Chapter 27

Antenna and Transmission-Line Measurements

The principal quantities to be measured on transmission lines are line current or voltage, and standing-wave ratio. Measurements of current or voltage are made for the purpose of determining the power input to the line. SWR measurements are useful in connection with the design of coupling circuits and the adjustment of the match between the antenna and transmission line, as well as in the adjustment of matching circuits.

For most practical purposes a *relative* measurement is sufficient. An uncalibrated indicator that shows when the largest possible amount of power is being put into the line is just as useful, in most cases, as an instrument that measures the power accurately. It is seldom necessary to know the actual number of watts going into the line unless the overall efficiency of the system is being investigated. An instrument that shows when the SWR is close to 1 to 1 is all that is needed for most impedance-matching adjustments. Accurate measurement of SWR is necessary only in studies of antenna characteristics such as bandwidth, or for the design of some types of matching systems, such as a stub match.

Quantitative measurements of reasonable accuracy demand good design and careful construction in the measuring instruments. They also require intelligent use of the equipment, including a knowledge not only of its limitations but also of stray effects that often lead to false results. Until the complete conditions of the measurements are known, a certain amount of skepticism regarding numerical data resulting from amateur measurements with simple equipment is justified. On the other hand, purely qualitative or relative measurements are easy to make and are reliable for the purposes mentioned above.

LINE CURRENT AND VOLTAGE

A current or voltage indicator that can be used with coaxial line is a useful piece of equipment. It need not be elaborate or expensive. Its principal function is to show when the maximum power is being taken from the transmitter; for any given set of line conditions (length, SWR, etc) this will occur when the transmitter coupling is adjusted for maximum current or voltage at the input end of the line. Although the final-amplifier plate or collector current meter is frequently used for this purpose, it is not always a reliable indicator. In many cases, particularly with a screen-grid tube in the final stage, minimum loaded plate current does not occur simultaneously with maximum power output.

RF VOLTMETER

A germanium diode in conjunction with a low-range milliammeter and a few resistors can be assembled to form an RF voltmeter suitable for connecting across the two conductors of a coaxial line, as shown in **Fig 1**. It consists of a voltage divider,

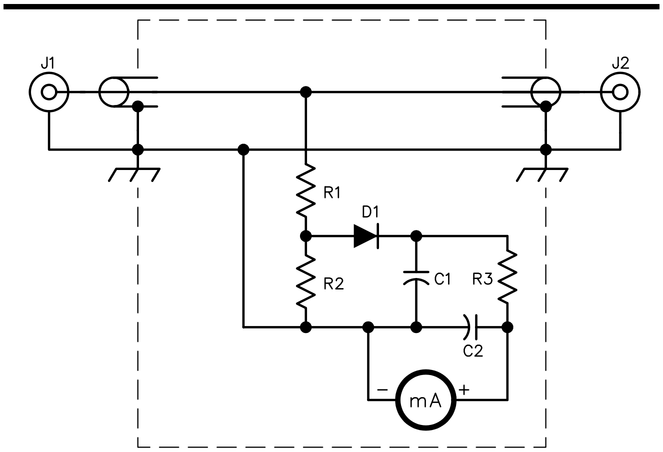


Fig 1—RF voltmeter for coaxial line.
C1, C2—0.005 or 0.01-µF ceramic.
D1—Germanium diode, 1N34A.
J1, J2—Coaxial fittings, chassis-mounting type.
M1—0-1 milliammeter (more sensitive meter may be used if desired; see text).
R1—6.8 kΩ, composition, 1 W for each 100 W of RF power.
R2—680 Ω, 1/2 or 1 W composition.
R3—10 kΩ, 1/2 W (see text).

R1-R2, having a total resistance about 100 times the Z_0 of the line (so the power consumed will be negligible) with a diode rectifier and milliammeter connected across part of the divider to read relative RF voltage. The purpose of R3 is to make the meter readings directly proportional to the applied voltage, as nearly as possible, by "swamping" the resistance of D1, since the diode resistance will vary with the amplitude of the current through the diode.

The voltmeter may be constructed in a small metal box, indicated by the dashed line in the drawing, and fitted with coax receptacles. R1 and R2 should be composition resistors. The power rating for R1 should be 1 W for each 100 W of carrier power in the matched line; separate 1 or 2-W resistors should be used to make up the total power rating required, to a total resistance as given. Any type of resistor can be used for R3; the total resistance should be such that about 10 V dc will be developed across it at full scale. For example, a 0-1 milliammeter would require 10 kΩ, a 0-500 microammeter would take 20 kΩ, and so on. For comparative measurements only, R3 may be a variable resistor so the sensitivity can be adjusted for various power levels.

In constructing such a voltmeter, care should be used to prevent inductive coupling between R1 and the loop formed by R2, D1 and C1, and between the same loop and the line conductors in the assembly. With the lower end of R1 disconnected from R2 and grounded to the enclosure, but without changing its position with respect to the loop, there should be no meter indication when full power is going through the line.

If more than one resistor is used for R1, the units should be arranged end to end with very short leads. R1 and R2 should be kept 1/2 inch or more from metal surfaces parallel to the body of the resistor. If these precautions are observed the voltmeter will give consistent readings at frequencies up to 30 MHz. Stray capacitance and stray coupling limit the accuracy at higher frequencies but do not affect the utility of the instrument for comparative measurements.

Calibration

The meter may be calibrated in RF voltage by comparison with a standard such as an RF ammeter. This requires that the line be well matched so the impedance at the point of measurement is equal to the actual Z_0 of the line. Since in that case $P = I^2 Z_0$, the power can be calculated from the current. Then $E = \sqrt{PZ_0}$. By making current and voltage measurements at a number of different power levels, enough points may be obtained to permit drawing a calibration curve for the voltmeter.

RF AMMETERS

An RF ammeter can be mounted in any convenient location at the input end of the transmission line, the principal precaution in its mounting being that the capacitance to ground, chassis, and nearby conductors should be low. A bakelite-case instrument can be mounted on a metal panel without introducing enough shunt capacitance to ground to cause serious error up to 30 MHz. When a metal-case instrument is installed on a metal panel it should be mounted on a separate sheet of insulating material in such a way that there is 1/8 inch or more separation between the edge of the case and the metal.

A 2-inch instrument can be mounted in a 2 × 4 × 4-inch metal box as shown in **Fig 2**. This is a convenient arrangement for use with coaxial line.

Installed this way, a good quality RF ammeter will measure current with an accuracy that is

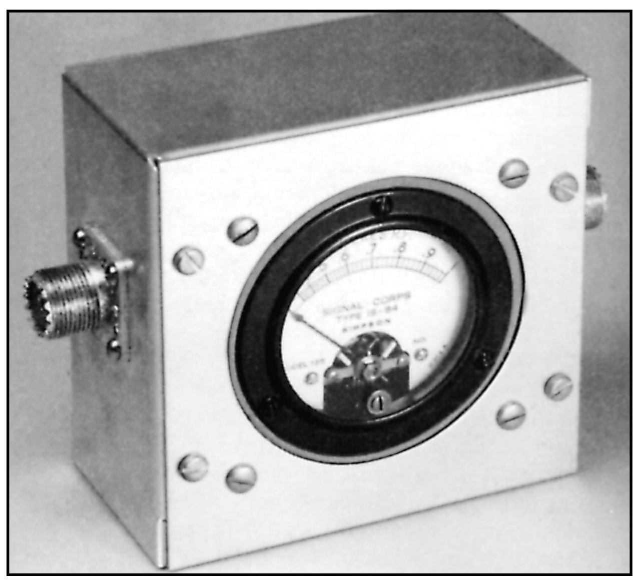


Fig 2—A convenient method of mounting an RF ammeter for use in a coaxial line. This is a metal-case instrument mounted on a thin bakelite panel. The cutout in the metal clears the edge of the meter by about 1/8 inch.

entirely adequate for calculating power in the line. As discussed above in connection with calibrating RF voltmeters, the line must be closely matched by its load so the actual impedance will be resistive and equal to Z_0. The scales of such instruments are cramped at the low end, however, which limits the range of power that can be measured by a single meter. The useful current range is about 3 to 1, corresponding to a power range of about 9 to 1.

SWR Measurements

On parallel-conductor lines it is possible to measure the standing-wave ratio by moving a current (or voltage) indicator along the line, noting the maximum and minimum values of current (or voltage) and then computing the SWR from these measured values. This cannot be done with coaxial line since it is not possible to make measurements of this type inside the cable. The technique is, in fact, seldom used with open lines, because it is not only inconvenient but sometimes impossible to reach all parts of the line conductors. Also, the method is subject to considerable error from antenna currents flowing on the line.

Present-day SWR measurements made by amateurs practically always use some form of "directional coupler" or RF bridge circuit. The indicator circuits themselves are fundamentally simple, but considerable care is required in their construction if the measurements are to be accurate. The requirements for indicators used only for the adjustment of impedance-matching circuits, rather than actual SWR measurement, are not so stringent and an instrument for this purpose can be made easily.

BRIDGE CIRCUITS

Two commonly used bridge circuits are shown in **Fig 3**. The bridges consist essentially of two voltage dividers in parallel, with a voltmeter connected between the junctions of each pair of "arms," as the individual elements are called. When the equations shown to the right of each circuit are satisfied there is no potential difference between the two junctions, and the voltmeter indicates zero voltage. The bridge is then said to be in "balance."

Taking Fig 3A as an illustration, if R1 = R2, half the applied voltage, E, will appear across each resistor. Then if $R_s = R_x$, ½ E will appear across each of these resistors and the voltmeter reading will be zero. Remember that a matched transmission line has a purely resistive input impedance, and suppose that the input terminals of such a line are substituted for R_x. Then if R_s is a resistor equal to the Z_0 of the line, the bridge will be balanced. If the line is not perfectly matched, its input impedance will not equal Z_0 and hence will not equal R_s, since the latter is chosen to equal Z_0. There will then be a difference in potential between points X and Y, and the voltmeter will show a reading. Such a bridge therefore can be used to show the presence of standing waves on the line, because the line input impedance will be equal to Z_0 only when there are no standing waves.

Considering the nature of the incident and reflected components of voltage that make up the actual voltage at the input terminals of the line, as discussed in Chapter 24, it should be clear that when $R_s = Z_0$, the bridge is always in balance for the incident component. Thus the voltmeter does not respond to the incident component at any time but reads only the reflected component (assum-

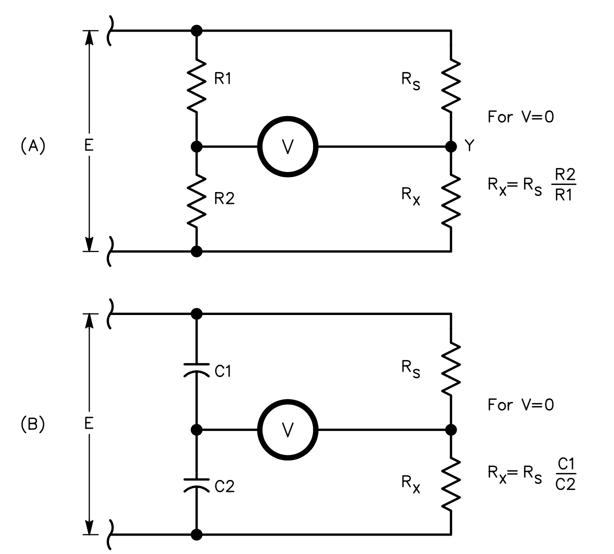


Fig 3—Bridge circuits suitable for SWR measurement. A—Wheatstone type using resistance arms. B—Capacitance-resistance bridge ("Micromatch"). Conditions for balance are independent of frequency in both types.

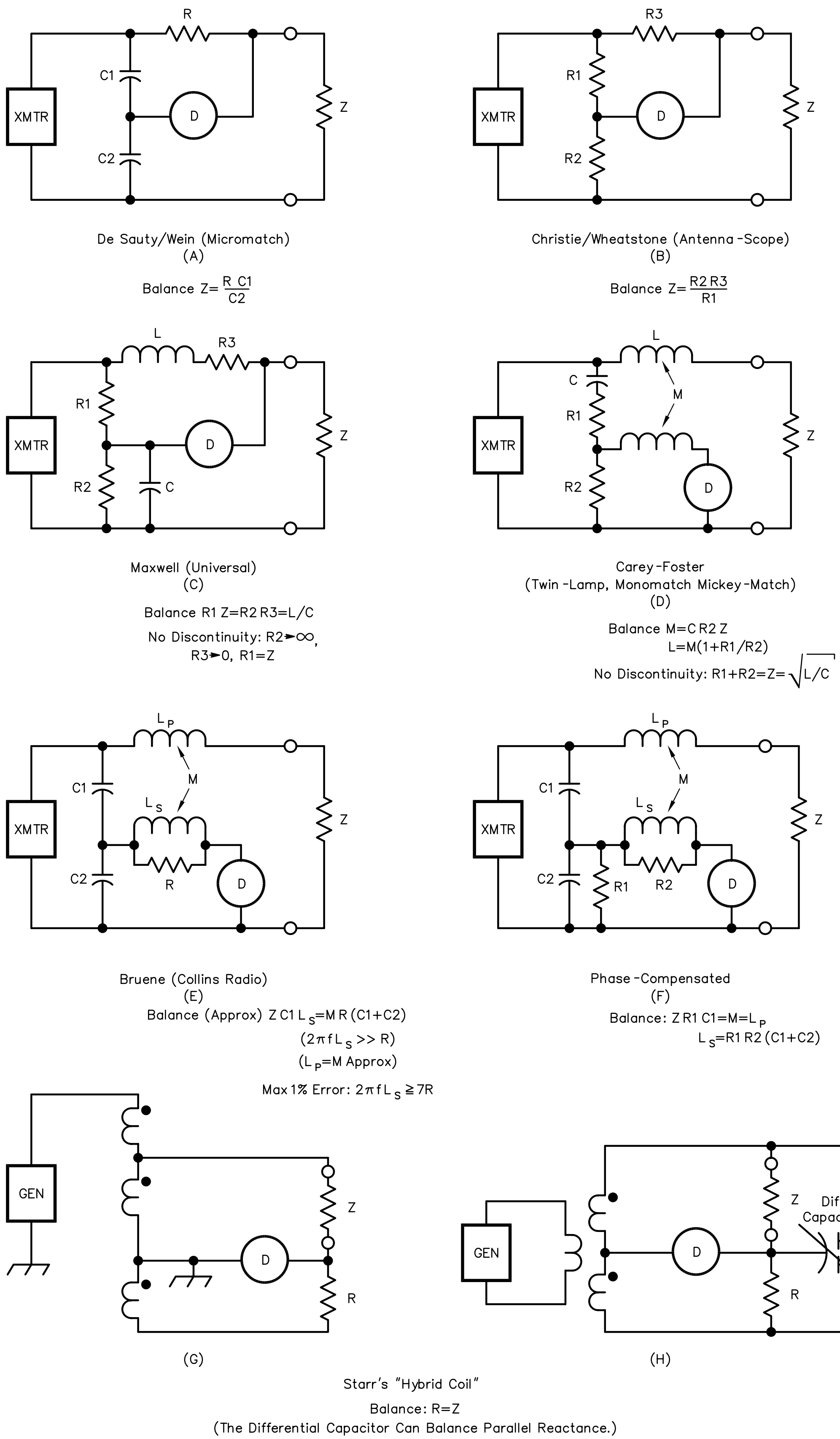


Fig 4—Various types of SWR indicator circuits and commonly known names of bridge circuits or devices in which they have been used. Detectors (D) are usually semiconductor diodes with meters, isolated with RF chokes and capacitors. However, the detector may be a radio receiver. In each circuit, Z represents the load being measured. (This information provided by David Geiser, WA2ANU)

ing that R2 is very small compared with the voltmeter impedance). The incident component can be measured across either R1 or R2, if they are equal resistances. The standing-wave ratio is then

$$\mathrm{SWR} = \frac{E1 + E2}{E1 - E2} \qquad \text{(Eq 1)}$$

where E1 is the incident voltage and E2 is the reflected voltage. It is often simpler to normalize the voltages by expressing E2 as a fraction of E1, in which case the formula becomes

$$\mathrm{SWR} = \frac{1 + k}{1 - k} \qquad \text{(Eq 2)}$$

where k = E2/E1.

The operation of the circuit in Fig 3B is essentially the same, although this circuit has arms containing reactance as well as resistance.

It is not necessary that R1 = R2 in Fig 3A; the bridge can be balanced, in theory, with any ratio of these two resistances provided R_S is changed accordingly. In practice, however, the accuracy is highest when the two are equal; this circuit is generally so used.

A number of types of bridge circuits appear in **Fig 4**, many of which have been used in amateur products or amateur construction projects. All except that at G can have the generator and load at a common potential. At G, the generator and detector are at a common potential. The positions of the detector and transmitter (or generator) may be interchanged in the bridge, and this may be an advantage in some applications.

Bridges shown at D, E, F and H may have one terminal of the generator, detector and load common. Bridges at A, B, E, F, G and H have constant sensitivity over a wide frequency range. Bridges at B, C, D and H may be designed to show no discontinuity (impedance lump) with a matched line, as shown in the drawing. Discontinuities with A, E and F may be small.

Bridges are usually most sensitive when the detector bridges the midpoint of the generator voltage, as in G or H, or in B when each resistor equals the load impedance. Sensitivity also increases when the currents in each leg are equal.

Resistance Bridge

The basic bridge type shown in Fig 3A may be home constructed and is reasonably accurate for SWR measurement. A practical circuit for such a bridge is given in **Fig 5** and a representative layout is shown in **Fig 6**. Properly built, a bridge of this design can be used for measurement of standing-wave ratios up to about 15 to 1 with good accuracy.

Important constructional points to be observed are:

1) Keep leads in the RF circuit short, to reduce stray inductance.

2) Mount resistors two or three times their body diameter away from metal parts, to reduce stray capacitance.

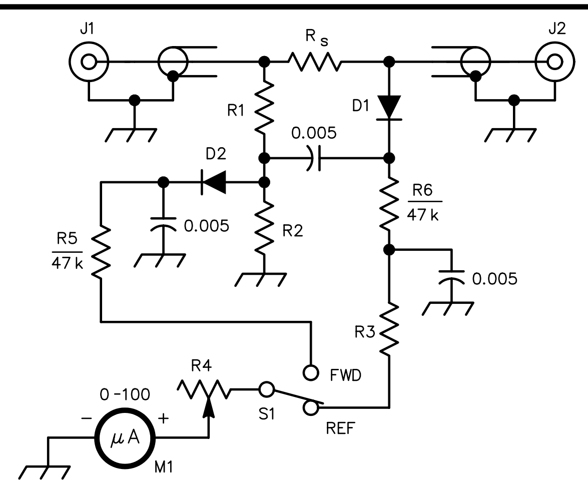


Fig 5—Resistance bridge for SWR measurement. Capacitors are disc ceramic. Resistors are 1/2-W composition except as noted below.

D1, D2—Germanium diode, high back resistance type (1N34A, 1N270, etc).
J1, J2—Coaxial connectors, chassis-mounting type.
M1—0-100 dc microammeter.
R1, R2—47 Ω, 1/2-W composition (see text).
R3—See text.
R4—50-kΩ volume control.
R_S—Resistance equal to line Z_0 (1/2 or 1 W composition).
S1—SPDT toggle.

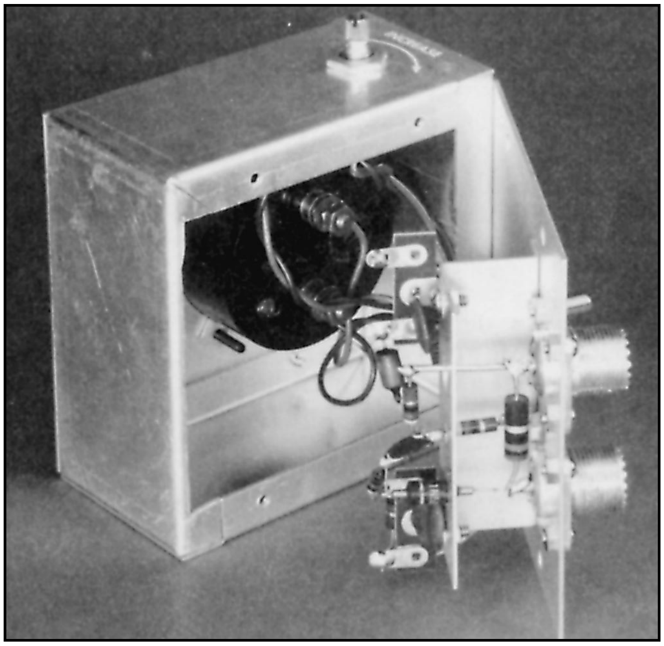


Fig 6—A 2 × 4 × 4-inch aluminum box is used to house this SWR bridge, which uses the circuit of Fig 5. The variable resistor, R4, is mounted on the side. The bridge components are mounted on one side plate of the box and a subchassis formed from a piece of aluminum. The input connector is at the top in this view. R_s is connected directly between the two center posts of the connectors. R2 is visible behind it and perpendicular to it. One terminal of D1 projects through a hole in the chassis so the lead can be connected to J2. R1 is mounted vertically to the left of the chassis in this view, with D2 connected between the junction of R1-R2 and a tie point.

3) Place the RF components so there is as little inductive and capacitive coupling as possible between the bridge arms.

In the instrument shown in Fig 6, the input and line connectors, J1 and J2, are mounted fairly close together so the standard resistor, R_s, can be supported with short leads directly between the center terminals of the connectors. R2 is mounted at right angles to R_s, and a shield partition is used between these two components and the others.

The two 47-kΩ resistors, R5 and R6 in Fig 5, are voltmeter multipliers for the 0-100 microammeter used as an indicator. This is sufficient resistance to make the voltmeter linear (that is, the meter reading is directly proportional to the RF voltage) and no voltage calibration curve is needed. D1 is the rectifier for the reflected voltage and D2 is for the incident voltage. Because of manufacturing variations in resistors and diodes, the readings may differ slightly with two multipliers of the same nominal resistance value, so a correction resistor, R3, is included in the circuit. Its value should be selected so that the meter reading is the same with S1 in either position, when RF is applied to the bridge with the line connection open. In the instrument shown, a value of 1000 Ω was required in series with the multiplier for reflected voltage; in other cases different values probably would be needed and R3 might have to be put in series with the multiplier for the incident voltage. This can be determined by experiment.

The value used for R1 and R2 is not critical, but the two resistors should be matched to within 1 or 2% if possible. The resistance of R_s should be as close as possible to the actual Z_0 of the line to be used (generally 52 or 75 Ω). The resistor should be selected by actual measurement with an accurate resistance bridge, if one is available.

R4 is for adjusting the incident-voltage reading to full scale in the measurement procedure described below. Its use is not essential, but it offers a convenient alternative to exact adjustment of the RF input voltage.

Testing

R1, R2 and R_s should be measured with a reliable ohmmeter or resistance bridge after wiring is completed, in order to make sure their values have not changed from the heat of soldering. Disconnect one side of the microammeter and leave the input and output terminals of the unit open during such measurements, in order to avoid stray shunt paths through the rectifiers.

Check the two voltmeter circuits as described above, applying enough RF (about 10 V) to the input terminals to give a full-scale reading with the line terminals open. If necessary, try different values for R3 until the reading is the same with S1 in either position.

With J2 open, adjust the RF input voltage and R4 for full-scale reading with S1 in the incident-voltage position. Then switch S1 to the reflected-voltage position. The reading should remain at full scale. Next, short-circuit J2 by touching a screwdriver between the center terminal and the frame of the connector to make a low-inductance short. Switch S1 to the incident-voltage position and readjust

R4 for full scale, if necessary. Then throw S1 to the reflected-voltage position, keeping J2 shorted, and the reading should be full scale as before. If the readings differ, R1 and R2 are not the same value, or there is stray coupling between the arms of the bridge. It is necessary that the reflected voltage read full scale with J2 either open or shorted, when the incident voltage is set to full scale in each case, in order to make accurate SWR measurements.

The circuit should pass these tests at all frequencies at which it is to be used. It is sufficient to test at the lowest and highest frequencies, usually 1.8 or 3.5 and 28 or 50 MHz. If R1 and R2 are poorly matched but the bridge construction is otherwise good, discrepancies in the readings will be substantially the same at all frequencies. A difference in behavior at the low and high ends of the frequency range can be attributed to stray coupling between bridge arms, or stray inductance or capacitance in the arms.

To check the bridge for balance, apply RF and adjust R4 for full scale with J2 open. Then connect a resistor identical with R_S (the resistance should match within 1 or 2%) to the line terminals, using the shortest possible leads. It is convenient to mount the test resistor inside a cable connector (PL-259), a method of mounting that also minimizes lead inductance. When the test resistor is connected, the reflected-voltage reading should drop to zero. The incident voltage should be reset to full scale by means of R4, if necessary. The reflected reading should be zero at any frequency in the range to be used. If a good null is obtained at low frequencies but some residual current shows at the high end, the trouble may be the inductance of the test resistor leads, although it may also be caused by stray coupling between the arms of the bridge itself. If there is a constant low (but not zero) reading at all frequencies the cause is poor matching of the resistance values. Both effects can be present simultaneously. A good null must be obtained at all frequencies before the bridge is ready for use.

Bridge Operation

The RF power input to a bridge of this type must be limited to a few watts at most, because of the power-dissipation ratings of the resistors. If the transmitter has no provision for reducing power output to a very low value—less than 5 W—a simple "power absorber" circuit can be made up as shown in **Fig 7**. The lamp DS1 tends to maintain constant current through the resistor over a fairly wide power range, so the voltage drop across the resistor also tends to be constant. This voltage is applied to the bridge, and with the constants given is in the right range for resistance-type bridges.

To make the measurement, connect the unknown to J2 and apply sufficient RF voltage to J1 to give a full-scale incident-voltage reading. Use R4 to set the indicator to exactly full scale. Then throw S1 to the reflected voltage position and note the meter reading. The SWR is then found by substituting the readings in Eq 1.

For example, if the full-scale calibration of the dc instrument is 100 µA and the reading with S2 in the reflected-voltage position is 40 µA, the SWR is

$$\text{SWR} = \frac{100 + 40}{100 - 40} = \frac{140}{60} = 2.33 \text{ to } 1$$

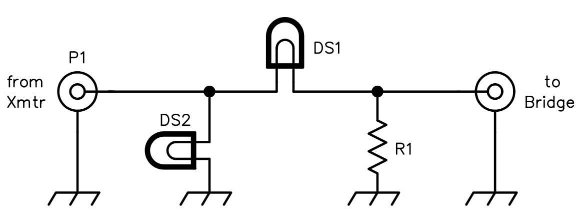


Fig 7—"Power absorber" circuit for use with resistance-type SWR bridges when the transmitter has no special provisions for power reduction. For RF powers up to 50 W, DS1 is a 117-V 40-W incandescent lamp and DS2 is not used. For higher powers, use sufficient additional lamp capacity at DS2 to load the transmitter to about normal output; for example, for 250 W output DS2 may consist of two 100-W lamps in parallel. R1 is made from three 1-W 68-Ω resistors connected in parallel. P1 and P2 are cable-mounting coaxial connectors. Leads in the circuit formed by the lamps and R1 should be kept short, but convenient lengths of cable may be used between this assembly and the connectors.

Instead of determining the SWR value by calculations, the *voltage* curve of **Fig 8** may be used. In this example the ratio of reflected to forward voltage is 40/100 = 0.4, and from Fig 8 the SWR value is seen to be about 2.3 to 1.

The meter scale may be calibrated in any arbitrary units as long as the scale has equal divisions, since it is the ratios of the voltages, and not the actual values, that determine the SWR.

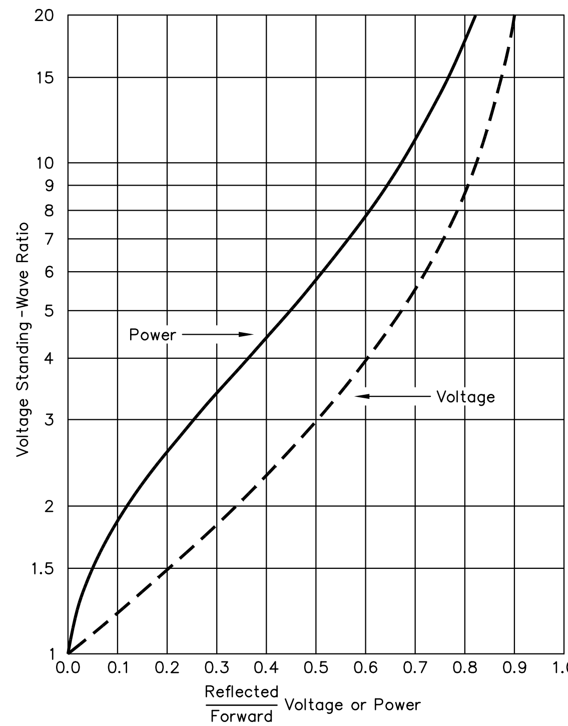


Fig 8—Chart for finding voltage standing-wave ratio when the ratio of reflected-to-forward voltage or reflected-to-forward power is known.

AVOIDING ERRORS IN SWR MEASUREMENTS

The principal causes of inaccuracies within the bridge are differences in the resistances of R1 and R2, stray inductance and capacitance in the bridge arms, and stray coupling between arms. If the checking procedure described above is followed carefully, the bridge of Fig 5 should be amply accurate for practical use. The accuracy is highest for low standing-wave ratios because of the nature of the SWR calculation; at high ratios the divisor in the equation above represents the difference between two nearly equal quantities, so a small error in voltage measurement may mean a considerable difference in the calculated SWR.

The standard resistor R_s must equal the actual Z_0 of the line. The actual Z_0 of a sample of line may differ by a few percent from the nominal figure because of manufacturing variations, but this has to be tolerated. In the 52 to 75-Ω range, the RF resistance of a composition resistor of $^1/_2$ or 1 W rating is essentially identical with its dc resistance.

"Antenna" Currents

As explained in Chapter 26, there are two ways in which "parallel" or "antenna" currents can be caused to flow on the *outside* of a coaxial line—currents induced on the line because of its spatial relationship to the antenna, and currents that result from the direct connection between the coax outer conductor and (usually) one side of the antenna. The induced current usually will not be troublesome if the bridge and the transmitter (or other source of RF power for operating the bridge) are shielded so that any RF currents flowing on the outside of the line cannot find their way into the bridge. This point can be checked by "cutting in" an additional section of line ($^1/_8$ to $^1/_4$ electrical wavelength, preferably) of the same Z_0. The SWR indicated by the bridge should not change except possibly for a slight decrease because of the additional line loss. If there is a marked change, better shielding may be required.

Parallel-type currents caused by the connection to the antenna will cause a change in SWR with line length, even though the bridge and transmitter are well shielded and the shielding is maintained throughout the system by the use of coaxial fittings. Often, merely moving the transmission line around will cause the indicated SWR to change. This is because the outside of the coax tends to become part of the antenna system, being connected to the antenna at the feed point, and so constitutes a load on the line, along with the desired load represented by the antenna itself. The SWR on the line then is determined by the composite load of the antenna and the outside of the coax, and since changing the line length (or position) changes one component of this composite load, the SWR changes too.

The remedy for such a situation is to use a good balun or to detune the outside of the line by proper choice of length. It is well to note that this is not a measurement error, since what the instrument reads is the actual SWR on the line. However, it is an undesirable condition since the line is operating at a higher SWR than it should—and would—if the parallel-type current on the outside of the coax were eliminated.

Spurious Frequencies

Off-frequency components in the RF voltage applied to the bridge may cause considerable error. The principal components of this type are harmonics and low-frequency subharmonics that may be fed through the final stage of the transmitter driving the bridge. The antenna is almost always a fairly selective circuit, and even though the system may be operating with a very low SWR at the desired frequency, it is practically always mismatched at harmonic and sub-harmonic frequencies. If such spurious frequencies are applied to the bridge in appreciable amplitude, the SWR indication will be very much in error. In particular, it may not be possible to obtain a null on the bridge with any set of adjustments of the matching circuit. The only remedy is to filter out the unwanted components by increasing the selectivity of the circuits between the transmitter final amplifier and the bridge.

MEASURING LINE LENGTH

The following material is taken from information in September 1985 *QST* by Charlie Michaels, W7XC (see Bibliography).

There is a popular myth that one may prepare an open quarter-wave line by connecting a loop of wire to one end and trimming the line to resonance (as indicated by a dip meter). This actually yields a line with capacitive reactance equal to the inductive reactance of the loop: A 4-inch wire loop yields a line 82.8° long at 18 MHz; a 2-inch loop yields an 86° line. As the loop size is reduced, line length approaches–but never equals–90°.

To make a quarter-wave open line, parallel connect a coil and capacitor that resonate at the required frequency (see Fig 9A). After adjusting the network to resonance, do not make further network adjustments. Open the connection between the coil and capacitor and series connect the line to the pair. Start with a line somewhat longer than required, and trim it until the circuit again resonates at the desired frequency. For a shorted quarter-wave line or an open half-wave line, connect the line in parallel with the coil and capacitor (see Fig 9B).

REFLECTOMETERS

Low-cost reflectometers that do not have a guaranteed wattmeter calibration are not ordinarily reliable for accurate numerical measurement of standing-wave ratio. They are, however, very useful as aids in the adjustment of matching networks, since the objective in such adjustment is to reduce the reflected voltage or power to zero. Relatively inexpensive devices can be used for this, since only good bridge balance is required, not actual calibration. Bridges of this type are usually "frequency- sensitive"—that is, the meter response becomes greater with increasing frequency, for the same applied voltage. When matching and line monitoring, rather than SWR measurement, is the principal use of the device, this is not a serious handicap.

Various simple reflectometers, useful for matching and monitoring, have been described from time to time in *QST* and in *The ARRL Handbook*. Because most of these are frequency sensitive, it is difficult to calibrate them accurately for power measurement, but their low cost and suitability for use at moderate power levels, combined with the ability to show accurately when a matching circuit has been properly adjusted, make them a worthwhile addition to the amateur station.

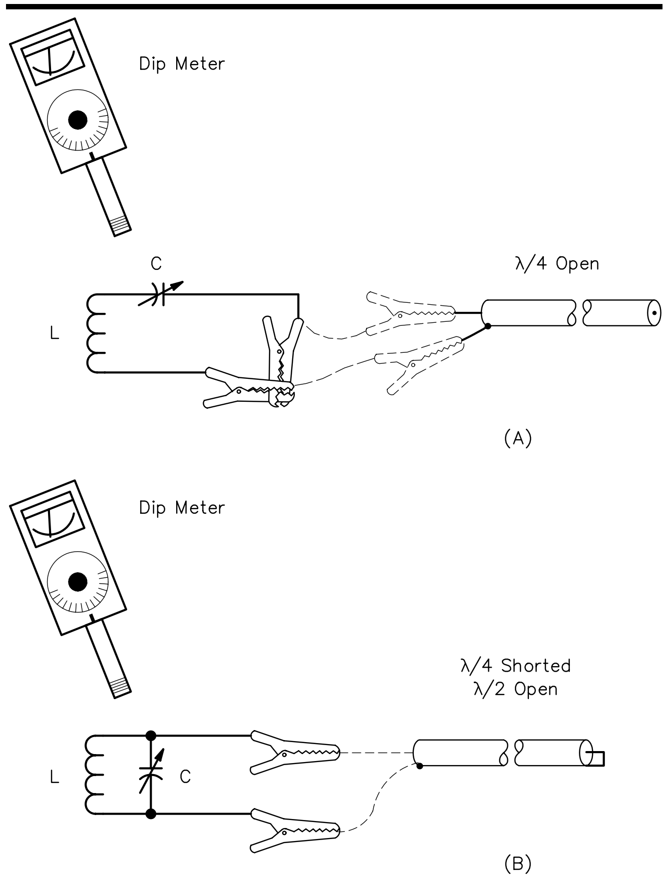


Fig 9—Methods of determining $^1/_4$ and $^1/_2$-λ line lengths. At A, $^1/_4$-λ open-circuited line; at B, $^1/_4$-λ shorted and $^1/_2$-λ open-circuited line.

The Tandem Match—An Accurate Directional Wattmeter

Most SWR meters are not very accurate at low power levels because the detector diodes do not respond to low voltage in a linear fashion. This design uses a compensating circuit to cancel diode nonlinearity. It also provides peak detection for SSB operation and direct SWR readout that does not vary with power level. The following information is condensed from an article by John Grebenkemper, KI6WX, in January 1987 *QST*.

DESIGN PRINCIPLES

Directional wattmeters for Amateur Radio use consist of three basic elements: a directional coupler, a detector and a signal-processing and display circuit. A directional coupler samples forward and reflected-power components on a transmission line. An ideal directional coupler would provide signals proportional to the forward and reflected voltages (independent of frequency), which could then be used to measure forward and reflected power over a wide frequency range. The best contemporary designs work over two decades of frequency.

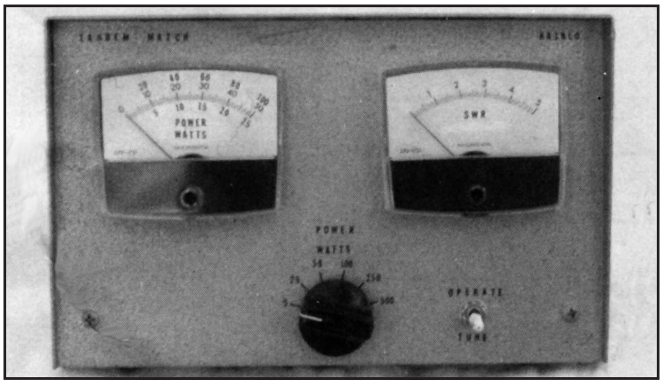


Fig 10—The Tandem Match uses a pair of meters to display net forward power and true SWR simultaneously.

The detector circuit provides a dc output voltage proportional to the ac input voltage. Most directional wattmeters use a single germanium diode as the detector element. A germanium, rather than silicon, diode is used to minimize diode nonlinearity at low power levels. Diode nonlinearity still causes SWR measurement errors

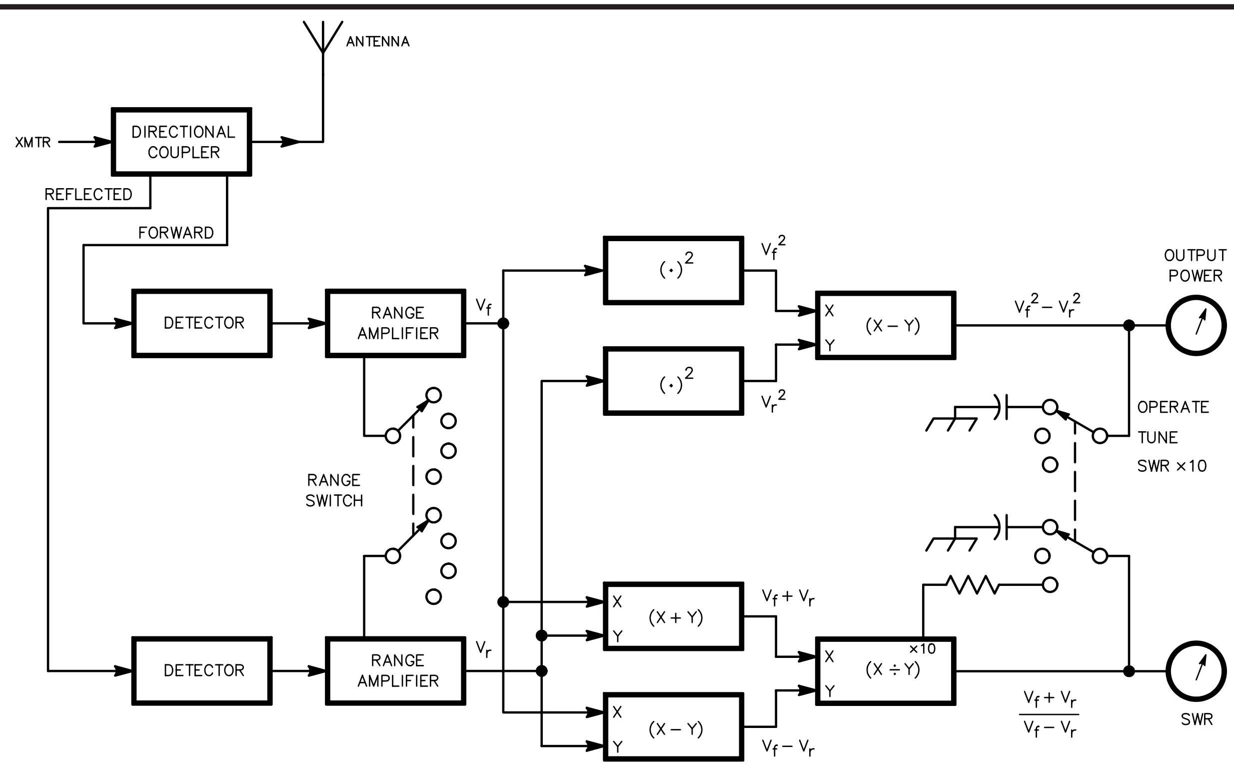


Fig 11—Block diagram of the Tandem Match.

unless it is compensated ahead of the display circuit. Most directional wattmeters do not work well at low power levels because of diode nonlinearity.

The signal-processing and display circuits compute and display the SWR. There are a number of ways to perform this function. Meters that display only the forward and reflected power require the operator to compute the SWR. Many instruments require that the operator adjust the meter to a reference level while measuring forward power, then switch to measure reflected power on a special scale that indicates SWR. Meters that directly compute the SWR using analog signal-processing circuits have been described by Fayman, Perras, Leenerts and Bailey (see the Bibliography at the end of this chapter).

Table 1

Performance Specifications for the Tandem Match

Power range:	1.5 to 1500 W
Frequency range:	1.8 to 54 MHz
Power accuracy:	Better than ± 10% (± 0.4 dB)
SWR accuracy:	Better than ± 5%
Minimum SWR:	Less than 1.05:1
Power display:	Linear, suitable for use with either analog or digital meters
	Analog or digital meters
Calibration:	Requires only an accurate voltmeter

The next section takes a brief look at several popular circuits that accomplish the above functions and compares them to the circuits used in the Tandem Match. The design specifications of the Tandem Match are shown in **Table l**, and a block diagram is shown in **Fig 11**.

CIRCUIT DESCRIPTION

A directional coupler consists of an input port, an output port and a coupled port. The device takes a portion of the power flowing from the input port to the output port and directs it to the coupled port, but *none* of the power flowing from the output port to the input port is directed to the coupled port. There are several terms that define the performance of a directional coupler:

1) *Insertion loss* is the amount of power that is lost as the signal flows from the input port to the output port. Insertion loss should be minimized so the coupler doesn't dissipate a significant amount of the transmitted power.

2) *Coupling factor* is the amount of power (or voltage) that appears at the coupled port relative to the amount of power (or voltage) transferred from the input port to the output port. The "flatness" (with frequency) of the coupling factor determines how accurately the directional wattmeter can determine forward and reflected power over a range of frequencies.

3) *Isolation* is the amount of power (or voltage) that appears at the coupled port relative to the amount of power (or voltage) transferred from the output port to the input port.

4) *Directivity* is the isolation less the coupling factor. Directivity dictates the minimum measurable SWR. A directional coupler with 20 dB of directivity measures a l: l SWR as l .22: l, but one with 30 dB measures a l: l SWR as 1.07:1.

The directional coupler most commonly used in Amateur Radio was first described in 1959 by Bruene in *QST* (see Bibliography). The coupling factor was fairly flat (± l dB), and the directivity was about 20 dB for a Bruene coupler measured from 3 to 30 MHz. Both factors limit the accuracy of the Bruene coupler for measuring low values of power and SWR. It is a simple directional coupler, however, and it works well over a wide frequency range if great precision is not required.

The coupler used in the Tandem Match (see **Fig 12**) consists of a pair of toroidal transformers connected in tandem. The configuration was patented by Carl G. Sontheimer and Raymond E. Fredrick (US Patent no. 3,426,298, issued February 4, 1969). It has been described by Perras, Spaulding (see Bibliography) and others. With coupling factors of 20 dB or greater, this coupler is suitable to sample both forward and reflected power.

The configuration used in the Tandem Match works well over the frequency range of 1.8 to 54 MHz, with a nominal coupling factor of 30 dB. Over this range, insertion loss is less than 0.1 dB. The coupling factor is flat to within ± 0.1 dB from 1.8 to 30 MHz, and increases to only ±0.3 dB at 50 MHz. Directivity exceeds 35 dB from 1.8 to 30 MHz and exceeds 26 dB at 50 MHz.

The low-frequency limit of this directional coupler is determined by the inductance of the transformer secondary windings. The inductive reactance should be greater than 150 Ω (three times the line

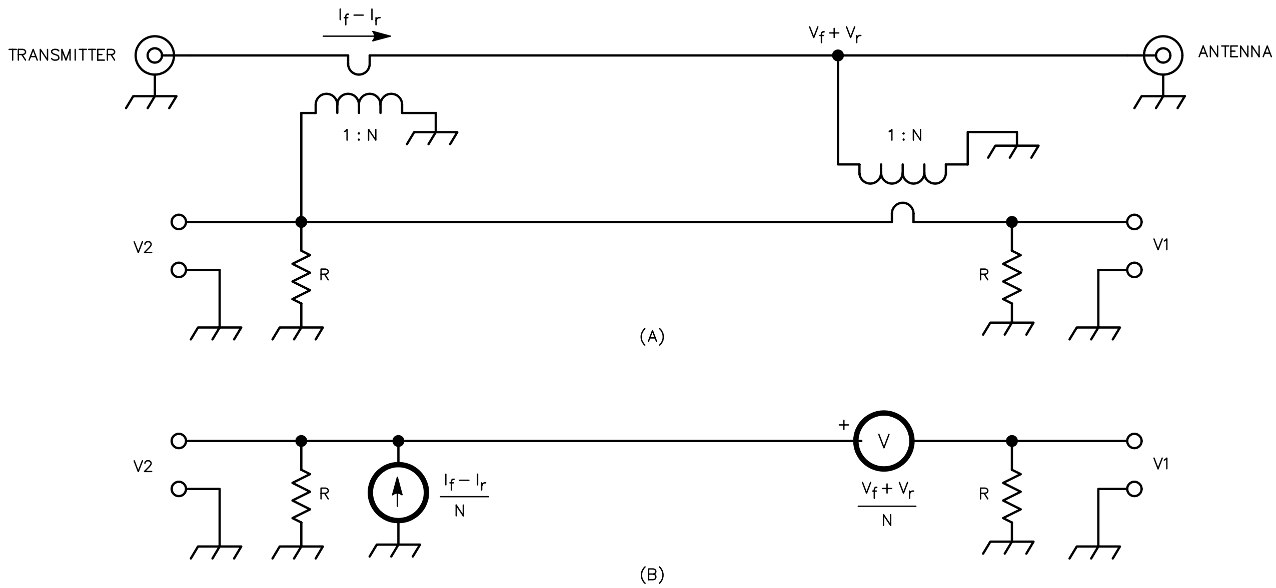


Fig 12—Simplified diagram of the Tandem Match directional coupler. At A, a schematic of the two transformers. At B, an equivalent circuit.

characteristic impedance) to reduce insertion loss. The high-frequency limit of this directional coupler is determined by the length of the transformer windings. When the winding length approaches a significant fraction of a wavelength, coupler performance deteriorates.

The coupler described here may overheat at 1500 W on 160 meters (because of the high circulating current in the secondary of T2). The problem could be corrected by using a larger core or one with greater permeability. A larger core would require longer windings; that option would decrease the high-frequency limit.

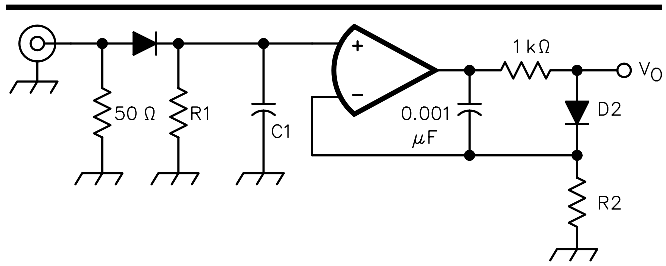


Fig 13—Simplified diagram of the detector circuit used in the Tandem Match. The output voltage, Vo is approximately equal to the input voltage. D1 and D2 must be a matched pair (see text). The op amp should have a low offset voltage (less than 1 mV), a low leakage current (less than 1 nA), and be stable over time and temperature. The resistor and capacitor in the feedback path assure that the op amp will be stable.

Detector Circuits

Most amateur directional wattmeters use a germanium diode detector to minimize the forward voltage drop. Detector voltage drop is still significant, however, and an uncompensated diode detector does not respond to small signals in a linear fashion. Many directional wattmeters compensate for diode nonlinearity by adjusting the meter scale.

The effect of underestimating detected power worsens at low power levels. Under these conditions, the ratio of the forward power to the reflected power is overestimated because the reflected power is always less than the forward power. This results in an instrument that underestimates SWR, particularly as power is reduced. A directional wattmeter can be checked for this effect by measuring SWR at several power levels: the SWR should be independent of power level.

The Tandem Match uses a feedback circuit to compensate for diode nonlinearity. A simplified diagram of the compensated detector is shown in **Fig 13**. When used with the 30-dB directional coupler, the output voltage of this circuit tracks the square root of power over a range from 10 mW to 1.5 kW. The compensated diode detector tracks the peak input voltage down to 30 mV, while an uncompensated germanium-diode detector shows sign)ficant errors at peak inputs of 1 V and less. More information about compensated detectors appears in Grebenkemper's *QEX* article, "Calibrating Diode Detectors" (see Bibliography).

The compensation circuit uses the voltage across a feedback diode, D2, to compensate for the voltage drop across the detector diode, D1. (The diodes must be a matched pair.) The average current through D1 is determined by the detector diode load resistor, R1. The peak current through this diode is several times larger than the average current; therefore, the current through D2 must be several times larger than the average current through D1 to compensate adequately for the peal voltage drop across D1. This is accomplished by making the feedback-diode load resistor, R2, several times smaller than R1. The voltage at the output of the compensated detector approximates the peak RF voltage at the input. For Schottky barrier diodes and a 1 MΩ detector-diode load resistor, a 5:1 ratio of R1 to R2 is nearly optimal.

Signal-Processing and Display Circuits

The signal-processing circuitry calculates and displays transmission-line power and SWR. When measuring forward power, most directional wattmeters display the actual forward power present in the transmission line, which is the sum of forward and reflected power if a match exists at the input end of the line. Transmission-line forward power is very close to the net forward power (the actual power delivered to the line) so long as the SWR is low. As the SWR increases, however, forward power becomes an increasingly poor measure of the power delivered to the load. At an SWR of 3:1, a forward power reading of 100 W implies that only 75 W is delivered to the load (the reflected power is 25 W), assuming the transmission-line loss is zero. The Tandem Match differs from most wattmeters in that *it displays the net forward power, rather than the sum of forward and reflected power.* This is the quantity which must be optimized to result in maximum radiated power (and which concerns the FCC).

The Tandem Match directly computes and displays the transmission-line SWR on a linear scale. As the displayed SWR is not affected by changes in transmitter power, a matching network can be simply adjusted to minimize SWR. Transmatch adjustment requires only a few watts.

The heart of the Tandem Match signal-processing circuit is the analog logarithm and antilogarithm circuitry shown in **Fig 14**. The circuit is based on the fact that collector current in a silicon transistor is proportional to the exponential (antilog) of its base-emitter voltage over a range of collector currents from a few nanoamperes to a few milliamperes when the collector-base voltage is zero (see Gibbons and Horn reference in the Bibliography). Variations of this circuit are used in the squaring circuits to convert voltage to power and in the divider circuit used to compute the SWR. With good op amps, this circuit will work well for input voltages from less than 100 mV to greater than 10 V. (For the Tandem Match, "good" op amps are quad-packaged, low-power-consumption, unity-gain-stable parts with input bias less than 1 nA and offset voltage less than 5 mV. Op amps that consume more power than those shown may require changes to the power supply.)

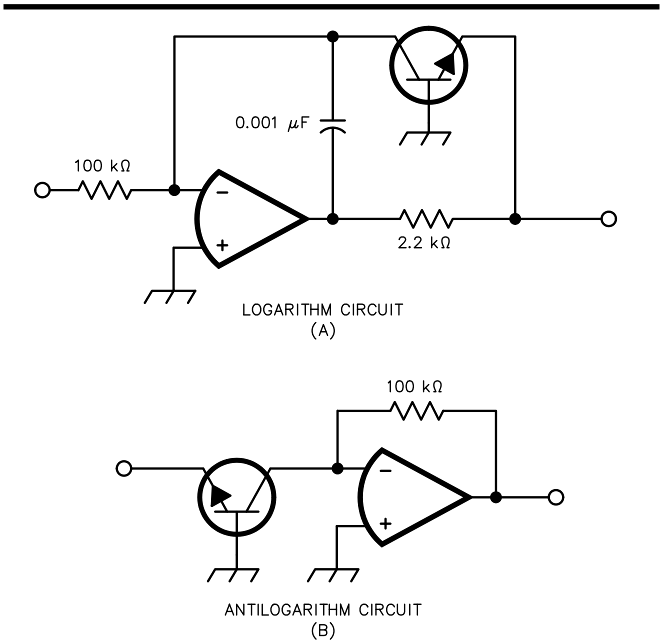


Fig 14—Simplified diagrams of the log circuit at A and the antilog circuit at B.

CONSTRUCTION

The schematic diagram for the Tandem Match is shown in **Fig 15**. The circuit is designed to operate from batteries and draw very little power. Much of the circuitry is of high impedance, so take care to isolate it from RF fields. House it in a metal case. Most problems in the prototype were caused by stray RF in the op-amp circuitry.

Directional Coupler

The directional coupler is constructed in its

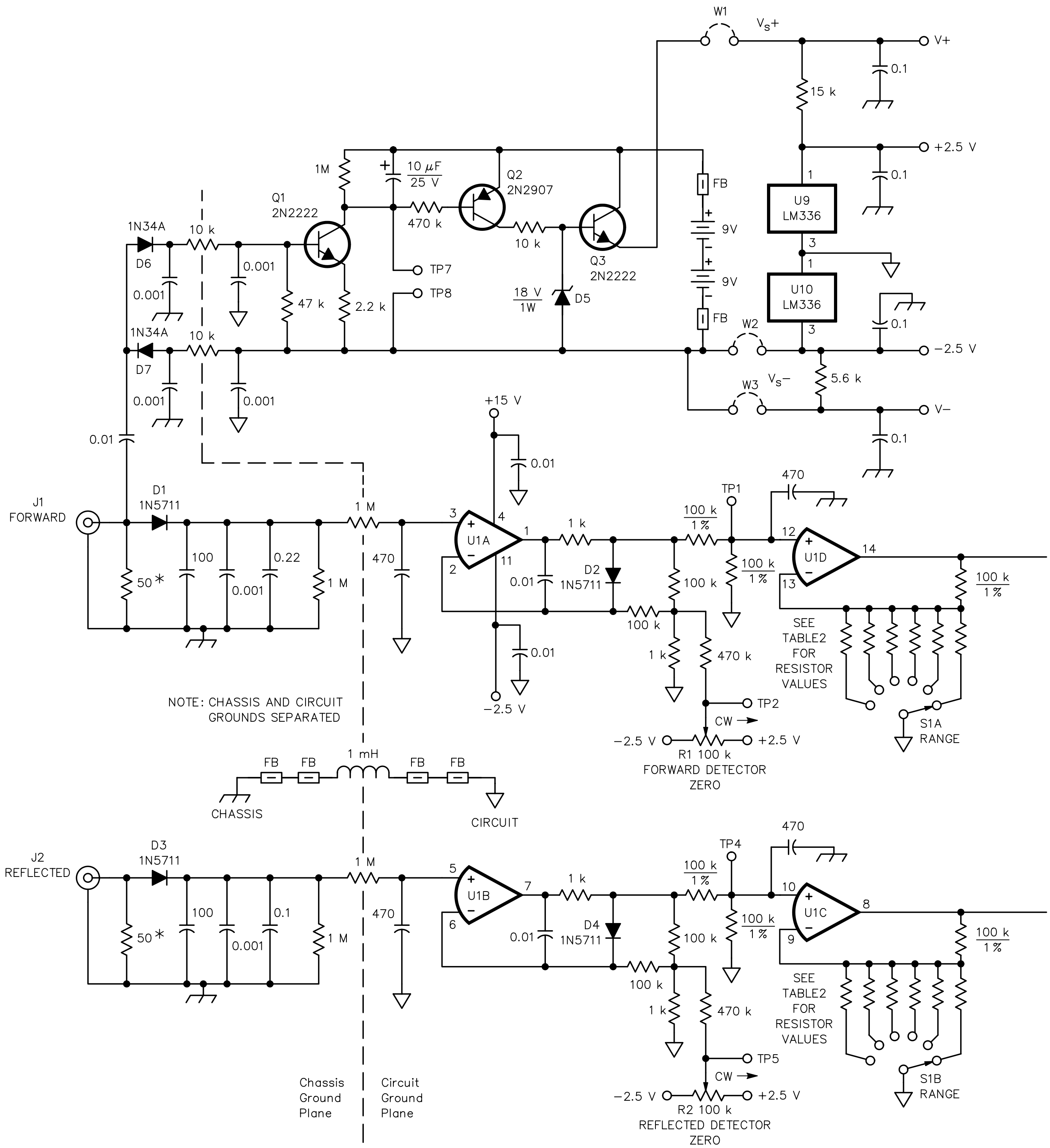
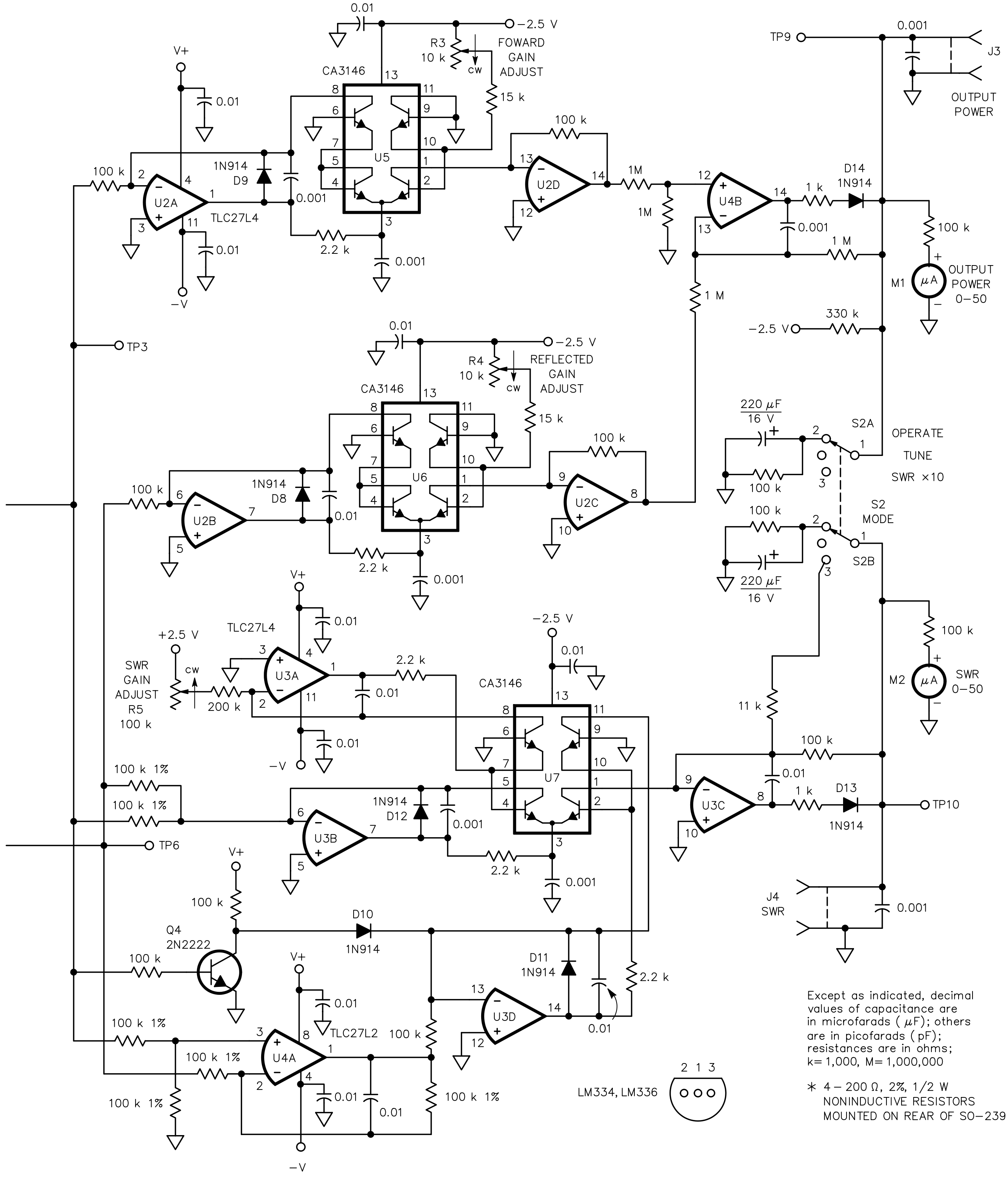


Fig 15—Schematic diagram for the Tandem Match directional wattmeter. Parts identified as RS are from Radio Shack. For other parts sources, see Table 3. See Fig 17 for construction of 50-Ω loads at J1 and J2.

D1, D2—Matched pair 1N5711, or equivalent.
D3, D4—Matched pair 1N5711, or equivalent.
D6, D7—1N34A.
D8-D14—1N914.
FB—Ferrite bead, Amidon FB-73-101 or equiv.
J1, J2—SO-239 connector.
J3, J4—Open-circuit jack.
M1, M2—50 μA panel meter, RS 270-1751.
Q1, Q3, Q4—2N2222 or equiv.
Q2—2N2907 or equiv.
R1, R2, R5—100 kΩ, 10-turn, cermet Trimpot.
R3, R4—10 kΩ, 10-turn, cermet Trimpot.
U1-U3—TLC27L4 or TLC27M4 quad op amp (Texas Instruments).
U4—TLC27L2 or TLC27M2 dual op amp.
U5-U7—CA3146 quad transistor array.
U8—LM334 adjustable current source.
U9-U10—LM336 2.5-V reference diode. See text.


0.01
−2.5 V
R3
10 k
CW
FOWARD
GAIN
ADJUST
CA3146
U5
15 k
V+
0.01
100 k
1N914
D9
U2A
TLC27L4
0.001
2.2 k
0.001
−V
TP3
100 k
U2D
1M
1M
U4B
1 k
D14
1N914
TP9
0.001
J3
OUTPUT
POWER
100 k
0.001
1 M
1 M
M1
OUTPUT
POWER
0–50
330 k
−2.5 V
0.01
−2.5 V
R4
10 k
CW
REFLECTED
GAIN
ADJUST
CA3146
U6
15 k
100 k
U2C
100 k
1N914
D8
U2B
0.01
2.2 k
0.001
220 µF
16 V
S2A
OPERATE
TUNE
SWR ×10
100 k
S2
MODE
S2B
220 µF
16 V
V+
0.01
TLC27L4
U3A
+2.5 V
SWR
GAIN
ADJUST
R5
100 k
200 k
2.2 k
0.01
0.01
−V
−2.5 V
0.01
CA3146
U7
100 k
M2
SWR
0–50
11 k
100 k
0.01
U3C
1 k
D13
1N914
TP10
100 k 1%
100 k 1%
TP6
U3B
1N914
D12
0.001
2.2 k
0.001
V+
100 k
Q4
2N2222
D10
1N914
D11
1N914
2.2 k
J4
SWR
0.001
100 k
V+
0.01
TLC27L2
U4A
U3D
100 k
0.01
100 k 1%
100 k 1%
100 k 1%
0.01
0.01
100 k 1%
−V
2 1 3
LM334, LM336
Except as indicated, decimal values of capacitance are in microfarads (µF); others are in picofarads (pF); resistances are in ohms; k= 1,000, M= 1,000,000
∗ 4 – 200 Ω, 2%, 1/2 W NONINDUCTIVE RESISTORS MOUNTED ON REAR OF SO–239

own small ($2^3/_4 \times 2^3/_4 \times 2^1/_4$-inch) aluminum box (see **Fig 16**). Two pairs of S0-239 connectors are mounted on opposite sides of the box. A piece of PC board is run diagonally across the box to improve coupler directivity. The pieces of RG-8X coaxial cable pass through holes in the PC board. (Note: Some brands of "mini 8" cable have extremely low breakdown voltage ratings and are unsuitable to carry even 100 W when the SWR exceeds 1:1. See the subsequent section, "High-Power Operation," for details of a coupler made with RG-8 cable.)

Begin by constructing T1 and T2, which are identical except for their end connections. Refer to Fig 16. The primary for each transformer is the center conductor of a length of RG-8X coaxial cable. Cut two cable lengths sufficient for mounting as shown in the figure. Strip the cable jacket, braid and dielectric as shown. The cable braid is used as a Faraday shield between the transformer windings, so it is only grounded at one end. *Important—connect the braid only at one end or the directional-coupler circuit will not work properly!* Wind two transformer secondaries, each 31 turns of #24 enameled wire on an Amidon T50-3 or equivalent powdered-iron core. Slip each core over one of the prepared cable pieces (including both the shield and the outer insulation). Mount and connect the transformers as shown in Fig 16, with the wire running through separate holes in the copper-clad PC board.

The directional coupler can be mounted separately from the rest of the circuitry if desired. If so, use two coaxial cables to carry the forward and reflected-power signals from the directional coupler to the detector inputs. Be aware, however that any losses in the cables will affect power readings.

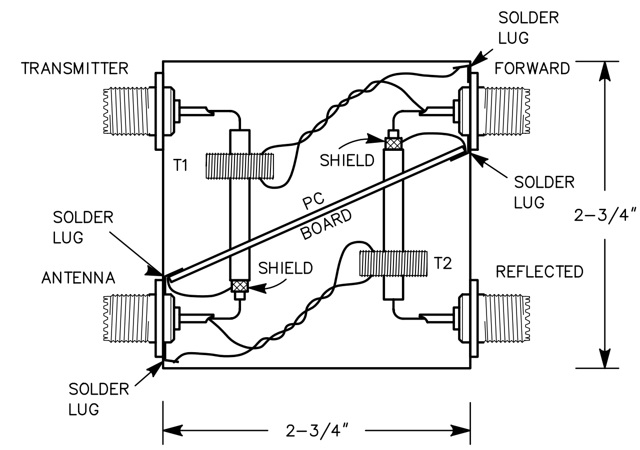


Fig 16—Construction details for the directional coupler.

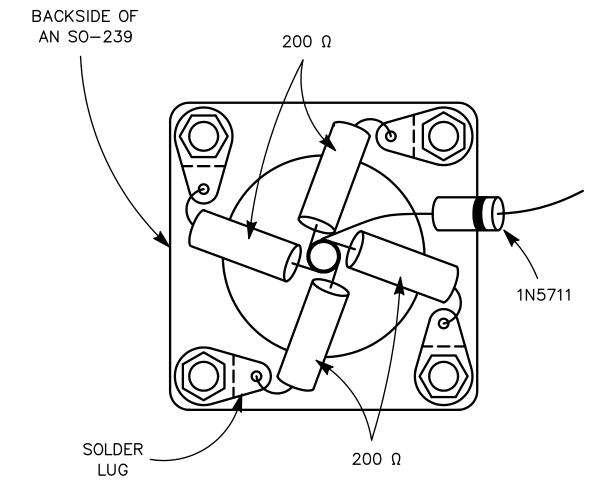


Fig 17—The parallel load resistors mounted on an SO-239 connector. Four 200-Ω, 2%, $^1/_2$-W resistors are mounted in parallel to provide a 50-Ω detector load.

This directional coupler has not been used at power levels in excess of 100 W. For more information about using the Tandem Match at high power levels, see the section, "High-Power Operation."

Detector and Signal-Processing Circuits

The detector and signal-processing circuits were constructed on a perforated, copper-clad circuit board. These circuits use two separate grounds—*it is extremely important that the grounds be isolated as shown in the circuit diagram.* Failure to do so may result in faulty circuit operation. Separate grounds prevent RF currents on the cable braid from affecting the op-amp circuitry.

The directional coupler requires good 50-Ω loads. They are constructed on the back of female UHF chassis connectors where the cables from the directional coupler enter the wattmeter housing. Each load consists of four 200-Ω resistors connected from the center conductor of the UHF connector to the four holes on the mounting flange, as shown in **Fig 17**. The detector diode is then run from the center conductor of the connector to the 100-pF and 1000-pF bypass capacitors, which are mounted next to the connector. The response of this load and detector combination measures flat to beyond 500 MHz.

Schottky-barrier diodes (type lN5711) were used in this design because they were readily available. Any RF-detector diode with a low forward voltage drop (less than 300 mV) and reverse breakdown voltage greater than 30 V could be used. (Germanium diodes could be used in this circuit, but performance will suffer. If germanium diodes are used, reduce the resistance values for the detector-diode and feedback-diode load resistors by a factor of 10.)

The detector diodes must be matched. This can be done with dc, using the circuit shown in **Fig 18**. Use a high-impedance voltmeter (10 MΩ or greater). For this project, diodes are matched when their forward voltage drops are equal (within a few millivolts). Diodes from the same batch will probably be sufficiently matched.

The rest of the circuit layout is not critical, but keep the lead lengths of the 0.001 and 0.01-pF bypass capacitors short. The capacitors provide additional bypassing for the op-amp circuitry.

D6 and D7 form a voltage doubler to detect the presence of a carrier. When the forward power exceeds 1.5 W, Q3 switches on and stays on until about 10 seconds after the carrier drops. (A connection from TP7 to TP9 forces the unit on, even with no carrier present.) The regulated references of +2.5 V and –2.5 V generated by the LM334 and two LM336s are critical. Zener- diode substitutes would significantly degrade performance.

The four op amps in U1 compensate for the nonlinearity of the detector diodes. D1-D2 and D3-D4 are the matched diode pairs discussed above. A RANGE switch selects the meter range. (A six-position switch was used here because it was handy.) The resistor values for the RANGE switch are shown in **Table 2**. Full-scale input power gives an output at U1C or U1D of 7.07 V. The forward and reflected-power detectors are zeroed with R1 and R2.

The forward and reflected-detector voltages are squared by U2, U5 and U6 so that the output voltages are proportional to forward and reflected power. The gain constants are adjusted using R3 and R4 so that an input of 7.07 V to the squaring circuit gives an output of 5 V. The difference between these two voltages is used by U4B to yield an output that is proportional to the power delivered to the transmission line. This voltage is peak detected (by an RC circuit connected to the OPERATE position of the MODE switch) to hold and indicate the maximum power during CW or SSB transmissions.

SWR is computed from the forward and reflected voltages by U3, U4 and U7. When no carrier is present, Q4 forces the SWR reading to be zero (that is, when the forward power is less than 2% of the full-scale setting of the RANGE switch). The SWR computation circuit gain is adjusted by R5. The output is peak detected in the OPERATE mode to steady the SWR reading during CW or SSB transmissions.

Transistor arrays (U5, U6 and U7) are used for the log and antilog circuits to guarantee that the transistors will be well matched. Discrete transistors may be used, but accuracy may suffer.

A three-position toggle switch selects the three operating modes. In the OPERATE mode, the power and SWR outputs are peak detected and held for a few seconds to allow meter reading during actual transmissions. In the TUNE mode, the meters display instantaneous output power and SWR.

A digital voltmeter is used to obtain more pre-

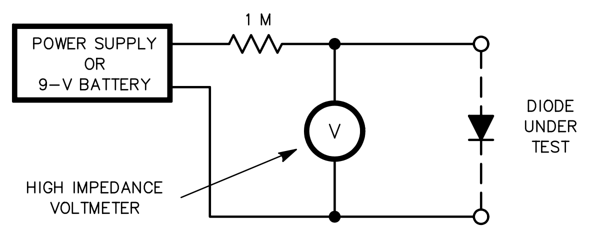


Fig 18—Diode matching test setup.

Table 2
Range-Switch Resistor Values

Full-Scale Power Level (W)	*Range Resistor (1% Precision) (kΩ)*
1	2.32
2	3.24
3	4.02
5	5.23
10	7.68
15	9.53
20	11.0
25	12.7
30	14.0
50	18.7
100	28.7
150	37.4
200	46.4
250	54 9
300	63.4
500	100.0
1000	237.0
1500	649.0
2000	Open

cise readings than are possible with analog meters. The output power range is 0 to 5 V (0 V = 0 W and 5 V = full scale). SWR output varies from 1 V (SWR = 1:1) to 5 V (SWR = 5:1). Voltages above 5 V are unreliable because of voltage limiting in some of the op amp circuits.

Calibration

The directional wattmeter can be calibrated with an accurate voltmeter. All calibration is done with dc voltages. The directional-coupler and detector circuits are inherently accurate if correctly built. To calibrate the wattmeter, use the following procedure:

1) Set the MODE switch to TUNE and the RANGE switch to 100 W or less.
2) Jumper TP7 to TP8. This turns the unit on.
3) Jumper TP1 to TP2. Adjust R1 for 0 V at TP3.
4) Jumper TP4 to TP5. Adjust R2 for 0 V at TP6.
5) Adjust R1 for 7.07 V at TP3.
6) Adjust R3 for 5.00 V at TP9, or a full-scale reading on Ml.
7) Adjust R2 for 7.07 V at TP6.
8) Adjust R4 for 0 V at TP9, or a zero reading on M1.
9) Adjust R2 for 4.71 V at TP6.
10) Adjust R5 for 5.00 V at TP10, or a full-scale reading on M2.
11) Set the RANGE switch to its most sensitive scale.
12) Remove the jumpers from TP1 to TP2 and TP4 to TP5.
13) Adjust R1 for 0 V at TP3.
14) Adjust R2 for 0 V at TP6.
15) Remove the jumper from TP7 to TP8.

This completes the calibration procedure. This procedure has been found to equal calibration with expensive laboratory equipment. The directional wattmeter should now be ready for use.

ACCURACY

Performance of the Tandem Match has been compared to other well-known directional couplers and laboratory test equipment, and it equals any amateur directional wattmeter I have tested. Power measurement accuracy compares well to a Hewlett-Packard HP-436A power meter. The HP meter has a specified measurement error of less than ± 0.05 dB. The Tandem Match tracked the HP436A within + 0.5 dB from 10 mW to 100 W, and within ± 0.1 dB from 1 W to 100 W. The unit was not tested above 100 W because a transmitter with a higher power rating was not available.

SWR performance was equally good when compared to the SWR calculated from measurements made with the HP436A and a calibrated directional coupler. The Tandem Match tracked the calcuflated SWR within ± 5% for SWR values from 1:1 to 5:1. SWR measurements were made at 8 W and 100 W.

OPERATION

Connect the Tandem Match in the 50-Ω line between the transmitter and the antenna matching network (or antenna if no matching network is used). Set the RANGE switch to a range greater than the transmitter output rating and the MODE switch to TUNE. When the transmitter is keyed, the Tandem Match automatically switches on and indicates both power delivered to the antenna and SWR on the transmission line. When no carrier is present, the OUTPUT POWER and SWR meters indicate zero.

The OPERATE mode includes RC circuitry to momentarily hold the peak-power and SWR readings during CW or SSB transmissions. The peak detectors are not ideal, so there could be about 10% variation from the actual power peaks and the SWR reading. The SWR × 10 mode increases the maximum readable SWR to 50:1. This range should be sufficient to cover any SWR vaIue that occurs in amateur use. (A 50-foot open stub of RG-8 yields a measured SWR of only 43:1, or less, at 2.4 MHz because of cable loss. Higher frequencies and longer cables exhibit a lesser maximum SWR.)

lt is easy to use the Tandem Match to adjust an antenna matching network: Adjust the transmitter for minimum output power (at least 1.5 W). With the carrier on and the MODE switch set to TUNE or SWR × 10, adjust the matching network for minimum SWR. Once the minimum SWR is obtained, set the transmitter to the proper operating mode and output power. Place the Tandem Match in the OPERATE mode.

DESIGN VARIATIONS

There are several ways in which this design could be enhanced. The most important is to add UHF capability. This would require a new directional-coupler design for the band of interest. (The existing detector circuit should work to at least 500 MHz.)

Those who desire a low-power directional wattmeter can build a directional coupler with a 20-dB coupling factor by decreasing the transformer turns ratio to 10:1. That version should be capable of measuring output power from 1 mW to about 150 W (and it should switch on at about 150 mW).

This change should also increase the maximum operating frequency to about 150 MHz (by virtue of the shorter transformer windings). If you desire 1.8-MHz operation, it may be necessary to change the toroidal core material for sufficient reactance (low insertion loss).

The Tandem Match circuit can accommodate coaxial cable with a characteristic impedance other than 50 Ω**.** The detector terminating resistors, transformer secondaries and range resistors must change to match the new design impedance.

The detector circuitry can be used (without the directional coupler) to measure low-level RF power in 50-Ω circuits. RF is fed directly to the forward detector (J1, Fig 15), and power is read from the output power meter. The detector is quite linear from 10 μW to 1.5 W.

HIGH-POWER OPERATION

This material was condensed from information by Frank Van Zant, KL71BA, in July 1989 *QST.* In April 1988, Zack Lau, W1VT, described a directional-coupler circuit (based on the same principle as

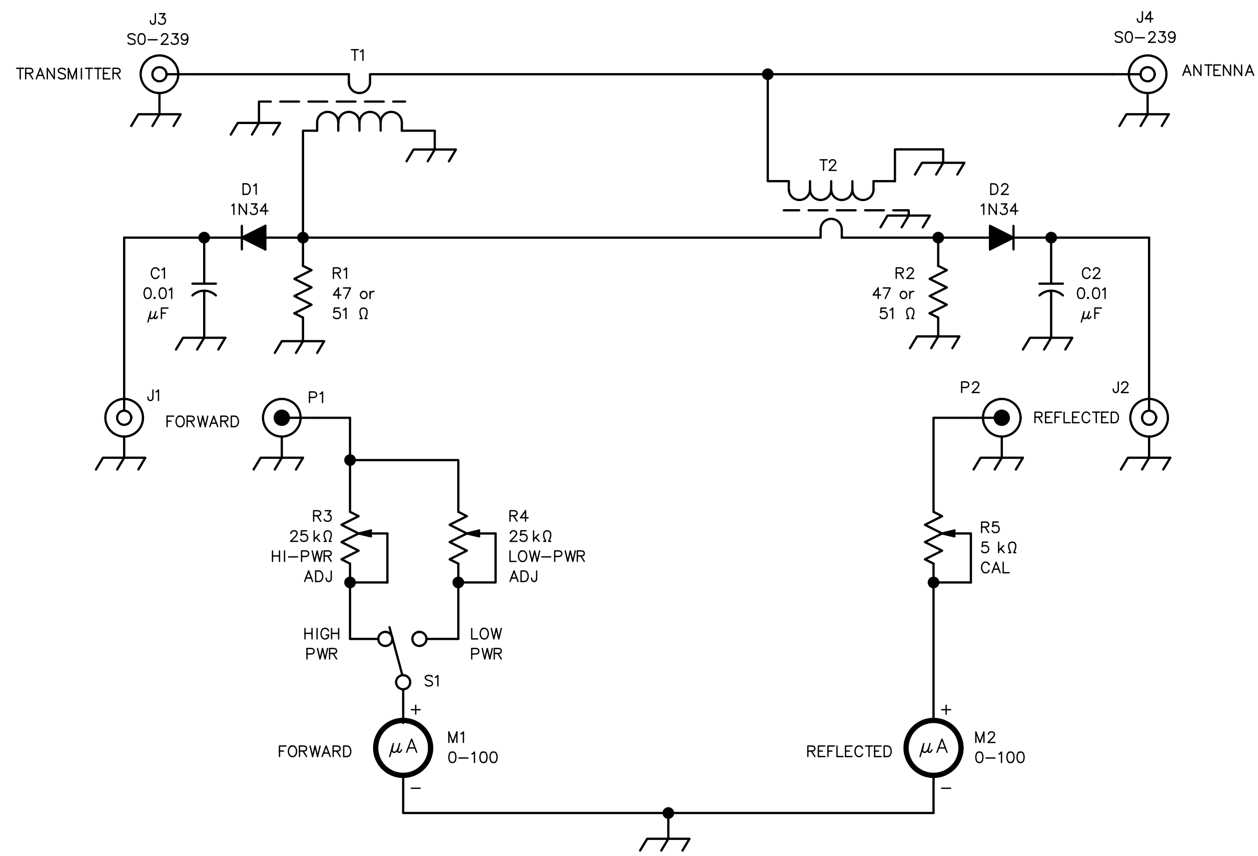


Fig 19—Schematic diagram of the high-power directional coupler. D1 and D2 are germanium diodes (1N34 or equiv). R1 and R2 are 47 or 51-Ω, 1/2-W resistors. C1 and C2 have 500-V ratings. The secondary windings of T1 and T2 each consist of 40 turns of #26 to #30 enameled wire on T-68-2 powdered-iron toroid cores. If the coupler is built into an existing antenna tuner, the primary of T1 can be part of the tuner coaxial output line. The remotely located meters (M1 and M2) are connected to the coupler box at J1 and J2 via P1 and P2.

Grebenkemper's circuit) for a QRP transceiver (see the Bibliography at the end of this chapter). The main advantage of Lau's circuit is very low parts count.

Grebenkemper used complex log-antilog amplifiers to provide good measurement accuracy. This application gets away from complex circuitry, but retains reasonable measurement accuracy over the 1 to 1500-W range. It also forfeits the SWR-computation feature.

Lau's coupler uses ferrite toroids. It works well at low power levels, but the ferrite toroids heat excessively with high power, causing erratic meter readings and the potential for burned parts.

The Revised Design

Powdered-iron loroids are used for the transformers in this version of Lau's basic circuit. The number of turns on the secondaries was increased to compensate for the lower permeability of powdered iron.

Two meters display reflected and forward power (see **Fig 19**). The germanium detector diodes (D1 and D2—1N34) provide fairly accurate meter readings, particularly if the meter is calibrated (using R3, R4 and R5) to place the normal transmitter output at mid scale. If the winding sense of the transformers is reversed, the meters are transposed (the forward-power meter becomes the reflected-power meter, and vice versa)

Construction

Fig 20 shows the physical layout of the coupler. The pickup unit is mounted in a $3^1/_2 \times 3^1/_2 \times$ 4-inch box. The meters, PC-mount potentiometers and HIGH/LOW power switch are mounted in a separate box or a compartment in an antenna tuner. Parts for this project are available from the suppliers listed in **Table 3**.

The primary windings of Tl and T2 are constructed much as Grebenkemper described, but use RG-8 with its jacket removed so that the core and secondary winding may fit over the cable. The braid is wrapped with fiberglass tape to insulate it from the secondary winding. An excellent alternative to fiberglass tape—with even higher RF voltage-breakdown characteristics—is ordinary plumber's Teflon pipe tape, available at most hardware stores.

The transformer secondaries are wound on T-68-2 powdered-iron toroid cores. They are 40 turns of #26 to #30 enameled wire spread evenly around each core. By using #26 to #30 wire on the cores, the cores slip over the tape-wrapped RG-8

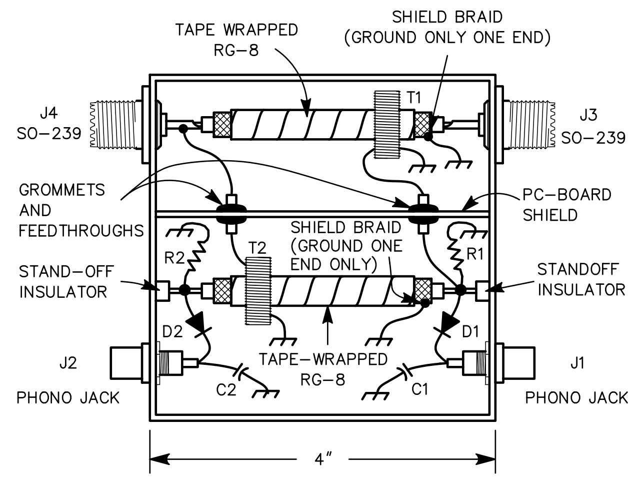


Fig 20—Directional-coupler construction details. Grommets or feedthrough insulators can be used to route the secondary winding of T1 and T2 through the PC board shield. A $3^1/_2 \times 3^1/_2 \times$ 4-inch box serves as the enclosure.

Table 3
Parts Sources
(Also see Chapter 21)

Components	*Source*
TLC-series and CA3146 ICs	Newark Electronics 4801 N Ravenswood St Chicago, IL 60640 312-784-5100
LM334, LM336, 1% resistors, trimmer potentiometers	Digi-Key Corporation 701 Brooks Ave S PO Box 677 Thief River Falls, MN 56701 800-344-4539
Toroid cores, Fiberglass tape	Amidon Associates PO Box 956 Torrance, CA 90508 213-763-5770
Meters	Fair Radio Sales PO Box 1105 Lima, OH 45802 419-227-6573
Toroid cores	Palomar Engineers PO Box 455 Escondido, CA 92033
Misc parts, toroid cores	Radiokit PO Box 973 Pelham NH 03076
0-150/1500-W-scale meters, A&M model no. 255-138, 1N5711 diodes	Surplus Sales of Nebraska 1315 Jones St Omaha, NE 68102

lines. With #26 wire on the toroids, a single layer of tape (slightly more with Teflon tape) over the braid provides an extremely snug fit for the core. Use care when fitting the cores onto the RG-8 assemblies. After the toroids are mounted on the RG-8 sections, coat the assembly with General Cement Corp Polystyrene Q Dope, or use a spot or two of RTV sealant to hold the windings in place and fix the transformers on the RG-8 primary windings.

Mount a PC-board shield in the center of the box, between T1 and T2, to minimize coupling between the transformers. Suspend T1 between the S0-239 connectors and T2 between two standoff insulators. The detector circuits (C1, C2, D1, D2, R1 and R2) are mounted inside the coupler box as shown.

Calibration, Tune Up and Operation

The coupler has excellent directivity. Calibrate the meters for various power levels with an RF ammeter and a 50-Ω dummy load. Calculate I^2R for each power level, and mark the meter faces accordingly. Use R3, R4 and R5 to adjust the meter readings within the ranges. Diode nonlinearities are thus taken into account, and Grebenkemper's signal-processing circuits are not needed for relatively accurate power readings.

Start the tune-up process using about 10 W, adjust the antenna tuner for minimum reflected power, and increase power while adjusting the tuner to minimize reflected power.

This circuit has been built into several antenna tuners with good success. The instrument works well at 1.5-kW output on 1.8 MHz. It also works well from 3.5 to 30 MHz with 1.2 and 1.5-kW output. The antenna is easily tuned for a 1:1 SWR using the null indication provided.

Amplifier settings for a matched antenna, as indicated with the wattmeter, closely agreed with those for a 50-Ω dummy load. Checks with a Palomar noise bridge and a Heath Antenna Scope also verified these findings. This circuit should handle more than 1.5 kW, as long as the SWR on the feed line through the wattmeter is kept at or near 1:1. (On one occasion high power was applied while the antenna tuner was not coupled to a load. Naturally the SWR was extremely high, and the output transformer secondary winding opened like a fuse. This resulted from the excessively high voltage across the secondary. The damage was easily and quickly repaired.)

An Inexpensive VHF Directional Coupler

Precision in-line metering devices capable of reading forward and reflected power over a wide range of frequencies are very useful in amateur VHF and UHF work, but their rather high cost puts them out of the reach of many VHF enthusiasts. The device shown in **Figs 21** through **24** is an inexpensive adaptation of their basic principles. It can be made for the cost of a meter, a few small parts, and bits of copper pipe and fittings that can be found in the plumbing stocks at many hardware stores. This project was originally described by Thomas McMullen, W1SL, in April 1972 *QST*.

Construction

The sampler consists of a short section of handmade coaxial line, in this instance, of 52 Ω impedance, with a reversible probe coupled to it. A small pickup loop built into the probe is terminated with a resistor at one end and a diode at the other. The resistor matches the impedance of the loop, not the impedance of the line section. Energy picked up

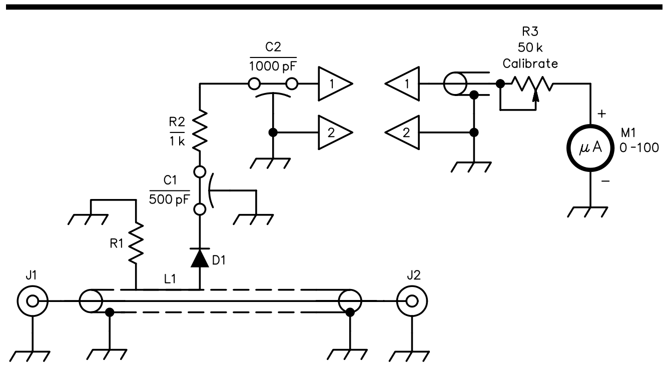


Fig 21—Circuit diagram for the line sampler.

C1—500-pF feedthrough capacitor, solder-in type.
C2—1000-pF feedthrough capacitor, threaded type.
D1—Germanium diode 1N34, 1N60, 1N270, 1N295, or similar.
J1, J2—Coaxial connector, type N (UG-58A).
L1—Pickup loop, copper strap 1-inch long × 3/16-inch wide. Bend into "C" shape with flat portion 5/8-inch long.
M1—0-100 μA meter.
R1—Composition resistor, 82 to 100 Ω. See text.
R3—50-kΩ composition control, linear taper.

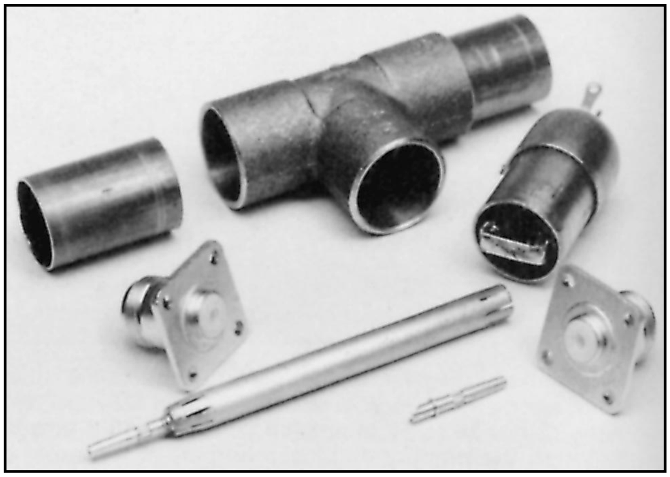


Fig 22—Major components of the line sampler. The brass T and two end sections are at the upper left in this picture. A completed probe assembly is at the right. The N connectors have their center pins removed. The pins are shown with one inserted in the left end of the inner conductor and the other lying in the right foreground.

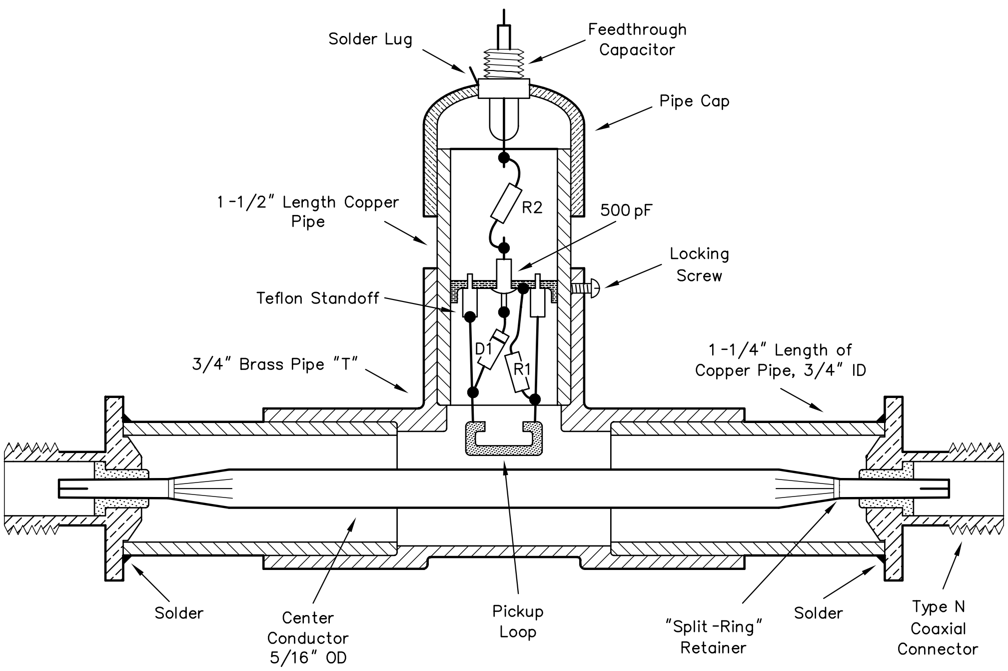


Fig 23—Cross-section view of the line sampler. The pickup loop is supported by two Teflon standoff insulators. The probe body is secured in place with one or more locking screws through holes in the brass T.

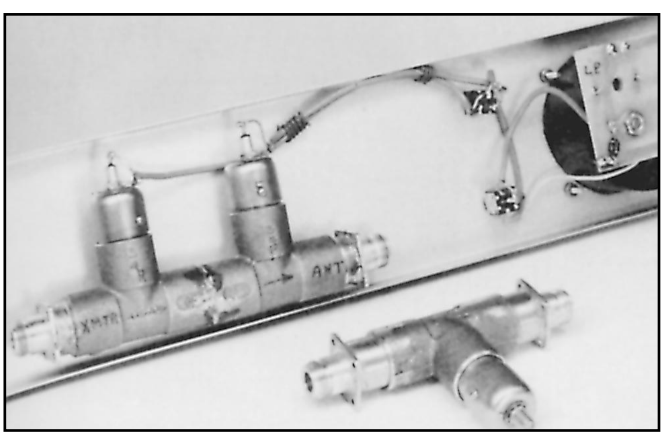


Fig 24—Two versions of the line sampler. The single unit described in detail here is in the foreground. Two sections in a single assembly provide for monitoring forward and reflected power without probe reversal.

by the loop is rectified by the diode, and the resultant current is fed to a meter equipped with a calibration control.

The principal metal parts of the device are a brass plumbing T, a pipe cap, short pieces of $^{3}/_{4}$-inch ID and $^{5}/_{16}$-inch OD copper pipe, and two coaxial fittings. Other available tubing combinations for 52-Ω line may be usable. The ratio of outer conductor ID to inner conductor OD should be 2.4/1. For a sampler to be used with other impedances of transmission line, see Chapter 24 for suitable ratios of conductor sizes. The photographs and Fig 23 show construction details.

Soldering of the large parts can be done with a 300-W iron or a small torch. A neat job can be done if the inside of the T and the outside of the pipe are tinned before assembling. When the pieces are reheated and pushed together, a good mechanical and electrical bond will result. If a torch is used, go easy with the heat, as an over-heated and discolored fitting will not accept solder well.

Coaxial connectors with Teflon or other heat-resistant insulation are recommended. Type N, with split-ring retainers for the center conductors, are preferred. Pry the split-ring washers out with a knife point or small screwdriver. Don't lose them, as they'll be needed in the final assembly.

The inner conductor is prepared by making eight radial cuts in one end, using a coping saw with a fine-toothed blade, to a depth of $^{1}/_{2}$ inch. The fingers so made are then bent together, forming a tapered end, as shown in Figs 22 and 23. Solder the center pin of a coaxial fitting into this, again being careful not to overheat the work.

In preparation for soldering the body of the coax connector to the copper pipe, it is convenient to use a similar fitting clamped into a vise as a holding fixture. Rest the T assembly on top, held in place by its own weight. Use the partially prepared center conductor to assure that the coax connector is concentric with the outer conductor. After being sure that the ends of the pipe are cut exactly perpendicular to the axis, apply heat to the coax fitting, using just enough so a smooth fillet of solder can be formed where the flange and pipe meet.

Before completing the center conductor, check its length. It should clear the inner surface of the connector by the thickness of the split ring on the center pin. File to length; if necessary, slot as with the other end, and solder the center pin in place. The fitting can now be soldered onto the pipe, to complete the 52-Ω line section.

The probe assembly is made from a $1^{1}/_{2}$ inch length of the copper pipe, with a pipe cap on the top to support the upper feedthrough capacitor, C2. The coupling loop is mounted by means of small Teflon standoffs on a copper disc, cut to fit inside the pipe. The disc has four small tabs around the edge for soldering inside the pipe. The diode, D1, is connected between one end of the loop and a 500-pF feedthrough capacitor, C1, soldered into the disc. The terminating resistor, R1, is connected between the other end of the loop and ground, as directly as possible.

When the disc assembly is completed, insert it into the pipe, apply heat to the outside, and solder the tabs in place by melting solder into the assembly at the tabs. The position of the loop with respect to the end of the pipe will determine the sensitivity of a given probe. For power levels up to 200 W the loop should extend beyond the face of the pipe about $^{5}/_{32}$ inch. For use at higher power levels the loop should protrude only $^{3}/_{32}$ inch. For operation with very low power levels the best probe position can be determined by experiment.

The decoupling resistor, R2, and feedthrough capacitor, C2, can be connected, and the pipe cap put in place. The threaded portion of the capacitor extends through the cap. Put a solder lug over it before tightening its nut in place. Fasten the cap with two small screws that go into threaded holes in the pipe.

Calibration

The sampler is very useful for many jobs even if it is not accurately calibrated, although it is desirable to calibrate it against a wattmeter of known accuracy. A good 52-Ω dummy load is required.

The first step is to adjust the inductance of the loop, or the value of the terminating resistor, for lowest reflected power reading. The loop is the easier to change. Filing it to reduce its width will increase its impedance. Increasing the cross section of the loop will lower the impedance, and this can be done by coating it with solder. When the reflected power reading is reduced as far as possible, reverse the probe and calibrate for forward power by increasing the transmitter power output in steps and making a graph of the meter readings obtained. Use the calibration control, R3, to set the maximum reading.

Variations

Rather than to use one sampler for monitoring both forward and reflected power by repeatedly reversing the probe, it is better to make two assemblies by mounting two T fittings end-to- end, using one for forward and one for reflected power. The meter can be switched between the probes, or two meters can be used.

The sampler described was calibrated at 146 MHz, as it was intended for repeater use. On higher bands the meter reading will be higher for a given power level, and it will be lower for lower frequency bands. Calibration for two or three adjacent bands can be achieved by making the probe depth adjustable, with stops or marks to aid in resetting for a given band. Of course more probes can be made, with each probe calibrated for a given band, as is done in some of the commercially available units.

Other sizes of pipe and fittings can be used by making use of information given in Chapter 24 to select conductor sizes required for the desired impedances. (Since it is occasionally possible to pick up good bargains in 75-Ω line, a sampler for this impedance might be desirable.)

Type N fittings were used because of their constant impedance and their ease of assembly. Most have the split-ring retainer, which is simple to use in this application. Some have a crimping method, as do apparently all BNC connectors. If a fitting must be+″sed and cannot be taken apart, drill a hole large enough to clear a soldering-iron tip in the copper-pipe outer conductor. A hole of up to $^3/_8$-inch diameter will have very little effect on the operation of the sampler.

A Calorimeter For VHF And UHF Power Measurements

A quart of water in a Styrofoam ice bucket, a roll of small coaxial cable and a thermometer are all the necessary ingredients for an accurate RF wattmeter. Its calibration is independent of frequency. The wattmeter works on the calorimeter principle: A given amount of RF energy is equivalent to an amount of heat, which can be determined by measuring the temperature rise of a known quantity of thermally insulated material. This principle is used in many of the more accurate high-power wattmeters. This procedure was developed by James Bowen, WA4ZRP, and was first described in December 1975 *QST*.

The roll of coaxial cable serves as a dummy load to convert the RF power into heat. RG-174 cable was chosen for use as the dummy load in this calorimeter because of its high loss factor, small size, and low cost. It is a standard 52-Ω cable of approximately 0.11 inch diameter. A prepackaged roll marked as 60 feet long, but measured to be 68 feet, was purchased at a local electronics store. A plot of measured RG-174 loss factor as a function of frequency is shown in **Fig 25**.

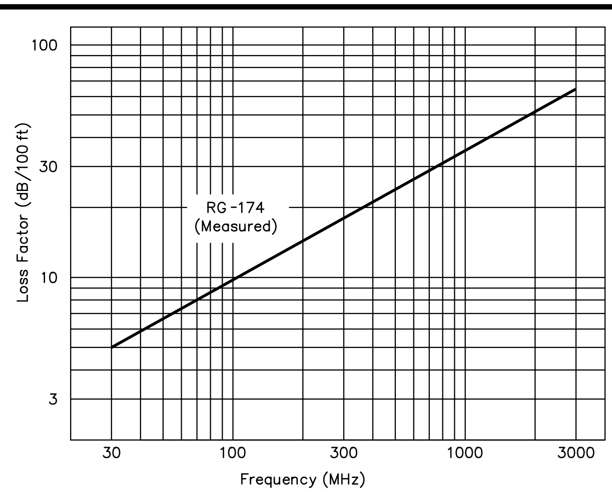


Fig 25—Loss factor of RG-174 coax used in the calorimeter.

In use, the end of the cable not connected to the transmitter is left open circuited. Thus, at 50 MHz, the reflected wave returning to the transmitter (after making a round trip of 136 feet through the cable) is 6.7 dB × 1.36 = 9.11 dB below the forward wave. A reflected wave 9.11 dB down represents an SWR to the transmitter of 2.08:1. While this value seems larger than would be desired, keep in mind that most 50-MHz transmitters can be tuned to match into an SWR of this magnitude efficiently. To assure accurate results, merely tune the transmitter for maximum power into the load before making the measurement. At higher frequencies the cable loss increases so the

SWR goes down. **Table 4** presents the calculated input SWR values at several frequencies for 68 feet of RG-174. At 1000 MHz and above, the SWR caused by the cable connector will undoubtedly exceed the very low cable SWR listed for these frequencies.

In operation, the cable is submerged in a quart of water and dissipated heat energy flows from the cable into the water, raising the water temperature. See **Fig 26**. The calibration of the wattmeter is based on the physical fact that one calorie of heat energy will raise one gram of liquid water 1° C. Since one quart of water contains 946.3 grams, the transmitter must deliver 946.3 calories of heat energy to the water to raise its temperature 1°C. One calorie of energy is equivalent to 4.186 joules and a joule is equal to 1 W for 1 second. Thus, the heat capacitance of 1 quart of water expressed in joules is 946.3 × 4.186 = 3961 joules/°C.

The heat capacitance of the cable is small with respect to that of the water, but nevertheless its effect should be included for best accuracy. The heat capacitance of the cable was determined in the manner described below.

The 68-foot roll of RG-174 cable was raised to a uniform temperature of 100°C by immersing it in a pan of boiling water for several minutes. A quart of tap water was poured into the Styrofoam ice bucket and its temperature was measured at 28.7°C. The cable was then transferred quickly from the boiling water to the water in the ice bucket. After the water temperature in the ice bucket had ceased to rise, it measured 33.0°C. Since the total heat gained by the quart of water was equal to the total heat lost from the cable, we can write the following equation:

$$(\Delta T_{WATER})(C_{WATER}) = -(\Delta T_{CABLE})(C_{CABLE})$$

where

ΔT_{WATER} = the change in water temperature
C_{WATER} = the water heat capacitance
ΔT_{CABLE} = the change in cable temperature
C_{CABLE} = the cable heat capacitance

Substituting and solving:

$$(33.0 - 28.7)(3961) = -(33.0 - 100)(C_{CABLE})$$

$$\frac{(4.3)(3961)}{67} = C_{CABLE}$$

$$254 \text{ joules/°} = C_{CABLE}$$

Thus, the total heat capacitance of the water and cable in the calorimeter is 3961 + 254 = 4215 joules/°C. Since 1°F = 5/9°C, the total heat capacitance can also be expressed as 4215 × 5/9 = 2342 joules/°F.

Table 4
Calculated Input SWR for 68 Feet of Unterminated RG-174 Cable

Freq (MHz)	*SWR*
50	2.08
144	1.35
220	1.20
432	1.06
1296	1.003
2304	1.0003

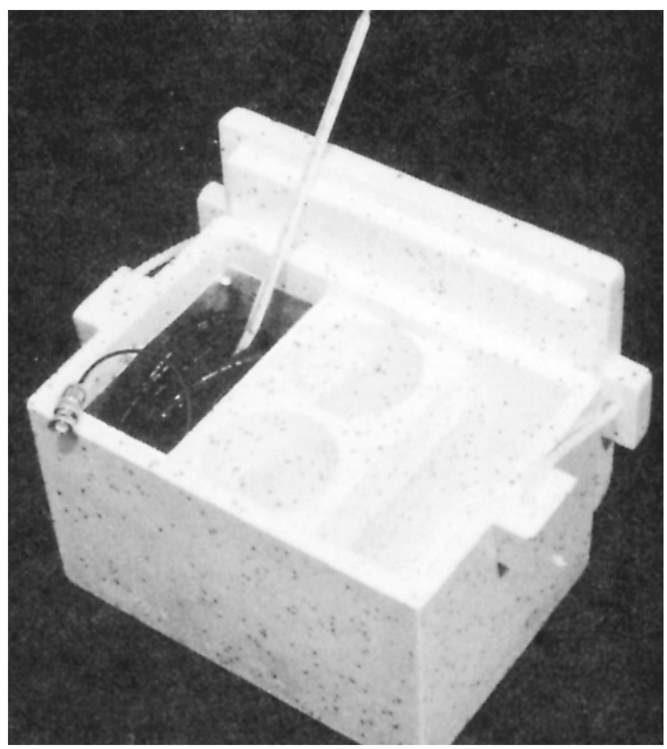


Fig 26—The calorimeter ready for use. The roll of coaxial cable is immersed in one quart of water in the left-hand compartment of the Styrofoam container. Also shown is the thermometer, which doubles as a stirring rod.

Materials and Construction

The quart of water and cable must be thermally insulated to assure that no heat is gained from or lost to the surroundings. A Styrofoam container is ideal for this purpose since Styrofoam has a very low thermal conductivity and a very low thermal capacitance. A local variety store was the source of a small Styrofoam cold chest with compartments for carrying sandwiches and drink cans. The rectangular compartment for sandwiches was found to be just the right size for holding the quart of water and coax.

The thermometer can be either a Celsius or Fahrenheit type, but try to choose one which has divisions for each degree spaced wide enough so that the temperature can be estimated readily to one-tenth degree. Photographic supply stores carry darkroom thermometers, which are ideal for this purpose. In general, glass bulb thermometers are more accurate than mechanical dial-pointer types.

The RF connector on the end of the cable should be a constant-impedance type. A BNC type connector especially designed for use on 0.11-inch diameter cable was located through surplus channels. If you cannot locate one of these, wrap plastic electrical tape around the cable near its end until the diameter of the tape wrap is the same as that of RG-58. Then connect a standard BNC connector for RG-58 in the normal fashion.

Carefully seal the opposite open end of the cable with plastic tape or silicone caulking compound so no water can leak into the cable at this point.

Procedure for Use

Pour 1 quart of water (4 measuring cups) into the Styrofoam container. As long as the water temperature is not very hot or very cold, it is unnecessary to cover the top of the Styrofoam container during measurements. Since the transmitter will eventually heat the water several degrees, water initially a few degrees cooler than air temperature is ideal because the average water temperature will very nearly equal the air temperature and heat transfer to the air will be minimized.

Connect the RG-174 dummy load to the transmitter through the shortest possible length of lower loss cable such as RG-8. Tape the connectors and adapter at the RG-8 to RG-174 joint carefully with plastic tape to prevent water from leaking into the connectors and cable at this point. Roll the RG-174 into a loose coil and submerge it in the water. Do not bind the turns of the coil together in any way, as the water must be able to freely circulate among the coaxial cable turns. All the RG-174 cable must be submerged in the water to ensure sufficient cooling. Also submerge part of the taped connector attached to the RG-174 as an added precaution.

On completing the above steps, quickly tune up the transmitter for maximum power output into the load. Cease transmitting and stir the water slowly for a minute or so until its temperature has stabilized. Then measure the water temperature as precisely as possible. After the initial temperature has been determined, begin the test “transmission,” measuring the total number of seconds of key-down time accurately. Stir the water slowly with the thermometer and continue transmitting until there is a significant rise in the water temperature, say 5 to 10 degrees. The test may be broken up into a series of short periods, as long as you keep track of the total key-down time. When the test is completed, continue to stir the water slowly and monitor its temperature. When the temperature ceases to rise, note the final indication as precisely as possible.

To compute the transmitter power output, multiply the calorimeter heat capacitance (4215 for C or 2342 for F) by the difference in initial and final water temperature. Then divide by the total number of seconds of key-down time. The resultant is the transmitter power in watts. A nomogram which can also be used to find transmitter power output is given in **Fig 27**. With a straight line, connect the total number of key-down seconds in the time column to the number of degrees change (F or C) in the temperature rise column, and read off the transmitter power output at the point where the straight line crosses the power-output column.

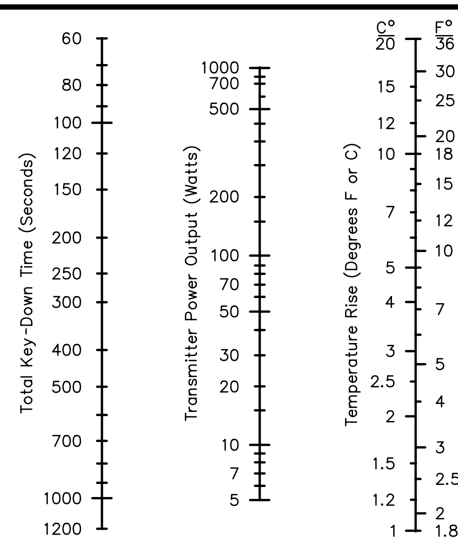


Fig 27—Nomogram for finding transmitter power output for the calorimeter.

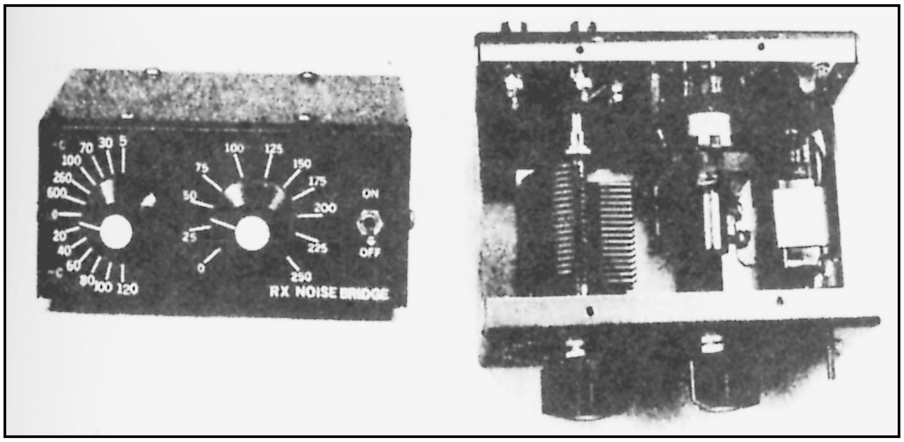


Fig 28—Exterior and interior views of the noise bridge. The unit is finished in red enamel. Press-on lettering is used for the calibration marks. Note that the potentiometer must be isolated from ground.

Power Limitation

The maximum power handling capability of the calorimeter is limited by the following. At very high powers the dielectric material in the coaxial line will melt because of excessive heating or the cable will arc over from excessive voltage. As the transmitter frequency gets higher, the excessive-heating problem is accentuated, as more of the power is dissipated in the first several feet of cable. For instance, at 1296 MHz, approximately 10% of the transmitting power is dissipated in the first foot of cable. Overheating can be prevented when working with high power by using a low duty cycle to reduce the average dissipated power. Use a series of short transmissions, such as two seconds on, 10 seconds off. Keep count of the total key-down time for power calculation purposes. If the cable arcs over, use a larger-diameter cable, such as RG-58, in place of the RG-174. The cable should be long enough to assure that the reflected wave will be down 10 dB or more at the input. It may be necessary to use more than one quart of water in order to submerge all the cable conveniently. If so, be sure to calculate the new value of heat capacitance for the larger quantity of water. Also you should measure the new coaxial cable heat capacitance using the method previously described.

A Noise Bridge For 1.8 Through 30 MHz

The noise bridge, sometimes referred to as an antenna (RX) noise bridge, is an instrument that will allow the user to measure the impedance of an antenna or other electrical circuits. The unit shown here, designed for use in the 1.8 through 30-MHz range, provides adequate accuracy for most measurements. Battery operation and small physical size make this unit ideal for remote-location use. Tone modulation is applied to the wide-band noise generator as an aid for obtaining a null indication. A detector, such as the station receiver, is required for operation.

The noise bridge consists of two parts—the noise generator and the bridge circuitry. See **Fig 29**. A 6.8-V Zener diode serves as the noise source. U1 generates an approximate 50% duty cycle, 1000-Hz square wave signal which is applied to the cathode of the Zener diode. The 1000-Hz modulation appears on the noise signal and provides a useful null detection enhancement effect. The broadband-noise signal is amplified by Q1, Q2 and associated components to a level that produces an approximate S9 signal in the receiver. Slightly more noise is available at the lower end of the frequency range, as no frequency compensation is applied to the amplifier. Roughly 20 mA of current is drawn from the 9-V battery, thus ensuring long battery life—providing the power is switched off after use!

The bridge portion of the circuit consists of T1, C1, C2 and R1. T1 is a trifilar wound transformer with one of the windings used to couple noise energy into the bridge circuit. The remaining two windings are arranged so that each one is in an arm of the bridge. C1 and R1 complete one arm and the UNKNOWN circuit, along with C2, comprise the remainder of the bridge. The terminal labeled RCVR is for connection to the detector.

The reactance range of a noise bridge is dependent on several factors, including operating frequency, value of the series capacitor (C3 or C3 plus C4 in Fig 29) and the range of the variable capacitor (C1 in Fig 29). The RANGE switch selects reactance measurements weighted toward either capacitance or inductance by placing C4 in parallel with C3. The zero-reactance point occurs when C1 is either nearly fully meshed or fully unmeshed. The RANGE switch nearly doubles the resolution of the reactance readings.

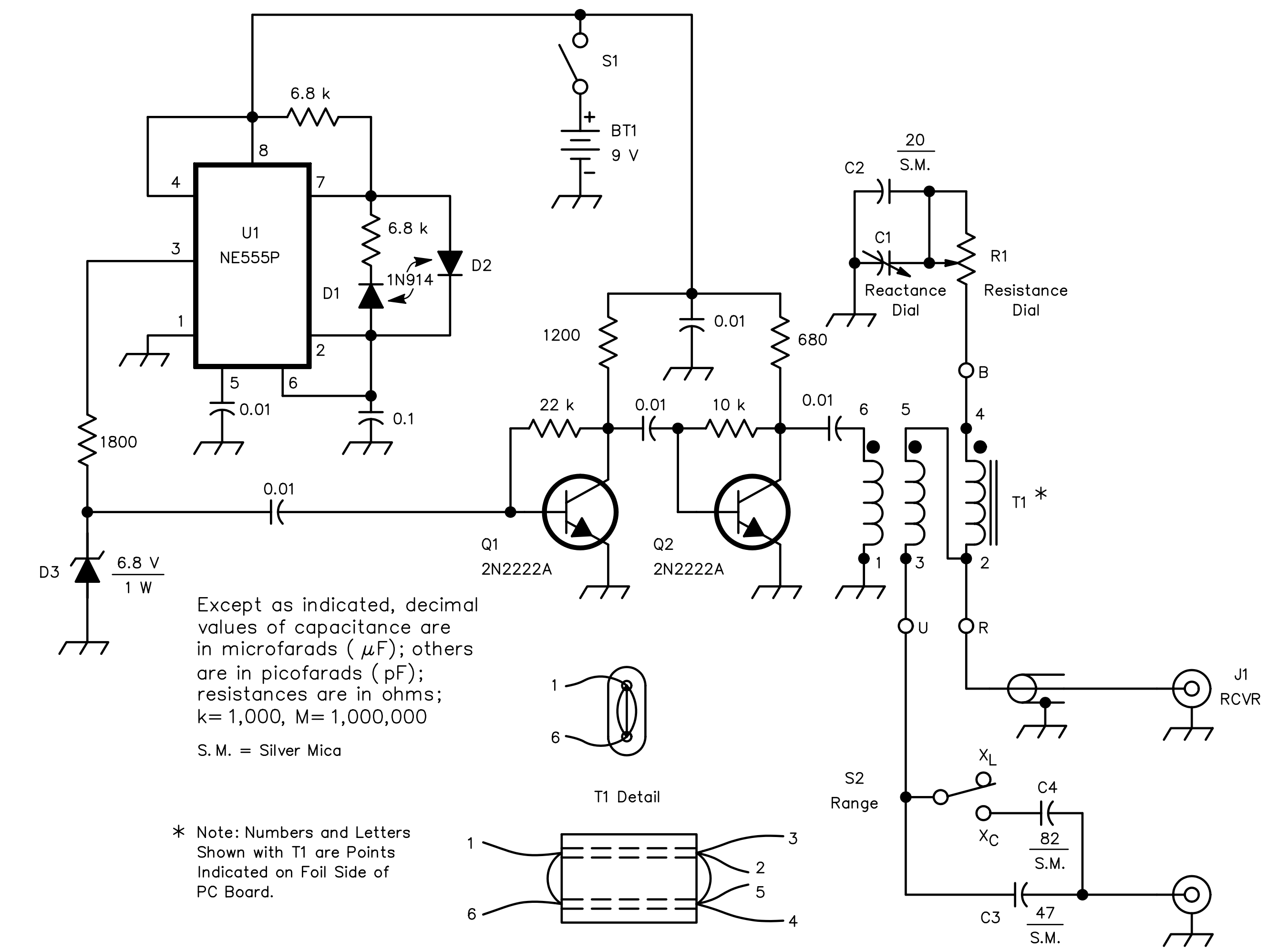


Fig 29—Schematic diagram of the noise bridge. Use 1/4-W composition resistors. Capacitors are miniature ceramic units unless indicated otherwise. Component designations indicated in the schematic but not called out in the parts list are for text and parts-placement reference only.

BT1—9-V battery, NEDA 1604A or equiv.
C1—15- to 150-pF variable
C2—20-pF mica.
C3—47-pF mica.
C4—82-pF mica.
J1, J2—Coaxial connector.
R1—Linear, 250 Ω, AB type. Use a good grade of resistor.
S1, S2—Toggle, SPST.
T1—Transformer; 3 windings on an Amidon BLN-43-2402 ferrite binocular core. Each winding is three turns of #30 enameled wire. One turn is equal to the wire passing once through both holes in the core. The primary winding starts on one side of the transformer, and the secondary and tertiary windings start on the opposite side.
U1—Timer, NE555 or equiv.

CONSTRUCTION

The noise bridge is contained in a homemade aluminum enclosure that measures 5 × 2 3/8 × 3 3/4 inches. Many of the circuit components are mounted on a circuit board that is fastened to the rear wall of the cabinet. The circuit-board layout is such that the lead lengths to the board from the bridge and coaxial connectors are at a minimum. An etching pattern and a parts-placement guide for the circuit board are shown in **Figs 30** and **31**.

Care must be taken when mounting the potentiometer, R1. For accurate readings the potentiometer must be well insulated from ground. In the unit shown this was accomplished by mounting the control on a piece of Plexiglas, which in turn was fastened to the chassis with a piece of aluminum angle stock. Additionally, a 1/4-inch control-shaft coupling and a length of phenolic rod were used to further isolate the control from ground where the shaft passes through the front panel. A high-quality potentiometer is required if good measurement results are to be obtained.

There is no such problem when mounting the variable capacitor because the rotor is grounded. Use a high-quality capacitor; do not try to save money on that component. Two RF connectors on the rear panel are connected to a detector (receiver) and to the UNKNOWN circuit. Do not use plastic-insulated phono connectors (they might influence bridge accuracy at higher frequencies). Use miniature coaxial cable (RG-174) between the RCVR connector and circuit board. Attach one end of C3 to the circuit board and the other directly to the UNKNOWN circuit connector.

Bridge Compensation

Stray capacitance and inductance in the bridge circuit can affect impedance readings. If a very accurate bridge is required, use the following steps to counter the effects of stray reactance. Because the physical location of the board, connectors and controls in the cabinet determine where compensation is needed, there is no provision for the compensation components on the printed circuit board.

Good calibration loads are necessary to check the accuracy of the noise bridge. Four are needed here: a 0-Ω (short-circuit) load, a 50-Ω load, a 180-Ω load, and a variable-resistance load. The short-circuit and fixed-resistance loads are used to check the accuracy of the noise bridge; the varaible-resistance load is used when measuring coaxial-cable loss.

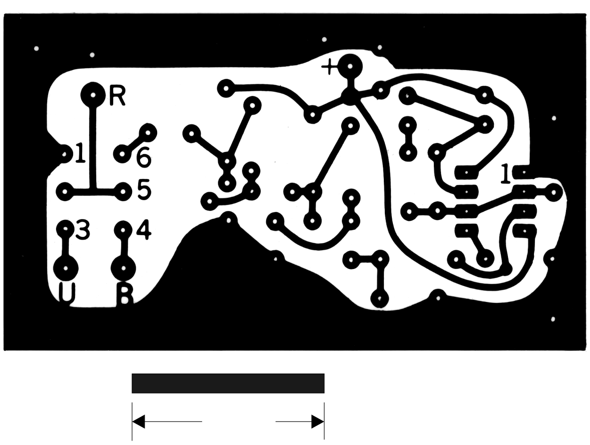


Fig 30—Etching pattern for the noise bridge PC board, at actual size. Black represents copper. This is the pattern for the bottom side of the board. The top side of the board is a complete ground plane with a small amount of copper removed from around the component holes.

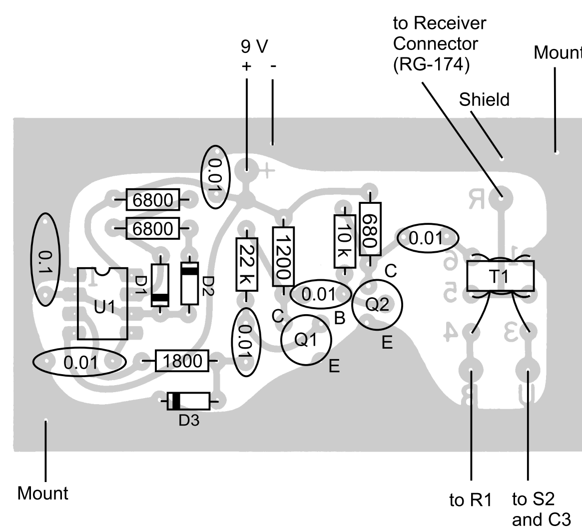


Fig 31—Parts-placement guide for the noise bridge as viewed from the component or top side of the board. Mounting holes are located in two corners of the board, as shown.

Construction details of the loads are shown in **Fig 32**. Each load is constructed inside a connector. When building the loads, keep leads as short as possible to minimize parasitic effects. The resistors must be noninductive (*not* wirewound). Quarter-watt, carbon-composition resistors should work fine. The potentiometer in the variable-resistance load is a miniature PC-mount unit with a maximum resistance of 100 Ω or less. The potentiometer wiper and one of the end leads are connected to the center pin of the connector; the other lead is connected to ground.

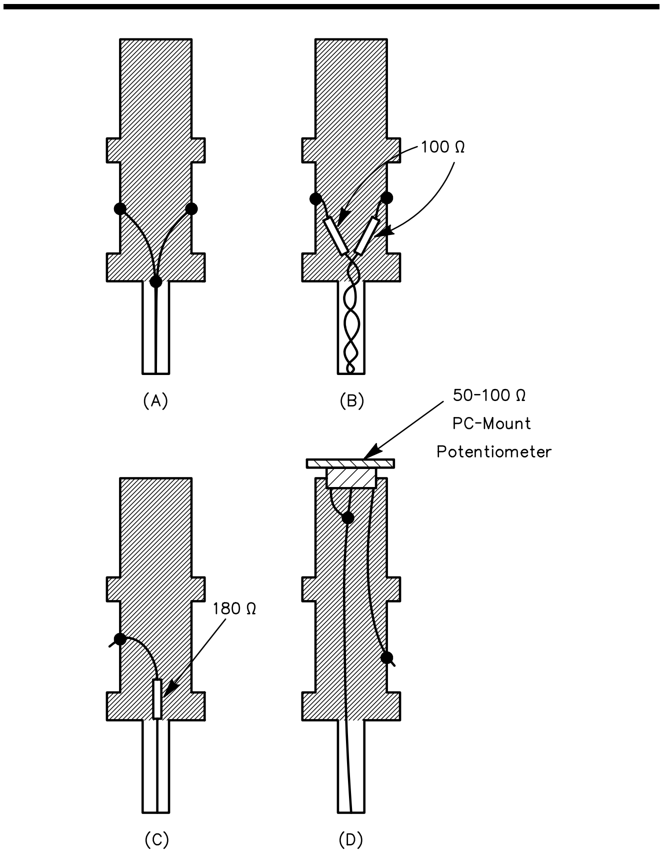


Fig 32—Construction details of the resistive loads used to check and calibrate the noise bridge. Each of the loads is constructed inside a coaxial connector that matches those on the bridge. (Views sh wn are cross-sections of PL-259 bodies; the sleeves are not shown.) Leads should be kept as short as possible to minimize parasitic inductance. A is a 0-Ω load; B depicts a 50-Ω load; C is a 180-Ω load; D shows a variable-resistance load used to determine the loss in a coaxial cable.

Stray Capacitance

Stray capcitance on the variable-resistor side of the bridge tends to be higher than that on the unknown side. This is so because the parasitic capacitance in the variable resistor, R1, is comparatively high.

The effect of parasitic capacitance is most easily detected using the 180-Ω load. Measure and record the actual resistance of the load, R_L. Connect the load to the UNKNOWN connector, place S2 in the X_L position, tune the receiver to 1.8 MHz, and null the bridge. (See the section, “Finding the Null” for tips.) Use an ohmmeter across R1 to measure its dc resistance. The magnitude of the stray capacitance can be calculated by

$$Cp = C3\left(\sqrt{\frac{R1}{R_L}} - 1\right) \qquad \text{(Eq 4)}$$

where

R_L = load resistance (as measured)
R1 = resistance of the variable resistor
C3 = series capacitance.

You can compensate for C_p by placing a variable capacitor, C_C, in the side of the bridge with lesser stray capacitance. If R1 is greater than R_L, stray capacitance is greater on the variable resistor side of the bridge: Place C_c between point U (on the circuit board) and ground. If R1 is less than R_L, stray capacitance is greater on the unknown side: Place C_c between point B and ground.

If the required compensating capacitance is only a few picofarads, you can use a gimmick capacitor (made by twisting two short pieces of insulated, solid wire together) for C_c. A gimmick capacitor is adjusted by trimming its length.

Stray Inductance

Parasitic inductance, if present, should be only a few tens of nanohenries. This represents a few ohms of inductive reactance at 30 MHz. The effect is best observed by reading the reactance of the 0-Ω test load at 1.8 and 30 MHz; the indicated reactance should be the same at both frequencies.

If the reactance reading decreases as frquency is increased, parasitic inductance is greater in the known arm, and compensating inductance is needed between point U and C3. If the reactance increases with frequency, the unknown-arm inductance is greater, and compensating inductance should be placed between point B and R1.

Compensate for stray inductance by placing a single-turn coil, made from a 1 to 2-inch length of solid wire, in the appropriate arm of the bridge. Adjust the size of this coil until the reactance reading remains constant from 1.8 to 30 MHz.

Calibration

Good calibration accuracy is necessary for accurate noise-bridge measurements. Calibration of the resistance scale is straightforward. To do this, tune the receiver to a frequency near 10 MHz. Attach the 0-Ω load to the UNKNOWN connector and null the bridge. This is the zero-resistance point; mark it on the front-panel resistance scale. The rest of the resistance range is calibrated by adjusting R1, measuring R1 with an accurate ohmmeter, calculating the increase from the zero point and marking the increase on the front panel.

Most bridges have the reactance scale marked in capacitance because capacitance does not vary with frequency. Unfortunately, that requires calibration curves or non-trivial calculations to arrive at the load reactance. An alternative method is to mark the reactance scale in *ohms* at a reference frequency of 10 MHz. This method calibrates the bridge near the center of its range and displays reactance directly, but it requires a simple calculation to scale the reactance reading for frequencies other than 10 MHz. The scaling equation is:

$$X_{u(f)} = X_{u(10)} \frac{10}{f} \tag{Eq 5}$$

where

f = frequency in MHz

$X_{u(10)}$ = reactance of the unknown load at 10 MHz

$X_{u(f)}$ = reactance of the unknown load at f.

A shorted piece of coaxial cable serves as a reactance source. (The reactance of a shorted, low-loss coaxial cable is dependent only on the cable length, the measurement frequency and the cable characteristic impedance.) Radio Shack RG-8M is used here because it is readily available, has relatively low loss and has an almost purely resistive characteristic impedance.

Prepare the calibration cable as follows:

1) Cut a length of coaxial cable that is slightly longer than $^1/_4$ λ at 10 MHz (about 20 feet for RG-8M). Attach a suitable connector to one end of the cable; leave the other end open-circuited.

2) Connect the 0-Ω load to the noise bridge UNKNOWN connector and set the receiver frequency to 10 MHz. Adjust the noise bridge for a null. Do not adjust the reactance control after the null is found.

3) Connect the calibration cable to the bridge UNKNOWN terminal. Null the bridge by adjusting *only* the varaible resistor and the receiver frequency. The receiver frequency should be less than 10 MHz; if it is above 10 MHz, the cable is too short, and you need to prepare a longer one.

4) Gradually cut short lengths from the end of the coaxial cable until you obtain a null at 10 MHz by adjusting only the resistance control. Then connect the cable center and shield conductors at the open end with a short length of braid. Verify that the bridge nulls with zero reactance at 20 MHz.

5) The reactance of the coaxial cable (normalized to 10 MHz) can be calculated from:

$$X_{i(10)} = R_0 \frac{f}{10} \tan\left(2\pi \frac{f}{40}\right) \tag{Eq 6}$$

where

$X_{i(10)}$ = cable reactance at 10 MHz

R_0 = characteristic resistance of the coaxial cable (52.5 Ω for Radio Shack RG-8M)

f = frequency in MHz

The results of Eq 6 have less than 5% error for reactances less than 500 Ω, as long as the test-cable loss is less than 0.2 dB. This error becomes significantly less at lower reactances (2% error at 300 Ω for a 0.2-dB-loss cable). The loss in 18 feet of RG-8M is 0.13 dB at 10 MHz. Reactance data for Radio Shack RG-8M is given in **Table 5**.

With the prepared cable and calibration values on hand, proceed to calibrate the reactance scale. Tune the receiver to the appropriate frequency for the desired reactance (given in Table 5, or found using Eq 6). Adjust the resistance and reactance controls to null the bridge. Mark the reactance reading on the front panel. Repeat this process until all desired reactance values have been marked. The resistance values needed to null the bridge during this calibration procedure may be significant (more than 100 Ω) at the higher reactances.

This calibration method is much more accurate than using fixed capacitors across the UNKNOWN connector. Also, you can calibrate a noise bridge in less than an hour using this method.

Table 5
Noise Bridge Calibration Data: Coaxial-Cable Method

This data is for Radio Shack RG-8M cable (R_0 = 52.5 Ω) cut to exactly ¹⁄₄ λ at 10 MHz; the reactances and capacitances shown correspond to this frequency.

Reactance				Capacitance	
X_i	f(MHz)	X_i	f(MHz)	C(pF)	f(MHz)
10	3.318	–10	19.376	10	9.798
20	4.484	–20	18.722	20	9.612
30	5.262	–30	18.048	30	9.440
40	5.838	–40	17.368	40	9.280
50	6.286	–50	16.701	50	9.130
60	6.647	–60	16.062	60	8.990
70	6.943	–70	15.471	70	8.859
80	7.191	–80	14.936	80	8.735
90	7.404	–90	14.462	90	8.618
100	7.586	–100	14.044	100	8.508
110	7.747	–110	13.682	110	8.403
120	7.884	–120	13.369	120	8.304
130	8.009	–130	13.097	130	8.209
140	8.119	–140	12.861	140	8.119
150	8.217	–150	12.654		
160	8.306	–160	12.473		
170	8.387	–170	12.313		
180	8.460	–180	12.172		
190	8.527	–190	12.045		
200	8.588	–200	11.932		
210	8.645	–210	11.831	–10	10.219
220	8.697	–220	11.739	–20	10.459
230	8.746	–230	11.655	–30	10.721
240	8.791	–240	11.579	–40	11.010
250	8.832	–250	11.510	–50	11.328
260	8.872	–260	11.446	–60	11.679
270	8.908	–270	11.387	–70	12.064
280	8.942	–280	11.333	–80	12.484
290	8.975	–290	11.283	–90	12.935
300	9.005	–300	11.236	–100	13.407
350	9.133	–350	11.045	–110	13.887
400	9.232	–400	10.905	–120	14.357
450	9.311	–450	10.798	–130	14.801
500	9.375	–500	10.713	–140	15.211

Finding the Null

In use, a receiver is attached to the RCVR connector and some load of unknown value is connected to the UNKNOWN terminal. The receiver allows us to hear the noise present across the bridge arms at the frequency of the receiver passband. The strength of the noise signal depends on the strength of the noise-bridge battery, the receiver bandwidth/sensitivity and the impedance difference between the known and unknown bridge arms. The noise is stronger and the null more obvious with wide receiver passbands. Set the receiver to the widest bandwidth AM mode available.

The noise-bridge output is heard as a 1000-Hz tone. When the impedances of the known and unknown bridge arms are equal, the voltage across the receiver is minimized; this is a null. In use, the null may be difficult to find because it appears only when both bridge controls approach the values needed to balance the bridge.

To find the null, set C1 to mid-scale, sweep R1 slowly through its range and listen for a reduction in noise (it's also helpful to watch the S meter). If no reduction is heard, set R1 to mid-range and sweep C1. If there is still no reduction, begin at one end of the C1 range and sweep R1. Increment C1 about 10% and sweep R1 with each increment until some noise reduction appears. Once noise reduction begins, adjust C1 and R1 alternately for minimum signal.

MEASURING COAXIAL-CABLE PARAMETERS WITH A NOISE BRIDGE

Coaxial cables have a number of properties that affect the transmission of signals through them. Generally, radio amateurs are concerned with cable attenuation and characteristic impedance. If you plan to use a noise bridge to make antenna-impedance measurements, however, you need to accurately determine not just cable impedance and attenuation, but also electrical length. Fortunately, all of these parameters are easy to measure with an accurate noise bridge.

Cable Electrical Length

With the noise bridge and a general-coverage receiver, you can easily locate frequencies at which the line in question is a multiple of $^1/_2$ λ, because a shorted $^1/_2$ λ line has a 0-Ω impedance (neglecting line loss). By locating two adjacent null frequencies, you can solve for the length of line in terms of $^1/_2$ λ at one of the frequencies and calculate the line length (overall accuracy is limited by noise-bridge accuracy and line loss, which broadens the nulls). As an interim variable, express cable length as the frequency at which a cable is 1 λ long. This length will be represented by f_λ. Follow these steps to determine f_λ for a coaxial cable.

1) Tune the receiver to the frequency range of interest. Attach the short-circuit load to the noise bridge UNKNOWN connector and null the bridge.

2) Disconnect the far end of the coaxial cable from its load (the antenna) and connect it to the 0-Ω test load. Connect the near end of the cable to the bridge UNKNOWN connector.

3) Adjust the receiver frequency and the noise-bridge resistance control for a null. *Do not change the noise bridge reactance-control setting during this procedure.* Note the frequency at which the null is found; call this frequency f_n. The noise-bridge resistance at the null should be relatively small (less than 20 Ω).

4) Tune the receiver upward in frequency until the next null is found. Adjust the resistance control, if necessary, to improve the null, *but do not adjust the reactance control*. Note the frequency at which this second null is found; this is f_{n+2}.

5) Solve Eq 7 for n and the electrical length of the cable.

$$n = \frac{2f_n}{f_{n+2} - f_n} \qquad \text{(Eq 7)}$$

$$f_\lambda = \frac{4f_n}{n} \qquad \text{(Eq 8)}$$

$$\ell = \frac{f_0}{f_\lambda} \qquad \text{(Eq 9)}$$

where

n = cable electrical length in quarter waves, at f_n

f_λ = frequency at which the cable is 1 λ

ℓ = cable elctrical length, in λ

For example, consider a 74-foot length of Columbia 1188 foam-dielectric cable (velocity factor = 0.78) to be used on the 10-m band. Based on the manufacturer's specification, the cable is 2.796 λ at 29 MHz. Nulls were found at 24.412 (f_n) and 29.353 (f_{n+2}) MHz. Eq 7 yields n = 9.88, which produces 9.883 MHz from Eq 8 and 2.934 λ for Eq 9. If the manufacturer's specification is correct, the measured length is off by less than 5%, which is very reasonable. Ideally, n would yield an integer. The difference between n and the closest integer indicates that there is some error.

This procedure also works for lines with an open circuit as the termination (n will be close to an odd number). End effects from the PL-259 increase the effective length of the coaxial cable, however; this decreases the calculated f_λ.

Cable Characteristic Impedance

The characteristic impedance of the coaxial cable is found by measuring its input impedance at two frequencies separated by $^1/_4$ f_λ. This must be done when the cable is terminated in a resistive load. Characteristic impedance changes slowly as a function of frequency, so this measurement must be done near the frequency of interest. The measurement procedure is as follows.

1) Place the 50-Ω load on the far end of the coaxial cable and connect the near end to the UNKNOWN connector of the noise bridge. (Measurement error is minimized when the load resistance is close to the characteristic impedance of the cable. This is the reason for using the 50-Ω load.)

2) Tune the receiver approximately $^1/_8$ f_λ below the frequency of interest. Adjust the bridge resis-

tance and reactance controls to obtain a null, and note their readings as R_{f1} and X_{f1}. Remember, the reactance reading must be scaled to the measurement frequency.

3) Increase the receiver frequency by exactly $^1/_4$ f_λ. Null the bridge again, and note the readings as R_{f2} and X_{f2}.

4) Calculate the characteristic impedance of the coaxial cable using Eqs 10 through 15. A scientific calculator is helpful for this.

$$R = R_{f1} \times R_{f2} - X_{f1} \times X_{f2} \quad \text{(Eq 10)}$$

$$X = R_{f1} \times X_{f2} + X_{f1} \times R_{f2} \quad \text{(Eq 11)}$$

$$Z = \sqrt{R^2 + X^2} \quad \text{(Eq 12)}$$

$$R_0 = \sqrt{Z} \cos\left[\frac{1}{2} \tan^{-1}\left(\frac{X}{R}\right)\right] \quad \text{(Eq 13)}$$

$$Z_0 = \sqrt{Z} \sin\left[\frac{1}{2} \tan^{-1}\left(\frac{X}{R}\right)\right] \quad \text{(Eq 14)}$$

$$Z_0 = R_0 + jX_0 \quad \text{(Eq 15)}$$

Let's continue with the example used earlier for cable length. The measurements are:

f1 = 29.000 – (9.883/8) = 27.765 MHz
R_{f1} = 64 Ω
X_{f1} = –22 Ω × (10/27.765) = –7.9 Ω
f2 = 27.765 + (9.88$^3/_4$) = 30.236 MHz
R_{f2} = 50 Ω
X_{f2} = –24 Ω × (10/30.236) = –7.9 Ω

When used in Eqs 10 through 15, these data yield:
R = 3137.59
X = –900.60
Z = 3264.28
R_0 = 56.58 Ω
X_0 = –7.96 Ω

Cable Attenuation

Cable loss can be measured once the cable electrical length and characteristic resistance are known. The measurement must be made at a frequency where the cable presents no reactance. Reactance is zero when the cable electrical length is an integer multiple of λ/4. You can easily meet that condition by making the measurement frequency an integer multiple of $^1/_4$ f_λ. Loss at other frequencies can be interpolated with reasonable accuracy. This procedure employs a resistor-substitution method that provides much greater accuracy than is achieved by directly rewading the resistance from the noise-bridge scale.

1) Determine the approximate frequency at which you want to make the loss measurement by using

$$n = \frac{4f_0}{f_\lambda} \quad \text{(Eq 16A)}$$

Round n to the nearest integer, then

$$f1 = \frac{n}{4} f_\lambda \quad \text{(Eq 16B)}$$

2) If n is odd, leave the far end of the cable open; if n is even, connect the 0-Ω load to the far end

of the cable. Attach the near end of the cable to the UNKNOWN connector on the noise bridge.

3) Set the noise bridge to zero reactance and the receiver to f1. Fine tune the receiver frequency and the noise-bridge resistance to find the null.

4) Disconnect the cable from the UNKNOWN terminal, and connect the variable-resistance calibration load in its place. Without changing the resistance setting on the bridge, adjust the load resistor and the bridge reactance to obtain a null.

5) Remove the variable-resistance load from the bridge UNKNOWN terminal and measure the load resistance using an ohmmeter that's accurate at low resistance levels. Refer to this resistance as R_i.

6) Calculate the cable loss in decibels using

$$\alpha \ell = 8.69 \frac{R_i}{R_0} \qquad \text{(Eq 17)}$$

To continue this example, Eq 16A gives n = 11.74, so measure the attenuation at n = 12. From Eq 16B, f1 = 29.649 MHz. The input resistance of the cable measures 12.1 Ω with 0-Ω load on the far end of the cable; this corresponds to a loss of 1.86 dB.

USING A NOISE BRIDGE TO MEASURE THE IMPEDANCE OF AN ANTENNA

The impedance at the end of a transmission line can be easily measured using a noise bridge. In many cases, however, you really want to measure the impedance of an antenna—that is, the impedance of the load at the far end of the line. There are several ways to handle this.

1) Measurements can be made with the noise bridge at the antenna. This is usually not practical because the antenna must be in its final position for the measurement to be accurate. Even if it can be done, making such a measurement is certainly not very convenient.

2) Measurements can be made at the source end of a coaxial cable—if the cable length is an exact integer multiple of $^1/_2$ λ. This effectively restricts measurements to a single frequency.

3) Measurements can be made at the source end of a coaxial cable and corrected using a Smith Chart as shown in Chapter 28. This graphic method can result in reasonable estimates of antenna impedance—as long as the SWR is not too high and the cable is not too lossy. However, it doesn't compensate for the complex impedance characteristics of real-world coaxial cables. Also, compensation for cable loss can be tricky to apply. These problems, too, can lead to significant errors.

4) Last, measurements can be corrected using the transmission-line equation, as discussed under "Transmission Line Equations" later. The transmission-line equation can be solved using only a scientific calculator, but it is best handled with a programmable calculator or personal computer.* This is the best method for calculating antenna impedances from measured parameters, but it requires that you measure the feed-line characteristics beforehand—measurements for which you need access to both ends of the feed line.

The procedure for determining antenna impedance is to first measure the electrical length, characteristic impedance, and attenuation of the coaxial cable connected to the antenna. After making these measurements, connect the antenna to the coaxial cable and measure the input impedance of the cable

*Listings for BASIC and HP-41C calculator programs are available from the Technical Department at ARRL HQ for an SASE. Request order no. 3495, and send a business-size stamped reply envelope.

Table 6
Impedance Data for Inverted-V Antenna

Freq (MHz)	*R_u (Ω)*	*X_u @ 10 MHz (Ω)*	*X_u (Ω)*	*R_L (Ω)*	*X_L (Ω)*
27.0	44	85	31.5	24	–65
27.2	60	95	34.9	26	–56
27.4	75	85	31.0	30	–51
27.6	90	40	14.5	32	–42
27.8	90	–20	–7.2	35	–34
28.0	75	–58	–20.7	38	–24
28.2	65	–65	–23.0	40	–19
28.4	56	–52	–18.3	44	–12
28.6	50	–40	–14.0	44	–6
28.8	48	–20	–6.9	47	1
29.0	50	0	0.0	52	8
29.2	55	20	6.8	57	15
29.4	64	30	10.2	63	21
29.6	78	20	6.8	75	26
29.8	85	0	0	78	30
30.0	90	–50	–16.7	89	33

at a number of frequencies. Then use these measurements in the trnasmission-line equation to determine the actual antenna impedance at each frequency.

Table 6 and **Fig 33** give an example of such a calculation. The antenna used for this example is a 10-meter inverted V about 30 feet above the ground. The arms of the antenna are separated by a 120° angle. Each arm is exactly 8 feet long, and the antenna is made of #14 wire. The feed line is the 74-foot length of Columbia 1188 characterized earlier.

See Fig 33A. From this plot of impedance measurements, it is very difficult to determine anything about the antenna. Resistance and reactance vary substantially over this frequency range, and the antenna appears to be resonant at 27.7, 29.0 and 29.8 MHz.

The plot in Fig 33B shows the true antenna impedance. This plot has been corrected for the effects of the cable using the transmission-line equation. The true antenna resistance and reactance both increase smoothly with frequency. The antenna is resonant at 28.8 MHz, with a radiation resistance at resonance of 47 Ω. This is normal for an inverted V.

When doing the conversions, be careful not to make measurement errors. Such errors introduce more errors into the corrected data. This problem is most significant when the transmission line is near an odd multiple of a $^1/_4\,\lambda$ and the line SWR and/or attenuation is high. Measurement errors are probably present if small changes in the input impedance or transmission-line characteristics appear as large changes in antenna impedance. If this effect is present, it can be minimized by making the measurements with a transmission line that is approximately an integer multiple of $^1/_2\,\lambda$.

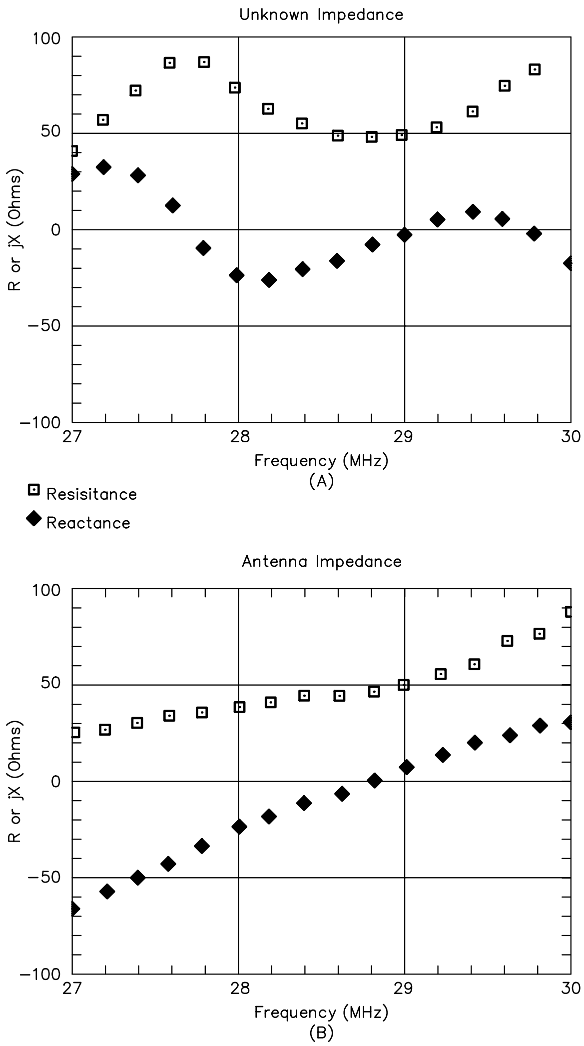


Fig 33–Impedance plot of an inverted-V antenna cut for 29 MHz. At A, a plot of resistances and reactances, measured using the noise bridge, at the end of a 74-foot length of Columbia 1188 coaxial cable. At B, the actual antenna-impedance plot (found using the transmission-line equation to remove the effects of the trnasmission line).

TRANSMISSION LINE EQUATIONS

The impedance transformation that occurs when a signal propagates on a transmission line can be solved either graphically (using a Smith Chart), or numerically, using the transmission-line equation.The transmission line may be either open-wire line or coaxial cable. With the advent of personal computers, impedance transformation in a transmission line is more easily and accurately claculated numerically than with the Smith Chart. The impedance transformation along a transmission line is given by Eq 18, which is taken from work by S. Ramo, J. Whinnery and T. Van Duzer (see Bibliography). All trig-based functions in Eqs 18 through 38 are expressed or calculated in radians. (One radian is 360/2π ≈ 57.29°.)

$$Z_i = Z_0\left(\frac{Z_L\cosh(\gamma\ell) + Z_0\sinh(\gamma\ell)}{Z_0\cosh(\gamma\ell) + Z_L\sinh(\gamma\ell)}\right) \qquad \text{(Eq 18)}$$

where

Z_i = input impedance of the transmission line

Z_0 = characteristic impedance of the transmission line
Z_L = load impedance at the end of the transmission line
ℓ = length of the transmission line in radians
γ = complex propagation constant ($\gamma = \alpha + j\beta$)
α = attenuation constant in nepers per unit length (1 neper = 8.69 dB)
β = phase constant in radians per unit length

The impedances and the propagation constant may be complex numbers. The complex hyperbolic sine and cosine may be found by

$$\sinh(\alpha\ell + j\beta\ell) = \cos(\beta\ell)\sinh(\alpha\ell) + j\sin(\beta\ell)\cosh(\alpha\ell) \qquad \text{(Eq 19)}$$

$$\cosh(\alpha\ell + j\beta\ell) = \cos(\beta\ell)\cosh(\alpha\ell) + j\sin(\beta\ell)\sinh(\alpha\ell) \qquad \text{(Eq 20)}$$

$$\sinh(\alpha\ell) = \frac{e^{\alpha\ell} - e^{-\alpha\ell}}{2} \qquad \text{(Eq 21)}$$

$$\cosh(\alpha\ell) = \frac{e^{\alpha\ell} + e^{-\alpha\ell}}{2} \qquad \text{(Eq 22)}$$

For finding the load impedance (with a known transmission-line input impedance), the transmission-line equation is best written as

$$Z_L = Z_0\left(\frac{Z_i\cosh(\gamma\ell) - Z_0\sinh(\gamma\ell)}{Z_0\cosh(\gamma\ell) - Z_i\sinh(\gamma\ell)}\right) \qquad \text{(Eq 23)}$$

Most antenna measurements are made through a fixed length of coaxial cable. Therefore, we'll assume that $\alpha\ell$ is a single unit that we'll call the attenuation of the cable. This is commonly measured in decibels, but must be converted to nepers for use in the transmission-line equation. The phase constant can be expressed as a function of frequency and the length of the transmission line by:

$$\beta\ell = 2\pi\frac{f}{f_\lambda} \qquad \text{(Eq 24)}$$

where

f = frequency of operation
f_λ = frequency at which the transmission line is 1 λ long electrically

A shorted transmission line is used to measure f_λ. To do this, find a frequency at which the transmission line has zero reactance and a low resistance (less than the characteristic resistance of the transmission line). Call this frequency f_n. Increase the frequency until the next zero-reactance, low-resistance point is found.We'll call this f_{n+2}. (The subscript indicates the number of quarter wavelengths that are present on the transmission line; n is always an integer.)

$$n = \frac{2f_n}{f_{n+2} - f_n} \qquad \text{(Eq 25)}$$

where n = 2, 4, 6, . . .

$$f_\lambda = \frac{4f_n}{n} \qquad \text{(Eq 26)}$$

The value of f_λ calculated in Eq 26 assumes that the transmission line has a nonreactive characteristic impedance. This is generally not true, but Eq 26 is accurate nonetheless; it yields an error of less than 2.5% for a transmission line with less than 3 dB loss and a reactive characteristic-impedance component of less than 10 Ω.

Transmission-line characteristic impedance is almost always complex. Good coaxial cable has a very small reactive characteristic-impedance component (on the order of a few ohms). Cable characteristic impedance is most easily calculated by placing a load at one end of the cable and measuring the impedance at the other end at two frequencies separated by $^1/_4\ f_\lambda$. The input impedance of the cable is then

$$Z_i(f) = Z_0\left(\frac{Z_L\cosh(\gamma\ell) + Z_0\sinh(\gamma\ell)}{Z_0\cosh(\gamma\ell) + Z_L\sinh(\gamma\ell)}\right) \qquad \text{(Eq 27)}$$

$$Z_i\left(f + f_\lambda / 4\right) = Z_0\left(\frac{Z_L\sinh(\gamma\ell) + Z_0\cosh(\gamma\ell)}{Z_0\sinh(\gamma\ell) + Z_L\cosh(\gamma\ell)}\right) \qquad \text{(Eq 28)}$$

Eqs 27 and 28 can be manipulated so that the characteristic impedance can be found by

$$Z_0 = \sqrt{Z_i(f)\ Z_i\left(f + f_\lambda/4\right)} \qquad \text{(Eq 29)}$$

The square root is complex, and may be calculated with a scientific calculator using Eqs 30 through 34.

$$Z = R + jX = Z_i(f)\ Z_i(f + f_\lambda/4) \qquad \text{(Eq 30)}$$

$$|Z| = \sqrt{R^2 + X^2} \qquad \text{(Eq 31)}$$

$$R_0 = \sqrt{|Z|}\ \cos\left[1/2\ \tan^{-1}\left(\frac{X}{R}\right)\right] \qquad \text{(Eq 32)}$$

$$X_0 = \sqrt{|Z|}\ \sin\left[1/2\ \tan^{-1}\left(\frac{X}{R}\right)\right] \qquad \text{(Eq 33)}$$

$$Z_0 = R_0 + jX_0 \qquad \text{(Eq 34)}$$

Transmission-line attenuation can be calculated after using this transmission-line impedance equation

$$Z_i = Z_0 \times \left\{\frac{Z_L\left[\cos(\beta\,\ell) + j\,\alpha\ell\sin(\beta\,\ell)\right] + Z_0\left[\alpha\ell\cos(\beta\,\ell) + j\sin(\beta\,\ell)\right]}{Z_0\left[\cos(\beta\,\ell) + j\,\alpha\ell\sin(\beta\,\ell)\right] + Z_L\left[\alpha\ell\cos(\beta\,\ell) + j\sin(\beta\,\ell)\right]}\right\} \qquad \text{(Eq 35)}$$

This equation yields an error of less than 5%—as long as the transmission-line loss is less than 3 dB. If the transmission line is an odd multiple of a quarter wavelength (n = 1, 3, 5, . . .) and is terminated by an open circuit, or if the transmission line is an even multiple of a quarter wavelength (n = 2, 4, 6, . . .) and is terminated by a short circuit, the input impedance is given by

$$Z_i = \alpha\ell\ Z_0 \qquad \text{(Eq 36)}$$

The attenuation of this transmission line can be found by

$$\alpha\ell = \frac{R_i}{R_0} \qquad \text{(Eq 37)}$$

where R_i and R_0 are the resistive parts of the input impedance and the characteristic impedance, respectively. The transmission-line attenuation increases with frequency. An estimate for this attenuation is given by

$$\alpha\ell(f) = \alpha\ell(f_\alpha)\left(\frac{f}{f_\alpha}\right)^\sigma \qquad \text{(Eq 38)}$$

where $0.5 < \sigma < 1$

This equation can be used to interpolate between measured values of attenuation. For most coaxial cables, $\sigma = 0.5$ works well.

A Practical Time-Domain Reflectometer

A time-domain reflectometer (TDR) is a simple but powerful tool used to evaluate transmission lines. When used with an oscilloscope, a TDR displays impedance "bumps" (open and short circuits, kinks and so on) in transmission lines. Commercially produced TDRs cost from hundreds to thousands of dollars each, but you can add the TDR described here to your shack for much less. This material is based on a *QST* article by Tom King, KD5HM (see Bibliography), and supplemented with information from the references.

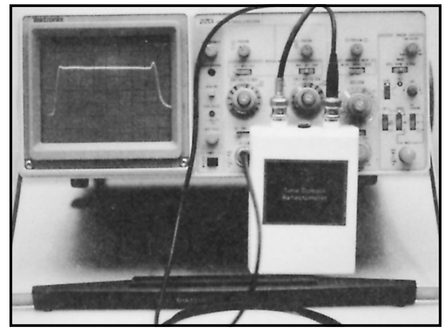


Fig 34—The time-domain reflectometer shown here is attached to a small portable oscilloscope.

How a TDR Works

A simple TDR consists of a square-wave generator and an oscilloscope. The generator sends a train of dc pulses down a transmission line, and the oscilloscope lets you observe the incident and reflected waves from the pulses (when the scope is synchronized to the pulses). A little analysis of the scope display tells the nature and location of any impedance changes along the line.

The nature of an impedance disturbance is identified by comparing its pattern to those in **Fig 35**. The patterns are based on the fact that the reflected wave from a disturbance is determined by the incident-wave magnitude and the reflection coefficient of the disturbance. (The patterns shown neglect losses; actual patterns may vary somewhat from those shown.)

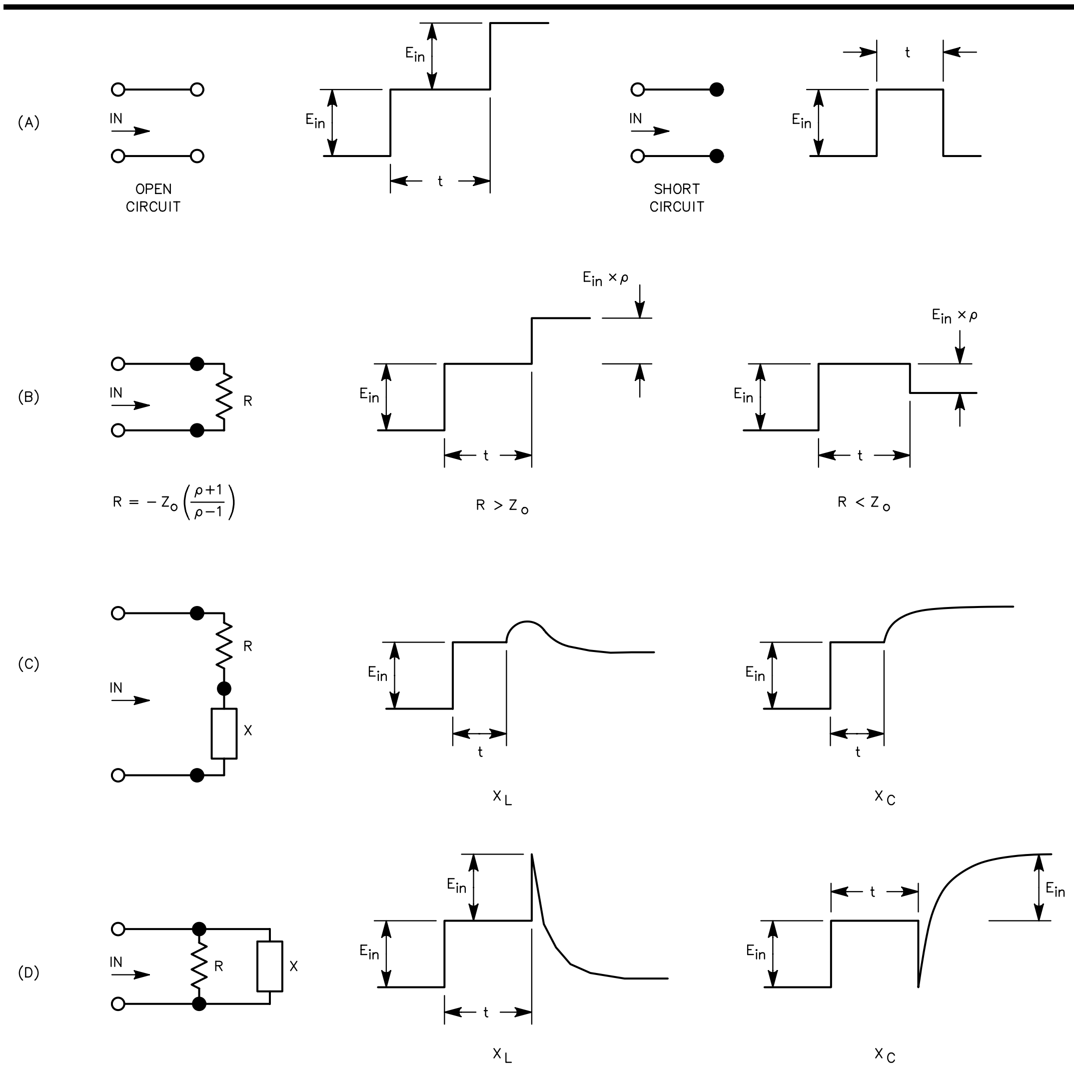


Fig 35—Characteristic TDR patterns for various loads. The location of the load can be calculated from the transit time, t, which is read from the oscilloscope (see text). R values can be calculated as shown (for purely resistive loads only—$\rho < 0$ when R < Z_0; $\rho > 0$ when R > Z_0). Values for reactive loads cannot be calculated simply.

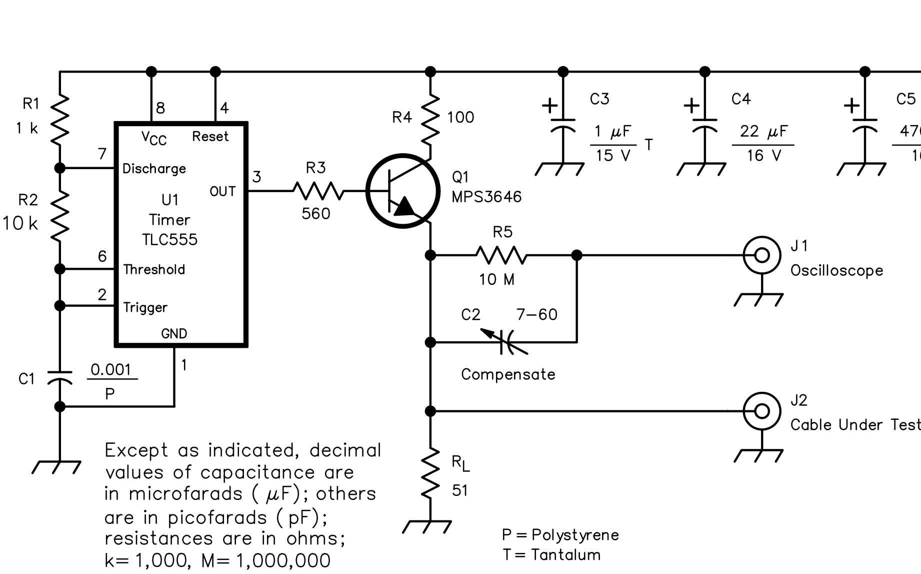


Fig 36—Schematic diagram of the time-domain reflectometer. All resistors are 1/4 W, 5% tolerance. U1 is a CMOS 555 timer. Circuit current drain is 10 to 25 mA. When building the TDR, observe the construction cautions discussed in the text. C2 is available trom Mouser Electronics, part no. ME242-8050. Right-angle BNC connectors for use at J1 and J2 can be obtained trom Newark Electronics, part no. 89N1578. S1 can be obtained from All Electronics, part no. NISW-1. An SPST toggle switch can also be used at S1.

The location of a disturbance is calculated with a simple proportional method: The round-trip time (to the disturbance) can be read from the oscilloscope screen (graticule). Thus, you need only read the time, multiply it by the velocity of the radio wave (the speed of light adjusted by the velocity factor of the transmission line) and divide by two. The distance to a disturbance is given by:

$$\ell = \frac{(983.6 \times \mathrm{VF} \times t)}{2} \qquad \text{(Eq 1)}$$

where

ℓ = line length in feet

VF = velocity factor of the transmission line (from 0 to 1.0)

t = time delay in microseconds (μs).

The Circuit

The time-domain reflectometer circuit in **Fig 36** consists of a CMOS 555 timer configured as an astable multivibrator, followed by an MPS3646 transistor acting as a 15-ns-risetime buffer. The timer provides a 71-kHz square wave. This is applied to the 50-Ω transmission line under test (connected at J2). The oscilloscope is connected to the circuit at J1.

Construction

An etching pattern for the TDR is shown in **Fig 37**. **Fig 38** is the part-placement diagram. The TDR is designed for a 4×3×1-inch enclosure (in-

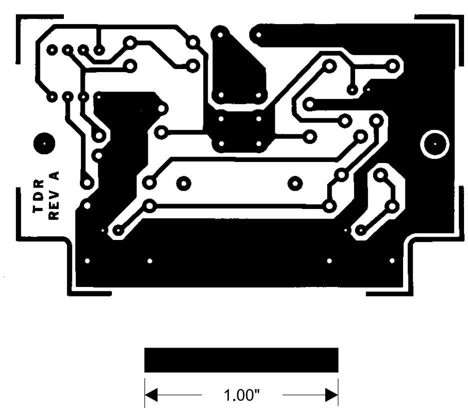
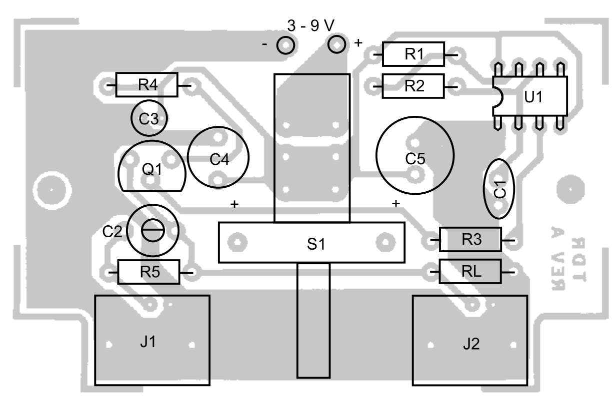


Fig 37—Full-size PC-board etching pattern tor the TDR. Black areas represent unetched copper foil.



Flg 38—Part-placement diagram for the TDR. Parts are mounted on the nonfoil side of the board; the shaded area represents an X-ray view of the copper pattern. Be sure to observe the polarity markings of C3, C4 and C5.

cluding the batteries). Sl, Jl and J2 are right-angle-mounted components.

Two aspects of construction are critical. First use *only* an MPS3646 for Q1! This type was chosen for its good performanee in this circuit. If you substitute another transistor, the circuit may not perform properly.

Second, for the TDR to provide accurate measurements, the cable connected to J1 (between the TDR and the oscilloscope) must not introduce impedance mismatches in the circuit. *Do not make this cable from ordinary coaxial cable.* Oscilloscope-probe cable is the best thing to use for this connection. (It took the author about a week and several phone calls to determine that scope-probe cable isn't "plain old coax." Probe cable has special characteristics that prevent undesired ringing and other problems.) Mount a binding post at J1 and connect a scope probe to the binding post when testing cables with the TDR. R5 and C2 form a compensation network—much like the networks in oscilloscope probes—to adjust for effects of the probe wire.

The TDR is designed to operate from dc between 3 and 9 V. Two C cells (in series—3 V) supply operating voltage in this version. The circuit draws only 10 to 25 mA, so the cells should last a long time (about 200 hours of operation). U1 can function with supply voltages as low as 2.25 to 2.5.

If you want to use the TDR in transmission-line systems with characteristic impedances other than 50 Ω, change the value of R_L to match the system impedance as closely as possible.

Calibrating and Using the TDR

Just about any scope with a bandwidth of at least 10 MHz should work fine with the TDR, but for tests in short-length cables, a 50-MHz scope provides for much more accurate measurements. To calibrate the TDR, terminate CABLE UNDER TEST connector, J2, with a 51-Ω resistor. Connect the scope vertical input to J1. Turn on the TDR, and adjust the scope timebase so that one square-wave cycle from the TDR fills as much of the scope display as possible (without uncalibrating the timebase). The waveform should resemble **Fig 39**. Adjust C2 to obtain maximum amplitude and sharpest corners on the observed waveform. That's all there is to the calibration process!

To use the TDR, connect the cable under test to J2, and connect the scope vertical input to J1. If the waveform you observe is different from the one you observed during calibration, there are impedance variations in the load you're testing. See **Fig 40**, showing an unterminated test cable connected to the TDR. The beginning of the cable is shown at point A. (AB represents the TDR output-pulse rise time.) Segment AC shows the portion of the transmission line that has a 50-Ω impedance. Between points C and D, there is a mismatch in the line. Because the scope trace is higher than the 50-Ω trace, the impedance of this part of the line is higher than 50 Ω—in this case, an open circuit.

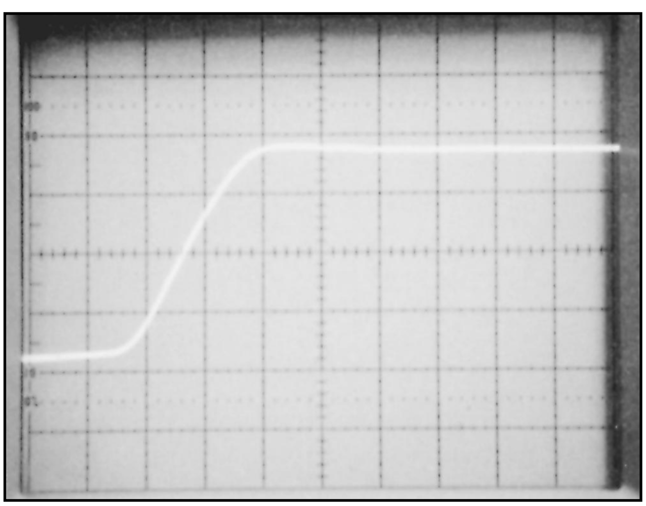


Fig 39—TDR calibration trace as shown on an oscilloscope. Adjust C2 (See Figs 36 and 38) for maximum deflection and sharpest waveform corners during calibration. See text.

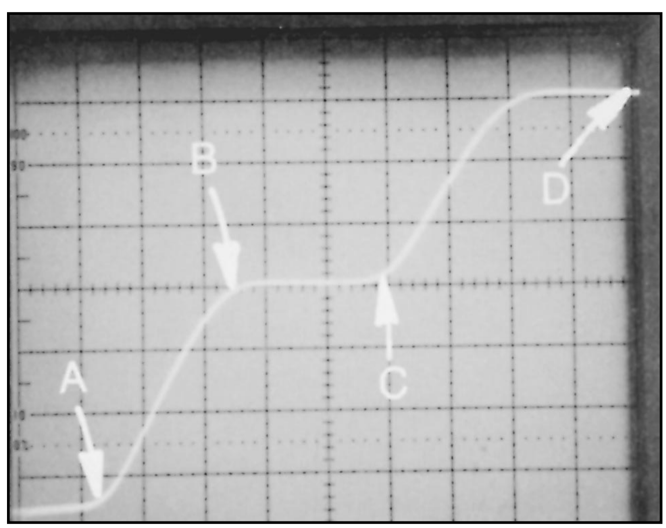


Fig 40—Open-circuited test cable. The scope is set for 0.01 µs per division. See text for interpretation of the waveform.

To determine the length of this cable, read the length of time over which the 50-Ω trace is displayed. The scope is set for 0.01 µs per division, so the time delay for the 50-Ω section is (0.01 µs × 4.6 divisions) = 0.046 µs. The manufacturer's specified velocity factor (VF) of the cable is 0.8. Eq 1 tells us that the 50-Ω section of the cable is

$$\ell = \frac{(983.6 \times 0.8 \times 0.046 \mu s)}{2} = 18.1 \text{ ft}$$

The TDR provides reasonable agreement with the actual cable length—in this case, the cable is really 16.5 feet long. (Variations in TDR-derived calculations and actual cable lengths can occur as a result of cable VFs that can vary considerably from published values. Many cables vary as much as 10% from the specified values!)

A second example is shown in **Fig 41**, where a length of 3/4-inch Hardline is being tested. The line feeds a 432-MHz vertical antenna at the top of a tower. Fig 41 shows that the 50-Ω line section has a delay of (6.6 divisions × 0.05 µs) = 0.33 µs. Because the trace is straight and level at the 50-Ω level, the line is in good shape. The trailing edge at the right-hand end shows where the antenna is connected to the feed line.

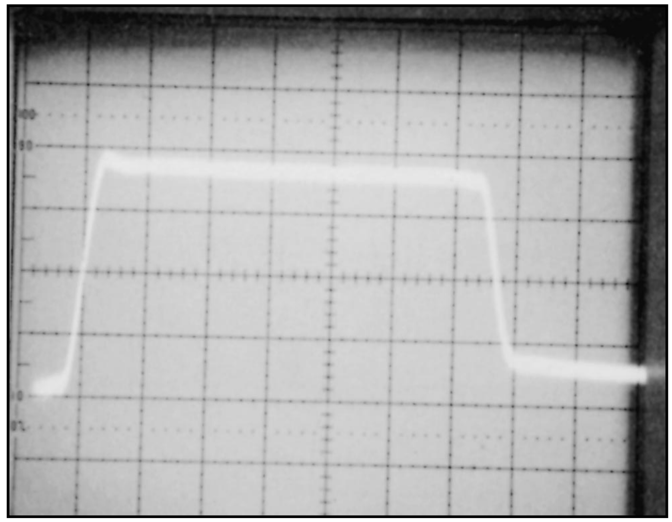


Fig 41—TDR display of the impedance characteristics of the 142-foot Hardline run to the 432-MHz antenna at KD5HM. The scope is set for 0.05 µs per division. See text for discussion.

To determine the actual length of the line, use the same procedure as before: Using the published VF for the Hardline (0.88) in Eq 1, the line length is

$$\ell = \frac{(983.6 \times 0.88 \times 0.33 \ \mu s)}{2} = 142.8 \ \text{ft}$$

Again, the TDR-derived measurement is in fairly close agreement with the actual cable length (142 feet).

Final Notes

The time-domain reflectometer described here is not frequency specific; its measurements are not made at the frequency at which a system is designed to be used. Because of this, the TDR cannot be used to verify the impedance of an antenna, nor can it be used to measure cable loss at a specific frequency. Just the same, in two years of use, it has never failed to help locate a transmission-line problem. The vast majority of transmission-line problems result from improper cable installation or connector weathering.

Limitations

Certain limitations are characteristic of TDRs because the signal used to test the line differs from the system operating frequency and because an oscilloscope is a broadband device. In the instrument described here, measurements are made with a 71-kHz square wave. That wave contains components at 71 kHz and odd harmonics thereof, with the majority of the energy coming from the lower frequencies. The leading edge of the trace indicates that the response drops quickly above 6 MHz. (The leading edge in Fig 40 is 0.042 µs, corresponding to a period of 0.168 µs and a frequency of 5.95 MHz.) The result is dc pulses of approximately 7 µs duration. The scope display combines the circuit responses to all of those frequencies. Hence, it may be difficult to interpret any disturbance which is narrowband in nature (affecting only a small range of frequencies, and thus a small portion of the total power), or for which the travel time plus pattern duration exceeds 7 µs.

The 432-MHz vertical antenna in Fig 41 illustrates a display error resulting from narrow-band response. The antenna shows as a major impedance disturbance because it is mismatched at the low frequencies that dominate the TDR display, yet it is matched at 432 MHz. For an event that exceeds the observation window, consider a l-µF capacitor across a 50-Ω line. You would see only part of the pattern shown in Fig 35C because the time constant ($1 \times 10^{-6} \times 50 = 50$ µs) is much larger than the 7-µs window.

In addition, TDRs are unsuitable for measurements where there are major impedance changes inside the line section to be tested. Such major changes mask reflections from additional changes farther down the line.

Because of these limitations, TDRs are best suited for spotting faults in dc-continuous systems that maintain a constant impedance from the generator to the load. Happily, most amateur stations would be ideal subjects for TDR analysis, which can conveniently check antenna cables and connectors for short and open-circuit conditions and locate the position of such faults with fair accuracy.

Measuring Soil Conductivity

An important parameter for both vertical and horizontal antennas is soil conductivity. For horizontal antennas, the energy reflected from the earth beneath it affects the antenna impedance, thereby affecting the SWR and the current flowing in the antenna elements, which in turn affects the distant signal strength. (This is discused in more detail in Chapter 3.) The conductivity of the ground within several wavelengths of the antenna also affects the ground reflection factors discussed in Chapter 3.

The conductivity of the soil under and in the near vicinity of a vertical antenna is most important in determining the extent of the radial system required and the overall performance. Short verticals with very small radial systems can be surprisingly efective—in the right location. The material in this section was prepared by Jerry Sevick, W2FMI.

Most soils are nonconductors of electricity when completely dry. Conduction through the soil results from conduction throught the water held in the soil. Thus, conduction is electrolytic. Dc techniques for measuring conductivity are impractical because they tend to deplete the carriers of electricity in the vicinity of the electricity in the vicinity of the electrodes. The main factors contributing to the conductivity of soil are

1) Type of soil.
2) Type of salts contained in the water.
3) Concentration of salts dissolved in the contained water.
4) Moisture content.
5) Grain size and distribution of material.
6) Temperature.
7) Packing density and pressure.

Although the type of soil is an important factor in determining its conductivity, rather large variations can take place between locations because of the other factors involved. Generally, loams and garden soils have the highest conductivities. These are followed in order by clays, sand and gravel. Soils have been classified according to conductivity, as discussed in Chapter 3.

Making Conductivity Measurements

Since conduction through the soil is almost entirely electrolytic, ac measurement techniques are preferable. Many commercial instruments using ac techniques are available and described in the literature. But rather simple ac measurement techniques can be used that provide accuracies on the order of 25% and are quite adequate for the radio amateur. Such a setup was developed by Jerry Sevick, W2FMI, and M. C. Waltz, W2FNQ, and was published by Sevick in April 1978 and March 1981 *QST*. It is shown in **Figs 42** through **44**.

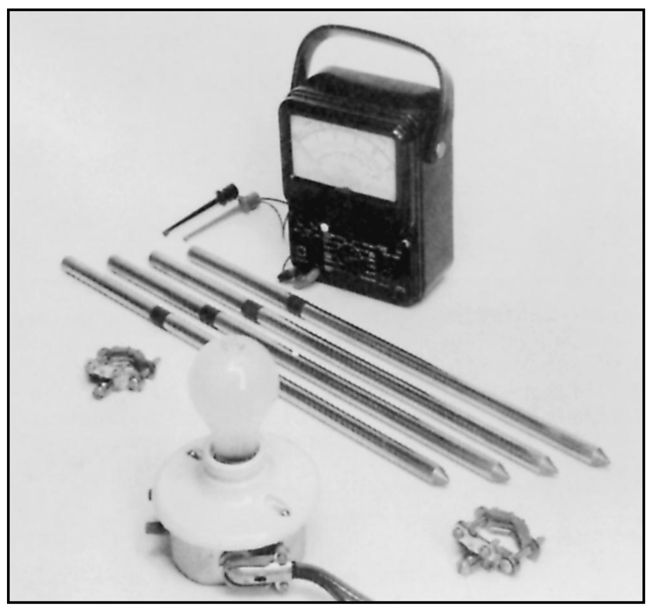


Fig 42—The complete soil conductivity measuring setup. The four probes are cut to 18-inch lengths from an 8-foot copper-coated steel ground rod. (This length provides a measuring stick for spacing the probes when driving them into the soil.) The tip of each probe is ground to a point, and black electricians' tape indicates the depth to which it is to be driven for measurements. Two ground clamps provide for connections to the driven probes.

Four probes are used. Each is 9/16 inch in diameter, and may be made of either iron or copper. The probes are inserted in a straight line at a spacing of 18 inches (dimension *d* in Fig 44). The penetration depth is 12 inches. *Caution:* Do not insert the probes with the power applied! A shock hazard exists! After applying power, measure the voltage drops V1 and V2, as shown in the diagram. Depending on soil conditions, readings should fall in the range from 2 to 10 volts.

Earth conductivity, c, may be determined from

$$c = 21 \times \frac{V1}{V2} \text{ millisiemens per meter}$$

For example, assume the reading across the resistor (V1) is 4.9 V, and the reading between the two center probes (V2) is 7.2 V. The conductivity is calculated as 21 × 4.9/7.2 = 14mS/m.

Soil conditions may not be uniform in different parts of your yard. A few quick measurements will reveal whether this is the case or not. **Fig 45** shows the conductivity readings taken in one location over a period of three months. It is interesting to note the general drop in conductivity over the three months, as well as the short-term changes from periods of rain.

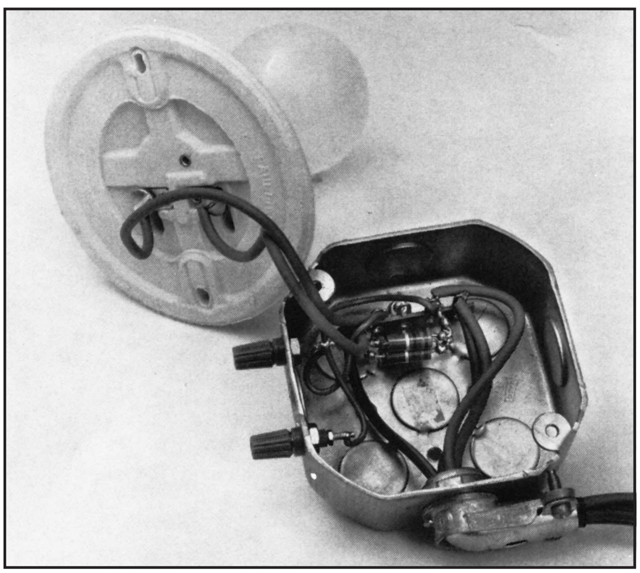


Fig 43—A standard 3½ inch electrical outlet box and a porcelain ceiling fixture may be used to construct the soil conductivity test set. The rersistors comprising R1 are mounted on a tie-point strip inside the box, and test-point jacks provide for measuring the voltage drop across the resistor combination. Leads exiting the box through the cable clamp are protected with several layers of electricians' tape. These leads run approximately 4 feet to the power plug and to small alligator clamps shown in Fig 42. Large clips such as for connecting to automotive battery posts may be used instead of ground clamps.

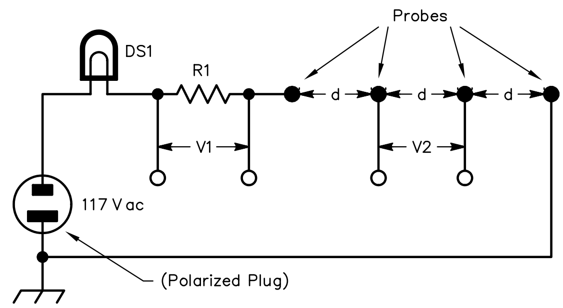


Fig 44—Schematic diagram, four-point probe method for measuring earth conductivity.

DS1—100-W electric light bulb.

R1—14.6 Ω, 5 W. A suitable resistance can be made by paralleling five 1-W resistors, three of 68 Ω and two of 82 Ω. (The dissipation rating of this combination will be 4.7 W.)

Probes—See text and Fig 42.

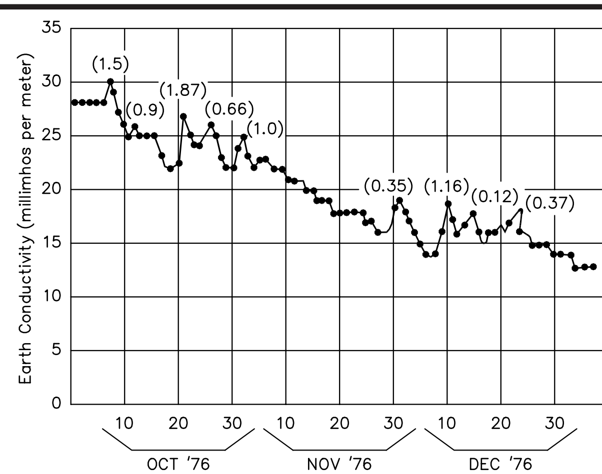


Fig 45—Earth conductivity at a central New Jersey location during a three-month period. Numbers in parentheses indicate inches of rainfall.

A Switchable RF Attenuator

A switchable RF attenuator is helpful in making antenna gain comparisons or plotting antenna radiation patterns; attenuation may be switched in or out of the line leading to the receiver to obtain an initial or reference reading on a signal strength meter. Some form of attenuator is also helpful for locating hidden transmitters, where the real trick is pinpointing the signal source from within a few hundred feet. At such a close distance, strong signals may overload the front end of the receiver, making it impossible to obtain any indication of a bearing.

The attenuator of **Figs 46** and **47** is designed for low power levels, not exceeding $^1/_4$ W. If for some reason the attenuator will be connected to a transceiver, a means of bypassing the unit during transmit periods must be devised. An attenuator of this type is commonly called a step attenuator, because any amount of attenuation from 0 dB to the maximum available (81 dB for this particular instrument) may be obtained in steps of 1 dB. As each switch is successively thrown from the OUT to the IN position, the attenuation sections add in cascade to yield the total of the attenuator steps switched in. The maximum attenuation of any single section is limited to 20 dB because leak-through would probably degrade the accuracy of higher values. The tolerance of resistor values also becomes more significant regarding accuracy at higher attenuation values.

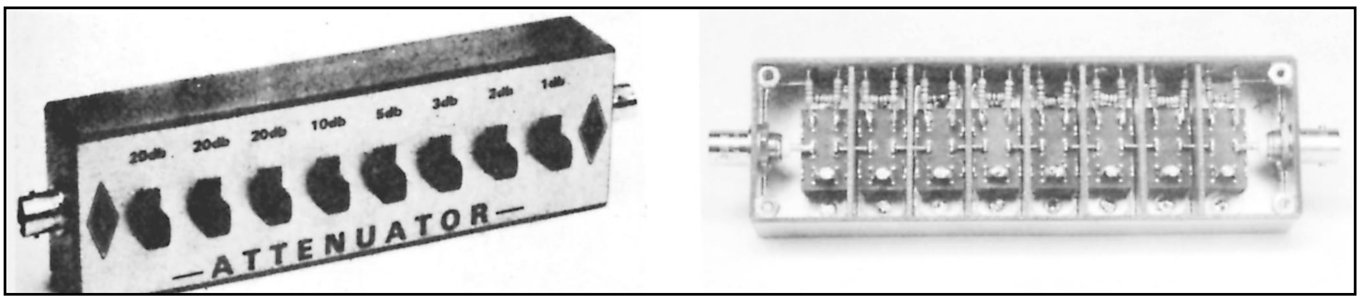


Fig 46—A construction method for a step attenuator. Double-sided circuit-board material, unetched (except for panel identification), is cut to the desired size and soldered in place. Flashing copper may also be used, although it is not as sturdy. Shielding partitions between sections are necessary to reduce signal leakage. Brass nuts soldered at each of the four corners allow machine screws to secure the bottom cover. The practical limit for total attenuation is 80 or 90 dB, as signal leakage around the outside of the attenuator will defeat attempts to obtain much greater amounts.

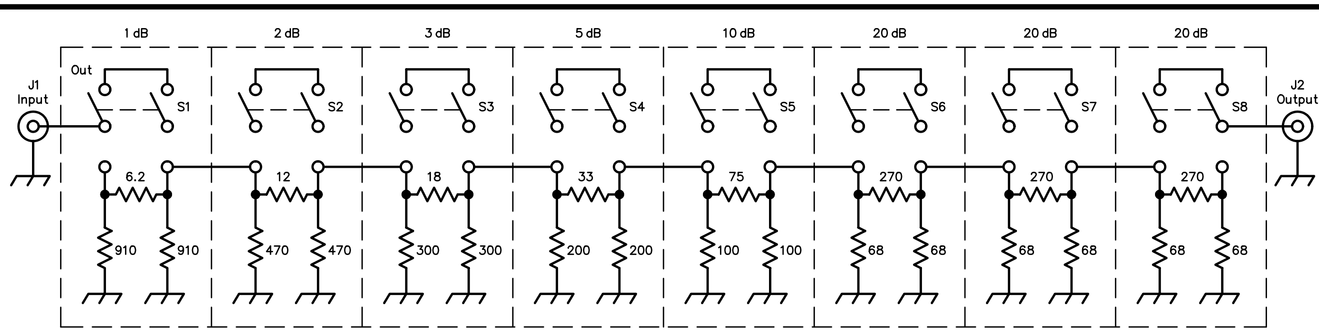


Fig 47—Schematic diagram of the step attenuator, designed for a nominal impedance of 52 ohms. Resistance values are in ohms. Resistors are $^1/_4$-watt, carbon-composition types, 5% tolerance. Broken lines indicate walls of circuit-board material. A small hole is drilled through each partition wall to route bus wire. Keep all leads as short as possible. The attenuator is bilateral; that is, the input and output ends may be reversed.

J1, J2—Female BNC connectors, Radio Shack 278-105 or equiv.

S1-S8, incl.—DPDT slide switches, standard size. (Avoid subminiature or toggle switches.) Stackpole S-5022CD03-0 switches are used here.

A good quality commercially made attenuator will cost upward from $150, but for less than $25 in parts and a few hours of work, an attenuator may be built at home. it will be suitable for frequencies up to 450 MHz. Double-sided PC board is used for the enclosure. The version of the attenuator shown in Fig 46 has identification lettering etched into the top surface (or front panel) of the unit. This adds a nice touch and is a permanent means of labeling. Of course rub-on transfers or Dymo tape labels could be used as well.

Female BNC single-hole, chassis-mount connectors are used at each end of the enclosure. These connectors provide a means of easily connecting and disconnecting the attenuator.

Construction

After all the box parts are cut to size and the necessary holes made, scribe light lines to locate the inner partitions. Carefully tack-solder all partitions in position. A 25-W pencil type of iron should provide sufficient heat. Dress any PC board parts that do not fit squarely. Once everything is in proper position, run a solder bead all the way around the joints. Caution! Do not use excessive amounts of solder, as the switches must later be fit flat inside the sections. The top, sides, ends and partitions can be completed. Dress the outside of the box to suit your taste. For instance, you might wish to bevel the box edges. Buff the copper with steel wool, add lettering, and finish off the work with a coat of clear lacquer or polyurethane varnish.

Using a little lacquer thinner or acetone (and a lot of caution), soak the switches to remove the grease that was added during their manufacture. When they dry, spray the inside of the switches lightly with a TV tuner cleaner/lubricant. Use a sharp drill bit (about 3/16 inch will do), and countersink the mounting holes on the actuator side of the switch mounting plate. This ensures that the switches will fit flush against the top plate. At one end of each switch, bend the two lugs over and solder them together. Cut off the upper halves of the remaining switch lugs. (A close look at Fig 46 will help clarify these steps.)

Solder the series-arm resistors between the appropriate switch lugs. Keep the lead lengths as short as possible and do not overheat the resistors. Now solder the switches in place to the top section of the enclosure by flowing solder through the mounting holes and onto the circuit-board material. Be certain that you place the switches in their proper positions; correlate the resistor values with the degree of attenuation. Otherwise, you may wind up with the 1-dB step at the wrong end of the box—how embarrassing!

Once the switches are installed, thread a piece of #18 bare copper wire through the center lugs of all the switches, passing it through the holes in the partitions. Solder the wire at each switch terminal. Cut the wire between the poles of each individual switch, leaving the wire connecting one switch pole to that of the neighboring one on the other side of the partition, as shown in Fig 46. At each of the two end switch terminals, leave a wire length of approximately 1/8 inch. Install the BNC connectors and solder the wire pieces to the connector center conductors.

Now install the shunt-arm resistors of each section. Use short lead lengths. Do not use excessive amounts of heat when soldering. Solder a no. 4-40 brass nut at each inside corner of the enclosure. Recess the nuts approximately 1/16-inch from the bottom edge of the box to allow sufficient room for the bottom panel to fit flush. Secure the bottom panel with four no. 4-40, 1/4-inch machine screws and the project is completed. Remember to use caution, always, when your test setup provides the possibility of transmitting power into the attenuator.

A Portable Field-Strength Meter

Few amateur stations, fixed or mobile, are without need of a field-strength meter. An instrument of this type serves many useful purposes during antenna experiments and adjustments. When work is to be done from many wavelengths away, a simple wavemeter lacks the necessary sensitivity. Further, such a device has a serious fault because its linearity leaves much to be desired. The information in this section is based on a January 1973 *QST* article by Lew McCoy, W1ICP.

The field-strength meter described here takes care of these problems. Additionally, it is small, measuring only 4 × 5 × 8 inches. The power supply consists of two 9-V batteries. Sensitivity can be set for practically any amount desired. However, from a usefulness standpoint, the circuit should not be too sensitive or it will respond to unwanted signals. This unit also has excellent linearity with regard to field strength. (The field strength of a received signal varies inversely with the distance from the source, all other things being equal.) The frequency range includes all amateur bands from 3.5 through 148 MHz, with band-switched circuits, thus avoiding the use of plug-in inductors. All in all, it is a quite useful instrument.

The unit is pictured in **Figs 48** and **49**, and the schematic diagram is shown in **Fig 50**. A type 741 op-amp IC is the heart of the unit. The antenna is connected to J1, and a tuned circuit is used ahead of a diode detector. The rectified signal is coupled as dc and amplified in the op amp. Sensitivity of the op amp is controlled by inserting resistors R3 through R6 in the circuit by means of S2.

With the circuit shown, and in its most sensitive setting, M1 will detect a signal from the antenna on the order of 100 µV. Linearity is poor for approximately the first $^{1}/_{5}$ of the meter range, but then is almost straight-line from there to full-scale deflection. The reason for the poor linearity at the start of the readings is because of nonlinearity of the diodes at the point of first conduction. However, if gain measurements are being made this is of no real importance, as accurate gain measurements can be made in the linear portion of the readings.

The 741 op amp requires both a positive and a negative voltage source. This is obtained by connecting two 9-V batteries in series and grounding the center. One other feature of the instrument is that it can be used remotely by connecting an external meter at J2. This is handy if you want to adjust an antenna and observe the results without having to leave the antenna site.

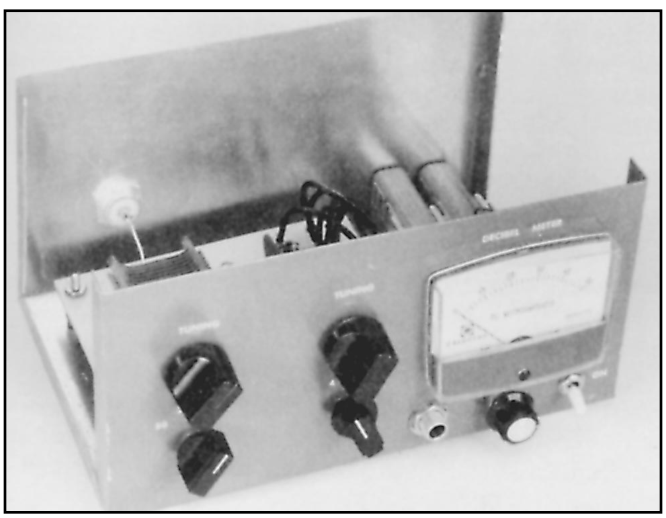


Fig 48—The linear field-strength meter. The control at the upper left is for C1 and the one to the right for C2. At the lower left is the band switch, and to its right the sensitivity switch. The zero-set control for M1 is located directly below the meter.

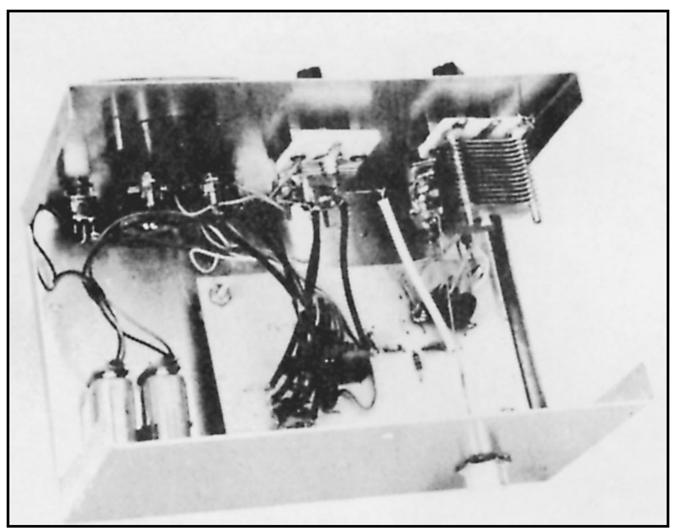


Fig 49—Inside view of the field-strength meter. At the upper right is C1 and to the left, C2. The dark leads from the circuit board to the front panel are the shielded leads described in the text.

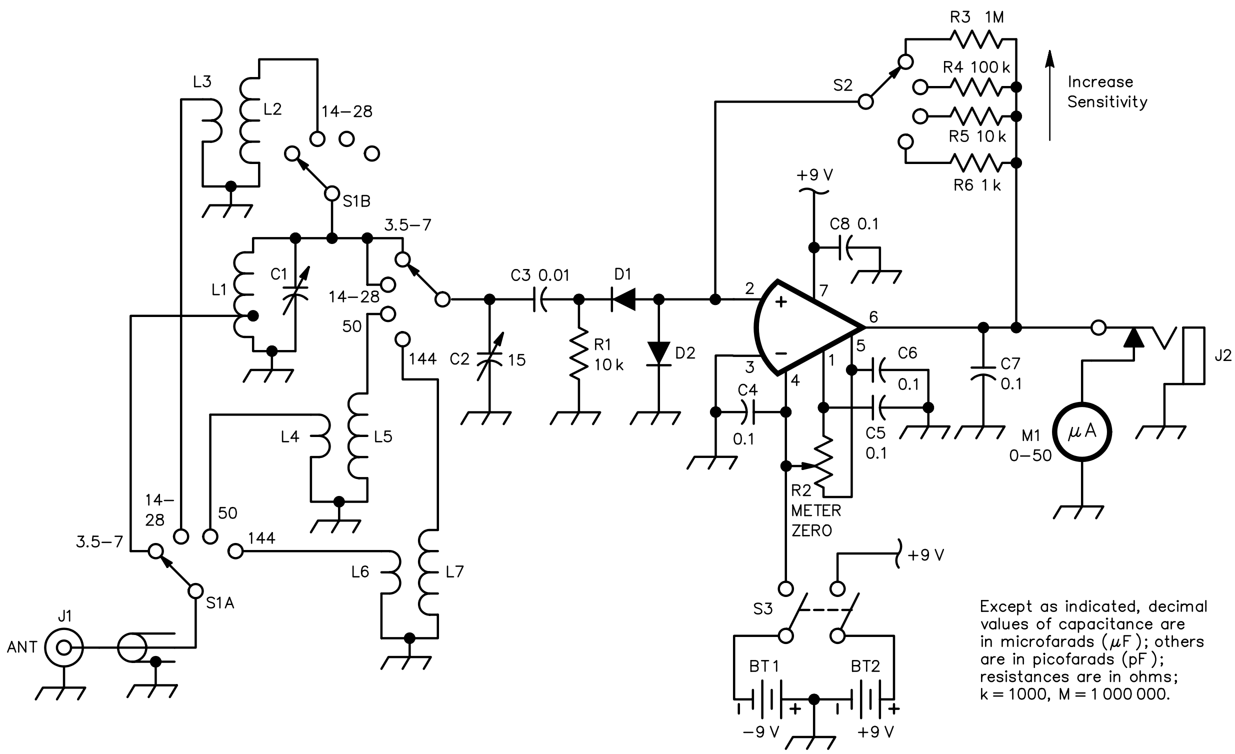


Fig 50—Circuit diagram of the linear field-strength meter. All resistors are 1/4 or 1/2-W composition types.

C1 — 140 pF variable.
C2 — 15-pF variable
D1, D2 — 1N914 or equiv.
L1 — 34 turns #24 enam. wire wound on an Amidon T-68-2 core, tapped 4 turns from ground end.
L2 — 12 turns #24 enam. wire wound on T-68-2 core.
L3 — 2 turns #24 enam. wire wound at ground end of L2.
L4 — 1 turn #26 enam. wire wound at ground end of L5.
L5 — 12 turns #26 enam. wire wound on T-25-12 core.
L6 — 1 turn #26 enam. wire wound at ground end of L7.
L7 — 1 turn #18 enam. wire wound on T-25-12 core.
M1 — 50 or 100 μA dc.
R2 — 10-kΩ control, linear taper.
S1 — Rotary switch, 3 poles, 5 positions, 3 sections.
S2 — Rotary switch, 1 pole, 4 positions.
S3 — DPST toggle.
U1 — Type 741 op amp. Pin nos. shown are for a 14-pin package.

L1 is the 3.5/7 MHz coil and is tuned by C1. The coil is wound on a toroid form. For 14, 21 or 28 MHz, L2 is switched in parallel with L1 to cover the three bands. L5 and C2 cover approximately 40 to 60 MHz, and L7 and C2 from 130 MHz to approximately 180 MHz. The two VHF coils are also wound on toroid forms.

Construction Notes

The majority of the components may be mounted on an etched circuit board. A shielded lead should be used between pin 4 of the IC and S2. The same is true for the leads from R3 through R6 to the switch. Otherwise, parasitic oscillations may occur in the IC because of its very high gain.

In order for the unit to cover the 144-MHz band, L6 and L7 should be mounted directly across the appropriate terminals of S1, rather than on a circuit board. The extra lead length adds too much stray capacitance to the circuit. It isn't necessary to use toroid forms for the 50 and 144-MHz coils. They were used in the version described here simply because they were available. Air-wound coils of the appropriate inductance can be substituted.

Calibration

The field-strength meter can be used "as is" for a relative-reading device. A linear indicator scale will serve admirably. However, it will be a much more useful instrument for antenna work if it is calibrated in decibels, enabling the user to check relative gain and front-to-back ratios. If one has access to a calibrated signal generator, it can be connected to the field-strength meter and different signal levels can be fed to the device for making a calibration chart. Signal-generator voltage ratios can be converted to decibels by using the equation,

$$dB = 20 \log (V1/V2)$$

where

V1/V2 is the ratio of the two voltages
log is the common logarithm (base 10)

Let's assume that M1 is calibrated evenly from 0 to 10. Next, assume we set the signal generator to provide a reading of 1 on M1, and that the generator is feeding a 100-μV signal into the instrument. Now we increase the generator output to 200 μV, giving us a voltage ratio of 2 to 1. Also let's assume M1 reads 5 with the 200-μV input. From the equation above, we find that the voltage ratio of 2 equals 6.02 dB between 1 and 5 on the meter scale. M1 can be calibrated more accurately between 1 and 5 on its scale by adjusting the generator and figuring the ratio. For example, a ratio of 126 μV to 100 μV is 1.26, corresponding to 2.0 dB. By using this method, all of the settings of S2 can be calibrated. In the instrument shown here, the most sensitive setting of S2 with R3, 1 MΩ, provides a range of approximately 6 dB for M1. Keep in mind that the meter scale for each setting of S1 must be calibrated similarly for each band. The degree of coupling of the tuned circuits for the different bands will vary, so each band must be calibrated separately.

Another method for calibrating the instrument is using a transmitter and measuring its output power with an RF wattmeter. In this case we are dealing with power rather than voltage ratios, so this equation applies:

$$dB = 10 \log (P1/P2)$$

where P1/P2 is the power ratio.

With most transmitters the power output can be varied, so calibration of the test instrument is rather easy. Attach a pickup antenna to the field-strength meter (a short wire a foot or so long will do) and position the device in the transmitter antenna field. Let's assume we set the transmitter output for 10 W and get a reading on M1. We note the reading and then increase the output to 20 W, a power ratio of 2. Note the reading on M1 and then use Eq 2. A power ratio of 2 is 3.01 dB. By using this method the instrument can be calibrated on all bands and ranges.

With the tuned circuits and coupling links specified in Fig 50, this instrument has an average range on the various bands of 6 dB for the two most sensitive positions of S2, and 15 dB and 30 dB for the next two successive ranges. The 30-dB scale is handy for making front-to-back antenna measurements without having to switch S2.

An RF Current Probe

The RF current probe of **Figs 51** through 53 operates on the magnetic component of the electromagnetic field, rather than the electric field. Since the two fields are precisely related, as discussed in Chapter 23, the relative field-strength measurements are completely equivalent. The use of the magnetic field offers certain advantages, however. The instrument may be made more compact for the same sensitivity, but its principal advantage is that it may be used near a conductor to measure the current flow without cutting the conductor.

In the average amateur location there may be substantial currents flowing in guy wires, masts and towers, coaxial-cable braids, gutters and leaders, water and gas pipes, and perhaps even drainage pipes. Current may be flowing in telephone and power lines as well. All of these RF currents may have an influence on antenna patterns or be of significance in the case of RFI.

The circuit diagram of the current probe appears in Fig 52, and construction is shown in the photo, Fig 53. The winding data given here apply only to a ferrite rod of the particular dimensions and material specified. Almost any microammeter can be used, but it is usually convenient to use a rather sensitive meter and provide a series resistor to "swamp out" nonlinearity arising from diode conduction characteristics. A control is also used to adjust instrument sensitivity as required during operation. The tuning capacitor may be almost anything that will cover the desired range.

As shown in the photos, the circuit is constructed in a metal box. This enclosure shields the detector circuit from the electric field of the radio wave. A slot must be cut with a hacksaw across the back of the box, and a thin file may be used to smooth the cut. This slot is necessary to prevent the box from acting as a shorted turn.

Fig 51—The RF current probe. The sensitivity control is mounted at the top of the instrument, with the tuning and band switches on the lower portion of the front panel. Frequency calibration of the tuning control was not needed for the intended use of this particular instrument, but marks identifying the various amateur bands would be helpful for general-purpose use. If the unit is provided with a calibrated dial, it can also be used as an absorption wavemeter.

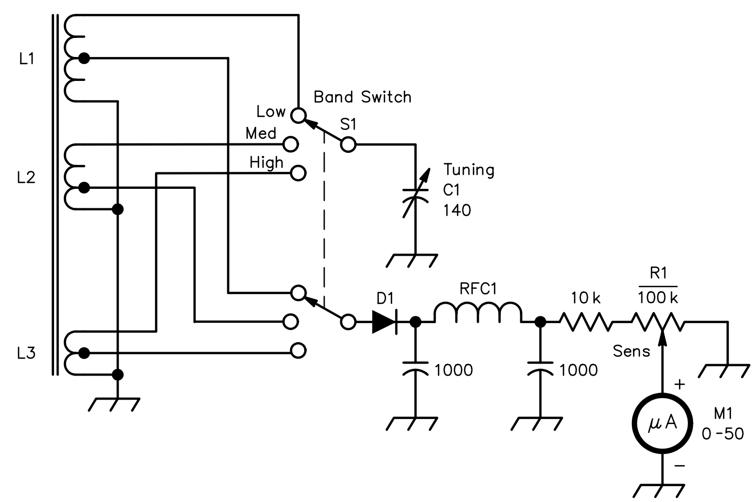


Fig 52—Schematic diagram of the RF current probe. Resistances are in ohms; k = 1000. Capacitances are in picofarads; fixed capacitors are silver mica. Be sure to ground the rotor of C1, rather than the stator, to avoid hand capacitance. L1, L2 and L3 are each close-wound with #22 enam. wire on a single ferrite rod, 4 inches long and 1/2 inch diameter, with μ = 125 (Amidon R61-50-400). Windings are spaced approximately 1/4 inch apart. The ferrite rod, the variable capacitor, and other components may be obtained from Radiokit (see Chapter 21).

C1—Air variable, 6-140 pF; Hammarlund HF140 or equiv.
D1—Germanium diode; 1N34A, 1N270 or equiv.
L1—1.6-5 MHz; 30 turns, tapped at 3 turns from grounded end.
L2—5-20 MHz; 8 turns, tapped at 2 turns from grounded end.
L3—17-39 MHz; 2 turns, tapped at 1 turn.
M1—Any microammeter may be used. The one pictured is a Micronta meter, Radio Shack no. 270-1751.
R1—Linear taper.
RFC1—1 mH; Miller no. 4642 or equiv. Value is not critical.
S1—Ceramic rotary switch, 1 section, 2 poles, 2 to 6 positions; Centralab PA2002 or PA2003 or equiv.

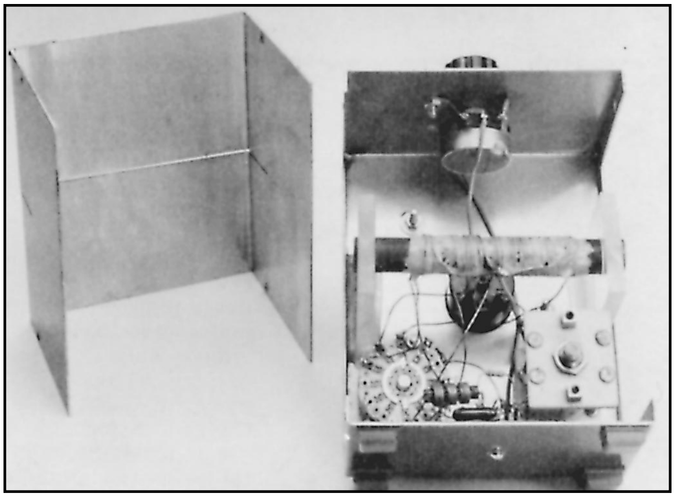


Fig 53—The current probe just before final assembly. Note that all parts except the ferrite rod are mounted on a single half of the 3×4×5-inch Minibox (Bud CU-2105B or equiv.). Rubber grommets are fitted in holes at the ends of the slot to accept the rod during assembly of the enclosure. Leads in the RF section should be kept as short as possible, although those from the rod windings must necessarily be left somewhat long to facilitate final assembly.

Using the Probe

In measuring the current in a conductor, the ferrite rod should be kept at right angles to the conductor, and at a constant distance from it. In its upright or vertical position, this instrument is oriented for taking measurements in vertical conductors. It must be laid horizontal to measure current in horizontal conductors.

Numerous uses for the instrument are suggested in an earlier paragraph. In addition, the probe is an ideal instrument for checking the current distribution in antenna elements. It is also useful for measuring RF ground currents in radial systems. A buried radial may be located easily by sweeping the ground. Current division at junctions may be investigated. “Hot spots” usually indicate areas where additional radials would be effective.

Stray currents in conductors not intended to be part of the antenna system may often be eliminated by bonding or by changing the physical lengths involved. Guy wires and other unwanted “parasitic” elements will often give a tilt to the plane of polarization and make a marked difference in front-to-back ratios. When the ferrite rod is oriented parallel to the electric field lines, there will be a sharp null reading that may be used to locate the plane of polarization quite accurately. When using the meter, remember that the magnetic field is at right angles to the electric field.

The current probe may also be used as a relative signal strength meter. In making measurements on a vertical antenna, the meter should be located at least two wavelengths away, with the rod in a horizontal position. For horizontal antennas, the instrument should be at approximately the same height as the antenna, with the rod vertical.

Antenna Measurements

Of all the measurements made in Amateur Radio systems, perhaps the most difficult and least understood are various measurements of antennas. For example, it is relatively easy to measure the frequency and CW power output of a transmitter, the response of a filter, or the gain of an amplifier. These are all what might be called bench measurements because, when performed properly, all the factors that influence the accuracy and success of the measurement are under control. In making antenna measurements, however, the "bench" is now perhaps the backyard. In other words, the environment surrounding the antenna can affect the results of the measurement. Control of the environment is not at all as simple as it was for the bench measurement, because now the work area may be rather spacious. This section describes antenna measurement techniques which are closely allied to those used in an antenna measuring event or contest. With these procedures the measurements can be made successfully and with meaningful results. These techniques should provide a better understanding of the measurement problems, resulting in a more accurate and less difficult task. The information in this section was provided by Dick Turrin, W2IMU, and originally published in November 1974 *QST*.

SOME BASIC IDEAS

An antenna is simply a transducer or coupler between a suitable feed line and the environment surrounding it. In addition to efficient transfer of power from feed line to environment, an antenna at VHF or UHF is most frequently required to concentrate the radiated power into a particular region of the environment.

To be consistent in comparing different antennas, it is necessary that the environment surrounding the antenna be standardized. Ideally, measurements should be made with the measured antenna so far removed from any objects causing environmental effects that it is literally in outer space—a very impractical situation. The purpose of the measurement techniques is therefore to simulate, under practical conditions, a controlled environment. At VHF and UHF, and with practical-size antennas, the environment *can* be controlled so that successful and accurate measurements can be made in a reasonable amount of space.

The electrical characteristics of an antenna that are most desirable to obtain by direct measurement are: (1) gain (relative to an isotropic source, which by definition has a gain of unity); (2) space-radiation pattern; (3) feed-point impedance (mismatch) and (4) polarization.

Polarization

In general the polarization can be assumed from the geometry of the radiating elements. That is to say, if the antenna is made up of a number of linear elements (straight lengths of rod or wire which are resonant and connected to the feed point) the polarization of the electric field will be linear and polarized parallel to the elements. If the elements are not consistently parallel with each other, then the polarization cannot easily be assumed. The following techniques are directed to antennas having polarization that is essentially linear (in one plane), although the method can be extended to include all forms of elliptic (or mixed) polarization.

Feed-Point Mismatch

The feed-point mismatch, although affected to some degree by the immediate environment of the antenna, does *not* affect the gain or radiation characteristics of an antenna. If the immediate environment of the antenna does not affect the feed-point impedance, then any mismatch intrinsic to the antenna tuning reflects a portion of the incident power back to the source. In a receiving antenna this reflected power is reradiated back into the environment, "free space," and can be lost entirely. In a transmitting antenna, the reflected power goes back to the final amplifier of the transmitter if it is not matched.

In general an amplifier by itself is *not* a matched source to the feed line, and, if the feed line has very low loss, the amplifier output controls are customarily altered during the normal tuning procedure to obtain maximum power transfer to the antenna. The power which has been reflected from the antenna combines with the source power to travel again to the antenna. This procedure is called conjugate matching, and the feed line is now part of a resonant system consisting of the mismatched antenna, feed line, and amplifier tuning circuits. It is therefore possible to use a mismatched antenna to its full gain potential, provided the mismatch is not so severe as to cause heating losses in the system, especially the feed line and matching devices. (See also the discussion of additional loss caused by SWR in Chapter 24.) Similarly, a mismatched receiving antenna may be conjugately matched into the receiver front end for maximum power transfer. In any case it should be clearly kept in mind that the feed-point mismatch does *not* affect the radiation characteristics of an antenna. It can only affect the system efficiency wherein heating losses are concerned.

Why then do we include feed-point mismatch as part of the antenna characteristics? The reason is that for efficient system performance, most antennas are resonant transducers and present a reasonable match over a relatively narrow frequency range. It is therefore desirable to design an antenna, whether it be a simple dipole or an array of Yagis, such that the final single feed-point impedance be essentially resistive and of magnitude consistent with the impedance of the feed line which is to be used. Furthermore, in order to make accurate, absolute gain measurements, it is vital that the antenna under test accept all the power from a matched-source generator, or that the reflected power caused by the mismatch be measured and a suitable error correction for heating losses be included in the gain calculations. Heating losses may be determined from information contained in Chapter 24.

While on the subject of feed-point impedance, mention should be made of the use of baluns in antennas. A balun is simply a device which permits a lossless transition between a balanced system—feed line or antenna—and an unbalanced feed line or system. If the feed point of an antenna is symmetric such as with a dipole and it is desired to feed this antenna with an unbalanced feed line such as coax, it is necessary to provide a balun between the line and the feed point. Without the balun, current will be allowed to flow on the outside of the coax. The current on the outside of the feed line will cause radiation and thus the feed line becomes part of the antenna radiation system. In the case of beam antennas where it is desired to concentrate the radiated energy is a specific direction, this extra radiation from the feed line will be detrimental, causing distortion of the expected antenna pattern.

ANTENNA TEST SITE SET-UP AND EVALUATION

Since an antenna is a reciprocal device, measurements of gain and radiation patterns can be made with the test antenna used either as a transmitting or as a receiving antenna. In general and for practical reasons, the test antenna is used in the receiving mode, and the source or transmitting antenna is located at a specified fixed remote site and unattended. In other words the source antenna, energized by a suitable transmitter, is simply required to illuminate or flood the receiving site in a controlled and constant manner.

As mentioned earlier, antenna measurements ideally should be made under "free-space" conditions. A further restriction is that the illumination from the source antenna be a plane wave over the effective aperture (capture area) of the test antenna. A plane wave by definition is one in which the magnitude and phase of the fields are uniform, and in the test-antenna situation, *uniform over the effective area plane of the test antenna*. Since it is the nature of all radiation to expand in a spherical manner at great distance from the source, it would seem to be most desirable to locate the source antenna as far from the test site as possible. However, since for practical reasons the test site and source location will have to be near the Earth and not in outer space, the environment must include the effects of the ground surface and other obstacles in the vicinity of both antennas. These effects almost always dictate that the test range (spacing between source and test antennas) be as short as possible consistent with maintaining a nearly error-free plane wave illuminating the test *aperture*.

A nearly error-free plane wave can be specified as one in which the phase and amplitude, from center to edge of the illuminating field over the test aperture, do not deviate by more than about 30° and 1 dB, respectively. These conditions will result in a gain-measurement error of no more than a few percent less that the true gain. Based on the 30° phase error alone, it can be shown that the minimum range distance is approximately

$$S_{min} = 2\frac{D^2}{\lambda}$$

where D is the largest aperture dimension and λ is the free-space wavelength in the same units as D. The phase error over the aperture D for this condition is 1/16 wavelength.

Since aperture size and gain are related by

$$\text{Gain} = \frac{4\pi A_e}{\lambda^2}$$

where A_e is the effective aperture area, the dimension D may be obtained for simple aperture configurations. For a square aperture

$$D^2 = G\frac{\lambda^2}{4\pi}$$

which results in a minimum range distance for a square aperture of

$$S_{min} = G\,\frac{\lambda}{2\pi}$$

and for a circular aperture of

$$S_{min} = G\,\frac{2\lambda}{\pi^2}$$

For apertures with a physical area that is not well defined or is much larger in one dimension that in other directions, such as a long thin array for maximum directivity in one plane, it is advisable to use the maximum estimate of D from either the expected gain or physical aperture dimensions.

Up to this point in the range development, only the conditions for minimum range length, S_{min}, have been established, as though the ground surface were not present. This minimum S is therefore a necessary condition even under "free-space" environment. The presence of the ground further complicates the range selection, not in the determination of S but in the exact location of the source and test antennas above the earth.

It is always advisable to select a range whose intervening terrain is essentially flat, clear of obstructions, and of uniform surface conditions, such as all grass or all pavement. The extent of the range is determined by the illumination of the source antenna, usually a beam, whose gain is no greater than the highest gain antenna to be measured. For gain measurements the range consists essentially of the region in the beam of the test antenna. For radiation-pattern measurements, the range is considerably larger and consists of all that area illuminated by the source antenna, especially around and behind the test site. Ideally a site should be chosen where the test-antenna location is near the center of a large open area and the source antenna located near the edge where most of the obstacles (trees, poles, fences, etc) lie.

The primary effect of the range surface is that some of the energy from the source antenna will be reflected into the test antenna while other energy will arrive on a direct line-of-sight path. This is illustrated in **Fig 54**. The use of a flat, uniform ground surface assures that there will be essentially a

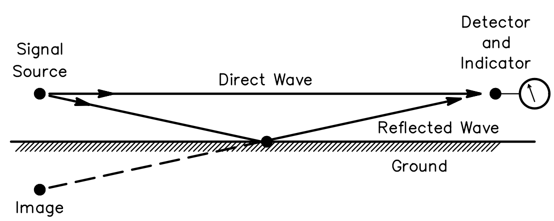


Fig 54—On an antenna test range, energy reaching the receiving equipment may arrive after being reflected from the surface of the ground, as well as by the direct path. The two waves may tend to cancel each other, or may reinforce one another, depending on their phase relationship at the receiving point.

mirror reflection even though the reflected energy may be slightly weakened (absorbed) by the surface material (ground). In order to perform an analysis it is necessary to realize that horizontally polarized waves undergo a 180° phase reversal upon reflection from the earth. The resulting illumination amplitude at any point in the test aperture is the vector sum of the electric fields arriving from the two directions, the direct path and the reflected path. If a perfect mirror reflection is assumed from the ground (it is nearly that for practical ground conditions at VHF/UHF) and the source antenna is isotropic, radiating equally in all directions, then a simple geometric analysis of the two path lengths will show that at various points in the vertical plane at the test-antenna site the waves will combine in different phase relationships. At some points the arriving waves will be in phase, and at other points they will be 180° out of phase. Since the field amplitudes are nearly equal, the resulting phase change caused by path length difference will produce an amplitude variation in the vertical test-site direction similar to a standing wave, as shown in **Fig 55**.

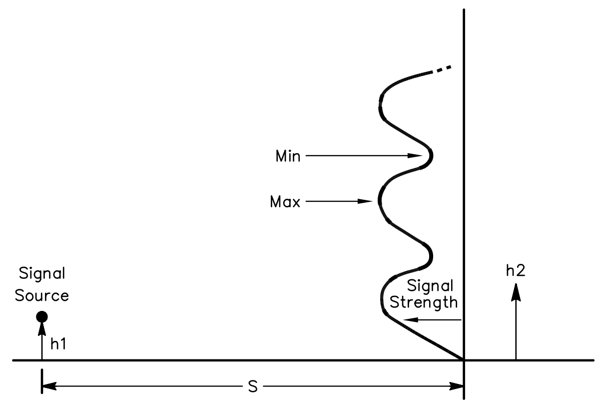


Fig 55—The vertical profile, or plot of signal strength versus test-antenna height, for a fixed height of the signal source above ground and at a fixed distance. See text for definitions of symbols.

The simplified formula relating the location of h2 for maximum and minimum values of the two-path summation in terms of h1 and S is

$$h2 = n\frac{\lambda}{4}\cdot\frac{S}{h1}$$

with n = 0, 2, 4, . . . for minimums and
n = 1, 3, 5, . . . for maximums, and S is much larger than either h1 or h2.

The significance of this simple ground reflection formula is that it permits the approximate location of the source antenna to be determined to achieve a nearly plane-wave amplitude distribution *in the vertical direction* over a particular test *aperture size*. It should be clear from examination of the height formula that as h1 is decreased, the vertical distribution pattern of signal at the test site, h2, expands. Also note that the signal level for h2 equal to zero is always zero on the ground regardless of the height of h1.

The objective in using the height formula then is, given an effective antenna aperture to be illuminated from which a minimum S (range length) is determined and a suitable range site chosen, to find a value for h1 (source antenna height). The required value is such that the *first* maximum of vertical distribution at the test site, h2, is at a practical distance above the ground and at the same time the signal amplitude over the aperture in the vertical direction does not vary more than about 1 dB. This last condition is not sacred but is closely related to the particular antenna under test. In practice these formulas are useful only to initialize the range set-up. A final check of the vertical distribution at the test site must be made by direct measurement. This measurement should be conducted with a small low-gain but unidirectional probe antenna such as a corner reflector or two-element Yagi that is moved along a vertical line over the intended aperture site. Care should be exercised to minimize the effects of local environment around the probe antenna and that the beam of the probe be directed at the source antenna at all times for maximum signal. A simple dipole is undesirable as a probe antenna because it is susceptible to local environmental effects.

The most practical way to instrument the vertical distribution measurement is to construct some kind of vertical track, preferably of wood, with a sliding carriage or platform which may be used to support and move the probe antenna. It is assumed of course that a stable source transmitter and calibrated receiver or detector are available so variations of the order of 1/2 dB can be clearly distinguished.

Once these initial range measurements are completed successfully, the range is now ready to accommodate any aperture size less in vertical extent than the largest for which S_{min} and the vertical field distribution were selected. The test antenna is placed with the center of its aperture at the height h2 where maximum signal was found. The test antenna should be tilted so that its main beam is pointed in the direction of the source antenna. The final tilt is found by observing the receiver output for maximum signal. This last process must be done empirically since the apparent location of the source is somewhere between the actual source and its image, below the ground.

An example will illustrate the procedure. Assume that we wish to measure a 7-foot diameter parabolic reflector antenna at 1296 MHz (λ = 0.75 foot). The minimum range distance, S_{min}, can be readily computed from the formula for a circular aperture.

$$S_{min} = 2\frac{D^2}{\lambda} = 2 \times \frac{49}{0.75} = 131 \text{ ft}$$

Now a suitable site is selected based on the qualitative discussion given before.

Next determine the source height, h1. The procedure is to choose a height h1 such that the first minimum above ground (n = 2 in formula) is at least two or three times the aperture size, or about 20 feet.

$$h1 = n\frac{\lambda}{4}\frac{S}{h2} = 2 \times \frac{0.75}{4} \times \frac{131}{20} = 2.5 \text{ ft}$$

Place the source antenna at this height and probe the vertical distribution over the 7-foot aperture location, which will be about 10 feet off the ground.

$$h2 = n\frac{\lambda}{4}\frac{S}{h1} = 1 \times \frac{0.75}{4} \times \frac{131}{2.5} = 9.8 \text{ ft}$$

The measured profile of vertical signal level versus height should be plotted. From this plot, empirically determine whether the 7-foot aperture can be fitted in this profile such that the 1- dB variation is not exceeded. If the variation exceeds 1 dB over the 7-foot aperture, the source antenna should be lowered and h2 raised. Small changes in h1 can quickly alter the distribution at the test site. **Fig 56** illustrates the points of the previous discussion.

The same set-up procedure applies for either horizontal or vertical linear polarization. However, it is advisable to check by direct measurement at the site for each polarization to be sure that the vertical distribution is satisfactory. Distribution probing in the horizontal plane is unnecessary as little or no variation in amplitude should be found, since the reflection geometry is constant. Because of this, antennas with apertures which are long and thin, such as a stacked collinear vertical, should be measured with the long dimension parallel to the ground.

A particularly difficult range problem occurs in measurements of antennas which have depth as well as cross- sectional aperture area. Long end-fire antennas such as long Yagis, rhombics, V-beams, or arrays of these antennas, radiate as volumetric arrays and it is therefore even more essential that the illuminating field from the source antenna be reasonably uniform in depth as well as plane wave in cross section. For measuring these types of antennas it is advisable to make several vertical profile measurements which cover the depth of the array. A qualitative check on the integrity of the illumination for long end-fire antennas can be made by moving the array or antenna axially (forward and backward) and noting the change in received signal level. If the signal level

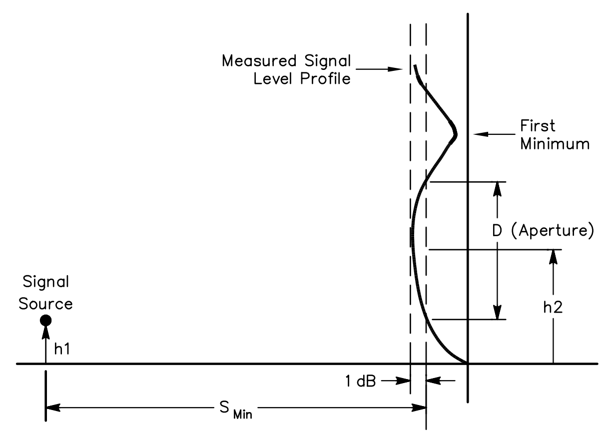


Fig 56—Sample plot of a measured vertical profile.

varies less than 1 or 2 dB for an axial movement of several wavelengths then the field can be considered satisfactory for most demands on accuracy. Large variations indicate that the illuminating field is badly distorted over the array depth and subsequent measurements are questionable. it is interesting to note in connection with gain measurements that any illuminating field distortion will always result in measurements that are lower than true values.

ABSOLUTE GAIN MEASUREMENT

Having established a suitable range, the measurement of gain relative to an isotropic (point source) radiator is almost always accomplished by direct comparison with a calibrated standard-gain antenna. That is, the signal level with the test antenna in its optimum location is noted. Then the test antenna is removed and the standard-gain antenna is placed with its aperture at the center of location where the test antenna was located. The difference in signal level between the standard and the test antennas is measured and appropriately added to or subtracted from the gain of the standard-gain antenna to obtain the absolute gain of the test antenna, absolute here meaning with respect to a point source which has a gain of unity by definition. The reason for using this reference rather than a dipole, for instance, is that it is more useful and convenient for system engineering. It is assumed that both standard and test antennas have been carefully matched to the appropriate impedance and an accurately calibrated and matched detecting device is being used.

A standard-gain antenna may be any type of unidirectional, preferably planar-aperture, antenna, which has been calibrated either by direct measurement or in special cases by accurate construction according to computed dimensions. A standard-gain antenna has been suggested by Richard F. H. Yang (see Bibliography). Shown in **Fig 57**, it consists of two in-phase dipoles $^1/_2\ \lambda$ apart and backed up with a ground plane 1 λ square.

In Yang's original design, the stub at the center is a balun formed by cutting two longitudinal slots of $^1/_8$-inch width, diametrically opposite, on a $^1/_4$-λ section of $^7/_8$-inch rigid 52-Ω coax. An alternative method of feeding is to feed RG-8 or RG-213 coax through slotted $^7/_8$-inch copper tubing. Be sure to leave the outer jacket on the coax to insulate it from the copper-tubing balun section. When constructed accurately to scale for the frequency of interest, this type of standard will have an absolute gain of 7.7 dBd (dB gain over a dipole) with an accuracy of plus or minus 0.25 dB.

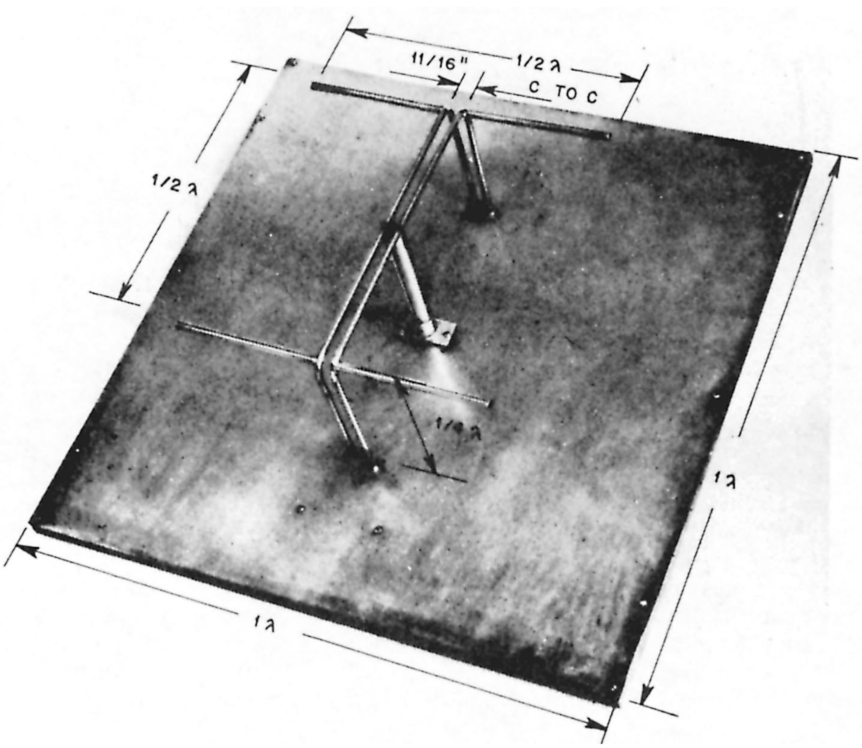


Fig 57—Standard-gain antenna. When accurately constructed for the desired frequency, this antenna will exhibit a gain of 7.7 dB over a dipole radiator, plus or minus 0.25 dB. In this model, constructed for 432 MHz, the elements are $^3/_8$-inch diameter tubing. The phasing and support lines are of $^5/_{16}$-inch diameter tubing or rod.

RADIATION-PATTERN MEASUREMENTS

Of all antenna measurements, the radiation pattern is the most demanding in measurement and most difficult to interpret. Any antenna radiates to some degree in all directions into the space surrounding it. Therefore, the radiation pattern of an antenna is a three-dimensional representation of the magnitude, phase and polarization. In general, and in practical cases for Amateur Radio communications, the polarization is well defined and only the magnitude of radiation is important. Furthermore, in many of these cases the radiation in one particular plane is of primary interest, usually the plane corresponding to that of the earth's surface, regardless of polarization.

Because of the nature of the range setup, measurement of radiation pattern can be successfully made only in a plane nearly parallel to the earth's surface. With beam antennas it is advisable and usually sufficient to take two radiation pattern measurements, one in the polarization plane and one at right angles to the plane of polarization. These radiation patterns are referred to in antenna literature as the principal E-plane and H-plane patterns, respectively, E plane meaning parallel to the electric field which is the polarization plane and H plane meaning parallel to the magnetic field. The electric field and magnetic field are always perpendicular to each other in a plane wave as it propagates through space.

The technique in obtaining these patterns is simple in procedure but requires more equipment and patience than does making a gain measurement. First, a suitable mount is required which can be rotated in the azimuth plane (horizontal) with some degree of accuracy in terms of azimuth angle positioning. Second, a signal-level indicator calibrated over at least a 20-dB dynamic range with a readout resolution of at least 2 dB is required. A dynamic range of up to about 40 dB would be desirable but does not add greatly to the measurement significance.

With this much equipment, the procedure is to locate first the area of maximum radiation from the beam antenna by carefully adjusting the azimuth and elevation positioning. These settings are then arbitrarily assigned an azimuth angle of zero degrees and a signal level of zero decibels. Next, without changing the elevation setting (tilt of the rotating axis), the antenna is carefully rotated in azimuth in small steps which permit signal-level readout of 2 or 3 dB per step. These points of signal level corresponding with an azimuth angle are recorded and plotted on polar coordinate paper. A sample of the results is shown on ARRL coordinate paper in **Fig 58**.

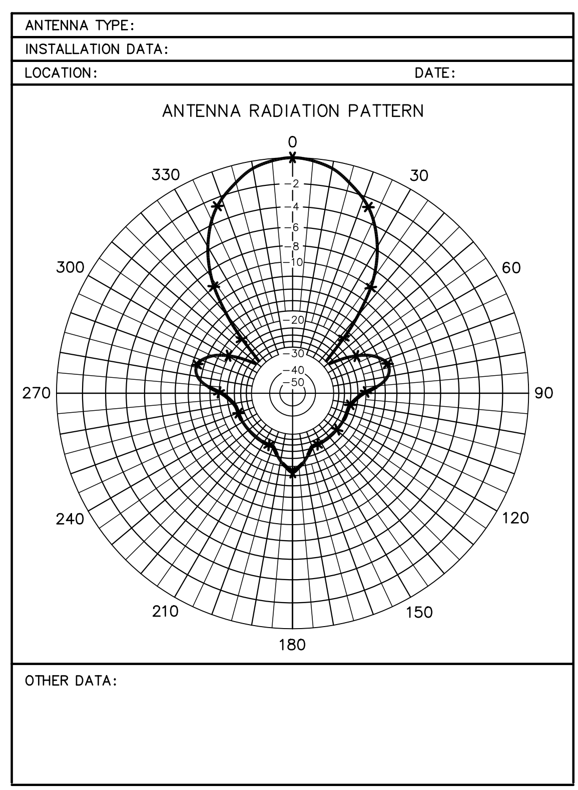


Fig 58—Sample plot of a measured radiation pattern, using techniques described in the text. The plot is on coordinate paper available from ARRL HQ. The form provides space for recording significant data and remarks.

On the sample radiation pattern the measured points are marked with an X and a continuous line is drawn in, since the pattern is a continuous curve. Radiation patterns should preferably be plotted on a logarithmic radial scale, rather than a voltage or power scale. The reason is that the log scale approximates the response of the ear to signals in the audio range. Also many receivers have AGC systems that are somewhat logarithmic in response; therefore the log scale is more representative of actual system operation.

Having completed a set of radiation-pattern measurements, one is prompted to ask, "Of what use are they?" The primary answer is as a diagnostic tool to determine if the antenna is functioning as it was intended to. A second answer is to know how the antenna will discriminate against interfering signals from various directions.

Consider now the diagnostic use of the radiation patterns. If the radiation beam is well defined, then there is an approximate formula relating the antenna gain to the measured half-power beamwidth of the E and H-plane radiation patterns. The half-power beamwidth is indicated on the polar plot where the radiation level falls to 3 dB below the main beam 0-dB reference on either side. The formula is

$$\text{Gain} \cong \frac{41,253}{\theta_E \phi_H}$$

where θ_E and ϕ_H are the half-power beamwidths in degrees of the E and H-plane patterns, respectively. This equation assumes a lossless antenna system.

To illustrate the use of this equation, assume that we have a Yagi antenna with a boom length of two wavelengths. From known relations (described in Chapter 11) the expected gain of a Yagi with a boom length of two wavelengths is about 12 dB; its gain, G, equals 15.8. Using the above relationship, the product of $\theta_E \times \phi_H \cong 2600$ square degrees. Since a Yagi produces a nearly symmetric beam shape in cross section, $\theta_E \cong \phi_H = 51°$. Now if the measured values of θ_E and ϕ_H are much larger than 51°, then the gain will be much lower than the expected 12 dB.

As another example, suppose that the same antenna (a 2- wavelength-boom Yagi) gives a measured gain of 9 dB but the radiation pattern half- power beamwidths are approximately 51°. This situation indicates that although the radiation patterns seem to be correct, the low gain shows inefficiency somewhere in the antenna, such as lossy materials or poor connections.

Large broadside collinear antennas can be checked for excessive phasing-line losses by comparing the gain computed from the radiation patterns with the direct-measured gain. It seems paradoxical, but it is indeed possible to build a large array with a very narrow beamwidth indicating high gain, but actually having very low gain because of losses in the feed distribution system.

In general, and for most VHF/UHF Amateur Radio communications, gain is the primary attribute of an antenna. However, radiation in other directions than the main beam, called sidelobe radiation, should be examined by measurement of radiation patterns for effects such as nonsymmetry on either side of the main beam or excessive magnitude of sidelobes. (Any sidelobe which is less than 10 dB below the main beam reference level of 0 dB should be considered excessive.) These effects are usually attributable to incorrect phasing of the radiating elements or radiation from other parts of the antenna which was not intended, such as the support structure or feed line.

The interpretation of radiation patterns is intimately related to the particular type of antenna under measurement. Reference data should be consulted for the antenna type of interest, to verify that the measured results are in agreement with expected results.

To summarize the use of pattern measurements, if a beam antenna is first checked for gain (the easier measurement to make) and it is as expected, then pattern measurements may be academic. However, if the gain is lower than expected it is advisable to make the pattern measurements as an aid in determining the possible cause of low gain.

Regarding radiation pattern measurements, remember that the results measured under proper range facilities will not necessarily be the same as observed for the same antenna at a home-station installation. The reasons may be obvious now in view of the preceding information on the range setup, ground reflections, and the vertical-field distribution profiles. For long paths over rough terrain where many large obstacles may exist, the effects of ground reflection tend to become diffused, although they still can cause unexpected results. For these reasons it is usually unjust to compare VHF/UHF antennas over long paths.

BIBLIOGRAPHY

Source material and more extended discussion of topics covered in this chapter can be found in the references given below.

A. Bailey, "The Antenna Lab, Parts 1 and 2," *Radio Communication*, Aug and Sep 1983.

L. Blake, *Transmission Lines and Waveguides* (New York: John Wiley & Sons, 1969), pp 244-251.

J. H. Bowen, "A Calorimeter for VHF and UHF Power Measurements, *QST*, Dec 1975, pp 11-13.

W. Bruene, "An Inside Picture of Directional Wattmeters," *QST*, April 1959.

J. Carr, "Fond Fault with Your Coax," *73*, Oct 1984, pp 10-14.

D. DeMaw, "In-Line RF Power Metering," *QST*, December 1969.

D. Fayman, "A Simple Computing SWR Meter," *QST*, Jul 1973, pp 23-33.

J. Gibbons and H. Horn, "A Circuit With Logarithmic Response Over Nine Decades," *IEEE Transactions on Circuit Theory*, Vol CT-11, No. 3, Sep 1964, pp 378-384.

J. Grebenkemper, "Calibrating Diode Detectors," *QEX*, Aug 1990.

J. Grebenkemper, "The Tandem Match—An Accurate Directional Wattmeter," *QST*, Jan 1987.

J. Grebenkemper, "Improving and Using R-X Noise Bridges," *QST*, Aug 1989; feedback, *QST*, Jan 1990, p 27.

T. King, "A Practical Time-Domain Reflectometer," *QST*, May 1989, pp 22-24.

Z. Lau and C. Hutchinson, "Improving the HW-9 Transceiver," *QST*, Apr 1988, pp 26-29.

V. G. Leenerts, "Automatic VSWR and Power Meter," *Ham Radio*, May 1980, pp 34-43.

J. Lenk, *Handbook of Oscilloscopes* (Englewood Cliffs, NJ: Prentice-Hall, 1982), pp 288-292.

L. McCoy, "A Linear Field-Strength Meter," *QST*, January, 1973.

T. McMullen, "The Line Sampler, an RF Power Monitor for VHF and UHF," *QST*, April 1972.

C. Michaels, "Determining Line Lengths," Technical Correspondence, *QST*, Sep 1985, pp 43-44.

H. Perras, "Broadband Power-Tracking VSWR Bridge," *Ham Radio*, Aug 1979, pp 72-75.

S. Ramo, J. Whinnery and T. Van Duzer, *Fields and Waves in Communication Electronics* (New York: John Wiley & Sons, 1967), Chap 1.

J. Sevick, "Short Ground-Radial Systems for Short Verticals," *QST*, Apr 1978, pp 30-33.

J. Sevick, "Measuring Soil Conductivity," *QST*, Mar 1981, pp 38- 39.

W. Spaulding, "A Broadband Two-Port S-Parameter Test Set," *Hewlett-Packard Journal*, Nov 1984.

D. Turrin, "Antenna Performance Measurements," *QST*, Nov 1974, pp 35-41.

F. Van Zant, "High-Power Operation with the Tandem Match Directional Coupler," Technical Correspondence, *QST*, Jul 1989, pp 42-43.

R. F. H. Yang, "A Proposed Gain Standard for VHF Antennas," *IEEE Transactions on Antennas and Propagation*, Nov 1966.