (Thermistor Mount) Reflection Coefficient
$e^{-j\theta}$ = Phase Shift Introduced by Transmission Line

Note: $\Gamma = \rho e^{-j\phi}$ where $e^{-j\phi}$ is the phasor associated with the Mount Reflection Coefficient.

Figure A-1. Flowgraph of Thermistor Mount and Power Meter Connected to a Microwave Generator transposing,

$$\frac{P_{indicated}}{\eta_e} = \frac{E_g^{\,2}}{Z_o} \text{ which is by definition } P_o \tag{8}$$

Therefore, if we want to evaluate the general equation in terms of power which can be delivered to a Z_o load, we use the equality $E_g{}^2/Z_o = P_o$ and evaluate (7) for the limit worst cases of Γ_m, Γ_g and $e^{-2j\theta}$ as follows:

$$P_{indicated} = \eta_e P_o \frac{\left(1 - \rho_m^{\,2}\right)}{\left(1 \pm \rho_m \rho_g\right)^2}$$

transposing:

$$P_o = P_{indicated} \frac{\left(1 \pm \rho_m \rho_g\right)^2}{\eta_e \left(1 - \rho_m^{\,2}\right)} \tag{9}$$

By definition,

$$\eta_e \left(1 - \rho_m^{\,2}\right) = K_b \text{ therefore,}$$

$$P_o = \frac{P_{indicated} \left(1 \pm \rho_m \rho_g\right)^2}{K_b} \tag{10}$$

which is the correction equation (3. 7) given in Section III for Z_o available power.

Conjugate Available Power

In the case of a conjugate load, the phase of $\Gamma_m \Gamma_g e^{-2j\theta}$ is zero degrees. Therefore, $\Gamma_m \Gamma_g e^{-2j\theta} = \rho_m \rho_g$, and the general equation (7) becomes, in this case,

$$P_{indicated} = \eta_e \frac{E_g^{\,2}}{Z_o} \frac{\left(1 - \rho_m^{\,2}\right)}{\left(1 - \rho_m \rho_g\right)^2} \tag{11}$$

A second condition of a conjugate match is that $\rho_m = \rho_g$. Therefore, ρ_m may be replaced by ρ_g in this case, resulting in

$$P_{indicated} = \eta_e \frac{E_g^{\,2}}{Z_o} \frac{\left(1 - \rho_g^{\,2}\right)}{\left(1 - \rho_g^{\,2}\right)^2}$$

$$= \eta_e \frac{E_g^{\,2}}{Z_o \left(1 - \rho_g^{\,2}\right)}$$

transposing,

$$\frac{E_g^{\,2}}{Z_o} = P_{indicated} \frac{\left(1 - \rho_g^{\,2}\right)}{\eta e} \tag{12}$$

For a conjugate load, $P_c = P_{indicated}/\eta_e$ so by substitution, equation (12) becomes,

$$\frac{E_g^{\,2}}{Z_o} = P_c \left(1 - \rho_g^{\,2}\right) \tag{13}$$

Returning to the general equation (7) we may evaluate the indicated power in terms of a conjugate load by substituting the special case of equation (13). Thus equation (7) becomes,

$$P_{indicated} = \eta_e P_c \frac{\left(1 - \rho_g^{\,2}\right)\left(1 - \rho_m^{\,2}\right)}{\left|1 - \Gamma_m \Gamma_g e^{-2j\theta}\right|^2} \tag{14}$$

taking the worst case limits,

$$P_{indicated} = \eta_e \frac{P_c \left(1 - \rho_g^{\,2}\right)\left(1 - \rho_m^{\,2}\right)}{\left(1 \pm \rho_m \rho_g\right)^2} \tag{15}$$

transposing, conjugate available power is therefore,

$$P_c = \frac{P_{indicated} \left(1 \pm \rho_m \rho_g\right)^2}{\eta_e \left(1 - \rho_g^{\,2}\right)\left(1 \ \rho_m^{\,2}\right)} =$$

$$\frac{P_{indicated} \left(1 \pm \rho_m \rho_g\right)^2}{K_b \left(1 - \rho_g^{\,2}\right)}, \tag{16}$$

which is the correction equation (3. 4) given in Section III for conjugate available power.

APPENDIX II

BOLOMETER MOUNT EFFICIENCY MEASUREMENT (BY THE IMPEDANCE METHOD)

B. P. Hand
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Palo Alto, California

References: 1) "Determination of Efficiency of Microwave Bolometer Mounts from Impedance Data," David M. Kerns, Jrnl. of Res. NBS, June 1949, p. 579.

2) "An Improved Method of Measuring Efficiencies of UHF and Microwave Bolometer Mounts," R.W. Beatty and Frank Reggia, Jrnl. of Res. NBS, June 1955, p. 321.

These references discuss the theoretical aspects of bolometer-mount efficiency measurement by the impedance method. This note is intended to discuss the actual techniques used in making the impedance measurement in X-band waveguide. The circuit is shown in Figure A-2.

The method requires the measurement of input reflection coefficient for three values of bolometer resistance. To simplify calculations, the reflection is made zero for one value of resistance by tuning the mount. A reference level is set on the 415E with a short on the line, and the reflection is read on the 415E for each of the extreme values of resistance. The relative phase of the two reflection coefficients may be determined with the X885A and X382A in the auxiliary arm.

Possible sources of error which must be accounted for are:

(1) The imperfection of the short used to set reference level. It must be calibrated for its own reflection coefficient, which will be in the order of 1% to 2% less than 100%. The reference level setting must be corrected for this. A VSWR measurement may be made using the X382A to determine the reflection of the short.

(2) The re-reflection between detector mount and short when setting the reference level. A signal

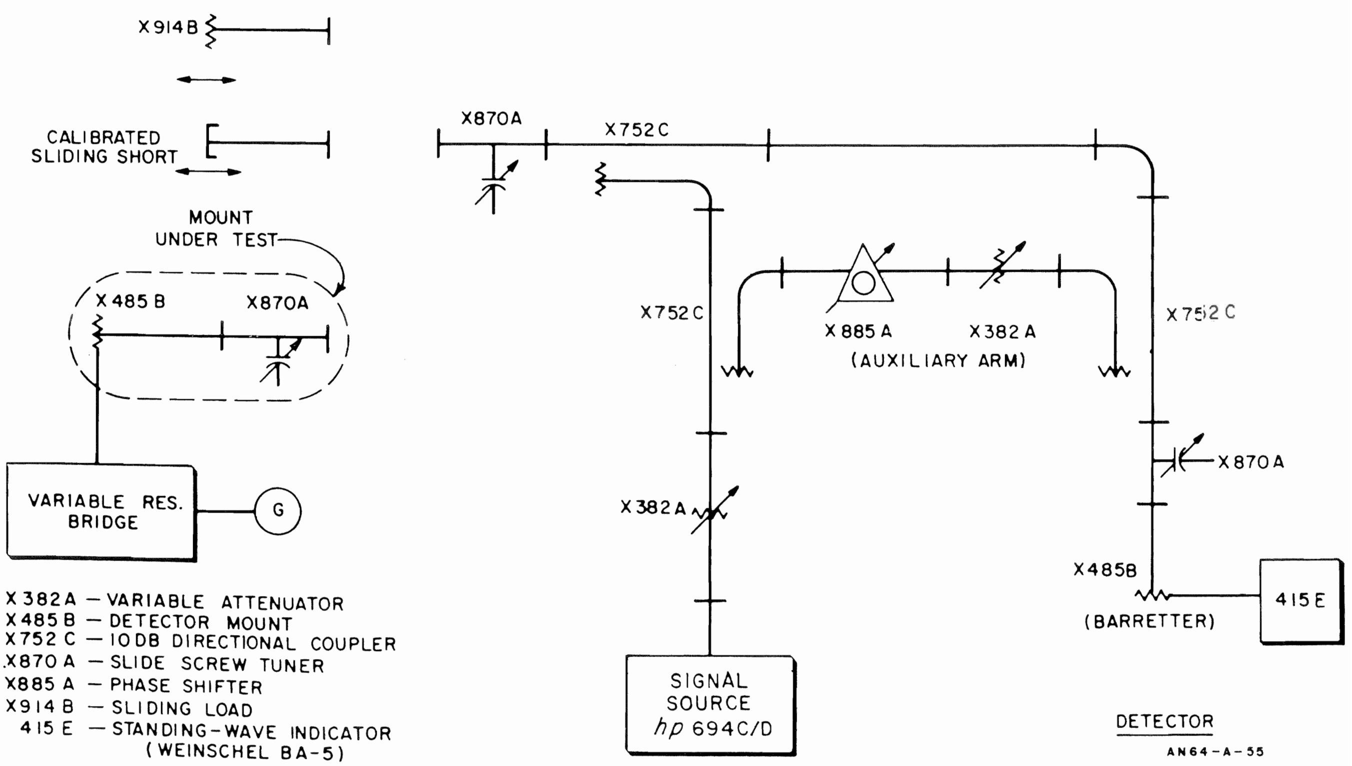

Figure A-2. Circuit Diagram

reflected from the detector will re-reflect from the short and add in random phase at the detector. Thus, an adjustable short must be used, and the detector must be tuned to give very small variation in output as the position of the short is varied. The correct reference level is the average of the maximum and minimum readings. When the output is read on a db scale, the average db reading does not correspond to average voltage. The necessary setting for average voltage is given in Figure A-3.

(3) The finite directivity of the coupler used to introduce the signal into the system. The X914B and adjacent X870A are used to cancel out the leakage signal.

(4) The calibration of the 415E. This may be calibrated at 1 kc using a ratio transformer. Typical readings on the 415E will be at full scale on the 30 and 50 ranges, corresponding to a reflection coefficient of -20 db or 0.1. To obtain greatest resolution the expanded scale should be used.

(5) The accuracy of the resistance measurements. These should be determined to 0.1% or better. A bridge made up of G.R. 0.05% resistors and decades is suitable. A circuit diagram is given Figure A-4.

With high-efficiency mounts (>95%), the additional complication of the phase-measurement equipment is not ordinarily justified in terms of the additional accuracy obtained. The second reference gives an evaluation of the error incurred in omitting the phase measurement.

Procedure is as follows:

(1) Tune out the directivity signal.

Put the X914B on the line. Set the signal source to the desired frequency and adjust for maximum output. Adjust the X914B for a maximum signal on the 415E. Adjust the X870A to obtain a null on the 415E. Back off probe penetration on the X870A, as the X914B is moved back and forth, to obtain a steady reading on the 415E. The position of the probe may have to be readjusted slightly. Reduce the signal generator output to about 2 milliwatts.

(2) Set the reference level.

Put the sliding short on the line. Adjust the X870A on the detector for a minimum of variation in the 415E reading, preferably less than 1.5 db. Set the short to give the correct average reading as determined from the curve. Set the level to give a reading at full scale on the 20 range. Set the 415E METER SCALE switch to EXPAND. Set the RANGE switch to 30. Adjust the GAIN control to give a full-scale reading.

(3) Tune the mount.

Put the mount and its tuner on the line. (The combination together must be regarded as the mount being calibrated.) Adjust the tuner for a null on the 415E while the variable-resistance bridge is balanced in the 200-ohm position. R_2 is now 200 ohms, while ρ_2 is zero.

(4) Measure reflection coefficients.

Change the resistance of the bolometer by means of the bridge to obtain a reading at full scale on the 50 range (expanded). (The GAIN control setting must not be changed from step 2.) Balance the bridge while maintaining the 415E reading. Read the bolometer resistance off the bridge dials. Repeat for the other value of resistance. Typical values for $\rho_1 = \rho_3 = 0.1$ are about 160 ohms for R_1 and 245 ohms for R_3.

(5) Calculate the efficiency.

Substitute values obtained in the formula

$$\eta = \frac{2R_2\,(R_3 - R_1)\,\rho_1\,\rho_3}{(R_3 - R_2)\,(R_2 - R_1)\,(\rho_1 + \rho_3)} \quad *$$

Read the calibrated short and apply the proper correction factor for its setting and the frequency. Since it will have a few percent loss, the actual reference level is higher than indicated, and therefore the reflection coefficients are less than calculated. Thus the efficiency is lower than calculated, and the correction must be subtracted.

The above procedure need not be followed rigorously, of course. Some barretters will depart from square law at a reading of full scale on the 20 range of the 415E. Here the X382A on the signal source can be calibrated against the 415E and used to vary the level a known 10 or 20 db between the reference level and reflection coefficient measurements. A monitor should be included in the system to check on source stability. The signal level should be as high as possible during all null adjustments. Other such refinements will naturally suggest themselves, with a little experience. The above is intended to indicate the basic technique and the major sources of error to be avoided.

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*For nominal values of $R_2 = 200$

$\rho_1 = \rho_3 = 0.1$

$\rho_2 = 0,$

this formula reduces to $\eta = \dfrac{20\,(R_3 - R_1)}{(R_3 - 200)\,(200 - R_1)}$

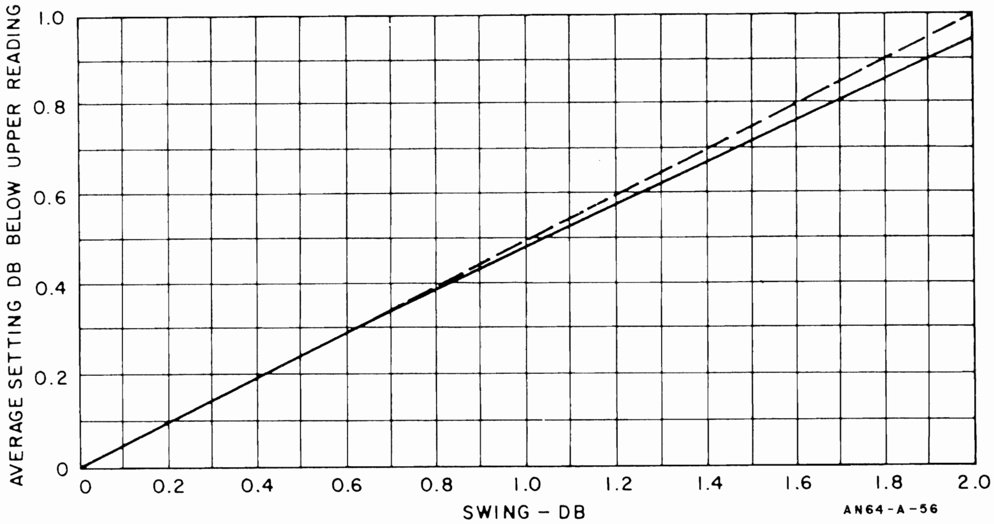
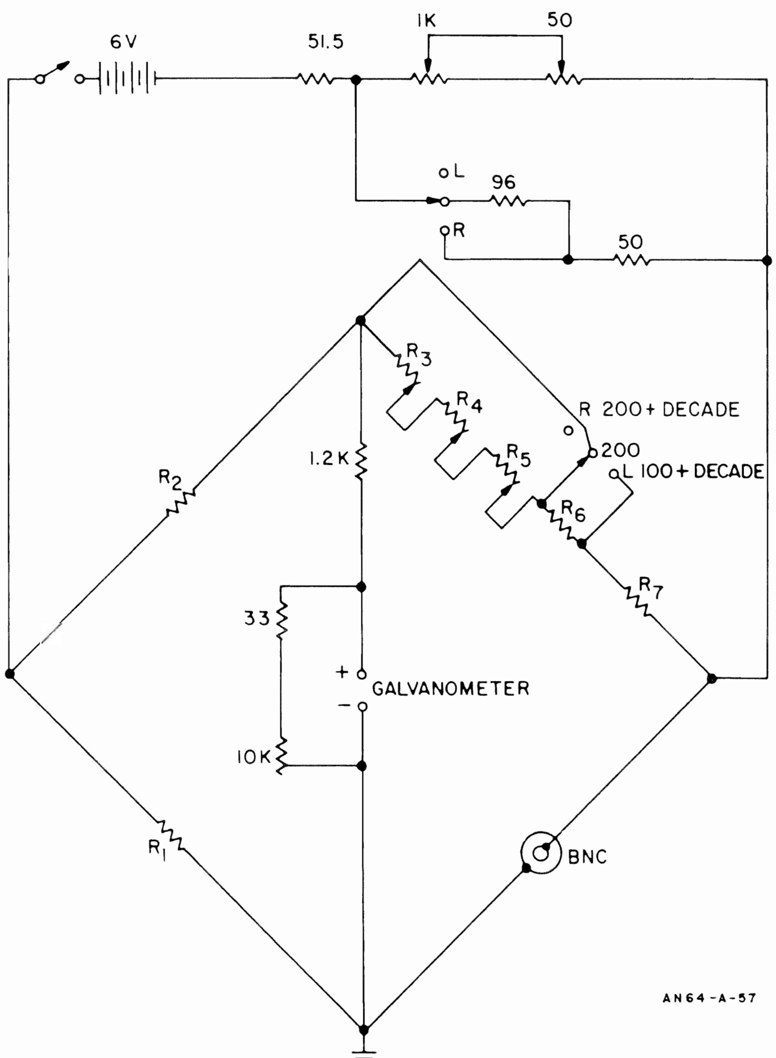


Figure A-3. DB Setting for Average Voltage



R1, R2 — 200.0Ω ± 0.05%

R3 − 0−100Ω DECADE 10Ω STEPS ± 0.05%

R4 − 0−10Ω DECADE 1Ω STEPS ± 0.15%

R5 − 0−1Ω DECADE 0.1Ω STEPS ± 0.50%

R6, R7 — 100.0Ω ± 0.05%

GALVANOMETER − USE LEEDS & NORTHROP 2400 SENSITIVITY − 0.3 μA / DIV.

SWITCH − L − 100−211Ω
CTR − 200Ω
R − 200−311Ω

R1 R2 — GENERAL RADIO 500 E
R3 — GENERAL RADIO 510 C
R4 — GENERAL RADIO 510 B
R5 — GENERAL RADIO 510 A
R6 R7 — GENERAL RADIO 500 D

CURRENT RANGES

LEFT: 4.7 ma AT 114.2Ω TO >21 ma AT >211Ω
CTR: 15.6 TO >21 ma
RIGHT: 18.2 ma AT 215.7Ω TO 21 ma AT 251Ω

BATTERY − 6V STORAGE OR 5 MERCURY CELLS (DO NOT USE FLASHLIGHT CELLS)

Figure A-4. Variable Resistance Bolometer Bridge

To make phase measurements on ρ_1 and ρ_3, install the auxiliary arm shown in the circuit diagram, and keep the X382A set for maximum attenuation, except during the phase measurement. After ρ_1 is determined (or R1 determined for a given ρ_1) adjust the X885A and X382A to null ρ_1 out, and read the X885A. After ρ_3 is obtained, null it out in similar fashion and again read the X885A. The difference in the X885A readings is the phase angle between ρ_1 and ρ_3. The necessary correction can then be calculated as shown on page 323 of Reference 2.

APPENDIX III TABLES

TABLE 1. DECIBELS VS. VOLTAGE AND POWER*

The Decibel Chart below indicates DB for any ratio of voltage or power up to 100 DB. For voltage ratios greater than 10 (or power ratios greater than 100) the ratio can be broken down into two products, the DB found for each separately, the two results then added. For example: To convert a voltage ratio of 200:1 to DB: 200:1 VR equals the product of 100:1 and 2:1. 100:1 equals 40 DB; 2:1 equals 6 DB. Therefore, 200:1 VR equals 40 DB + 6 DB or 46 DB.

Voltage Ratio	Power Ratio	−db +	Voltage Ratio	Power Ratio	Voltage Ratio	Power Ratio	−db +	Voltage Ratio	Power Ratio	Voltage Ratio	Power Ratio	−db +	Voltage Ratio	Power Ratio
1.0000	1.0000	0	1.000	1.000	.4467	.1995	7.0	2.239	5.012	.1995	.03981	14.0	5.012	25.12
.9886	.9772	.1	1.012	1.023	.4416	.1950	7.1	2.265	5.129	.1972	.03890	14.1	5.070	25.70
.9772	.9550	.2	1.023	1.047	.4365	.1905	7.2	2.291	5.248	.1950	.03802	14.2	5.129	26.30
.9661	.9333	.3	1.035	1.072	.4315	.1862	7.3	2.317	5.370	.1928	.03715	14.3	5.188	26.92
.9550	.9120	.4	1.047	1.096	.4266	.1820	7.4	2.344	5.495	.1905	.03631	14.4	5.248	27.54
.9441	.8913	.5	1.059	1.122	.4217	.1778	7.5	2.371	5.623	.1884	.03548	14.5	5.309	28.18
.9333	.8710	.6	1.072	1.148	.4169	.1738	7.6	2.399	5.754	.1862	.03467	14.6	5.370	28.84
.9226	.8511	.7	1.084	1.175	.4121	.1698	7.7	2.427	5.888	.1841	.03388	14.7	5.433	29.51
.9120	.8318	.8	1.096	1.202	.4074	.1660	7.8	2.455	6.026	.1820	.03311	14.8	5.495	30.20
.9016	.8128	.9	1.109	1.230	.4027	.1622	7.9	2.483	6.166	.1799	.03236	14.9	5.559	30.90
.8913	.7943	1.0	1.122	1.259	.3981	.1585	8.0	2.512	6.310	.1778	.03162	15.0	5.623	31.62
.8810	.7762	1.1	1.135	1.288	.3936	.1549	8.1	2.541	6.457	.1758	.03090	15.1	5.689	32.36
.8710	.7586	1.2	1.148	1.318	.3890	.1514	8.2	2.570	6.607	.1738	.03020	15.2	5.754	33.11
.8610	.7413	1.3	1.161	1.349	.3846	.1479	8.3	2.600	6.761	.1718	.02951	15.3	5.821	33.88
.8511	.7244	1.4	1.175	1.380	.3802	.1445	8.4	2.630	6.918	.1698	.02884	15.4	5.888	34.67
.8414	.7079	1.5	1.189	1.413	.3758	.1413	8.5	2.661	7.079	.1679	.02818	15.5	5.957	35.48
.8318	.6918	1.6	1.202	1.445	.3715	.1380	8.6	2.692	7.244	.1660	.02754	15.6	6.026	36.31
.8222	.6761	1.7	1.216	1.479	.3673	.1349	8.7	2.723	7.413	.1641	.02692	15.7	6.095	37.15
.8128	.6607	1.8	1.230	1.514	.3631	.1318	8.8	2.754	7.586	.1622	.02630	15.8	6.166	38.02
.8035	.6457	1.9	1.245	1.549	.3589	.1288	8.9	2.786	7.762	.1603	.02570	15.9	6.237	38.90
.7943	.6310	2.0	1.259	1.585	.3548	.1259	9.0	2.818	7.943	.1585	.02512	16.0	6.310	39.81
.7852	.6166	2.1	1.274	1.622	.3508	.1230	9.1	2.851	8.128	.1567	.02455	16.1	6.383	40.74
.7762	.6026	2.2	1.288	1.660	.3467	.1202	9.2	2.884	8,318	.1549	.02399	16.2	6.457	41.69
.7674	.5888	2.3	1.303	1.698	.3428	.1175	9.3	2.917	8.511	.1531	.02344	16.3	6.531	42.66
.7586	.5754	2.4	1.318	1.738	.3388	.1148	9.4	2.951	8.710	.1514	.02291	16.4	6.607	43.65
.7499	.5623	2.5	1.334	1.778	.3350	.1122	9.5	2.985	8,913	.1496	.02239	16.5	6.683	44.67
.7413	.5495	2.6	1.349	1.820	.3311	.1096	9.6	3.020	9.120	.1479	.02188	16.6	6.761	45.71
.7328	.5370	2.7	1.365	1.862	.3273	.1072	9.7	3.055	9.333	.1462	.02138	16.7	6.839	46.77
.7244	.5248	2.8	1.380	1.905	.3236	.1047	9.8	3.090	9.550	.1445	.02089	16.8	6.918	47.86
.7161	.5129	2.9	1.396	1.950	.3199	.1023	9.9	3.126	9.772	.1429	.02042	16.9	6.998	48.98
.7079	.5012	3.0	1.413	1.995	.3162	.1000	10.0	3.162	10.000	.1413	.01995	17.0	7.079	50.12
.6998	.4898	3.1	1.429	2.042	.3126	.09772	10.1	3.199	10.23	.1396	.01950	17.1	7.161	51.29
.6918	.4786	3.2	1.445	2.089	.3090	.09550	10.2	3.236	10.47	.1380	.01905	17.2	7.244	52.48
.6839	.4677	3.3	1.462	2.138	.3055	.09333	10.3	3.273	10.72	.1365	.01862	17.3	7.328	53.70
.6761	.4571	3.4	1.479	2.188	.3020	.09120	10.4	3.311	10.96	.1349	.01820	17.4	7.413	54.95
.6683	.4467	3.5	1.496	2.239	.2985	.08913	10.5	3.350	11.22	.1334	.01778	17.5	7.499	56.23
.6607	.4365	3.6	1.514	2.291	.2951	.08710	10.6	3.388	11.48	.1318	.01738	17.6	7.586	57.54
.6531	.4266	3.7	1.531	2.344	.2917	.08511	10.7	3.428	11.75	.1303	.01698	17.7	7.674	58.88
.6457	.4169	3.8	1.549	2.399	.2884	.08318	10.8	3.467	12.02	.1288	.01660	17.8	7.762	60.26
.6383	.4074	3.9	1.567	2.455	.2851	.08128	10.9	3.508	12.30	.1274	.01622	17.9	7.852	61.66
.6310	.3981	4.0	1.585	2.512	.2818	.07943	11.0	3.548	12.59	.1259	.01585	18.0	7.943	63.10
.6237	.3890	4.1	1.603	2.570	.2786	.07762	11.1	3.589	12.88	.1245	.01549	18.1	8.035	64.57
.6166	.3802	4.2	1.622	2.630	.2754	.07586	11.2	3.631	13.18	.1230	.01514	18.2	8.128	66.07
.6095	.3715	4.3	1.641	2.692	.2723	.07413	11.3	3.673	13.49	.1216	.01479	18.3	8.222	67.61
.6026	.3631	4.4	1.660	2.754	.2692	.07244	11.4	3.715	13.80	.1202	.01445	18.4	8.318	69.18
.5957	.3548	4.5	1.679	2.818	.2661	.07079	11.5	3.758	14.13	.1189	.01413	18.5	8.414	70.79
.5888	.3467	4.6	1.698	2.884	.2630	.06918	11.6	3.802	14.45	.1175	.01380	18.6	8.511	72.44
.5821	.3388	4.7	1.718	2.951	.2600	.06761	11.7	3.846	14.79	.1161	.01349	18.7	8.610	74.13
.5754	.3311	4.8	1.738	3.020	.2570	.06607	11.8	3.890	15.14	.1148	.01318	18.8	8.710	75.86
.5689	.3236	4.9	1.758	3.090	.2541	.06457	11.9	3.936	15.49	.1135	.01288	18.9	8.811	77.62
.5623	.3162	5.0	1.776	3.162	.2512	06310	12.0	3.981	15.85	.1122	.01259	19.0	8.913	79.43
.5559	.3090	5.1	1.799	3.236	.2483	.06166	12.1	4.027	16.22	.1109	.01230	19.1	9.016	81.28
.5495	.3020	5.2	1.820	3.311	.2455	.06026	12.2	4.074	16.60	.1096	.01202	19.2	9.120	83.18
.5433	.2951	5.3	1.841	3.388	.2427	.05888	12.3	4.121	16.98	.1084	.01175	19.3	9.226	85.11
.5370	.2884	5.4	1.862	3.467	.2399	.05754	12.4	4.169	17.38	.1072	.01148	19.4	9.333	87.10
.5309	.2818	5.5	1.884	3.548	.2371	.05623	12.5	4.217	17.78	.1059	.01122	19.5	9.441	89.13
.5248	.2754	5.6	1.905	3.631	.2344	.05495	12.6	4.266	18.20	.1047	.01096	19.6	9.550	91.20
.5188	.2692	5.7	1.928	3.715	.2317	.05370	12.7	4.315	18.62	.1035	.01072	19.7	9.661	93.33
.5129	.2630	5.8	1.950	3.802	.2291	.05248	12.8	4.365	19.05	.1023	.01047	19.8	9.772	95.50
.5070	.2570	5.9	1.972	3.890	.2265	.05129	12.9	4.416	19.50	.1012	.01023	19.9	9.886	97.72
.5012	.2512	6.0	1.995	3.981	.2239	.05012	13.0	4.467	19.95	.1000	.01000	20.0	10.000	100.00
.4955	.2455	6.1	2.018	4.074	.2213	.04898	13.1	4.519	20.42					
.4898	.2399	6.2	2.042	4.169	.2188	.04786	13.2	4.571	20.89		10^{-3}	30		10^{3}
.4842	.2344	6.3	2.065	4.266	.2163	.04677	13.3	4.624	21.38	10^{-2}	10^{-4}	40	10^{2}	10^{4}
.4786	.2291	6.4	2.089	4.365	.2138	.04571	13.4	4.677	21.88			50		10^{5}
											10^{-5}	60		10^{6}
.4732	.2239	6.5	2.113	4.467	.2113	.04467	13.5	4.732	22.39	10^{-3}	10^{-6}	70	10^{3}	10^{7}
.4677	.2188	6.6	2.138	4.571	.2089	.04365	13.6	4.786	22.91		10^{-7}	80		10^{8}
.4624	.2138	6.7	2.163	4.677	.2065	.04266	13.7	4.842	23.44	10^{-4}	10^{-8}	90	10^{4}	10^{9}
.4571	.2089	6.8	2.188	4.786	.2042	.04169	13.8	4.898	23.99	10^{-5}	10^{-9}			
.4519	.2042	6.9	2.213	4.898	.2018	.04074	13.9	4.955	24.55		10^{-10}	100	10^{5}	10^{10}

*Reprinted from the 1965 Microwave Engineers Handbook and buyers' guide, courtesy Horizon House-Microwave, Inc., Dedham, Mass.

TABLE 2. COMMON LOGARITHMS

FOUR-PLACE LOGARITHMS

N	0	1	2	3	4	5	6	7	8	9	Proportional Parts								
											1	2	3	4	5	6	7	8	9
10	0000	0043	0086	0128	0170	0212	0253	0294	0334	0374	*4	8	12	17	21	25	29	33	37
11	0414	0453	0492	0531	0569	0607	0645	0682	0719	0755	4	8	11	15	19	23	26	30	34
12	0792	0828	0864	0899	0934	0969	1004	1038	1072	1106	3	7	10	14	17	21	24	28	31
13	1139	1173	1206	1239	1271	1303	1335	1367	1399	1430	3	6	10	13	16	19	23	26	29
14	1461	1492	1523	1553	1584	1614	1644	1673	1703	1732	3	6	9	12	15	18	21	24	27
15	1761	1790	1818	1847	1875	1903	1931	1959	1987	2014	*3	6	8	11	14	17	20	22	25
16	2041	2068	2095	2122	2148	2175	2201	2227	2253	2279	3	5	8	11	13	16	18	21	24
17	2304	2330	2355	2380	2405	2430	2455	2480	2504	2529	2	5	7	10	12	15	17	20	22
18	2553	2577	2601	2625	2648	2672	2695	2718	2742	2765	2	5	7	9	12	14	16	19	21
19	2788	2810	2833	2856	2878	2900	2923	2945	2967	2989	2	4	7	9	11	13	16	18	20
20	3010	3032	3054	3075	3096	3118	3139	3160	3181	3201	2	4	6	8	11	13	15	17	19
21	3222	3243	3263	3284	3304	3324	3345	3365	3385	3404	2	4	6	8	10	12	14	16	18
22	3424	3444	3464	3483	3502	3522	3541	3560	3579	3598	2	4	6	8	10	12	14	15	17
23	3617	3636	3655	3674	3692	3711	3729	3747	3766	3784	2	4	5	7	9	11	13	15	17
24	3802	3820	3838	3856	3874	3892	3909	3927	3945	3962	2	4	5	7	9	11	12	14	16
25	3979	3997	4014	4031	4048	4065	4082	4099	4116	4133	2	3	5	7	9	10	12	14	15
26	4150	4166	4183	4200	4216	4232	4249	4265	4281	4298	2	3	5	7	8	10	11	13	15
27	4314	4330	4346	4362	4378	4393	4409	4425	4440	4456	2	3	5	6	8	9	11	13	14
28	4472	4487	4502	4518	4533	4548	4564	4579	4594	4609	2	3	5	6	8	9	11	12	14
29	4624	4639	4654	4669	4683	4698	4713	4728	4742	4757	1	3	4	6	7	9	10	12	13
30	4771	4786	4800	4814	4829	4843	4857	4871	4886	4900	1	3	4	6	7	9	10	11	13
31	4914	4928	4942	4955	4969	4983	4997	5011	5024	5038	1	3	4	6	7	8	10	11	12
32	5051	5065	5079	5092	5105	5119	5132	5145	5159	5172	1	3	4	5	7	8	9	11	12
33	5185	5198	5211	5224	5237	5250	5263	5276	5289	5302	1	3	4	5	6	8	9	10	12
34	5315	5328	5340	5353	5366	5378	5391	5403	5416	5428	1	3	4	5	6	8	9	10	11
35	5441	5453	5465	5478	5490	5502	5514	5527	5539	5551	1	2	4	5	6	7	9	10	11
36	5563	5575	5587	5599	5611	5623	5635	5647	5658	5670	1	2	4	5	6	7	8	10	11
37	5682	5694	5705	5717	5729	5740	5752	5763	5775	5786	1	2	3	5	6	7	8	9	10
38	5798	5809	5821	5832	5843	5855	5866	5877	5888	5899	1	2	3	5	6	7	8	9	10
39	5911	5922	5933	5944	5955	5966	5977	5988	5999	6010	1	2	3	4	5	7	8	9	10
40	6021	6031	6042	6053	6064	6075	6085	6096	6107	6117	1	2	3	4	5	6	8	9	10
41	6128	6138	6149	6160	6170	6180	6191	6201	6212	6222	1	2	3	4	5	6	7	8	9
42	6232	6243	6253	6263	6274	6284	6294	6304	6314	6325	1	2	3	4	5	6	7	8	9
43	6335	6345	6355	6365	6375	6385	6395	6405	6415	6425	1	2	3	4	5	6	7	8	9
44	6435	6444	6454	6464	6474	6484	6493	6503	6513	6522	1	2	3	4	5	6	7	8	9
45	6532	6542	6551	6561	6571	6580	6590	6599	6609	6618	1	2	3	4	5	6	7	8	9
46	6628	6637	6646	6656	6665	6675	6684	6693	6702	6712	1	2	3	4	5	6	7	7	8
47	6721	6730	6739	6749	6758	6767	6776	6785	6794	6803	1	2	3	4	5	5	6	7	8
48	6812	6821	6830	6839	6848	6857	6866	6875	6884	6893	1	2	3	4	5	5	6	7	8
49	6902	6911	6920	6928	6937	6946	6955	6964	6972	6981	1	2	3	4	4	5	6	7	8
50	6990	6998	7007	7016	7024	7033	7042	7050	7059	7067	1	2	3	3	4	5	6	7	8
51	7076	7084	7093	7101	7110	7118	7126	7135	7143	7152	1	2	3	3	4	5	6	7	8
52	7160	7168	7177	7185	7193	7202	7210	7218	7226	7235	1	2	2	3	4	5	6	7	7
53	7243	7251	7259	7267	7275	7284	7292	7300	7308	7316	1	2	2	3	4	5	6	6	7
54	7324	7332	7340	7348	7356	7364	7372	7380	7388	7396	1	2	2	3	4	5	6	6	7
N	0	1	2	3	4	5	6	7	8	9	1	2	3	4	5	6	7	8	9

* Interpolation in this section of the table is inaccurate.

FOUR-PLACE LOGARITHMS

N	0	1	2	3	4	5	6	7	8	9	Proportional Parts								
											1	2	3	4	5	6	7	8	9
55	7404	7412	7419	7427	7435	7443	7451	7459	7466	7474	1	2	2	3	4	5	5	6	7
56	7482	7490	7497	7505	7513	7520	7528	7536	7543	7551	1	2	2	3	4	5	5	6	7
57	7559	7566	7574	7582	7589	7597	7604	7612	7619	7627	1	2	2	3	4	5	5	6	7
58	7634	7642	7649	7657	7664	7672	7679	7686	7694	7701	1	1	2	3	4	4	5	6	7
59	7709	7716	7723	7731	7738	7745	7752	7760	7767	7774	1	1	2	3	4	4	5	6	7
60	7782	7789	7796	7803	7810	7818	7825	7832	7839	7846	1	1	2	3	4	4	5	6	6
61	7853	7860	7868	7875	7882	7889	7896	7903	7910	7917	1	1	2	3	4	4	5	6	6
62	7924	7931	7938	7945	7952	7959	7966	7973	7980	7987	1	1	2	3	3	4	5	6	6
63	7993	8000	8007	8014	8021	8028	8035	8041	8048	8055	1	1	2	3	3	4	5	5	6
64	8062	8069	8075	8082	8089	8096	8102	8109	8116	8122	1	1	2	3	3	4	5	5	6
65	8129	8136	8142	8149	8156	8162	8169	8176	8182	8189	1	1	2	3	3	4	5	5	6
66	8195	8202	8209	8215	8222	8228	8235	8241	8248	8254	1	1	2	3	3	4	5	5	6
67	8261	8267	8274	8280	8287	8293	8299	8306	8312	8319	1	1	2	3	3	4	5	5	6
68	8325	8331	8338	8344	8351	8357	8363	8370	8376	8382	1	1	2	3	3	4	4	5	6
69	8388	8395	8401	8407	8414	8420	8426	8432	8439	8445	1	1	2	2	3	4	4	5	6
70	8451	8457	8463	8470	8476	8482	8488	8494	8500	8506	1	1	2	2	3	4	4	5	6
71	8513	8519	8525	8531	8537	8543	8549	8555	8561	8567	1	1	2	2	3	4	4	5	5
72	8573	8579	8585	8591	8597	8603	8609	8615	8621	8627	1	1	2	2	3	4	4	5	5
73	8633	8639	8645	8651	8657	8663	8669	8675	8681	8686	1	1	2	2	3	4	4	5	5
74	8692	8698	8704	8710	8716	8722	8727	8733	8739	8745	1	1	2	2	3	4	4	5	5
75	8751	8756	8762	8768	8774	8779	8785	8791	8797	8802	1	1	2	2	3	3	4	5	5
76	8808	8814	8820	8825	8831	8837	8842	8848	8854	8859	1	1	2	2	3	3	4	5	5
77	8865	8871	8876	8882	8887	8893	8899	8904	8910	8915	1	1	2	2	3	3	4	4	5
78	8921	8927	8932	8938	8943	8949	8954	8960	8965	8971	1	1	2	2	3	3	4	4	5
79	8976	8982	8987	8993	8998	9004	9009	9015	9020	9025	1	1	2	2	3	3	4	4	5
80	9031	9036	9042	9047	9053	9058	9063	9069	9074	9079	1	1	2	2	3	3	4	4	5
81	9085	9090	9096	9101	9106	9112	9117	9122	9128	9133	1	1	2	2	3	3	4	4	5
82	9138	9143	9149	9154	9159	9165	9170	9175	9180	9186	1	1	2	2	3	3	4	4	5
83	9191	9196	9201	9206	9212	9217	9222	9227	9232	9238	1	1	2	2	3	3	4	4	5
84	9243	9248	9253	9258	9263	9269	9274	9279	9284	9289	1	1	2	2	3	3	4	4	5
85	9294	9299	9304	9309	9315	9320	9325	9330	9335	9340	1	1	2	2	3	3	4	4	5
86	9345	9350	9355	9360	9365	9370	9375	9380	9385	9390	1	1	2	2	3	3	4	4	5
87	9395	9400	9405	9410	9415	9420	9425	9430	9435	9440	0	1	1	2	2	3	3	4	4
88	9445	9450	9455	9460	9465	9469	9474	9479	9484	9489	0	1	1	2	2	3	3	4	4
89	9494	9499	9504	9509	9513	9518	9523	9528	9533	9538	0	1	1	2	2	3	3	4	4
90	9542	9547	9552	9557	9562	9566	9571	9576	9581	9586	0	1	1	2	2	3	3	4	4
91	9590	9595	9600	9605	9609	9614	9619	9624	9628	9633	0	1	1	2	2	3	3	4	4
92	9638	9643	9647	9652	9657	9661	9666	9671	9675	9680	0	1	1	2	2	3	3	4	4
93	9685	9689	9694	9699	9703	9708	9713	9717	9722	9727	0	1	1	2	2	3	3	4	4
94	9731	9736	9741	9745	9750	9754	9759	9763	9768	9773	0	1	1	2	2	3	3	4	4
95	9777	9782	9786	9791	9795	9800	9805	9809	9814	9818	0	1	1	2	2	3	3	4	4
96	9823	9827	9832	9836	9841	9845	9850	9854	9859	9863	0	1	1	2	2	3	3	4	4
97	9868	9872	9877	9881	9886	9890	9894	9899	9903	9908	0	1	1	2	2	3	3	4	4
98	9912	9917	9921	9926	9930	9934	9939	9943	9948	9952	0	1	1	2	2	3	3	4	4
99	9956	9961	9965	9969	9974	9978	9983	9987	9991	9996	0	1	1	2	2	3	3	3	4
N	0	1	2	3	4	5	6	7	8	9	1	2	3	4	5	6	7	8	9

*Reprinted from CRC Standard Math Tables, 11th Ed., pp. 18, 19 Courtesy Chemical Rubber Publishing Co., Cleveland, Ohio.

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