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Chapter 27

Antenna and Transmission Line Measurements

The principal quantities measured on transmission lines are current or voltage, including phase. From these measurements, forward and reflected power and standing wave ratio (SWR) may be obtained. For antennas, the primary measurement of interest to amateurs is field strength in order to determine an antenna's radiation pattern or to compare relative antenna performance. It is important to note that 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 1:1 is all you need for most impedance-matching adjustments.

Absolute quantitative measurements of amplitude or time (phase) become increasingly difficult at frequencies above a few MHz with numerous sources of error becoming more and more significant. Quantitative measurements of reasonable accuracy demand good design and careful construction in the measuring instruments. They also require intelligent use of the equipment, including knowledge not only of its limitations but also of stray effects in the instrument and also of the test configuration that often lead to false results. Until you know the complete conditions of the measurements, a certain amount of skepticism regarding numerical data resulting from amateur measurements with simple equipment is justified.

Accurate measurement of SWR, for example, 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. If such measurements are required, surplus lab equipment of very high quality is commonly available although usually not "in calibration."

On the other hand, purely qualitative or relative measurements, such as comparing one antenna to another, before-and-after, or max/min adjustments are easy to make and quite useful. This chapter presents methods and devices for making these measurements.

27.1 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 you adjust the transmitter output circuits for maximum current or voltage into the transmission line. Although a 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.

27.1.1 RF VOLTMETERS

You can combine a germanium or Schottky diode in conjunction with a low-range milliammeter and a few resistors to form an RF voltmeter suitable for connecting across the two conductors of a coaxial line, as shown in **Figure 27.1**. It

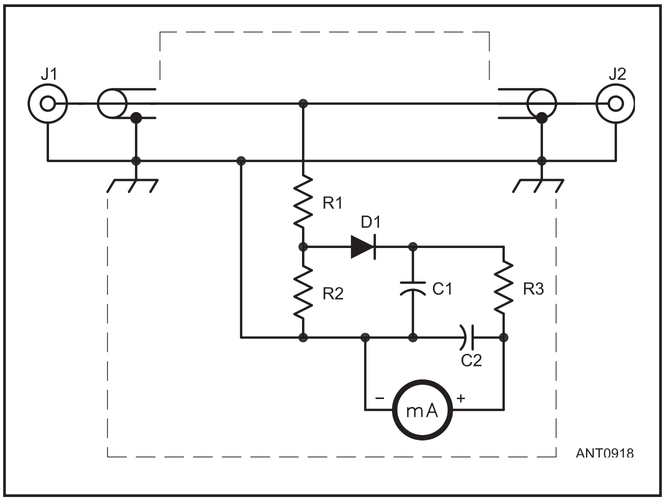


Figure 27.1 — RF voltmeter for coaxial line.
C1, C2 — 0.005- or 0.01-µF disc ceramic.
D1 — 1N34A germanium or 1N5817 Schottky diode.
J1, J2 — Coaxial fittings, chassis-mounting type.
M1 — 0-1 milliammeter (more sensitive meter may be used if desired; see text).
R1 — 6.8 kΩ, metal-oxide, 1 W for each 100 W of RF power.
R2 — 680 Ω, ½ or 1 W carbon-film or metal-oxide.
R3 — 10 kΩ, ½ W (see text).

consists of a voltage divider, 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.

You may construct the voltmeter in a small metal box, indicated by the dashed line in the drawing, and fitted with coax receptacles. R1 and R2 should be noninductive resistors such as carbon-film or metal-oxide types. 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 the 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. R1 may be a variable resistor so the sensitivity can be adjusted for various power levels.

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 ½ inch or more from metal surfaces parallel to the body of the resistor. These precautions must be observed if consistent measurements are to be obtained above a few 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

You may calibrate the meter for RF voltage at low frequencies by comparison with a standard such as an RF ammeter, wattmeter or oscilloscope. If a wattmeter is used, the line must be well matched so the impedance at the point of measurement is equal to the actual Z_0 of the line. The power can be calculated as $E = \sqrt{P Z_0}$. By making voltage measurements at a number of different power levels, you can obtain enough points to draw a calibration curve for your particular setup. Be advised that stray effects and nonlinearities inherent in this simple circuit make a true calibration questionable above a few MHz.

27.1.2 RF CURRENT METERS

The following project was designed by Tom Rauch, W8JI (**w8ji.com/building_a_current_meter.htm**). The circuit of **Figure 27.2** is based on a current transformer (T1) consisting of a T157-2 powdered-iron toroid core with a 20-turn winding. The meter is used with the current-carrying wire or antenna inserted through the middle of the core as a one-turn primary

When 1 A is flowing in the single-turn primary, the secondary current will be 50 mA (equal to primary current

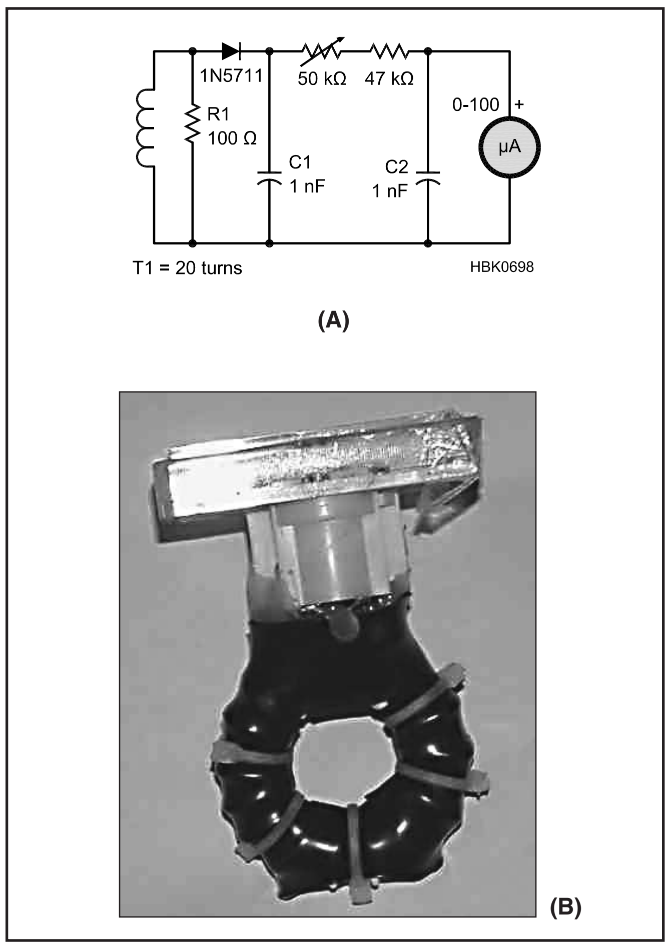


Figure 27.2 — The schematic of the RF current probe (A) and assembly of the RF current probe (B). Use an all-plastic meter and mount the circuits and toroid directly on the back of the meter case.

divided by the turns ratio of 20:1). R1 across the transformer flattens the frequency response and limits the output voltage. The RF voltage is then detected and filtered by D1 (a low-threshold Schottky diode for minimum voltage drop) and C1. The adjustable sum of R2 and R3 allow for full-scale (FS) calibration of the 100 µA meter. C2 provides additional filtering. The toroid core and all circuitry are glued to the back of the meter case with only R2 exposed — a screwdriver-adjustable calibration pot.

It is important to minimize stray capacitance by using a meter with all-plastic construction except for the electrical parts. The meter in Figure 27.2B has an all-plastic case including the meter scale. The meter movement and all metallic areas are small. The lack of large metallic components minimizes stray capacitance from the proximity of the meter. Low stray capacitance ensures the instrument has the least possible effect on the circuit being tested.

A value of 100 Ω for R1 gave the flattest response from 1.8 to 30 MHz. With 50 mA of secondary current, the voltage across R1 is $0.05 \times 100 = 5\ V_{RMS}$. The peak voltage is then $1.414 \times 5 = 7.1$ V. At full current, power dissipation in R1 = 50 mA × 5 V_{RMS} = 0.25 W so a ½-W or larger resistor should be used.

The meter used here was a 10,000 Ω/V model so for full-scale deflection from a primary current of 1 A producing a secondary voltage of ~7 V, the sum of R2 and R3 must be set to 7 × 10,000 = 70 kΩ. The low-current meter combined with high detected voltage improves detector linearity.

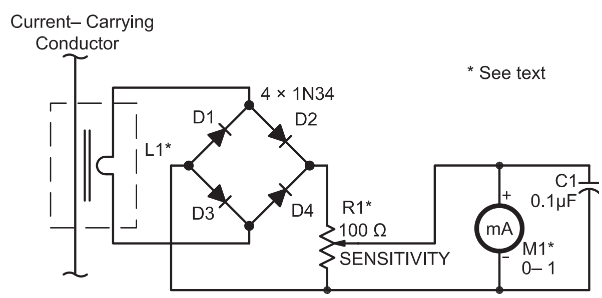
Calibration of the meter can be performed by using a calibrated power meter and a test fixture consisting of two RF connectors with a short piece of wire between them and through the transformer core. With 50 W applied to a 50-Ω load, the wire will be carrying 1 A of current. Full-scale accuracy is not required in comparison measurements, since the meter references against itself, but linearity within a few percent is important.

This transformer-based meter is much more reliable and linear than thermocouple RF ammeters and perturbs systems much less. Stray capacitance added to the system being tested is very small because of the proximity of the meter and the compact wiring area. Compared to actually connecting a meter with its associated lead lengths and capacitance in line with the load, the advantages of a transformer-coupled meter become apparent.

Clamp-on RF Current Probe

Sometimes it is not practical to disconnect a wire in order to sense RF current on it, such as a power cord or speaker wire. In such cases a *clamp-on* probe can be used as described in the February 1999 *QST* article by Steve Sparks, N5SV, and shown in **Figure 27.3**. The core is a split-core type — any common material will suffice (type 31, 75, 61, 43, etc) for HF use. If the enclosure is hand-held size, the instrument can be used as a handy detector and "sniffer" for RFI trouble-shooting. Because the split core does not close completely and consistently every time, this is not a precision instrument but is effective for relative comparison.

(A)



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(B)

Figure 27.3 — The RF current probe at A is small enough to be hand-carried as an effective RF "sniffer." At B is the schematic of the RF current probe. Use a metal enclosure for the probe.
C1 — 0.1 µF disc ceramic.
D1-D4 — 1N34A germanium or 1N5817 Schottky diodes.
L1 — Single turn of #14 AWG wire through a snap-on ferrite split-core, type 31, 43, 61, 73, or 75 material will work. Glue core to top of metal enclosure.
M1 — 1 mA analog meter. Substitute lower full-scale current for higher sensitivity.
R1 — 100 to 500 Ω panel-mount potentiometer.

27.1.3 RF AMMETERS

An RF ammeter is a good way to gauge output power. You can mount an RF ammeter in any convenient location at the input end of the transmission line, the principal precaution 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 installing a metal-case instrument on a metal panel, you should mount it on a separate sheet of insulating material so that there is ⅛ 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 **Figure 27.4**. 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 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 is 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.

Figure 27.4 — 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 ⅛ inch.

New RF ammeters are expensive and even surplus pricing can vary widely between $10 and $100 in today's market. AM radio stations are the main users of new units. The FCC defines the output power of AM stations based on the RF current in the antenna, so new RF ammeters are made mainly for that market. They are quite accurate, and their prices reflect that!

The good news is that used RF ammeters are often available. For example, Fair Radio Sales in Lima, Ohio, has been a consistent source of RF ammeters. Ham flea markets are also worth trying. Some grubbing around in your nearest surplus store or an older ham's junk box may provide just the RF ammeter you need.

Before buying a used RF ammeter, check to be sure it is actually an RF ammeter — it is common to find meters labeled "RF Amps" that are simple current meters intended to be used with an external RF current sensing unit.

RF Ammeter Substitutes

Don't despair if you can't find a used RF ammeter. It's possible to construct your own. Both hot-wire and thermocouple units can be homemade. Pilot lamps in series with antenna wires, or coupled to them in various ways, can indicate antenna current or even forward and reflected power. (See the Bibliography entries for Sutter and Wright.)

Another approach is to use a small low-voltage lamp as the heat/light element and use a photodetector driving a meter as an indicator. (Your eyes and judgment can serve as the indicating part of the instrument.) A feed line balance checker could be as simple as a couple of lamps with the right current rating and the lowest voltage rating available. You should be able to tell fairly well by eye which bulb is brighter or if they are about equal. You can calibrate a lamp-based RF ammeter with 60-Hz or dc power.

The optical approach is often taken in QRP portable equipment where an LED is used to replace a meter in an SWR bridge as described by Phil Salas, AD5X, in his Z-Match antenna tuner described in the **Transmission Line System Techniques** chapter.

As another alternative, you can build an RF ammeter that uses a dc meter to indicate rectified RF from a current transformer that you clamp over a transmission line wire as described by Zack Lau, W1VT. (See Bibliography.)

27.2 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 they require considerable care in construction to ensure accurate measurements. The requirements for indicators used only for the adjustment of impedance-matching circuits, rather than actual SWR measurement, are not so stringent, and you can easily make an instrument for this purpose.

27.2.1 BRIDGE CIRCUITS

Two commonly used bridge circuits are shown in **Figure 27.5**. The bridges consist essentially of two voltage dividers in parallel, with an RF 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 RF voltmeter indicates zero voltage. The bridge is then said to be in *balance*.

Taking Figure 27.5A 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 RF voltmeter reading will be zero. Remember that a matched transmission line has essentially a purely resistive input impedance. 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.

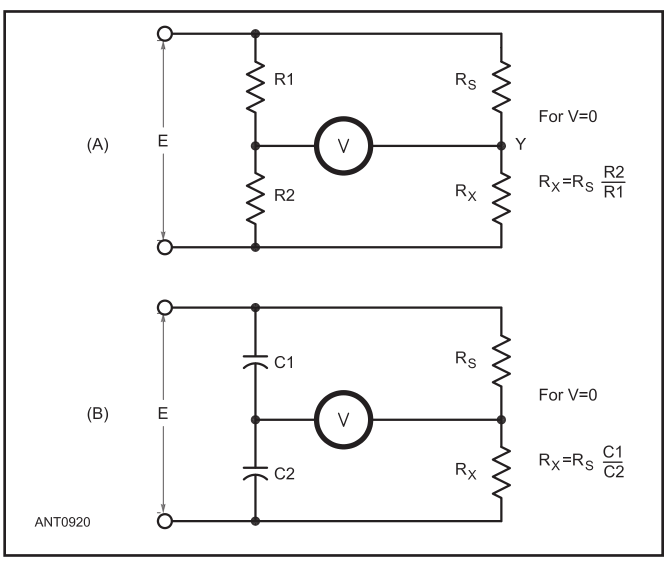


Figure 27.5 — Bridge circuits suitable for SWR measurement. At A, Wheatstone type using resistance arms. At B, capacitance-resistance bridge ("Micromatch"). Conditions for balance are independent of frequency in both types. The voltmeter must be an RF voltmeter.

If the line is not perfectly matched, its input impedance will not equal Z_0 and hence will not equal R_S, since you chose the latter to be equal to Z_0. There will then be a difference in potential between points X and Y, and the RF 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 the **Transmission Lines** chapter, it should be clear that when $R_S = Z_0$, the bridge is always in balance for the incident component. Thus the RF voltmeter does not respond to the incident component at any time but reads only the reflected component. The incident component can be measured across either R1 or R2, if they are equal resistances. The standing wave ratio is then

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

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

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

where k = E2/E1.

The operation of the circuit in Figure 27.5B is essentially the same, although this circuit has arms containing reactance as well as resistance.

It is not necessary that R1 = R2 in Figure 27.5A; 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 most commonly used.

A number of types of bridge circuits appear in **Figure 27.6**, many of which have been used in amateur products or amateur construction projects. The bridge at E is most often used in common low-cost SWR meters. (See the Bibliography entry for Silver for a description of how these meters work.) All circuits except that at G can have the ground returns of the generator and load at a common potential. At G, the generator and detector ground returns are at a common potential. You may interchange the positions of the detector and transmitter (or generator) in the bridge, and this may be advantageous in some applications.

The bridges shown at D, E, F and H may have one terminal of the generator, detector and load common. Bridges

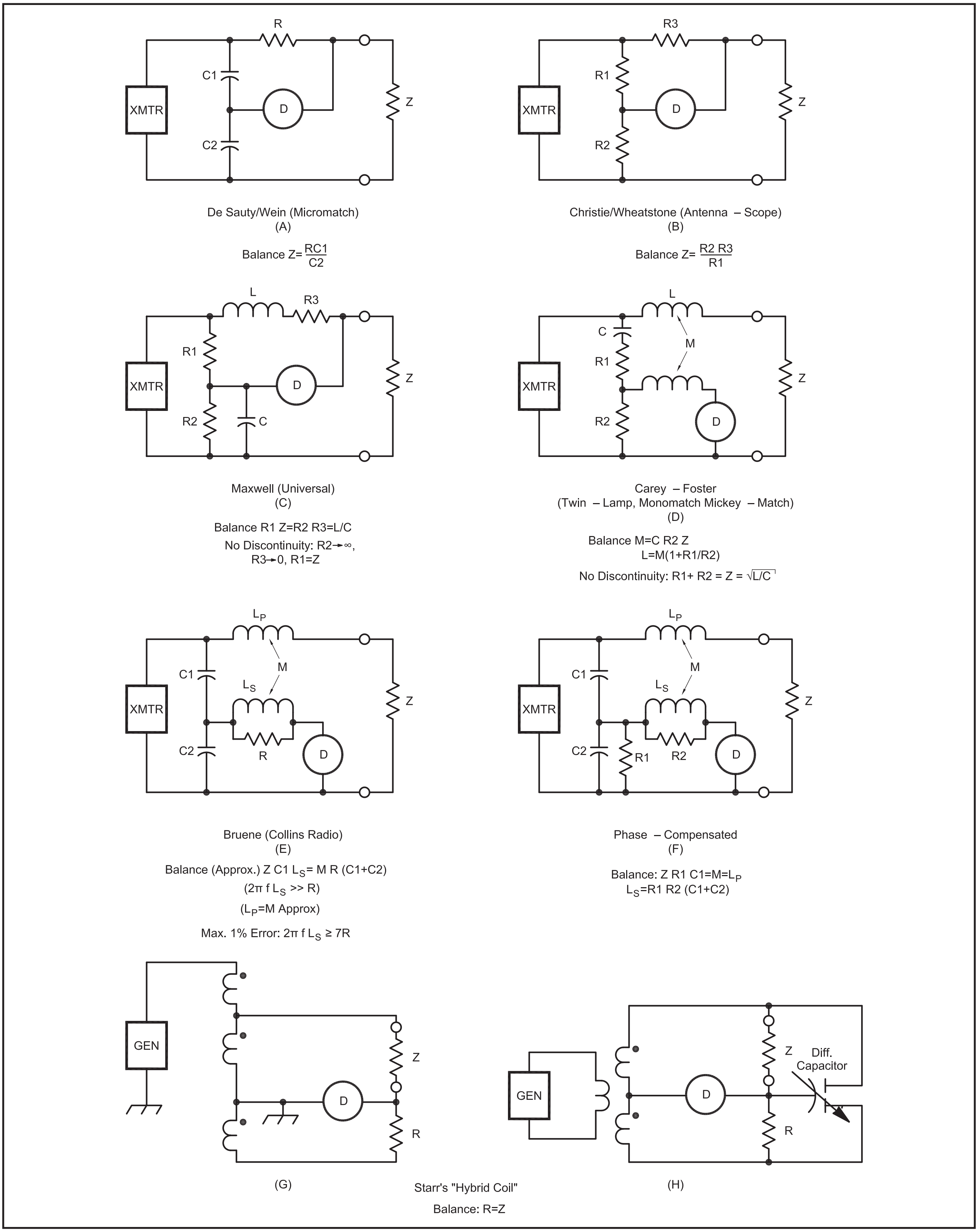


Figure 27.6 — Various types of SWR bridge indicator circuits and commonly known names of bridge circuits or devices in which they have been used. Detectors (D) are usually semiconductor diodes with meters, isolated with RF chokes and capacitors. However, the detector may be a radio receiver. In each circuit, Z represents the load being measured. (*This information provided by David Geiser, WA2ANU*)

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 bump) 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.

27.2.2 SWR RESISTANCE BRIDGE

The basic bridge configuration shown in Figure 27.5B may be home constructed and is reasonably accurate for SWR measurement at HF. A practical circuit for such a bridge is given in **Figure 27.7A** and a representative layout is shown in Figure 27.7B. Properly built, a bridge of this design can be used for measurement of SWRs up to about 15:1 with good accuracy. Resistance bridges cannot be left in the transmission during regular operation due to the power dissipation limits of the resistors. This bridge should be used for test purposes only.

You must also observe these important construction points:

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.

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 Figure 27.7B, 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 Figure 27.7A, are voltmeter multipliers for the 0-100 microammeter used as an indicator. This is sufficient resistance to make the voltmeter approximately 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. You should select its value 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. You can determine this by experiment.

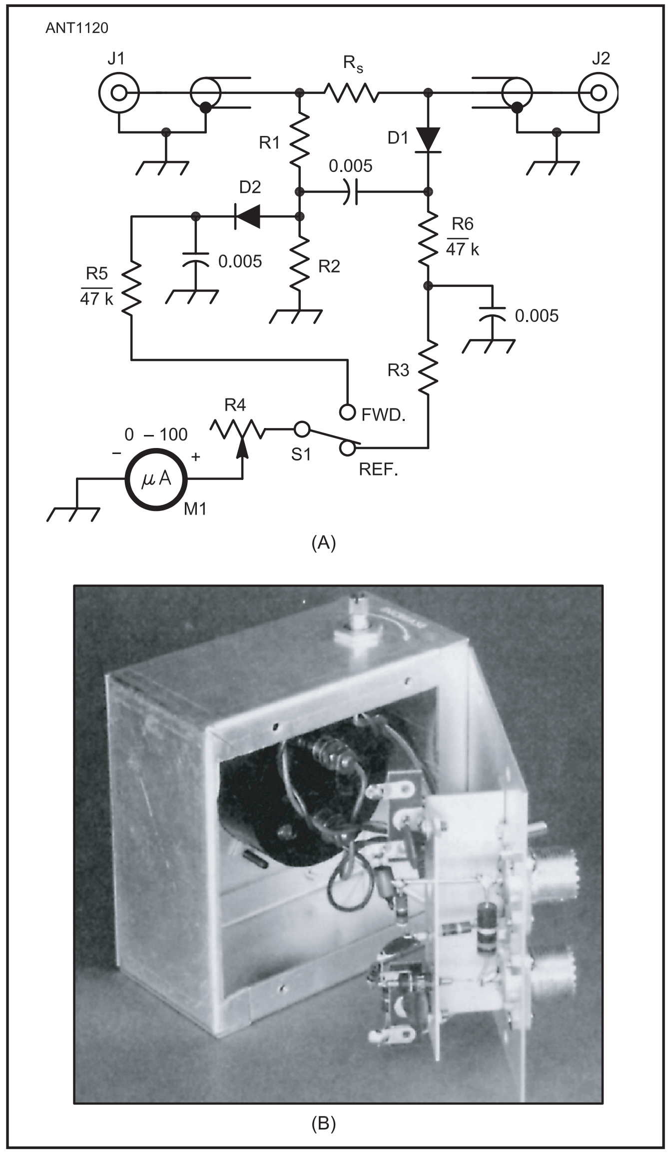


Figure 27.7 — At A, schematic of the resistance bridge for SWR measurement. Capacitors are disc ceramic. Resistors are ½-W composition except as noted below.

D1, D2 — 1N34A germanium or 1N5817 Schottky diodes.
J1, J2 — Coaxial connectors, chassis-mounting type.
M1 — 0-100 dc microammeter.
R1, R2 — 47 Ω, ½-W carbon-film or metal-oxide (see text).
R3 — See text.
R4 — 50-kΩ volume control.
R_s — Resistance equal to line Z_0 (½ or 1 W composition).
S1 — SPDT toggle.

At B, a 2 × 4 × 4-inch aluminum box is used to house this SWR bridge. 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.

The value used for R1 and R2 is not critical, but you should match the two resistors within 1% or 2% if possible. Keep the resistance of R_S as close as possible to the actual Z_0 of the line you use (generally 50 or 75 Ω). Select the resistor by actual measurement with an accurate resistance bridge, if you have one 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

Measure R1, R2 and R_S with a reliable digital ohmmeter or resistance bridge after completing the wiring. This will ensure that 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 to avoid stray shunt paths through the rectifiers.

Check the two voltmeter circuits as described above, applying enough RF (about 10 V peak, 1 V RMS) to the input terminals to give a full scale reading with R4 set for maximum deflection and 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. You must read the reflected voltage at full scale with J2 either open or shorted, when the incident voltage is set to full scale in each case, 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 you connect the test resistor 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 problem is poor matching of the resistance values. Both effects can be present simultaneously. You should make sure you obtain a good null at all frequencies before using your bridge.

Bridge Operation

You must limit the RF power input to a bridge of this type to 2 W at most, because of the power-dissipation ratings of the resistors. If the transmitter has no provision for reducing power output to this value a simple power-absorber circuit can be made up, as shown in **Figure 27.8**. Lamp DS1 changes resistance as it heats up — from a few ohms when cold to more than 100 Ω at full power. This increasing resistance 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 a measurement, connect the unknown load 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 using these 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

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

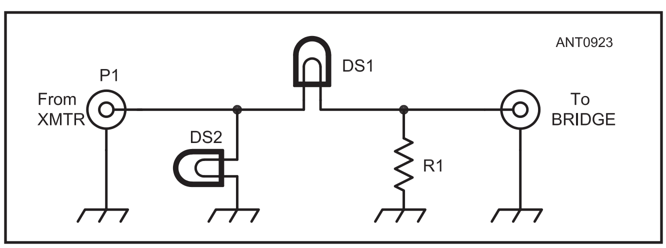


Figure 27.8 — Power-absorber circuit for use with resistance-type SWR bridges when the transmitter has no special provisions for power reduction. For RF powers up to 50 W, DS1 is a 120-V 40-W incandescent lamp and DS2 is not used. For higher powers, use sufficient additional lamp capacity at DS2 to load 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.

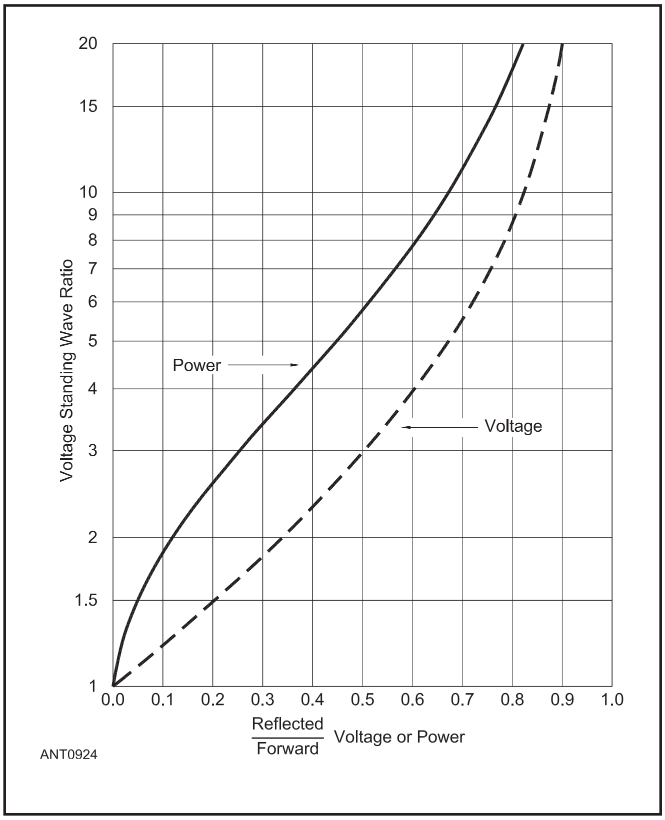


Figure 27.9 — Chart for finding voltage standing wave ratio when the ratio of reflected-to-forward voltage or reflected-to-forward power is known.

Instead of calculating the SWR value, you could use the voltage curve in **Figure 27.9**. In this example the ratio of reflected to forward voltage is 40/100 = 0.4, and the SWR value is about 2.3:1.

You may calibrate the meter scale in any arbitrary units, so long as the scale has equal divisions. It is the ratios of the voltages, and not the actual values, that determine the SWR.

27.2.3 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 checkout procedure described above is followed carefully, the bridge in Figure 27.6 should be sufficiently accurate for practical use. The accuracy is highest for low standing wave ratios because of the nature of the SWR calculation; at high SWR 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.

Detector nonlinearity is another source of error. A diode peak detector is approximately linear if the load impedance is high enough, and the signal is much greater than the diode forward-conduction voltage, but it will still have significant nonlinearity at the low end of the scale.

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 50- to 75-Ω range, the RF resistance of a noninductive resistor of ½- or 1-W rating is essentially identical with its dc resistance at VHF and below.

Common-Mode Currents

As explained in the **Transmission Line System Techniques** chapter, there are two ways in which unwanted *common-mode* (sometimes called *antenna*) currents can flow on the outside of a coaxial line — currents induced onto 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. Such currents can cause significant SWR measurements error and SWR that changes with line length but for different reasons.

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 inserting an additional section of line (⅛ to ¼ electrical wavelength preferably) of the same Z_0. The SWR indicated by the bridge should not change except for a slight decrease because of the additional line loss. If there is a marked change, you may need better shielding.

Common-mode currents can also flow on the outside of coaxial transmission lines if the outside surface of the shield is connected directly to one side of the antenna. Even if the antenna itself is balanced, this "extra" conductor will unbalance the system and common-mode current will flow on the outside of the line. In such cases, the SWR will vary 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 has become part of the antenna system by being connected to the antenna at the feed point. The outside shield of the line thus constitutes a load, 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. Since changing the line length (or position) changes one component of this composite load, the SWR changes too. This is an undesirable condition since the line is usually operating at a higher SWR than it should — and would if the common-mode current on the outside of the coax were eliminated.

The remedy for both situations is generally to use a common-mode choke balun as described in the **Transmission Line System Techniques** chapter or to detune the outside of the line by proper choice of length so that it presents a high impedance at the frequency of operation. Note that this is not a *measurement error*, since what the instrument reads is the actual SWR on the line.

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 almost always mismatched at harmonic and subharmonic 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.

27.2.4 REFLECTOMETERS

A reflectometer consists of a coupled pair of transmission-line impedance bridges (see Figure 27.6) operated back-to-back so that the generator and load are reversed in one bridge. The resulting imbalance between bridges is displayed on a meter with a scale calibrated so as to convert the imbalance to SWR. Various simple reflectometers have been described from time to time in *QST* and in *The ARRL Handbook*. (See Bibliography.)

Bridges of this type are usually frequency-sensitive — that is, the meter response increases with increasing frequency for the same applied voltage so that a CAL (calibration) potentiometer is required to set the meter sensitivity for each use.

Because most of these designs are frequency sensitive, it is difficult to calibrate them accurately for power measurement. Similarly, without a guaranteed power calibration, they cannot make accurate quantitative measurements of SWR 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.

A pair of typical reflectometer designs are given in **Figures 27.10** and **27.11**. (Both original *QST* articles are included on this book's CD-ROM.) The classic circuit by DeMaw in Figure 27.10 is a very useful circuit at low power levels. It can be scaled up to be used at higher power levels by reducing the number of turns on the toroid primary and increasing the voltage ratings of the voltage sensing capacitors, C1 and C2. (The article by Bruene in the Bibliography should be consulted, as well.) The design by Brown in Figure 27.11 is for use with 300-Ω twin-lead and may be used with parallel-wire feed lines of other characteristic impedances such as 450-Ω window line by changing R1 and R2 to match the impedance of the line.

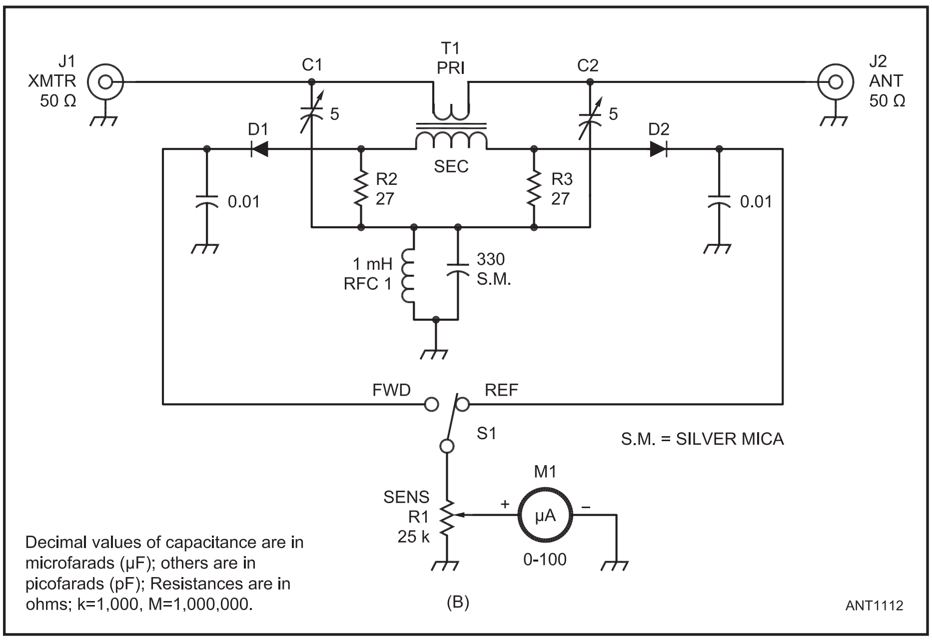


Figure 27.10 — Schematic diagram of the QRP VSWR indicator. Fixed-value capacitors are disc ceramic except those marked with S.M., which are silvered-mica. R2 and R3 and ¼-W carbon-film or metal-film units.

C1, C2 — Miniature PC board mount air trimmer.
D1, D2 — Silicon switching diode, 1N4148 type, matched for equivalent forward resistance by using an ohmmeter.
J1, J2 — RF connector receptacle (phono jack, BNC, UHF).
M1 — Miniature 50- or 100-μA dc meter.
R1 — Linear-taper miniature control, 25 kΩ.
RFC1 — Miniature 1-mH RF choke.
S1 — Miniature SPDT slide or toggle switch.
T1 — Toroidal transformer. Secondary: 60 turns #30 AWG enameled wire on a T68-2 powdered-iron core. Primary is two turns over secondary winding (see text).

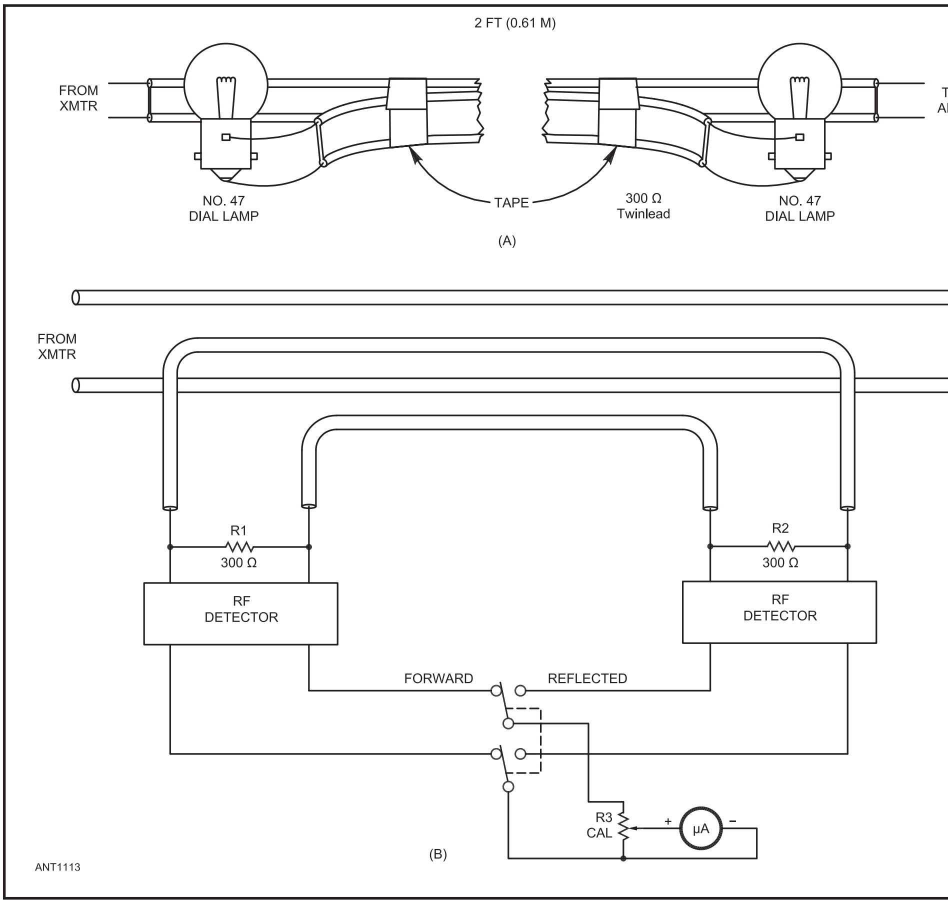


Figure 27.11 — The twin-lamp SWR indicator, shown at A, is a crude form of reflectometer. With a high SWR, both lamps will light with equal brightness. With low SWR, only the lamp on the left side will light. The reflectometer circuit at B yields more accurate results that can be calibrated to read SWR directly.

27.3 RF POWER MEASUREMENT

The standard commercial instruments amateurs use to measure RF power are the various models of the Thruline *directional wattmeters* from Bird Technologies (**www.bird-rf.com**) such as the popular Model 43. The meter consists of a section of transmission into which is inserted a selectable power-sensing element, popularly referred to as a "slug." The transmission line in the wattmeter is designed to have an element inserted without disrupting normal power flow through the meter.

The element consists of a pickup loop and terminating resistor that form a *directional coupler* — a circuit that couples to a transmission line and extracts a small amount of power flowing in one direction. (See the Bibliography entry by Wade for a tutorial on directional couplers, also include on this book's CD-ROM.) The construction of the Bird transmission line and sensing element are shown as Figure 3 in the Model 43 operating manual, available for download at **www.bird-rf.com** (look for the Model 43 product information). The element can be rotated so that the directional coupler picks up either forward or reflected power.

The energy from the directional coupler in the element then passes through a rectifying diode and filter capacitor that form an RF detector as described earlier in this chapter. The output of the RF detector then drives a meter that is calibrated in watts. The standard series of elements covers from 2 to 1000 MHz and from 5 W to 5000 W full-scale. A variety of specialty elements are also available.

For a close look at the construction of the Bird Thruline wattmeter and a typical power sensing element, see the Repeater Builders website article, "Photo Tour of a Bird Wattmeter Element," by Robert Meister, WA1MIK at **www.repeater-builder.com/projects/bird-element-tour/bird-element-tour.html**. A number of excellent white papers and application notes on the use of these ubiquitous instruments are available on the Bird Technologies website under "Resources and Tools."

27.3.1 DIRECTIONAL POWER/SWR METER

The following section is an overview of the January 2011 *QST* article by Bill Kaune, W7IEQ, "A Modern Directional Power/SWR Meter." The complete article including firmware and printed circuit board artwork is available on the CD-ROM included with this book.

The primary use for this unit is to monitor the output power and tuning of a transceiver. The author's station configuration is shown in **Figure 27.12**. RF power generated by the transmitter is routed via RG-8 coaxial cable through a

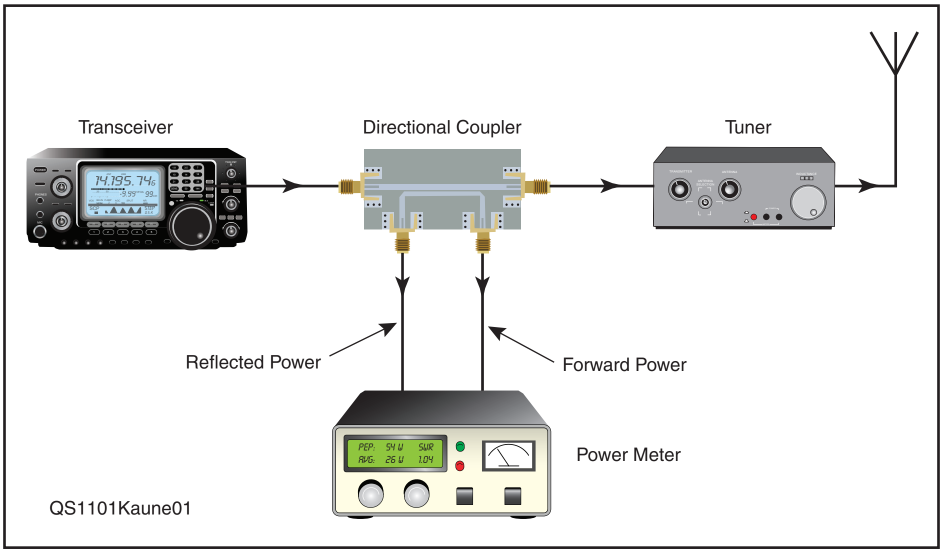


Figure 27.12 — W7IEQ station setup, including the power meter being described here.

directional coupler to an antenna tuner, which is connected to the antenna with RG-8. The directional coupler contains circuits that sample the RF power flowing from the transmitter to the tuner (the forward power) and the RF power reflected back from the tuner to the transmitter (the reflected power). These samples are sent via RG-58 cable to the two input channels of the power meter. This project includes the directional coupler and the power meter. Enough detail is provided in the full article so that an amateur can duplicate the device or modify the design.

Directional Coupler

The directional coupler is based on the unit described in "The Tandem Match" by John Grebenkemper, KI6WX in the January 1987 issue of *QST* and also included on this book's CD-ROM. A pair of FT-82-67 toroids with 31 turns of #26 AWG magnet wire over lengths of RG-8 form the basis of the directional coupler shown in **Figure 27.13**.

The forward and reflected power samples coupled are reduced by a factor of $1/N^2$, where N = 31 is the number of turns of wire on each toroid. Thus the forward and reflected power samples are reduced by about 30 dB. For example, if a transceiver were delivering a power of 100 W to a pure 50 Ω load, the forward power sample from the directional coupler would be about 0.1 W (20 dBm).

The directivity of a directional coupler is defined as the ratio of the forward power sample divided by the reflected power sample when the coupler is terminated in 50 Ω. In this coupler, the directivity measured using an inexpensive network analyzer is at least 35 dB at 3.5 MHz and 28 dB at 30 MHz.

Power/SWR Meter — Circuit Description

Figure 27.14 shows a front panel view of the power meter. An LCD displays the measured peak (PEP) and average (AEP) envelope powers as well as the standing wave ratio (SWR). The power meter calculates either the peak and average envelope power traveling from the transceiver to load (the forward power) or the peak and average envelope powers actually delivered to the load (the forward power minus reflected power). The average envelope power (AEP) represents an average of the forward or load powers over an averaging period of either 1.6 or 4.8 seconds.

A 1 mA-movement analog meter on the front panel facilitates antenna tuning. This meter continuously displays the quantity 1 – 1/SWR, where SWR is the standing wave ratio on the line. Thus, an SWR of 1.0 corresponds to a meter reading of 0 — no deflection of the meter. An SWR of 2 results in a 50% deflection of the meter, while an SWR of 5 produces an 80% deflection of the meter.

The forward and reflected power samples from the directional coupler are applied to a pair of Analog Devices

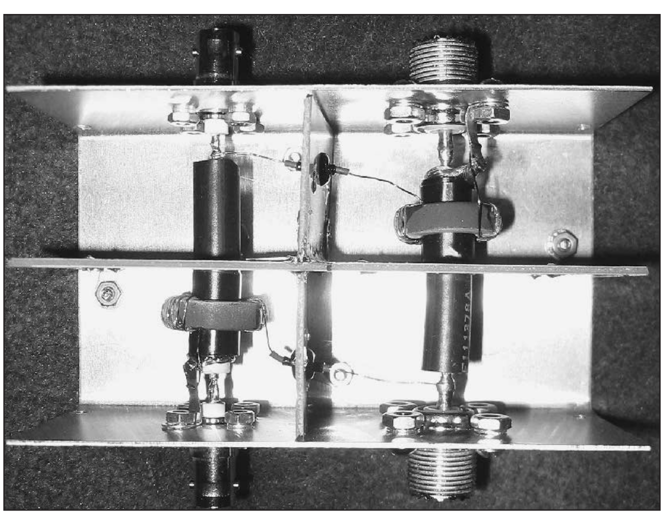


Figure 27.13 — Completed directional coupler.

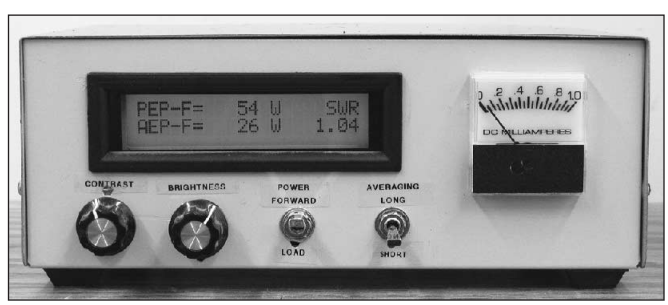


Figure 27.14 — Front panel of power meter. The LCD shows the peak envelope power (PEP), the average envelope power (AEP) and the SWR. The two knobs control the contrast and back lighting of the LCD. One toggle switch determines whether forward or load powers are displayed. A second switch sets the averaging time for the AEP calculation. The meter displays SWR and is used for tuning purposes.

AD8307 logarithmic detectors. External 20 dB attenuators (Mini-Circuits HAT-20) reduce the signals from the directional coupler to levels compatible with the AD8307. As noted earlier, the directional coupler has an internal attenuation of about 30 dB, so the total attenuation in each channel is about 50 dB. Thus, a rig operating at a power level of 1 kW (60 dBm) will result in an input to the forward power channel of about 10 dBm. (The schematic diagram and parts list of the power meter are provided on the CD-ROM version of the article.) The detectors are configured such that the time constant of their output follows the modulation envelope of the RF signal.

LF398 sample-and-hold ICs stabilize the voltages from the forward and reflected power logarithmic detectors. In this way both voltages can be sampled at the exact same time and held for subsequent analog-to-digital conversion and calculation of power and SWR by the PIC16F876A microprocessor (**www.microchip.com**). The processor also includes a pulse-width-modulated (PWM) output used to drive the analog SWR meter on the front panel.

27.3.2 HIGH-POWER RF SAMPLERS

If one wants to measure characteristics of a transmitter or high-power amplifier, some means of reducing the power of the device to 10 or 20 dBm must be used. The most straightforward way to do this is to use a 30 or 40 dB attenuator capable of handling the high power. A 30 dB attenuator will reduce a 100 W transmitter to 20 dBm. A 40 dB attenuator will reduce a 1 kW amplifier to 20 dBm. If further attenuation is needed, a simple precision attenuator may be used after the signal has been reduced to the 20 dBm level.

The problem with high-power attenuators is that they are expensive to buy or build because the front end of the attenuator must handle the output power of the transmitter or amplifier. If one already has a dummy load, an RF sampler may be used to produce a replica of the signal at a reduced power level. The sampler described here was originally presented in *QST* Technical Correspondence for May 2011 by Tom Thompson, WØIVJ. (The original article with construction information plus some supplemental information is available on this book's CD-ROM.)

A transformer sampler passes a single conductor (usually the insulated center conductor from a piece of coaxial cable) from the transmitter or amplifier to the dummy load through a toroidal inductor forming a transformer with a single turn primary. The secondary of the transformer is connected to a resistor network and then to the test equipment as shown in **Figure 27.15** The source, whether a transmitter or amplifier, is assumed to be a pure voltage source in series with a 50-Ω resistor. This most likely is not exactly the case but is sufficient for analysis.

If a current, I, flows into the dummy load, then a current, I / N flows in the secondary of the transformer, where N is the number of turns on the secondary. Figure 27.15 also shows the equivalent circuit, substituting a current source for the transformer. 40 dB is selected for the attenuation and 15 turns for the secondary of the transformer. If R_{SHUNT} = 15 Ω, and R_{SERIES} = 35 Ω, then the voltage across a 50-Ω load resistor, R_{SAMPLE}, is 1/100 of the voltage across the dummy load, which is 40 dB of attenuation.

Reflecting this resistor combination back through the transformer yields 0.06 Ω in series with the 50-Ω dummy load impedance. This is an insignificant change. Furthermore, reflecting 100 Ω from the primary to the secondary places 22.5 kΩ in parallel with R_{SHUNT}, which does not significantly affect its value. The test equipment sees a 50-Ω load looking back into the sampler. Even at low frequencies, where the reactance of the secondary winding is lower than 15 Ω, the impedance looking back into the sample port remains close to 50 Ω.

The samplers described here use an FT37-61 ferrite core followed by two resistors as described above. The through-line SWR is good up to 200 MHz, the SWR is fair looking into the sampled port, and the useful bandwidth extends

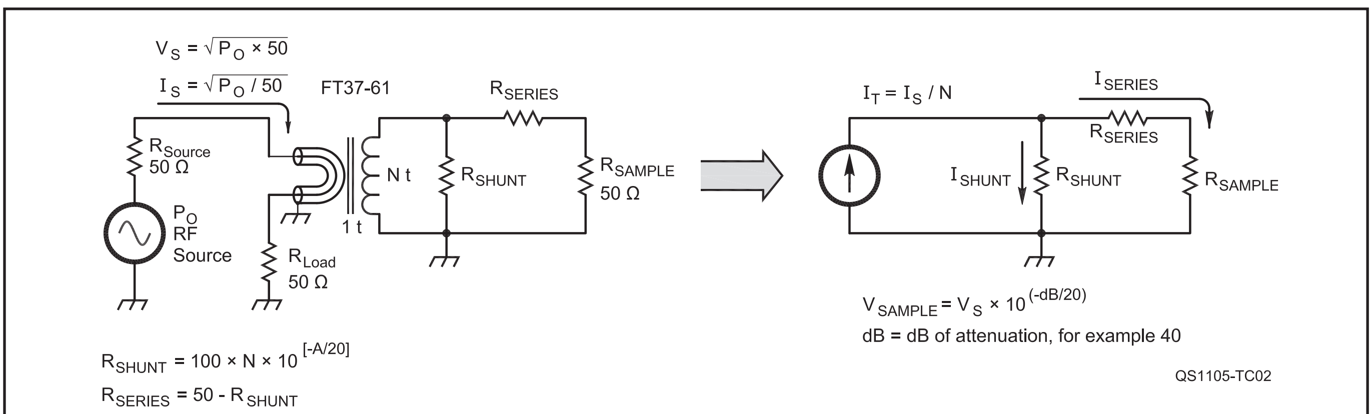


Figure 27.15 — RF sampler circuit diagram and equivalent circuit showing calculations.

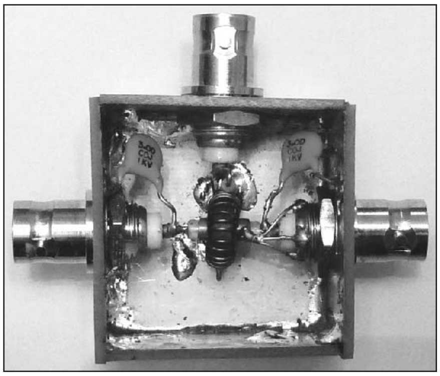


Figure 27.16 — RF sampler using box construction.

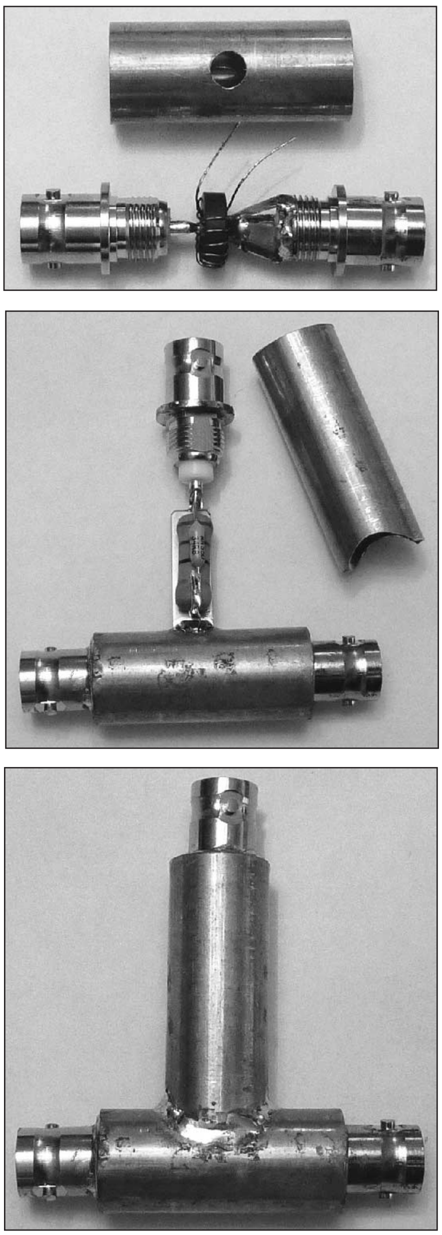


Figure 27.17 — RF sampler using tube construction.

from 0.5 MHz to about 100 MHz. If you are interested in an accurate representation of the third harmonic of your HF transmitter or amplifier, it is important for the sampler to give accurate attenuation into the VHF range.

Figure 27.16 shows a photo of a sampler built into a 1.3 × 1.3 × 1 inch (inside dimensions) box constructed from single-sided circuit board material. The through-line connection is made with a short piece of UT-141 semi-rigid coax with the shield grounded only on one side to provide electrostatic shielding between the toroid and the center conductor of the coax. (Do not ground both ends of the shield or a shorted turn is created.) R_{SHUNT} is hidden under the toroid, and R_{SERIES} is shown connected to the sample port. This construction technique looks like a short piece of 200-Ω transmission line in the through-line which affects the SWR at higher frequencies. This can be corrected by compensating with two 3 pF capacitors connected to the through-line input and output connectors as shown in the photo. The through-line SWR was reduced from 1.43:1 to 1.09:1 at 180 MHz by adding the capacitors. This compensation, however, causes the attenuation to differ at high frequencies depending on the direction of the through-line connection. A sampler constructed using the box technique is useable from below 1 MHz through 30 MHz.

Figure 27.17 shows a different approach using 9⁄16 inch diameter, 0.014 inch wall thickness, hobby brass tubing. This lowers the impedance of the through-line so that no compensation is needed. The through-line SWR for the tube sampler is 1.08:1 at 180 MHz which is as good as the box sampler and the sensitivity to through-line direction is reduced. Although the high frequency attenuation is not as good as the box sampler, the construction technique provides a more consistent result. A sampler constructed using the tube technique should be usable through 200 MHz.

27.3.3 AN INEXPENSIVE VHF DIRECTIONAL COUPLER

Precision inline 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 **Figures 27.18** through **27.20** 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 section at many hardware stores.

The sampler consists of a short section of handmade coaxial line, in this instance, of 50-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 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 ¾-inch ID and 5⁄16-inch OD copper pipe, and two coaxial fittings. Other available tubing

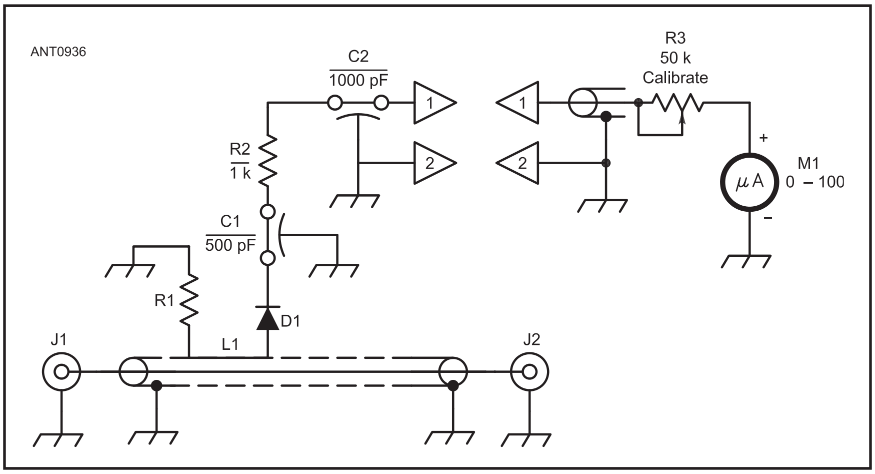


Figure 27.18 — Circuit diagram for the line sampler.
C1 — 500-pF feedthrough capacitor, solder-in type.
C2 — 1000-pF feedthrough capacitor, threaded type.
D1 — 1N34A germanium or 1N5817 Schottky diode.
J1, J2 — Coaxial connector, type N (UG-58A).
L1 — Pickup loop, copper strap 1-inch long × 3⁄16-inch wide. Bend into "C" shape with flat portion 5⁄8-inch long.
M1 — 0-100 μA meter.
R1 — 82 to 100 Ω, carbon-film or metal film.
R3 — 50-kΩ composition control, linear taper.

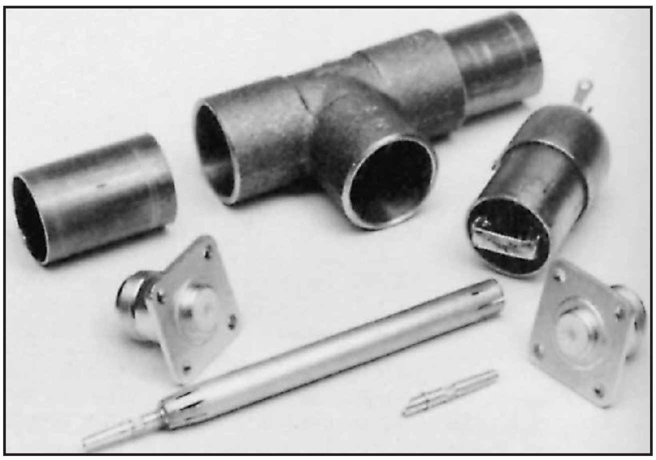


Figure 27.19 — Major components of the line sampler. The brass T and two end sections are at the upper left in this picture. A completed probe assembly is at the right. The N connectors have their center pins removed. The pins are shown with one inserted in the left end of the inner conductor and the other lying in the right foreground.

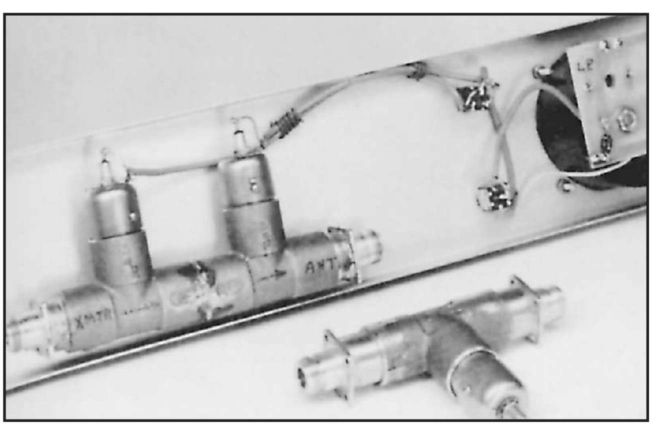


Figure 27.20 — Two versions of the line sampler. The single unit described in detail here is in the foreground. Two sections in a single assembly provide for monitoring forward and reflected power without probe reversal.

combinations for 50-Ω line may be usable. The ratio of outer conductor ID to inner conductor OD should be 2.4/1. (The complete article with more detail is available on this book's CD-ROM.)

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 as described in the complete article.

27.3.4 RF STEP ATTENUATOR

A good RF step attenuator is one of the key pieces of equipment that belongs on your workbench. The attenuator in this project offers good performance yet can be built with a few basic tools. The attenuator is designed for use in 50-Ω systems, provides a total attenuation of 71 dB in 1-dB steps, offers respectable accuracy and insertion loss through 225 MHz and can be used at 450 MHz as shown in **Table 27.1**. This material was originally published as "An RF Step Attenuator" by Denton Bramwell, K7OWJ, in the June 1995 *QST*.

The attenuator consists of 10 resistive pi (π) attenuator sections such as the one in **Figure 27.21**. Each section consists of a DPDT slide switch and three ¼-W, 1%-tolerance metal-film resistors. The complete unit contains single 1, 2,

Table 27.1
Step Attenuator Performance at 148, 225, and 250 MHz
Measurements made in the ARRL Laboratory

Attenuator set for Maximum attenuation (71 dB)		*Attenuator set for minimum attenuation (0 dB)*	
Frequency (MHz)	*Attenuation (dB)*	*Frequency (MHz)*	*Attenuation (dB)*
148	72.33	148	0.4
225	73.17	225	0.4
450	75.83	450	0.84

Note: Laboratory-specified measurement tolerance of ±1 dB

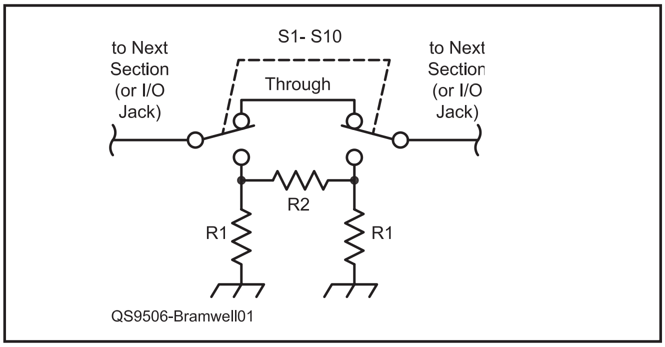


Figure 27.21 — Schematic of one section of the attenuator. All resistors are ¼-W, 1%-tolerance metal-film units. See Table 27.2 for the resistor values required for each attenuator section. There are six 10 dB sections and one each of 1, 2, 3, and 5 dB.

Table 27.2
Closest 1%-Tolerance Resistor Values

Attenuation (dB)	*R1 (Ω)*	*R2 (Ω)*
1.00	866.00	5.60
2.00	436.00	11.50
3.00	294.00	17.40
5.00	178.00	30.10
10.00	94.30	71.50

3 and 5-dB sections, and six 10-dB sections. **Table 27.2** lists the resistor values required for each section.

The enclosure is made of brass sheet stock, readily available at hardware and hobby stores. By selecting the right stock, you can avoid having to bend the metal and need only perform a minimum of cutting.

Construction

The enclosure can be built using only a nibbling tool, drill press, metal shears, and a soldering gun or heavy soldering iron. (Use a regular soldering iron on the switches and resistors.) One method of cutting the small pieces of rectangular tubing to length is to use a drill press equipped with a small abrasive cutoff wheel.

Brass is easy to work and solder. For the enclosure, you'll need two precut 2 × 12 × 0.025-inch sheets and two 1 × 12 × 0.025-inch sheets. The 2-inch-wide stock is used for the front and back panels; the 1-inch-wide stock is used for the ends and sides. For the internal wiring, you need a piece of 5⁄32 × 5⁄16-inch rectangular tubing, a ¼ × 0.032-inch strip, and a few small pieces of 0.005-inch-thick stock to provide interstage shields and form the 50-Ω transmission lines that run from the BNC connectors to the switches at the ends of the step attenuator.

For the front panel, nibble or shear a piece of 2-inch-wide brass to a length of about 9½ inches. Space the switches from each other so that a piece of the rectangular brass tubing lies flat and snugly between them. See **Figure 27.22**. Drill holes for the #4-40 mounting screws and nibble or punch rectangular holes for the bodies of the slide switches.

Before mounting any parts, solder in place one of the 1-inch-wide chassis side pieces to make the assembly more rigid. Solder the side piece to the edge of the top plate that faces the "through" side of the switches; this makes later assembly easier (see **Figure 27.23**). Although the BNC input and output connectors are shown mounted on the top (front) panel, better lead dress and high-frequency performance may result from mounting the connectors at the ends of the enclosure.

DPDT slide switches designed for sub-panel mounting often have mounting holes tapped for #4-40 screws. Enlarge the holes to allow a #4-40 screw to slide through. Before mounting the switches, make the "through" switch connection (see Figure 27.21) by bending the two lugs at one end of each switch toward each other and soldering the lugs together or solder a small strip of brass between the lugs and clip off the lug ends. Mount the switches above the front panel, using 5⁄32-inch-high by 7⁄32-inch-OD spacers. Use the same size spacer on the inside. On the inside, the spacer creates a small post that helps reduce capacitive coupling from one side of the attenuator to the other. The spacers position the switch so that the 50-Ω stripline can be formed later.

The trick to getting acceptable insertion loss in the "through" position is to make the attenuator look as much as possible like 50-Ω coax. That's where the rectangular tubing and the ¼ × 0.032-inch brass strip come into the picture (see Figure 27.22); they form a 50-Ω stripline. (See the **Transmission Lines** chapter for information on stripline.)

Cut pieces of the rectangular tubing about ¾-inch long, and sweat solder them to the front panel between each of the slide switches. Next, cut lengths of the ¼-inch strip long enough to conveniently reach from switch to switch, then cut one more piece. Drill 1⁄16-inch holes near both ends of all but one of the ¼-inch strips. The undrilled piece is used as a temporary spacer, so make sure it is flat and deburred.

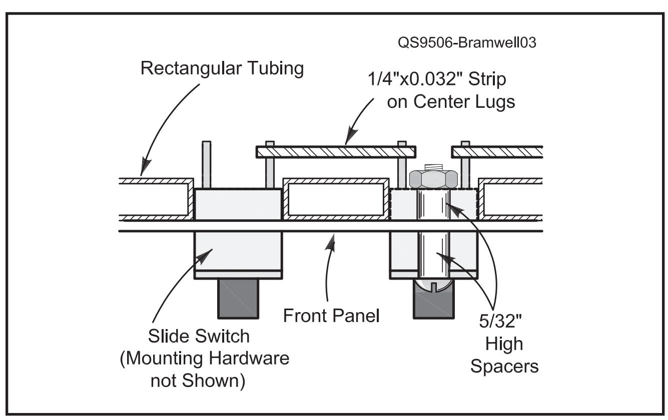


Figure 27.22 — Key to obtaining acceptable insertion loss in the "through" position is to make the whole device look as much as possible like 50-Ω coax. The rectangular tubing and the ¼ × 0.032-inch brass strip between the switch sections form a 50-Ω stripline.

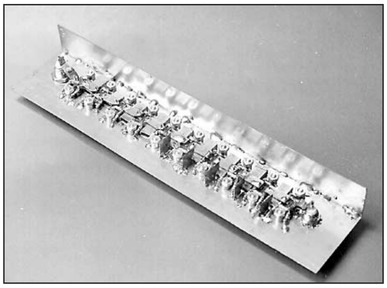


Figure 27.23 — Solder one of the 1-inch-wide chassis side pieces in place to make the assembly more rigid during construction. Solder the side piece to the edge of the top plate that faces the "through" side of the switches; this makes the rest of the assembly easier.

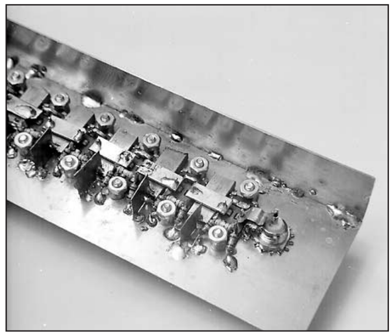


Figure 27.24 — The attenuator before final mechanical assembly. The ¼-inch strips are spaced 0.033 inch apart to form a 50-Ω connection from the BNC connector to the stripline. There are ½-inch square shields between 10-dB sections. The square shields have a notch in one corner to accommodate the end of the rectangular tubing.

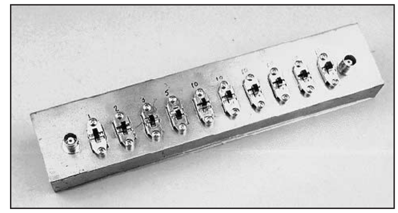


Figure 27.25 — The completed step attenuator in the enclosure of brass sheet. The BNC connectors may be mounted on the front panel at the end of the switches or on the end panels.

Lay the temporary spacer on top of the rectangular tubing between the first two switches, then drop one of the drilled ¼-inch pieces over it, with the center switch lugs through the 1⁄16-inch holes. Before soldering, check the strip to make sure there's sufficient clearance between the ¼-inch strip and the switch lugs; trim the corners if necessary. Use a screwdriver blade to hold the strip flat and solder the lugs to the strip. Remove the temporary spacer. Repeat this procedure for all switch sections. This creates a 50-Ω stripline running the length of the attenuator.

Next, solder in place the three 1%-tolerance resistors of each section, keeping the leads as short as possible. Use a generous blob of solder on ground leads to make the lead less inductive. Install a ½-inch-square brass shield between each 10-dB section to ensure that signals don't couple around the sections at higher frequencies.

Use parallel ¼-inch strips of 0.005-inch-thick brass spaced 0.033 inch apart to form 50-Ω feed lines from the BNC connectors to the switch contacts at each end of the stripline as shown in Figure 27.23. (Use the undrilled piece of 0.032-inch-thick brass to insure the proper line spacing.) The attenuator with all switches and shields in place is shown ready for final mechanical assembly in **Figure 27.24**

Finally, solder in place the remaining enclosure side, cut and solder the end pieces, and solder brass #4-40 nuts to the inside walls of the case to hold the rear (or bottom) panel. Drill and attach the rear panel and round off the sharp corners to prevent scratching or cutting anyone or anything. Add stick-on feet and labels and your step attenuator of **Figure 27.25** is ready for use.

Remember that the unit is built with ¼-W resistors, so it can't dissipate a lot of power. Remember, too, that for the attenuation to be accurate, the input to the attenuator must be a 50-Ω source and the output must be terminated in a 50-Ω load.

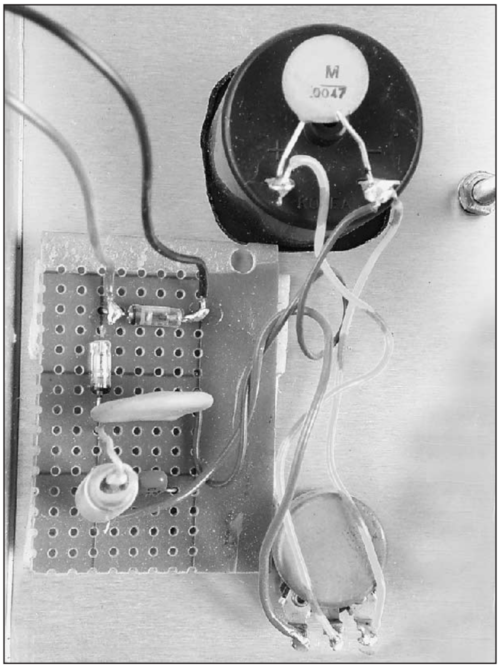
27.4 FIELD STRENGTH METERS

Few amateur stations, fixed or mobile, are without need of a *field-strength meter* (*FSM*). An instrument of this type serves many useful purposes during antenna experiments and adjustments. In its simplest form, the field strength meter is simply a diode detector and a sensitive meter with a potentiometer wired as a resistive divider to act as a sensitivity control as in **Figure 27.26**. This type of meter is commonly and inexpensively available both new and used. (See the Bibliography and this book's CD-ROM for the *QST* article describing how to build this simple FSM.)

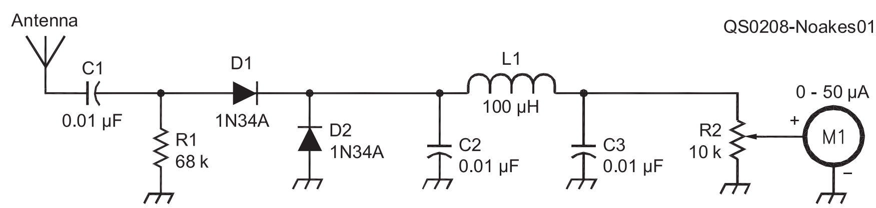
When work is to be done from many wavelengths away from the antenna, however, such a simple instrument lacks the necessary sensitivity. Further, such a device has a serious fault because its linearity leaves much to be desired and it is very wideband so that the measurement may be upset from any other strong nearby transmitter, such as an AM broadcast station. Thus, a more capable instrument is needed.

27.4.1 PORTABLE FIELD STRENGTH METER

The field-strength meter described here takes care of the problems associated with a simple FSM. Additionally, it is small, measuring only 4 × 5 × 8 inches. The power



(A)



(B)

Figure 27.26 — With the sensitivity control at mid-range, the simple FSM shown at A can easily detect a 1-W, 2 meter signal. B gives the schematic and parts list for the FSM. A metal enclosure is mandatory.
C1-C3 — 0.01 µF disc ceramic capacitor.
D1, D2 — 1N34A (germanium) or 1N5817 (Schottky) diode.
L1 — 100 µH inductor.
M1 — 50 µA analog meter.
R1 — 68 kΩ, ¼-W carbon-film or metal-film resistor.
R2 — 10 kΩ linear or audio taper potentiometer.
Antenna — BNC chassis-mount receptacle.

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 information in this section is based on a January 1973 *QST* article by Lew McCoy, W1ICP. (See the Bibliography.)

The unit is pictured in **Figures 27.27** and **27.28**, and the schematic diagram is shown in **Figure 27.29**. A type 741 op-amp IC is the heart of the unit. (Any general-purpose op-amp can be substituted for the 741.) 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.

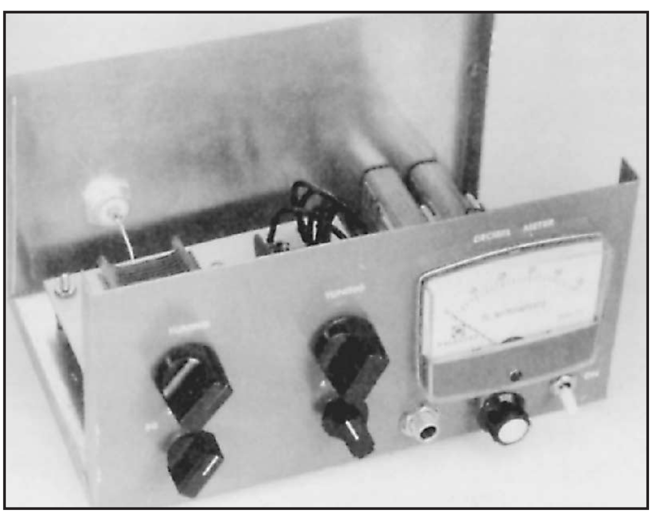


Figure 27.27 —The linear field strength meter. The control at the upper left is for C1 and the one to the right for C2. At the lower left is the band switch, and to its right the sensitivity switch. The zero-set control for M1 is located directly below the meter.

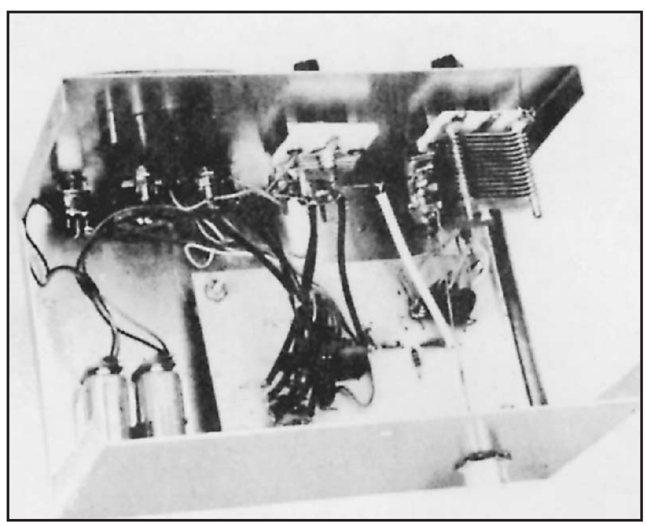


Figure 27.28 — Inside view of the field-strength meter. At the upper right is C1 and to the left, C2. The dark leads from the circuit board to the front panel are the shielded leads described in the text.

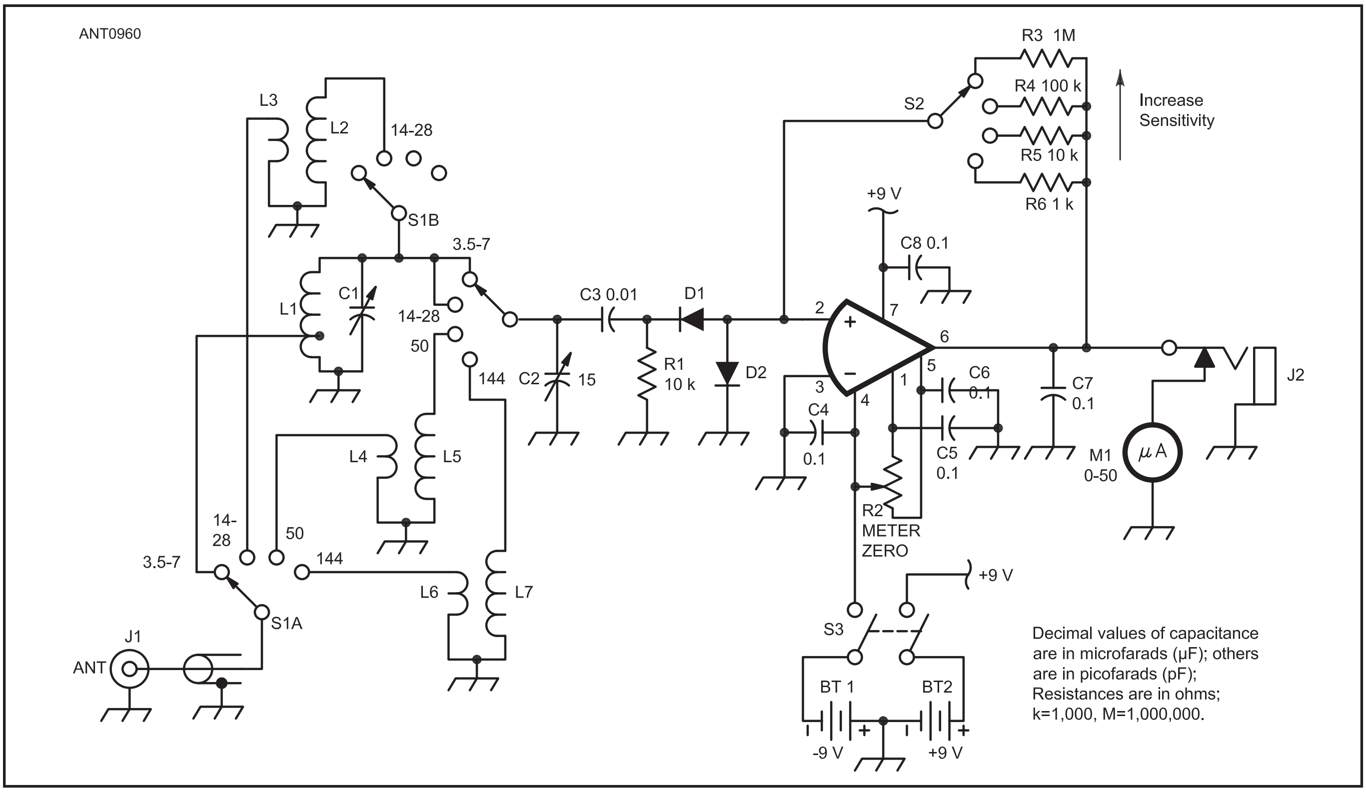


Figure 27.29 — Circuit diagram of the linear field strength meter. All resistors are ¼- or ½-W carbon-film or metal-film types.

C1 — 140 pF variable.
C2 — 15-pF variable
D1, D2 — 1N4148 or equiv.
L1 — 34 turns #24 AWG enameled wire wound on an T-68-2 core, tapped 4 turns from ground end.
L2 — 12 turns #24 AWG enameled wire wound on T-68-2 core.
L3 — 2 turns #24 AWG enameled wire wound at ground end of L2.
L4 — 1 turn #26 AWG enameled wire wound at ground end of L5.
L5 — 12 turns #26 AWG enameled wire wound on T-25-12 core.
L6 — 1 turn #26 AWG enameled wire wound at ground end of L7.
L7 — 1 turn #18 AWG enameled 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 or equivalent. Pin numbers shown are for a 14-pin package.

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 ⅓ 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.

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. You may substitute air-wound coils of the appropriate inductance.

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 you have access to a calibrated signal generator, connect it to the field-strength meter and use different signal levels fed to the device to make a calibration chart. Convert signal-generator voltage ratios to decibels by using the equation

$$dB = 20 \log (V1/V2) \qquad \text{(Eq 3)}$$

where V1/V2 is the ratio of the two voltages and 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 2:1. 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:

$$dB = 10 \log (P1/P2) \qquad \text{(Eq 4)}$$

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 4. 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 Figure 27.29, 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.

27.5 NOISE BRIDGE AND ANTENNA ANALYZER MEASUREMENTS

27.5.1 USING NOISE BRIDGES

The *noise bridge*, sometimes referred to as an *antenna (R-X) noise bridge*, is an instrument for measuring the impedance of an antenna or other electrical circuits. The unit shown here in **Figure 27.30**, 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. With an unknown impedance connected, the R and X controls are

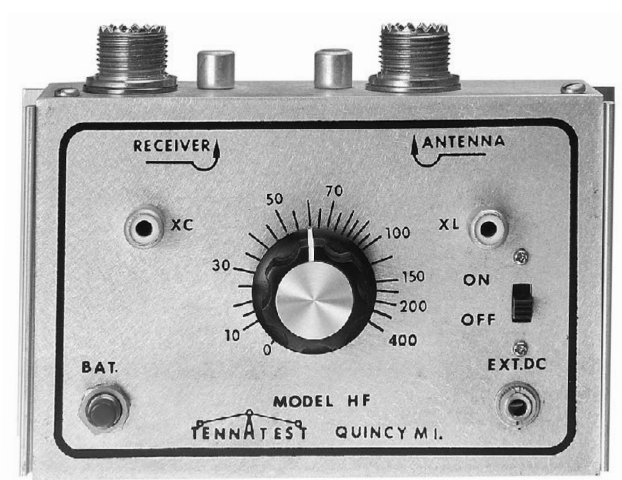


Figure 27.30 — A noise bridge contains a noise source and an external receiver serves as the detector. The bridge is adjusted for minimum noise in the receiver and values for R and X are read from calibrated dials.

adjusted for minimum noise in a receiver that acts as the bridge detector. (See the Bibliography and this book's CD-ROM for an article by Grebenkemper on using R-X noise bridges.) The values of resistance and reactance are then read from the R and X dials.

Measuring Cable Electrical Length with a Noise Bridge

With a noise bridge and a general-coverage receiver, you can easily locate frequencies at which the line in question is a multiple of ½ λ, because a shorted ½ 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 ½ λ at one of the frequencies and calculate the line length (overall accuracy is limited by bridge accuracy and line loss, which broadens the nulls). As an interim variable, you can 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. You will need calibrated test loads as shown in **Figure 27.31**.

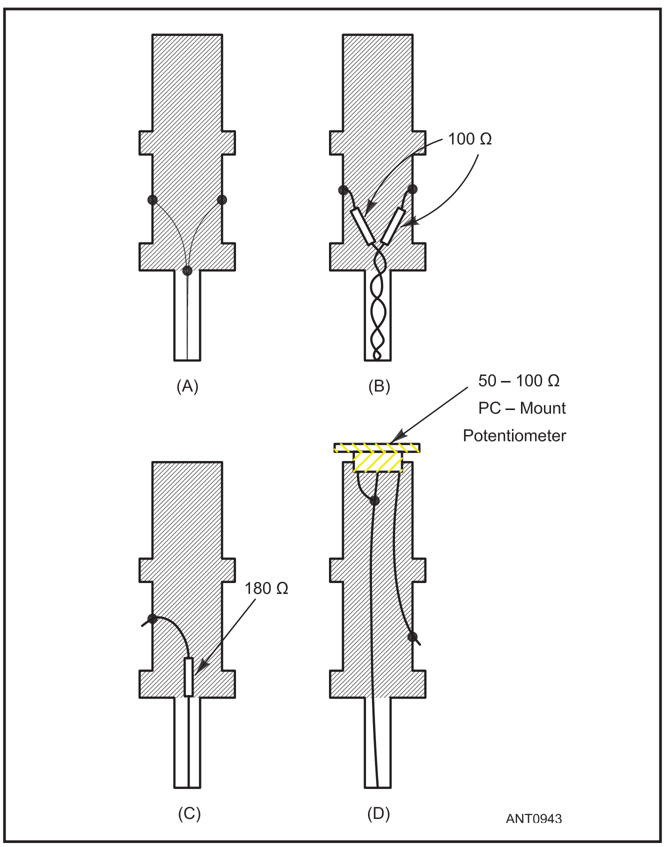


Figure 27.31 — Construction details of the resistive loads used to check and calibrate the noise bridge. Each of the loads is constructed inside a coaxial connector that matches those on the bridge. (Views shown are cross-sections of PL-259 bodies; the shells are not shown.) Leads should be kept as short 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. Use noninductive carbon-film or metal-film resistors.

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 5 for n and the electrical length of the cable.

$$n = \frac{2f_0}{f_{n+2} - f_0} \qquad \text{(Eq 5)}$$

$$f_\lambda = \frac{4f_n}{n} \qquad \text{(Eq 6)}$$

$$\ell = \frac{f_0}{f_\lambda} \qquad \text{(Eq 7)}$$

where

n = cable electrical length in quarter waves, at f_n

f_λ = frequency at which the cable is 1λ

f_0 = frequency at which electrical length is to be determined

ℓ = cable electrical length, in λ

For example, consider a 74-foot length of Carol C1188 foam-dielectric cable (velocity factor = 0.78) to be used on the 10 meter band. Based on the manufacturer's specification, the cable is 2.796 λ at 29 MHz (f_0). Nulls were found at 24.412 (f_n) and 29.353 (f_{n+2}) MHz. Eq 5 yields n = 9.88, which produces f_λ = 9.883 MHz from Eq 6 and ℓ = 2.934 λ from Eq 7. 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.

27.5.2 USING ANTENNA ANALYZERS

Antenna or SWR analyzers that employ a low-level variable frequency signal source and wideband RF detectors have become very popular for antenna and transmission line measurements, largely displacing the noise bridge and dip meter as the preferred tool for antenna system measurements. The basic operation of the instrument is well-covered by the instrument user's manual and supplemented by descriptions of its use in several measurement techniques described in this chapter.

Peter Shuch, WB2UAQ contributes methods for the first three common analyzer tasks. The Sep 1996 *QST* article by George Badger, W6TC, and others, "SWR Analyzer Tips, Tricks and Techniques: SWR Analyzer Hints" provides other interesting applications of SWR analyzers and the *QST* article by Frederick Hauff, W3NZ, "The Gadget — An SWR Analyzer Add-on" describes a useful test accessory. (Articles are included on this book's CD-ROM and listed in the Bibliography.)

Amateur antenna analyzers are not intended to be precision instruments — values for impedance and reactance should be considered to have accuracy of a few percent. If precision measurements are required use calibrated, laboratory-grade test instrumentation.

Common-mode currents and load imbalance can also cause errors in measurements involving lengths of cable. Use a good-quality choke balun when measuring antenna characteristics so that the outside surface of the cable does not influence the measurement. It may also be necessary to use RF choke techniques if the cable is long enough to pick up significant levels of RF from any nearby transmitters, such as broadcast stations or paging transmitters.

Measuring Line Length

In addition to the analyzer, you'll need a coaxial tee adaptor and a 50-Ω load (see Figure 27.31). Connect the tee adapter to the analyzer. To one arm of the tee, connect the 50-Ω load. Connect the cable under test (CUT) to the other arm. Short the far end of the line with the minimum length connection. Starting at a frequency too low for the line to be λ/4 long, slowly tune the analyzer frequency upward until the SWR decreases to a minimum or reaches 1:1. (The lossier the cable, the higher SWR will be at the minimum.) At that frequency the CUT will be λ/4 long because a shorted λ/4 line is an open-circuit at the other end and the analyzer will only see the 50-Ω load, regardless of the line's characteristic impedance.

Measuring Velocity Factor

Start by determining the frequency at which the line is λ/4 long as described above. To find the velocity factor, divide the line's physical length by the free-space wavelength at the frequency at which the line is λ/4 long. For example, if an 86-foot piece of line is λ/4 long at 7.58 MHz, the velocity factor (VF) = 86 / (984 / 7.58) = (86 × 7.58) / 984 = 0.662.

Measuring Characteristic Impedance

Characteristic impedance changes slowly as a function of frequency, so this measurement must be done near the frequency of interest, f_λ. The characteristic impedance of the coaxial cable is found by measuring its input impedance at two frequencies separated by ¼ f_λ. This must be done when the cable is terminated in a resistive load.

If your analyzer can measure the magnitude of impedance, a λ/4 line's characteristic impedance, Z_0, can be measured by using the formula

$$Z_0 = \sqrt{Z_i \times Z_L}$$

where
Z_i = the input impedance to the line
Z_L = the load impedance.

Terminate the line with a 50-Ω load. At the frequency for which the line is λ/4 long, measure input impedance, Z_i. If the input impedance is 50 Ω, so is the line's characteristic impedance. If the input impedance is some other value, use the equation above. For example, if Z_i = 100 Ω, then

$$Z_0 = \sqrt{100 \times 50} = 70.7\Omega$$

The preceding procedure yields only the magnitude of the characteristic impedance which actually includes some reactance. The measurement procedure to determine complex characteristic impedance follows.

1) Place the 50-Ω load on the far end of the coaxial cable and connect the near end to the analyzer. (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 analyzer approximately ⅛ f_λ below the frequency of interest. Call this frequency f1. Read R_{f1} and X_{f1}. Remember, the reactance reading must be scaled to the measurement frequency.

3) Increase the frequency by exactly ¼ f_λ. Call this frequency f2 and note the readings as R_{f2} and X_{f2}.

4) Calculate the characteristic impedance of the coaxial cable using Eqs 8 through 13. A scientific calculator or spreadsheet is helpful for this.

$$R = R_{f1} \times R_{f2} - X_{f1} \times X_{f2} \qquad \text{(Eq 8)}$$

$$X = R_{f1} \times X_{f2} + X_{f1} \times R_{f2} \qquad \text{(Eq 9)}$$

$$Z = \sqrt{R^2 + X^2} \qquad \text{(Eq 10)}$$

$$R_0 = \sqrt{Z}\,\cos\left[\frac{1}{2}\tan^{-1}\left(\frac{X}{R}\right)\right] \qquad \text{(Eq 11)}$$

$$X_0 = \sqrt{Z}\,\tan\left[\frac{1}{2}\tan^{-1}\left(\frac{X}{R}\right)\right] \qquad \text{(Eq 12)}$$

$$Z_0 = R_0 + jX_0 \qquad \text{(Eq 13)}$$

where Z_0 is the characteristic impedance of the transmission line.

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 + (9.883/4) = 30.236 MHz

R_{f2} = 50 Ω

X_{f2} = –24 Ω × (10/30.236) = –7.9 Ω

When used in Eqs 8 through 13, these data yield:

R = 3137.59 Ω

X = –900.60 Ω

Z = 3264.28 Ω

R_0 = 56.58 Ω

X_0 = –7.96 Ω

Remember the limitations on accuracy for inexpensive test equipment and be skeptical of data or calculations beyond two significant figures. The level of precision implied here is for illustration purposes only.

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 ¼ f_λ . Loss at other frequencies can be interpolated with reasonable accuracy. This procedure employs a resistor-substitution method using the test loads in Figure 27.31 that provides much greater accuracy than is achieved by directly reading the resistance from the analyzer. (You can also measure loss directly by using a wattmeter to measure power into and out of the line.)

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

$$n = \frac{4f_0}{f_\lambda} \quad \text{(Eq 14A)}$$

where f_0 is the nominal frequency.

Round n to the nearest integer, then

$$f1 = \frac{n}{4} f_\lambda \quad \text{(Eq 14B)}$$

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 analyzer and read resistance and reactance.

3) Disconnect the cable from the analyzer and connect the variable-resistance calibration load in its place. Without changing analyzer frequency, adjust the load resistor to obtain the same resistance and reactance.

5) Remove the variable-resistance load from the analyzer 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

$$\text{loss} = 8.69 \frac{R_i}{R_0} \quad \text{(Eq 15)}$$

To continue this example, Eq 14A gives n = 11.74, so measure the attenuation at n = 12. From Eq 14B, 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.

Antenna Impedance

The impedance at the end of a transmission line can be easily measured using an SWR analyzer, as shown in **Figure 29.32**. 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 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 ½ 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. (See this book's CD-ROM for an article on the Smith Chart.) 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

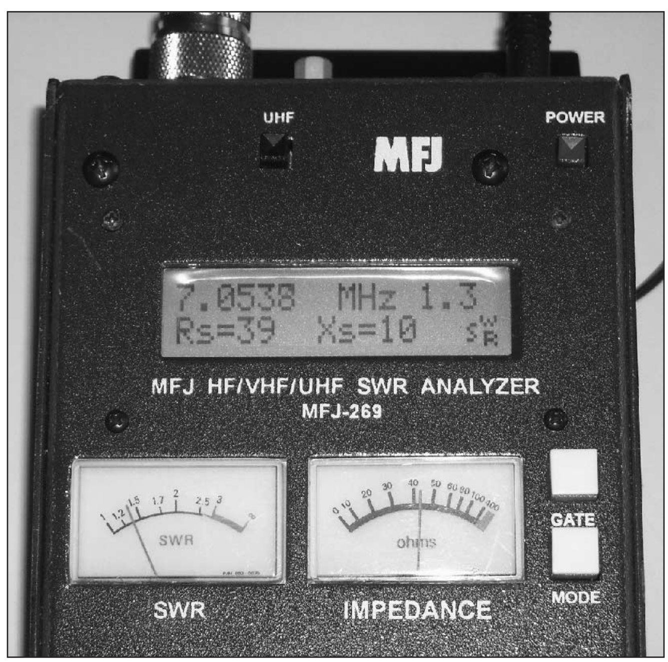


Figure 27.32 — When using an antenna analyzer to measure impedance data, use the numeric display and not the analog meters. In this example, the load impedance is 39 Ω (R_S) in series with a reactance (X_S) of 10 Ω. The sign of the reactance is not given by this instrument.

apply. These problems, too, can lead to significant errors.

4) Last, measurements can be corrected using the transmission-line equation. The *TLW* program included on this book's CD-ROM can do these complicated computations for you. 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 at a number of frequencies. Then use these measurements in the transmission-line equation to determine the actual antenna impedance at each frequency.

When doing the conversions, be careful not to introduce measurement errors as discussed earlier. Such errors will be carried through into the corrected data. This problem is most significant when the transmission line is near an odd multiple of a ¼ λ 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 or if changing the physical orientation of the line or instruments cause significant changes in the measured data. These effects can be minimized by decoupling the line from the antenna and by making the measurements with a transmission line that is approximately an integer multiple of ½ λ. Another clue that there is something amiss with the system is final data that changes erratically with frequency or in ways that aren't typical of antennas or transmission lines. If the data "looks funny," apply extra scrutiny before accepting it as fact.

27.6 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.

A handy "free" TDR is sometimes available from oscilloscopes with a trigger output pulse synchronized with the start of the sweep. The output pulse generally has a very fast rise time and although not adjustable, does serve the same purpose as an external TDR pulse generator.

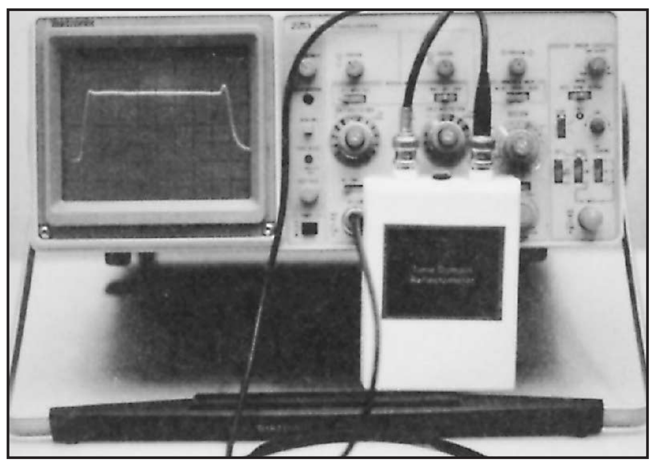


Figure 27.33 — The time-domain reflectometer shown here is attached to a small portable oscilloscope.

27.6.1 HOW A TDR WORKS

A simple TDR consists of a square-wave generator and an oscilloscope. See **Figure 27.33**. 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 **Figure 27.34**. 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.)

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:

$$\ell = \frac{983.6 \times VF \times t}{2} \qquad \text{(Eq 16)}$$

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 **Figure 27.35** consists of a CMOS 555 timer configured as an astable

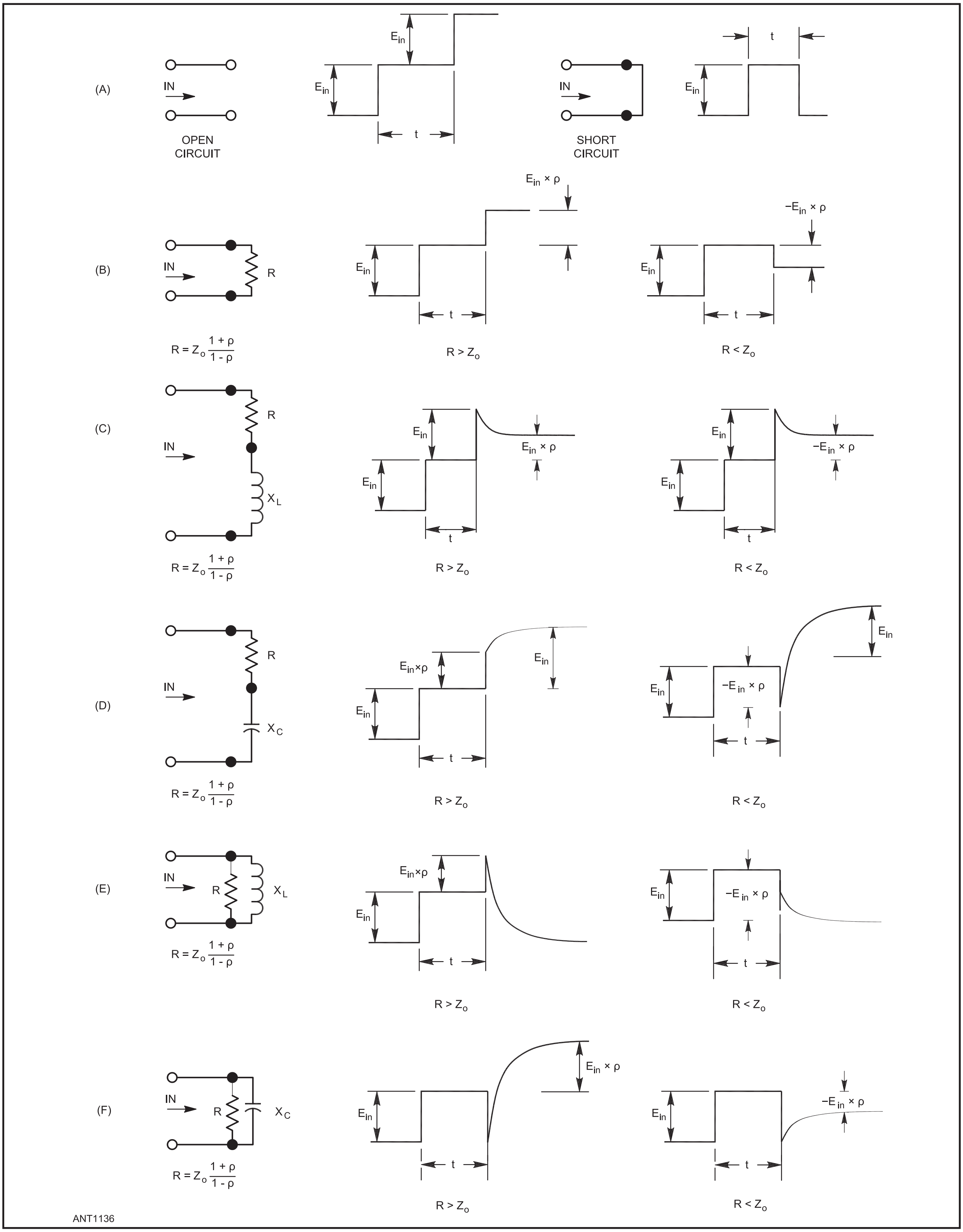


Figure 27.34 — Characteristic TDR patterns for various loads. The location of the load can be calculated from the transit time, t, which is read from the oscilloscope (see text). R values can be calculated as shown (for purely resistive loads only — $\rho < 0$ when $R < Z_0$; $\rho > 0$ when $R > Z_0$). Values for reactive loads cannot be calculated simply.

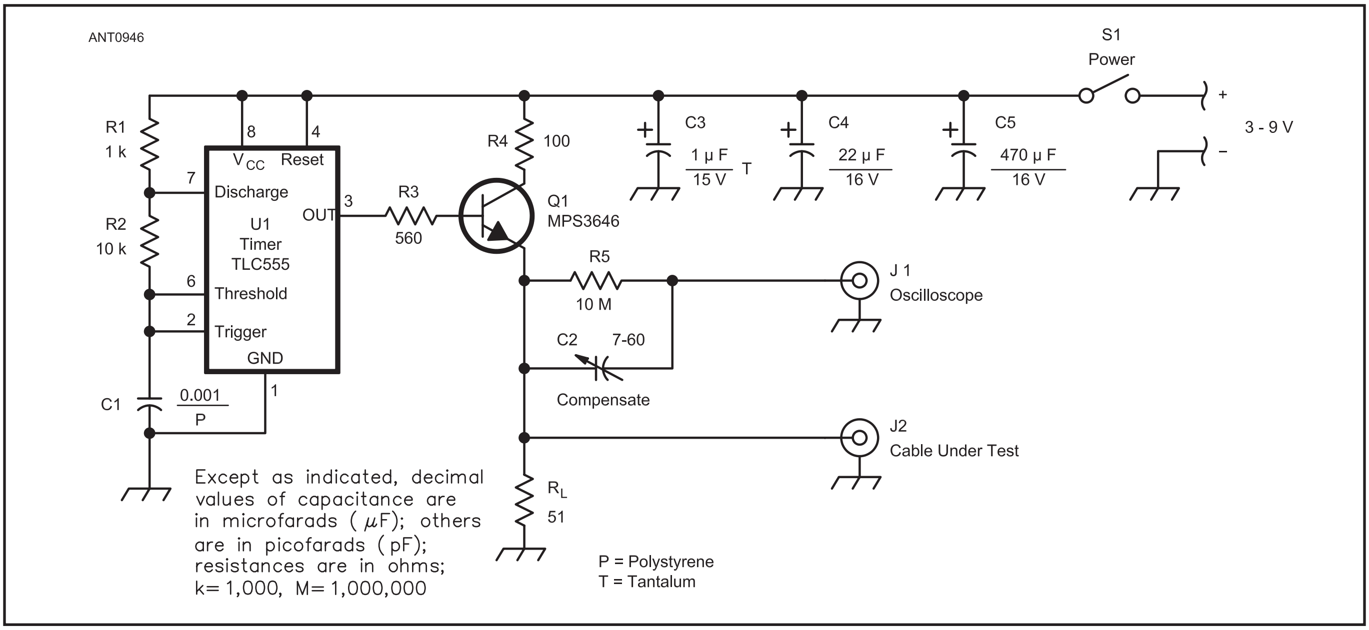


Figure 27.35 — Schematic diagram of the time-domain reflectometer. All resistors are ¼-W, 5% tolerance. U1 is a CMOS 555 timer. Circuit current drain is 10 to 25 mA. When building the TDR, observe the construction cautions discussed in the text.

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 **Figure 27.36**. **Figure 27.37** is the part-placement diagram. The TDR is designed for a 4 × 3 × 1-inch enclosure (including the batteries). S1, J1 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 performance 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 KD5HM 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

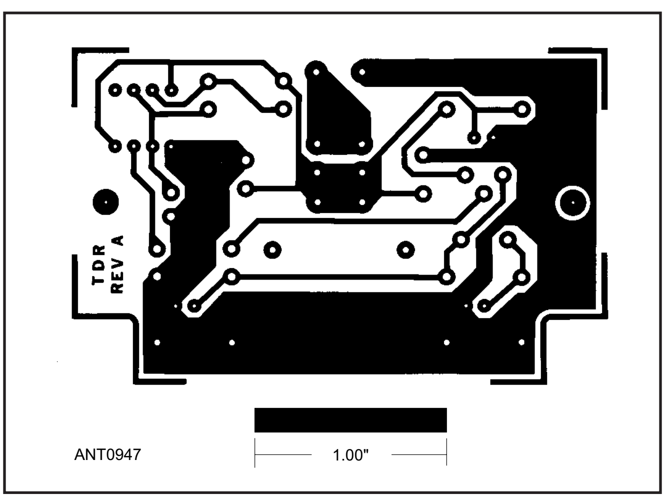


Figure 27.36 — Full-size PC-board etching pattern for the TDR. Black areas represent unetched copper foil. (A PC-board is available from FAR Circuits at www.farcircuits.net.)

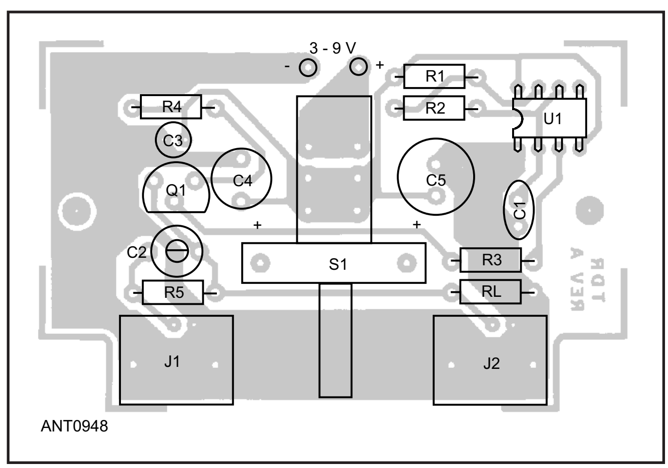


Figure 27.37 — Part-placement diagram for the TDR. Parts are mounted on the nonfoil side of the board; the shaded area represents an X-ray view of the copper pattern. Be sure to observe the polarity markings of C3, C4 and C5.

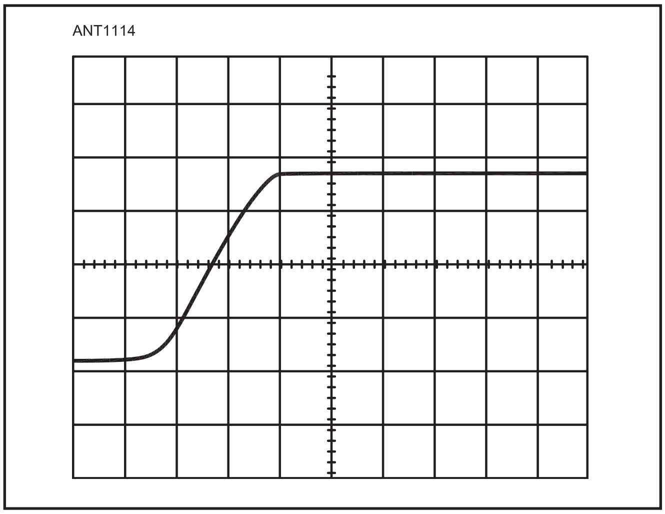


Figure 27.38 — TDR calibration trace as shown on an oscilloscope. Adjust C2 (See Figures 27.35 and 27.37) for maximum deflection and sharpest waveform corners during calibration. See text.

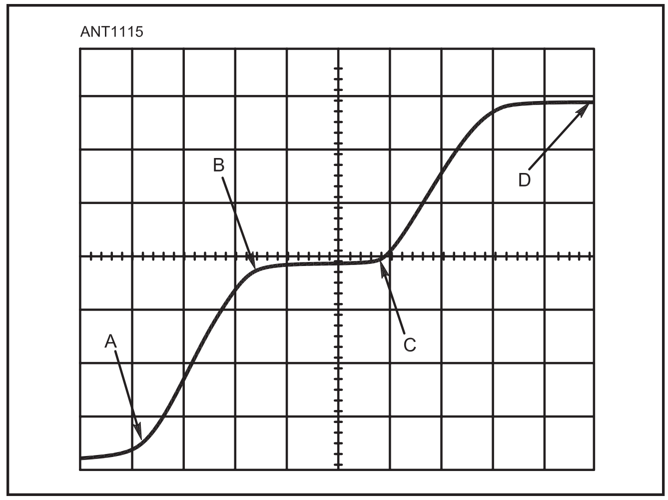


Figure 27.39 — Open-circuited test cable. The scope is set for 0.01 ms per division. See text for interpretation of the waveform.

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.

An alternative that avoids issues associated with loading of the TDR's OSCILLOSCOPE output is to attach the TDR's CABLE UNDER TEST output and the cable under test to a BNC tee connector mounted on the scope input. (In this configuration the OSCILLOSCOPE output is not used.)

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.

27.6.2 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 scope with at least 50 MHz bandwidth 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 the rise time from the TDR fills as much of the scope display as possible (without uncalibrating the timebase). The waveform should resemble **Figure 27.38**. 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 **Figure 27.39**, 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.

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 16 tells us that the 50-Ω section of the cable is

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

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 **Figure 27.40**, where a length of ¾-inch hardline is being tested. The line feeds a 432-MHz vertical antenna at the top of a tower. Figure 27.40 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.

To determine the actual length of the line, use the same

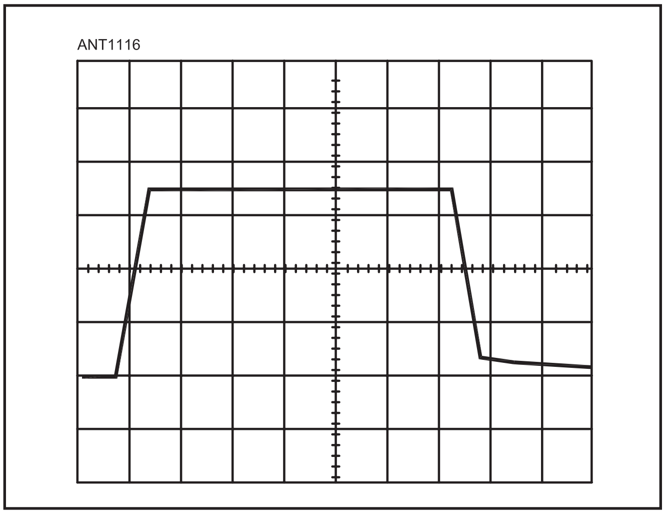


Figure 27.40 — TDR display of the impedance characteristics of the 142-foot hardline run to the 432-MHz antenna at KD5HM. The scope is set for 0.05 ms per division. See text for discussion.

procedure as before: Using the published VF for the hardline (0.88) in Eq 16, the line length is

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

Again, the TDR-derived measurement is in 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.

27.6.3 TDR 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 Figure 27.40 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 Figure 27.40 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 Figure 27.34C because the time constant ($1 \times 10^{-6} \times 50 = 50$ ms) 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.

27.7 VECTOR NETWORK ANALYZER

Professionals make transmission line measurements by employing a *vector network analyzer* (VNA) or the somewhat simpler *reflection-transmission test set* (more on this later). These instruments can make all the necessary measurements quickly and with great accuracy. However, in the past VNAs have been very expensive, out of reach for general amateur use. But thanks to modern digital technology VNAs that work with a laptop computer are now becoming available at prices an amateur might consider. As Paul Kiciak, N2PK (**n2pk.com**) and others have demonstrated, it's even possible to homebrew a VNA with performance that approaches a professional instrument. In addition, the data taken in the frequency domain by the VNA can be transformed to the time domain as a type of time-domain reflectometry as described in the Agilent application note "Comparison of Measurement Performance between Vector Network Analyzer and TDR Oscilloscope." (See Bibliography.)

VNAs are based on reflection and transmission measurements. To use a VNA it is very helpful to have a basic understanding of *scattering parameters* (S-parameters). Microwave engineers have long used these because they have to work with circuits that are large in terms of wavelength, where measurements of forward and reflected power are not easy.

27.7.1 S-PARAMETERS

In the chapter **Transmission Lines**, the reflection coefficient rho (ρ) is defined as the ratio of the reflected voltage (V_r) to the incident voltage (V_i):

$$\rho = \frac{V_r}{V_i} \qquad \text{(Eq 17)}$$

If we know the load impedance (Z_L) and the transmission line impedance (Z_0) we can calculate ρ from:

$$\rho = \frac{Z_L - Z_0}{Z_L + Z_0} \qquad \text{(Eq 18)}$$

Keep in mind that ρ is a complex number (a vector), which we represent by either amplitude and phase ($|Z|$, θ) or by real and imaginary parts ($R \pm jX$). The two representations are equivalent. Neglecting phase, from $|\rho|$ (the magnitude of ρ) we can then calculate SWR. That's very handy, but here we want to do something different. If we have an instrument that measures ρ and we know Z_0 then we can determine Z_L from:

$$Z_L = Z_0\left(\frac{1+\rho}{1-\rho}\right) \qquad \text{(Eq 19)}$$

Measuring ρ is one of the things that VNAs do very well. With a VNA, the measurement can be made at one end of a long transmission line with the load at the other end. The effect of the line can be calibrated out, as mentioned above, so that we are in effect measuring right at the load. Note that the symbol Γ (gamma) is also used to represent the reflection coefficient. The two symbols may be used interchangeably.

This approach can be used directly to measure the impedance and resonant frequency of a single antenna element. By open and short circuiting elements in an array of antenna elements we can determine the mutual as well as self impedances for, and between, all the elements. We can also use this approach to measure component values, inductor Q, etc.

This is an example of a *one-port* measurement; that is, a load at the end of a transmission line. However, to get the most out of a VNA, you need to generalize the above procedure. This is where S-parameters come into play.

VNAs usually have at least two RF connections: the transmit port (T) and the receive port (R). Professional units may have more RF connections. The T output provides a signal from a 50-Ω source and the R port is a detector with a 50-Ω input impedance. Basically we have a transmitter and a receiver. The transmit port uses a directional coupler to provide measurements of the forward and reflected signals at that output. The receive port measures the signal transmitted through the network.

Using incident and reflected voltages, the two-port network representation is now changed, as shown in **Figure 27.41**, where:

V_{1i} = incident voltage at port 1
V_{1r} = reflected voltage at port 1

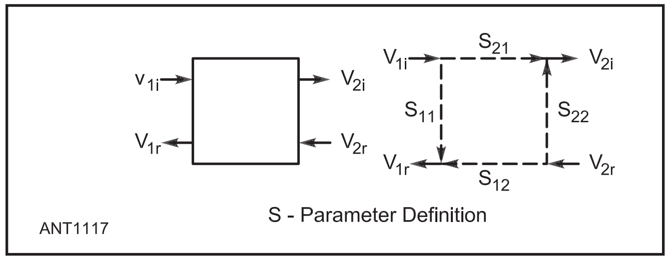


Figure 27.41 — Two-port network with incident and reflected waves.

V_{2i} = incident voltage at port 2
V_{2r} = reflected voltage at port 2

We can write an expression in terms of the incident and reflected voltages:

$$\begin{aligned} b_1 &= S_{11}a_1 + S_{12}a_2 \\ b_2 &= S_{21}a_1 + S_{22}a_2 \end{aligned} \qquad \text{(Eq 20)}$$

where:

$$a_1 - \frac{V_{1i}}{\sqrt{Z_0}} \qquad b_1 - \frac{V_{1r}}{\sqrt{Z_0}}$$
$$a_2 - = \frac{V_{2i}}{\sqrt{Z_0}} \qquad b_2 - \frac{V_{2r}}{\sqrt{Z_0}} \qquad \text{(Eq 21)}$$

We see that a_n and b_n are simply the incident and reflected voltages at the two ports divided by $\sqrt{Z_0}$. Because this is a linear network, $S_{21} = S_{12}$.

What are the S_{ij} quantities? These are called the *S-parameters*, which are defined by:

$$\begin{aligned} S_{11} &\equiv \left.\frac{b_1}{a_1}\right|_{a_2=0} = \left.\frac{V_{1r}}{V_{1i}}\right|_{V_{2i}=0} \\ S_{21} &\equiv \left.\frac{b_2}{a_1}\right|_{a_2=0} = \left.\frac{V_{2r}}{V_{1i}}\right|_{V_{2i}=0} \\ S_{12} &\equiv \left.\frac{b_1}{a_2}\right|_{a_1=0} = \left.\frac{V_{1r}}{V_{2i}}\right|_{V_{1i}=0} \\ S_{22} &\equiv \left.\frac{b_2}{a_2}\right|_{a_1=0} = \left.\frac{V_{2r}}{V_{2i}}\right|_{V_{1i}=0} \end{aligned} \qquad \text{(Eq 22)}$$

Note that the S_{ij} parameters are all ratios of reflected and incident voltages, and they are usually complex numbers. The condition that $a_2 = 0 = V_{2i}$ is the same as saying that port 2 is terminated in a load equal to Z_0 and the network is excited at port 1. This means there is no reflection from the load on port 2, which makes $V_{2i} = 0$. Similarly, if we terminate port 1 with Z_0 and excite port 2, then $V_{1i} = 0 = a_1$.

If we compare Eq 17 to the first line of Eq 22 we see that $S_{11} = \rho_1$, the reflection coefficient at port 1. We can now restate Eq 19 in terms of S_{11}:

$$Z = Z_0\left(\frac{1+S_{11}}{1-S_{11}}\right) \quad \text{(Eq 23)}$$

where Z is the impedance looking into port 1 with port 2 terminated in Z_0. In the case where port 2 does not exist — that is, you are measuring a single impedance (for example, measuring an impedance at port 1 with port 2 open-circuited) or a component, then Z is simply the impedance at that port. Since S_{11} is a standard measurement for VNAs you can calculate Z using Eq 23. In many cases the VNA software will do this calculation for you automatically. You can also measure an impedance at port 2 with port 1 open and determine Z_{22}.

S_{21} represents the ratio of the signal coming out of port 2 (V_{2r}) to the input signal on port 1 (V_{1i}) and is another standard VNA measurement. S_{21} is a measurement of the signal transmission between the ports through the network with port 2 terminated in Z_0; forward gain in most applications.

A full-feature VNA will measure all the S_{ij} parameters at once, but most of the lower-cost units of interest to amateurs are what we call reflection-transmission test sets. What this means is that they only measure S_{11} and S_{21}. To obtain S_{22} and S_{12} we have to interchange the test cables at the ports and run the measurements again. Normally the software will accommodate this as a second entry and we end up with the full set of S_{ij} parameters.

If we do run a full set of S_{ij} parameters then we can transform these to Z_{ij} using the following expressions, assuming that $S_{21} = S_{12}$:

$$Z_{11} = \frac{(1+S_{11})(1-S_{22})+S_{12}^2}{(1-S_{11})(1-S_{22})-S_{12}^2}$$

$$Z_{22} = \frac{(1-S_{11})(1+S_{22})+S_{12}^2}{(1-S_{11})(1-S_{22})-S_{12}^2} \quad \text{(Eq 24)}$$

$$Z_{12} = \frac{2S_{12}}{(1-S_{11})(1-S_{22})-S_{12}^2}$$

27.7.2 RETURN LOSS

Return loss (RL) is another term for S_{11}, the ratio of the reflected voltage to the incident voltage, usually expressed in dB.

$$RL = -20\log\left(\frac{V_{1i}}{V_{1r}}\right) = -10\log\left(\frac{P_{1i}}{P_{1r}}\right)$$

RL is measured as S_{11} by a VNA.

The name stems from measuring how much voltage is returned from a transmission line wave encountering a termination or impedance discontinuity. If the line is terminated in its characteristic impedance, the entire wave is absorbed and none is returned so the "loss" at the reflection is total and RL is infinite. If the line is open- or short-circuited, the entire wave is returned to the source and RL is 0. Note that RL is a positive quantity range from 0 (no transfer of wave energy at the termination) to infinite (all wave energy transferred to the termination). Negative RL would describe a voltage gain.

To convert from return loss to SWR, the following formulas are used:

$$|\rho| = 10^{-\frac{RL}{20}}$$

and

$$SWR = \frac{1+|\rho|}{1-|\rho|}$$

So, for example, a return loss of 20 dB is a reflection coefficient of 0.1, and an SWR of 1.22. A return loss of 10 dB is a reflection coefficient of 0.316, and an SWR of 1.92.

One of the most common measurements made is the standing wave ratio of an antenna. A low SWR means that the antenna input impedance is close to that of the measuring reference impedance. These measurements are from the July/August 2004 *QEX* article by McDermott and Ireland (see Bibliography) and show the magnitude of the return loss versus frequency for a KT34XA triband Yagi antenna at the end of 300 feet of hardline cable.

The resonance points are clearly visible. **Figure 27.42** shows the return loss of the antenna swept from 1 MHz to 50 MHz. The 20 meter, 15 meter and 10 meter band resonances are easily seen. (RL increases toward the bottom of the chart.) **Figure 27.43** shows a close-up of the return loss from 13.5 to 14.5 MHz. **Figure 27.44** shows this same close-up on a Smith Chart.

27.7.3 USING A VECTOR NETWORK ANALYZER

Discussing how to use a VNA is beyond the scope of this chapter but the user's manual for a VNA will explain the technique of calibrating that particular VNA and using it to make measurements. The second step will be the use of the measurements by computer software to display and transform the measurements into the desired parameters.

In addition, there are many online tutorials and application notes (see the Bibliography entries for Agilent) and the text *Microwave Electronics* by Pozar also goes into some detail about what the various measurements are and how they are made.

Array Measurement Example

The process of building and properly tuning a phased array often involves making a number of different measurements to achieve a desired level of performance, as was pointed out in the **Multielement Arrays** chapter. This section is adapted from material written by Rudy Severns, N6LF.

After erecting an array we would like to measure the resonant frequency of each element, the self-impedances of each element and the mutual impedances between the elements. We will also want to know these impedances over the whole operating band to help design a feed network. When building the feed network, we may need to check the values and Qs

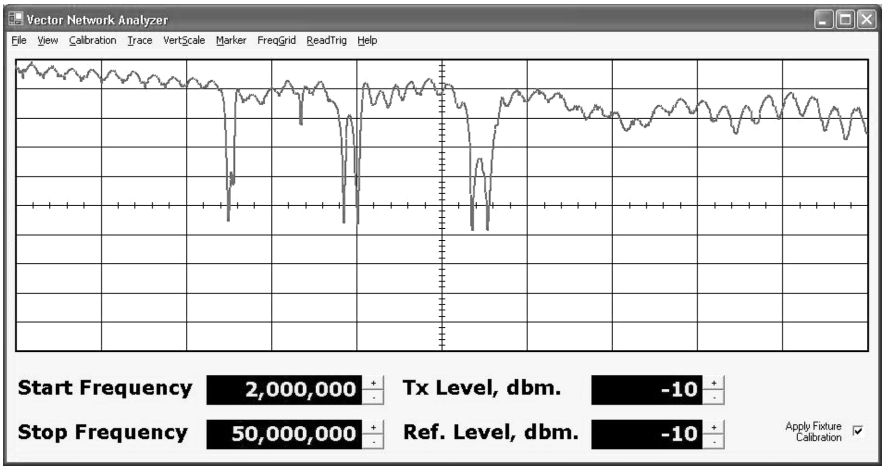


Figure 27.42 — Return loss of KT34XA antenna through 300 feet of hardline. Vertical scale is 5 dB/div. The three resonances at 20, 15, and 10 meters are clearly visible.

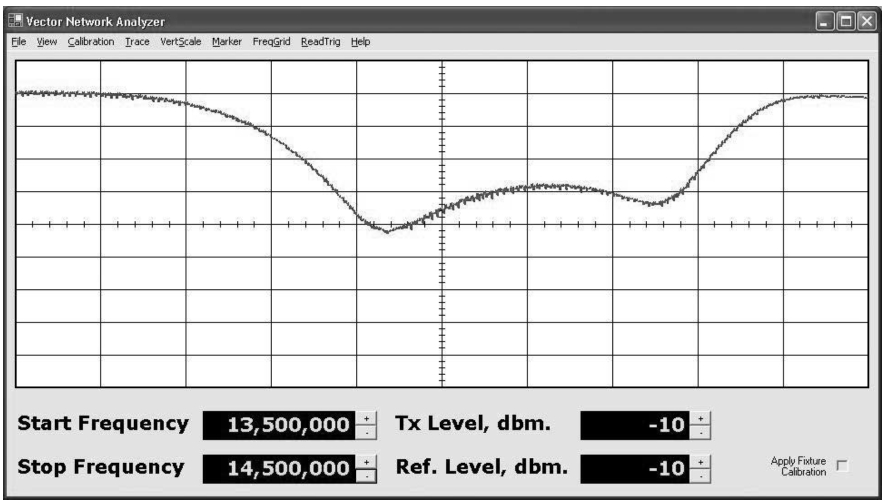


Figure 27.43 — Return loss of Figure 27.42 from 13.5 MHz to 14.5 MHz. A 26 dB return loss (best case at 13.94 MHz) is an SWR of 1.105 (at the ham-shack end of the feed line). Vertical scale is 5 dB/div.

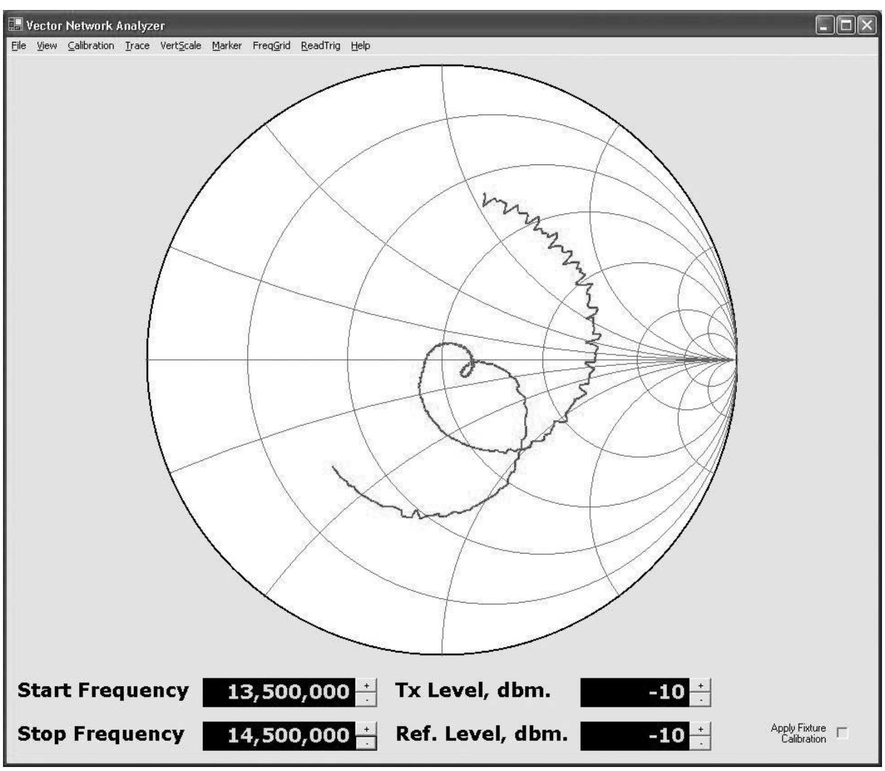


Figure 27.44 — Return loss of Figure 27.43 on a Smith Chart.

of the network elements and we will want to determine the electrical lengths of transmission lines.

Final tuning of the array requires that the relative current amplitudes and phases in each element be measured and adjusted, if necessary. We also will want to determine the SWR at the feed point. Doing all of this even moderately well can require quite a bit of equipment, some of which is heavy and requires ac line power. This can be a nuisance in the field, especially if the weather is not cooperating.

In a completed array with its feed network, the network can be excited by the VNA at the feed point and the relative current amplitudes and phases at each element can be measured over a frequency band. Then, adjustments can be made as needed. When the final values for the current amplitudes and phases are known, these values can be put back into an array model in a program like *EZNEC* to determine the pattern of the array across the whole frequency band. A multielement array actually behaves as a multiport network, so using a VNA is a natural solution to the measurement problem.

HF arrays are also large in terms of wavelength. The techniques for measuring forward and reverse powers work well even at 160 meters. For example, even though the array elements may be 100 feet apart, you can place your instruments in a central location and run cables out to each element. The effect of the cables from the VNA to the elements can be absorbed in the initial calibration procedure so the measurements read out at the VNA are effectively those at each element. In other words, the measurement reference points can be placed electrically at the base of the element, regardless of the physical location of the instrumentation and the interconnecting cables.

The discussion of S-parameters in the previous section can be viewed as measuring the characteristics of a 2-element array (see **Figure 27.45**) if one element is attached to the end of the transmission lines at each port as in **Figure 27.46**.

In the case of such an array, S_{21} represents the signal transmission due to the coupling between the elements. i.e. the signal coupled to element 2 as a result of a signal applied to element 1. Transmission lines are assumed to have Z_0 = 50 Ω (or the characteristic impedance of the overall system) and may be of any length required by the size of the array.

S-parameters can be determined for an array with any number of elements. In an n-port S-parameter measurement, all ports are terminated in Z_0 at the same time. Measurements are made between one set of ports at a time and repeated until all pairs of ports are measured.

To illustrate the principles of using a VNA we will use a simple 2-element array like that shown in Figure 27.45. To design a feed network to drive this array we need to know the input impedance of each element (Z_1 and Z_2) as a function of the drive currents (I_1 and I_2). The input impedances will depend on the self impedance of each element, the coupling

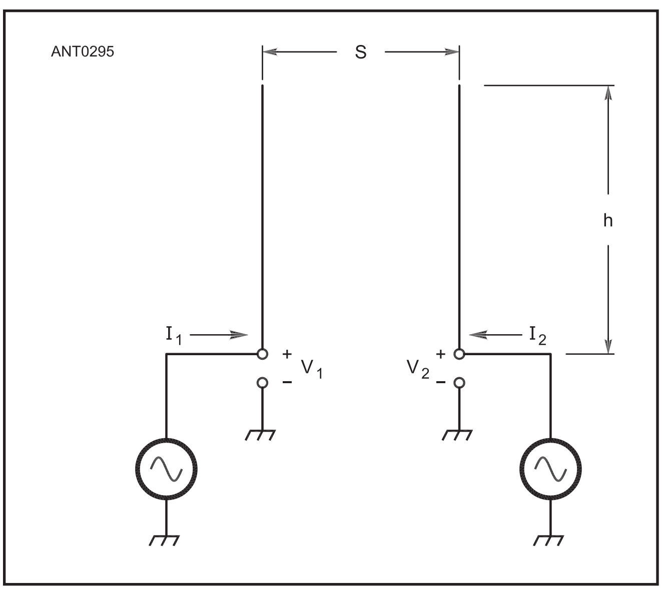


Figure 27.45 — A 2-element array, where h is the element height and S is the spacing between the elements.

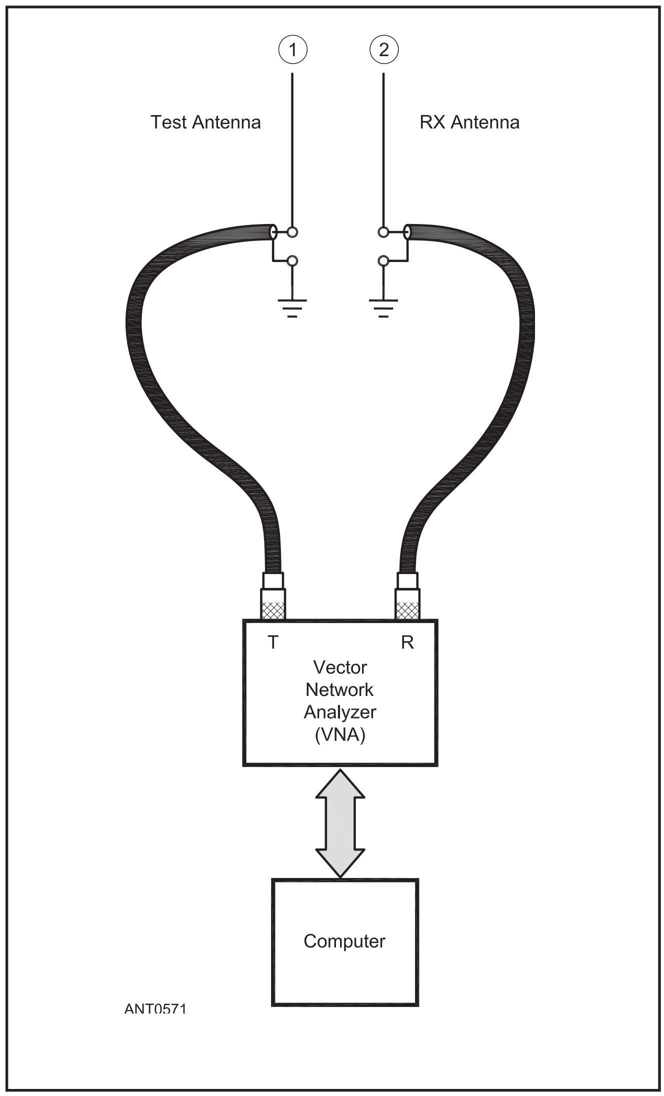


Figure 27.46 — Test setup to measure a 2-element array using a VNA.

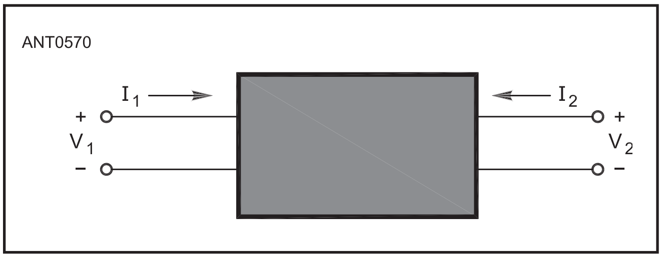


Figure 27.47 — Two-port representation of currents and voltages in the 2-element array in Figure 27.45.

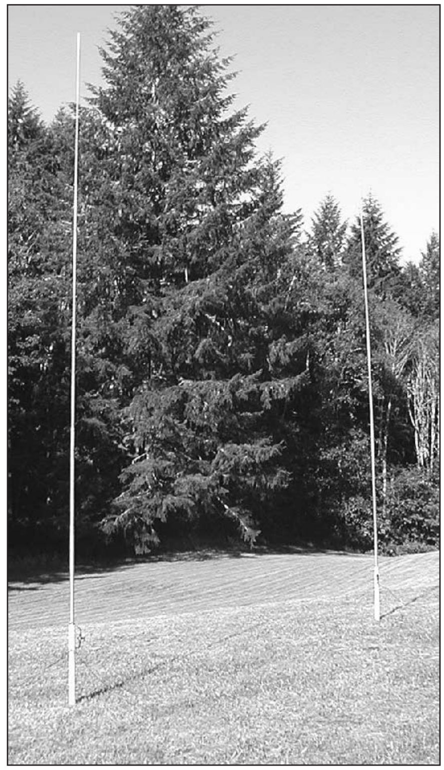


Figure 27.48 — 2-element 20 meter phased array (*Photo courtesy N7MQ*).

between them (the mutual impedance) and the drive currents in each element. To manage this problem we can represent a 2-element array as a two-port network, as shown in **Figure 27.47**. And we can relate the port voltages, currents and impedances with Eq 25:

$$V_1 = Z_{11}I_1 + Z_{12}I_2$$

$$V_2 = Z_{21}I_1 + Z_{22}I_2 \qquad \text{(Eq 25)}$$

Normally we know I_1 and I_2 from the design of the array, but we need to determine the resulting element impedances. That's the challenge. Fortunately, an array is a linear network, so $Z_{12} = Z_{21}$, which means we need only determine three variables: the self impedances Z_{11} and Z_{22} and the mutual impedance, Z_{12}.

Once we know Z_{11}, Z_{12}, Z_{22} and are given I_1 and I_2, we can determine the feed point impedances at each element from:

$$Z_1 = Z_{11} + \left(\frac{I_2}{I_1}\right) Z_{12}$$

$$Z_2 = Z_{22} + \left(\frac{I_1}{I_2}\right) Z_{12} \qquad \text{(Eq 26)}$$

This is the conventional approach. However, there are some problems here. We have to be able to accurately measure either voltages and currents or impedances in multiple elements that may be separated by large fractions of a wavelength. In addition, accurate measurements of current, voltage and impedance become increasingly more difficult as we go up in frequency.

It turns out that we can get the information more easily by measuring incident and reflected voltages at the ports and from those measurements determine the feed point impedances. A VNA is an instrument for measuring these voltages. It turns out to be easier to measure the ratios of two voltages rather than their absolute values.

The measurement setup using a VNA for a 2-element array is shown in Figure 27.46. A good way to illustrate the use of a VNA for array measurements is to work through an example with a real array. **Figure 27.48** is a picture of a 2-element 20 meter phased array built by Mark Perrin, N7MQ.

Each element is λ/4 (self resonant at 14.150 MHz) and spaced λ/4 (17 feet 5 inches). In the ideal case, both elements would have the same current amplitude with a 90° phase difference. This gives the cardioid pattern shown in the **Multielement Arrays** chapter. There are many schemes for correctly feeding such an array. The one used in this example uses two different 75-Ω transmission lines (one λ/4 and the other λ/2, electrically), as described by Roy Lewallen, W7EL and in Orr and Cowan. (See Bibliography.)

The first task is to resonate the elements individually. With the VNA set to measure S_{11} phase, we will get a graph like that shown in **Figure 27.49**.

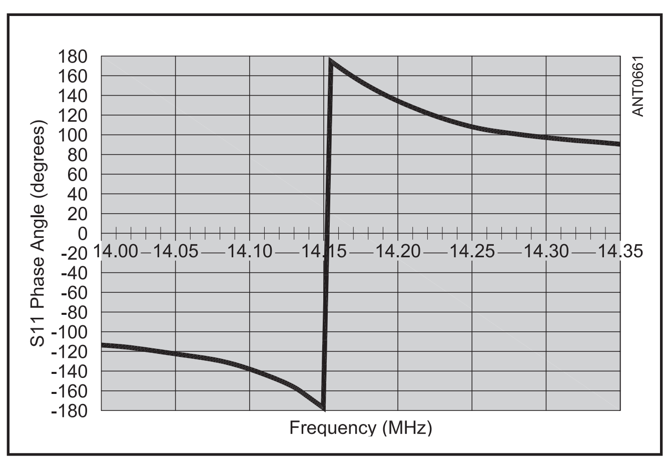


Figure 27.49 — S_{11} phase plot for an individual element.

At the λ/4 resonant frequency (f_r) we will see a sharp phase transition as we go from –180° to +180°. This is typical of any series resonant circuit. The length of each element is adjusted until the desired f_r is achieved. This is a very sensitive measurement. You can see the shift in f_r due to the wind blowing, the length of the element changing as it heats up in the sun or any interactions between the feed line and the antenna as you move the feed line around. In fact this is a very good point in the process to make sure everything is mechanically stable and free of unexpected couplings. Usually you will find it necessary place choke baluns on each element to reduce stray coupling.

The next step is to determine the self (Z_{11} and Z_{22}) and mutual (Z_{12}) impedances from which the actual driving point impedances present when the array is excited can be determined. There are two ways to go.

First, we can simply use the VNA as an impedance bridge — ie, make two S_{11} measurements at one element, first with the other element open (Z_{11} or Z_{22}) and then with it shorted (Z_1 or Z_2). We can convert the S_{11} measurements to impedances using Eq 23. The value for Z_{12} can be obtained from:

$$Z_{12} = \pm\sqrt{Z_{11}(Z_{11} - Z_1)}$$
$$Z_{12} = \pm\sqrt{Z_{22}(Z_{22} - Z_2)} \quad \text{(Eq 27)}$$

The second approach is to do a full two-port S-parameter measurement (S_{11}, S_{21}, S_{12} and S_{22}) and derive the impedances using Eq 23. Both approaches will work but the second approach has the advantage that the ± ambiguity in Eq 27 is eliminated.

For this example, the impedance values from the measurements at 14.150 MHz, turn out to be:

$$Z_{11} = 51.4 + j0.35$$

$$Z_{22} = 50.3 + j0.299 \quad \text{(Eq 28)}$$

$$Z_{12} = 15.06 + j19.26$$

With these values we can now determine the feed point impedances from:

$$Z_1' = Z_{11} + \frac{I_2}{I_1} Z_{12}$$
$$Z_2' = Z_{22} + \frac{I_1}{I_2} Z_{12} \quad \text{(Eq 29)}$$
$$\frac{I_1}{I_2} = -j$$

Note that $-j$ represents the 90° phase shift between the currents. Substituting the values from Eq 28 into Eq 29:

$$Z_1' = 32.09 - j14.7 \quad \text{(Eq 30)}$$

$$Z2' = 69.61 + j15.32$$

With these impedances in hand we can now design the feed network. In this particular example however, we have decided to use the λ/4 and λ/2 cables as described by Lewallen and accept the results. So we now proceed to cut and trim the two cables to length.

Again, there are two ways to go. First we can determine the frequency at which each cable is λ/4 long. At this point the input impedance of the cable will be equivalent to a series-resonant circuit and we can simply measure the phase of S_{11} as we did earlier for f_r and get a plot like that shown in Figure 27.49. In this example the λ/4 resonant frequencies of the two cables are 7.075 MHz and 14.150 MHz.

The second approach would be to measure S_{21} for each cable at 14.150 MHz. The phase shift in S_{21} tells you how long the cable is, in degrees, at a given frequency. Because there is a small variation in cable characteristics with frequency (dispersion) this approach is slightly more accurate since it is done at the desired operating frequency. But this is not very large effect at HF.

This brings us down to the final measurements, which are to check that the relative current amplitudes and phases between the two elements are correct. We can then determine the feed point SWR. The phase and amplitude ratios are made using the S_{12} capability of the VNA and the test setup shown in **Figure 27.50**.

The VNA transmit port is connected to the normal feed point. A current sensor (see the **Multielement Arrays** chapter for a discussion of current sensors) is inserted at the base of element 1 and the output of the sensor is returned to the

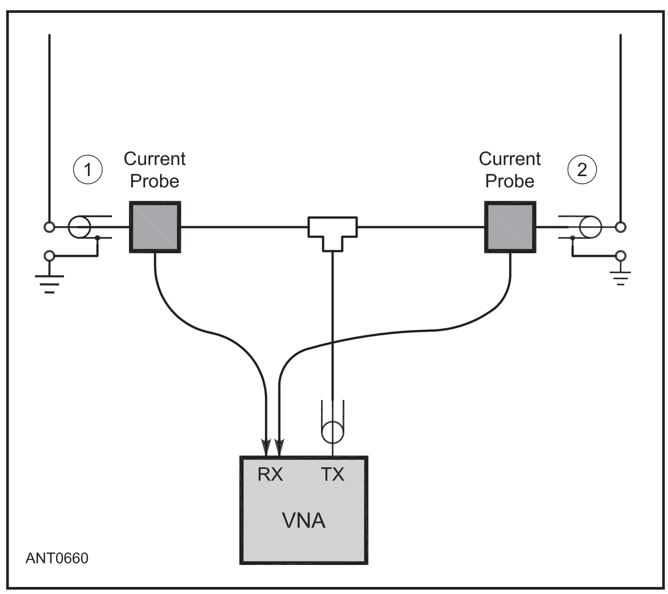


Figure 27.50 — Current phase and amplitude ratio test setup.

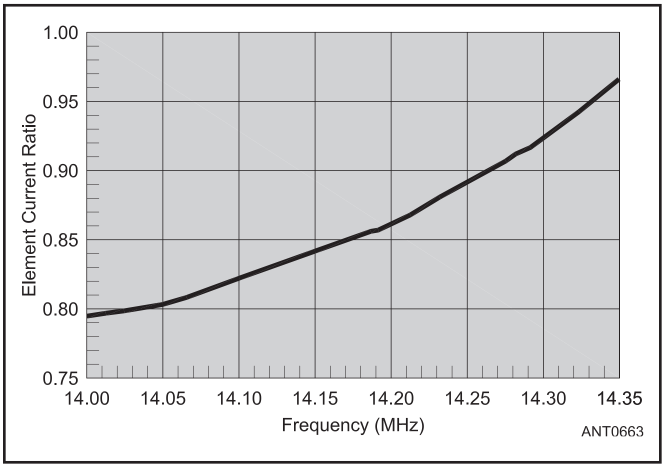


Figure 27.51 — Measured element current ratio over the 20 meter band.

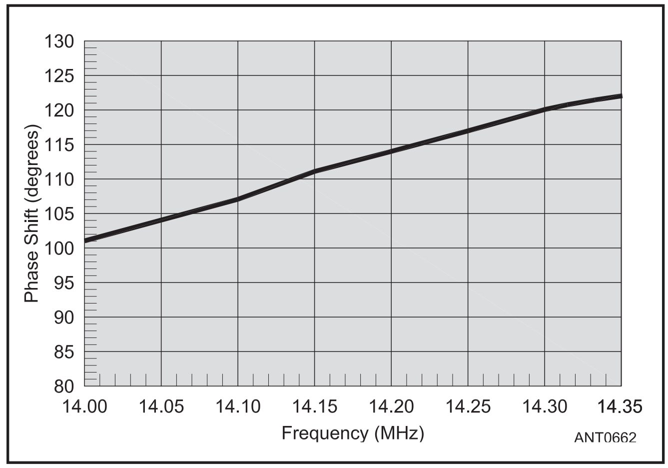


Figure 27.52 — Measured relative current phase shift over the 20 meter band.

detector or receive port of the VNA. A calibration run is then made to normalize this path. That makes it the reference.

Next, the current sensor is shifted to element 2. The amplitude and phase plots for S_{12} obtained at this point will be the desired relative phase shift and amplitude ratio between the currents in the array when driven at the normal feed point. **Figures 27.51** and **27.52** show the behavior of the example array over the 20 meter band. Note that the amplitude ratio has been converted from dB. We can now use these values in a *EZNEC* model of the array to determine the actual radiation pattern.

Obviously the W7EL feed scheme is not perfect, but it has a definite advantage of simplicity. If better performance is desired we can use the values of Z_1' and Z_2' determined earlier to design and fabricate a new feed network and then proceed to evaluate its performance in the same way.

The final measurement is to connect the transmit port of the VNA to the feed point and measure S_{11}. From this we can calculate the SWR:

$$SWR = \frac{1+|S_{11}|}{1-|S_{11}|}$$

In this example, the return loss, $|S_{11}|$, is about 19 dB over the entire 20 meter band. This corresponds to SWR= 1.25:1.

27.8 ANTENNA FIELD MEASUREMENTS

Of all the measurements made in Amateur Radio systems, perhaps the most difficult and least understood are various measurements of the radiated field from 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 probably your 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 that are closely allied to those used in an antenna measuring event or contest. With these procedures you can make measurements 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 was originally published in November 1974 *QST*. The conventions used by amateurs to plot radiation patterns and antenna measurements are covered in the **Antenna Fundamentals** chapter.

27.8.1 FIELD MEASUREMENT BASICS

An antenna is simply a transducer or coupler between a suitable feed line and the environment surrounding it. In addition to the 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 while comparing different antennas, you must standardize the environment surrounding the antenna. Ideally, you want to make measurements 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 consists of a number of linear elements (straight lengths of rod or wire that 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, and can be lost entirely.

In a transmitting antenna, the reflected power travels back down the feed line to the transmitter, where it changes the load impedance presented to that transmitter. The amplifier output controls are customarily altered during the normal tuning procedure to obtain maximum power transfer to the antenna. You can still 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 the **Transmission Lines** chapter.)

Similarly, a mismatched receiving antenna may be matched into the receiver front end for maximum power transfer. In any case you should clearly keep in mind that the feed point mismatch does not affect the radiation characteristics of an antenna. It can only affect the system efficiency when heating losses are considered.

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 is essentially resistive and matched to the feed line. 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 the **Transmission Lines** chapter.

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 that 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, you should 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 will become 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. See the **Transmission Line System Techniques** for additional details on this problem.

27.8.2 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 than the true gain. Based on the 30° phase error alone, it can be shown that the minimum range distance is approximately

$$S_{min} = 2\frac{D^2}{\lambda} \qquad \text{(Eq 31)}$$

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

$$\text{Gain} = \frac{4\pi A_e}{\lambda^2} \qquad \text{(Eq 32)}$$

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

$$D^2 = G\frac{\lambda^2}{4\pi} \qquad \text{(Eq 33)}$$

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

$$S_{min} = G\frac{\lambda}{2\pi} \qquad \text{(Eq 34)}$$

and for a circular aperture of

$$S_{min} = G\frac{2\lambda}{\pi^2} \qquad \text{(Eq 35)}$$

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 Yagi, whose gain is no greater than the lowest 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 you should choose a site where the test-antenna location is near the center of a large open area and the source antenna is 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 **Figure 27.53**. The use of a flat, uniform ground surface assures that there will be essentially a mirror reflection, even though the reflected energy may be slightly weakened (absorbed) by the surface material (ground). In order to perform an analysis you should 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.

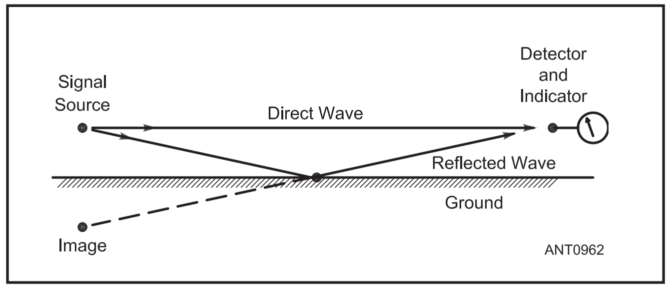


Figure 27.53 — On an antenna test range, energy reaching the receiving equipment may arrive after being reflected from the surface of the ground, as well as by the direct path. The two waves may tend to cancel each other, or may reinforce one another, depending on their phase relationship at the receiving point.

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 **Figure 27.54**.

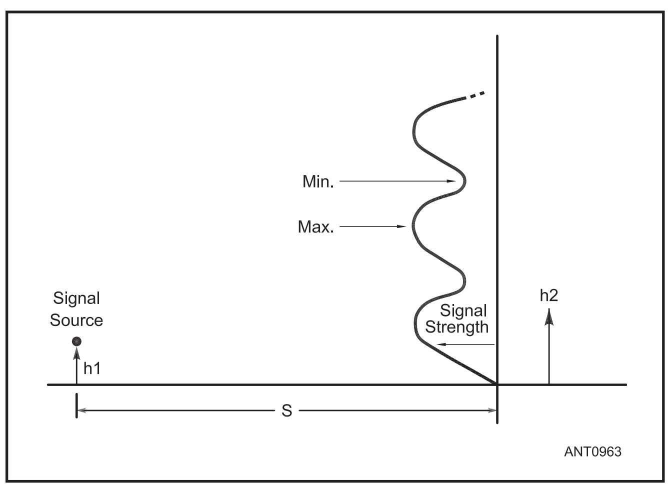


Figure 27.54 — The vertical profile, or plot of signal strength versus test-antenna height, for a fixed height of the signal source above ground and at a fixed distance. See text for definitions of symbols.

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

$$h2 = n\frac{\lambda}{4}\times\frac{S}{h1} \qquad \text{(Eq 36)}$$

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 you to determine the approximate location of the source antenna 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 absolutely necessary but is closely related to the particular antenna under test.

In practice these formulas are useful only to initialize the range setup. 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 2-element Yagi that you move 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 that 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 ½ dB can be clearly distinguished.

Once you conduct these initial range measurements 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. Place the test antenna with the center of its aperture at the height h2 where maximum signal was found. Tilt the test antenna 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.

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

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.

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

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.

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

Plot the measured profile of vertical signal level versus height. 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. **Figure 27.55** 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

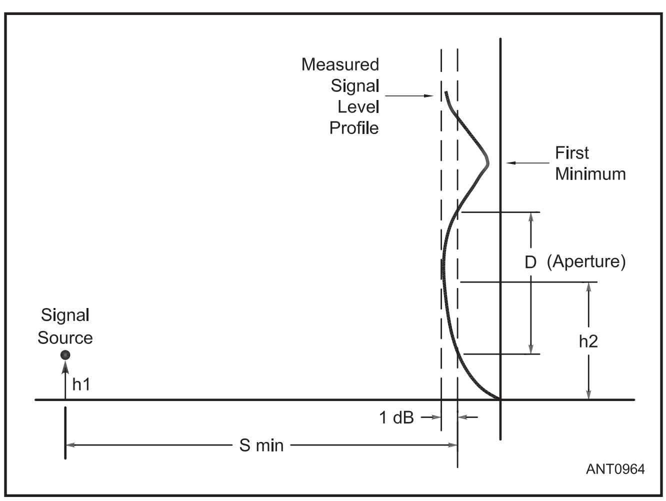


Figure 27.55 — Sample plot of a measured vertical profile.

with apertures that 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 that 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 that 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 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.

27.8.3 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 you remove the test antenna and place the standard-gain antenna with its aperture at the center of location where the test antenna was located. Measure the difference in signal level between the standard and the test antennas and add to or subtract from the gain of the standard-gain antenna to obtain the absolute gain of the test antenna. Here, *absolute* means with respect to a point source with 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. We assume 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 for VHF and low-UHF bands has been suggested by Richard F. H. Yang (see Bibliography). Shown in **Figure 27.56**, it consists of two in-phase dipoles ½ λ apart and backed up with a ground plane 1 λ square. (It is recommended that the builder cut the dipoles close to their free-space length and trim to resonance.)

In Yang's original design, the stub at the center is a balun formed by cutting two longitudinal slots of ⅛-inch width, diametrically opposite, on a ¼-λ section of ⅞-inch rigid 50-Ω coax. An alternative method of feeding is to feed RG-8 or RG-213 coax through slotted ¾-inch copper tubing with a

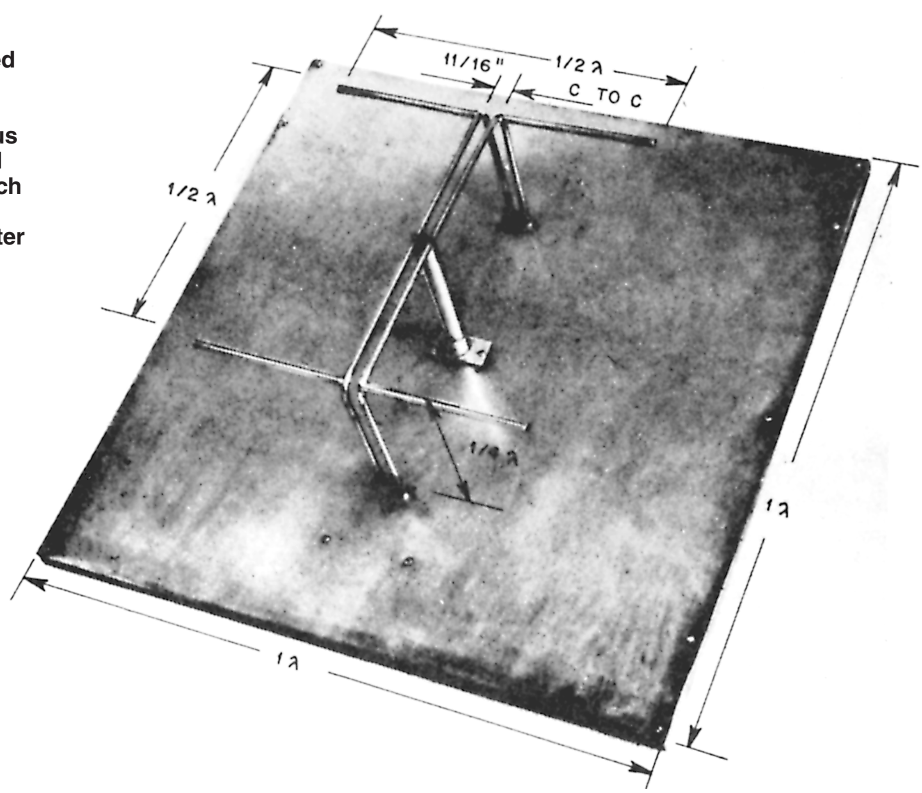


Figure 27.56 — Standard-gain antenna. When accurately constructed for the desired frequency, this antenna will exhibit a gain of 7.7 dB over a dipole radiator, plus or minus 0.25 dB. In this model, constructed for 432 MHz, the elements are ⅜-inch diameter tubing. The phasing and support lines are of 5⁄16-inch diameter tubing or rod.

⅞-inch OD. (Due to variations in ID/OD for stock copper tubing, either bring a section of the rigid coax or take careful measurements with a caliper to check for fit between the coax and the balun 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 9.85 dBi (7.7 dBd gain over a dipole in free space) with an accuracy of ± 0.25 dB. (The balun is described in detail in the original article.)

At 1296 MHz it may be more practical to build a reference horn out of sheet metal as described by Paul Wade, W1GHZ on his website. (**www.w1ghz.com**) The waveguide section can also be made out of sheet metal.

27.8.4 RADIATION PATTERN MEASUREMENTS

Of all antenna measurements, the radiation pattern is the most demanding in measurement and the 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* means parallel to the electric field that is the polarization plane and *H-plane* means parallel to the magnetic field in free space. The electric field and magnetic field are always perpendicular to each other in a plane wave as it propagates through space.

When the antenna is located over real Earth, the terms *azimuth* and *elevation* planes are commonly used, since the frame of reference is the Earth itself, rather than the electric and magnetic fields in free space. For a horizontally polarized antenna such as a Yagi mounted with its elements parallel to the ground, the azimuth plane is the E-plane and the elevation plane is the H-plane.

The technique to obtain these patterns is simple in procedure but requires more equipment and patience than does making a gain measurement. First, a suitable mount is required that 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 that 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 **Figure 27.57**. (See the **Antenna Fundamentals** chapter for more information on coordinate scales.)

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

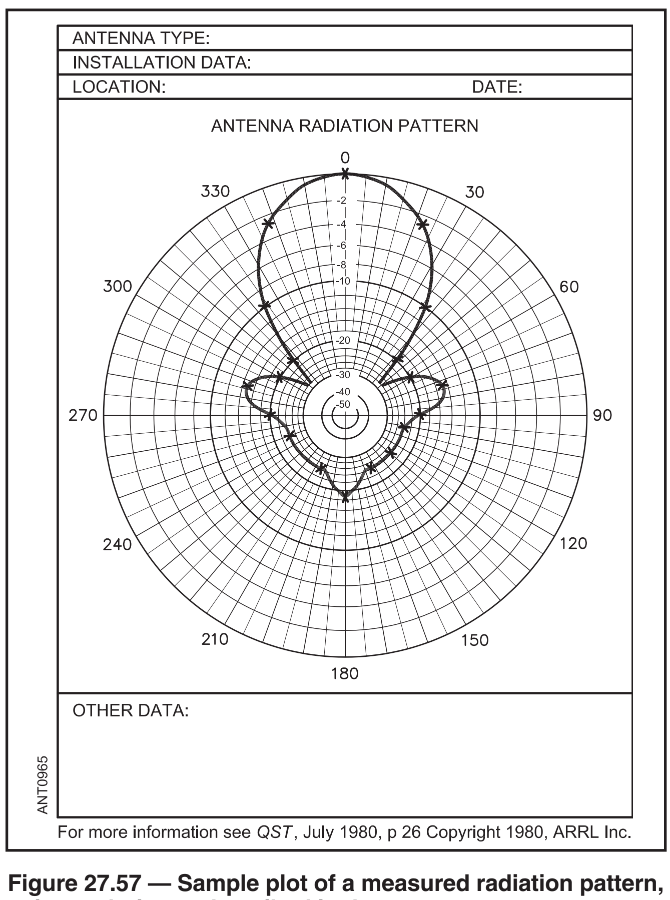


Figure 27.57 — Sample plot of a measured radiation pattern, using techniques described in the text.

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

$$\text{Gain (isotropic)} \cong \frac{41{,}253}{\theta_E \phi_H} \qquad \text{(Eq 37)}$$

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, where any side-lobes are well suppressed. (To obtain gain in dBi, take the log of isotropic gain and multiply by 10.)

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 the **HF Yagi and Quad Antennas** chapter) the expected free-space gain of a Yagi with a boom length of 2 λ is about 13 dBi; its gain, G, equals 20. Using the above relationship, the product of $\theta_E \times \phi_H \approx 2062$ square degrees. Since a Yagi produces a nearly symmetric beam shape in cross section, $\theta_E = \phi_H = 45°$. Now if the measured values of θ_E and ϕ_H are much larger than 45°, then the gain will be much lower than the expected 13 dBi.

As another example, suppose that the same antenna (a 2-wavelength-boom Yagi) gives a measured gain of 9 dBi but the radiation pattern half power beamwidths are approximately 45°. 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 directions other 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 that 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 that 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 pattern measurements to help determine the possible causes for 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.

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