Test Procedures and Projects

26

This chapter, written by ARRL Technical Advisor Doug Millar, K6JEY, covers the test equipment and measurement techniques common to Amateur Radio. With the increasing complexity of amateur equipment and the availability of sophisticated test equipment, measurement and test procedures have also become more complex. There was a time when a simple bakelite cased volt-ohm meter (VOM) could solve most problems. With the advent of modern circuits that use advanced digital techniques, precise readouts and higher frequencies, test requirements and equipment have changed. In addition to the test procedures in this chapter, other test procedures appear in Chapters 14 and 15.

TEST AND MEASUREMENT BASICS

The process of testing requires a knowledge of what must be measured and what accuracy is required. If battery voltage is measured and the meter reads 1.52 V, what does this number mean? Does the meter always read accurately or do its readings change over time? What influences a meter reading? What accuracy do we need for a meaningful test of the battery voltage?

A Short History of Standards and Traceability

Since early times, people who measured things have worked to establish a system of consistency between measurements and measurers. Such consistency ensures that a measurement taken by one person could be duplicated by others — that measurements are reproducible. This allows discussion where everyone can be assured that their measurements of the same quantity would have the same result. In most cases, and until recently, consistent measurements involved an artifact: a physical object. If a merchant or scientist wanted to know what his pound weighed, he sent it to a laboratory where it was compared to the official pound. This system worked well for a long time, until the handling of the standard pound removed enough molecules so that its weight changed and measurements that compared in the past no longer did so.

Of course, many such measurements depended on an accurate value for the force of gravity. This grew more difficult with time because the outside environment — such things as a truck going by in the street — could throw the whole procedure off. As a result, scientists switched to physical constants for the determination of values. As an example, a meter was defined as a stated fraction of the circumference of the Earth over the poles.

Generally, each country has an office that is in charge of maintaining the integrity of the standards of measurement and is responsible for helping to get those standards into the field. In the United States that office is the National Institute of Standards and Testing (NIST), formerly the National Bureau of Standards. The NIST decides what the volt and other basic units should be and coordinates those units with other countries. For a modest fee, NIST will compare its volt against a submitted sample and report the accuracy of the sample. In fact, special batteries arrive there each day to be certified and returned so laboratories and industry can verify that their test equipment really does mean 1.527 V when it says so.

Basic Units: Frequency and Time

Frequency and time are the most basic units for many purposes and the ones known to the best accuracy. The formula for converting one to the other is to divide the known value into 1. Thus the time to complete a single cycle at 1 MHz = 0.000001 s.

The history of the accuracy of time keeping, of course, begins with the clock. Wooden clocks, water clocks and mechanical clocks were ancestors to our current standard: the electronic clock based on frequency. In the 1920s, quartz crystal controlled clocks were developed in the laboratory and used as a standard. With the advent of radio communication time intervals could be transmitted by radio, and a very fundamental standard of time and frequency could be used locally with little effort. Today transmitters in several countries broadcast time signals on standard calibrated frequencies. **Table 26.1** contains the locations and frequencies of some of these stations.

The Sounds of Amateur Radio — **Listen to transmissions from WWV.**

The Sounds of Amateur Radio — **Listen to transmissions from CHU.**

In the 1960s, Hewlett-Packard began selling self-contained time and frequency standards called cesium clocks. In a cesium clock a crystal frequency is generated and multiplied to microwave frequencies. That energy is passed through a chamber filled with cesium gas. The gas acts as a very narrow bandpass filter. The output signal is detected and the crystal oscillator frequency is adjusted automatically so that a maximum of energy is detected. The output of the crystal is thus linked to the stability of the cesium gas and is usually accurate to several parts in 10^{-12}. This is much superior to a crystal oscillator alone; but at close to $40,000 each, cesium frequency standards are a bit extravagant for amateur use.

A rubidium frequency standard is an alternative to the cesium clock. They are not quite as accurate as the cesium, but they are much less expensive, relatively quick to warm up and can be quite small. Older models occasionally appear surplus. As with any precision instrument, it should be checked over and calibrated before use.

Most hams do not have access to cesium or rubidium standards—or need them. Instead we use crystal oscillators. Crystal oscillators provide three levels of stability. The least accurate is a single crystal mounted on a circuit board. The crystal frequency is affected by the temperature environment of the equipment, to the extent of a few parts per million (ppm) per degree Celsius. For example, the frequency of a 10-MHz crystal with temperature stability rated at 3 ppm might vary 60 Hz when temperature of the crystal changes by 2°C. If the crystal oscillator is followed by a frequency multiplier, any variation in the crystal frequency is also multiplied. Even so, the accuracy of a simple crystal

Table 26.1
Standard Frequency Stations
(Note: In recent years, frequent changes in these schedules have been common.)

Call Sign	*Location*	*Frequency (MHz)*
BSF	Taiwan	5, 15
CHU	Ottawa, Canada	3.330, 7.335, 14.670
FFH	France	2.500
IAM/IBF	Italy	5.000
JJY	Japan	2.5, 5, 8, 10, 15
LOL	Argentina	5, 10
RID	Irkutsk	5.004, 10.004, 15.004
RWM	Moscow	5, 4.996, 9.996, 14.996
WWV/WWVH	USA	2.5, 5, 10, 15, 20
VNG	Australia	2.5, 5
ZSC	South Africa	4.291, 8.461, 12.724 (part time)

oscillator is sufficient for most of our needs and most amateur equipment relies on this technique. For a discussion of crystal oscillators and temperature compensation, look in the **Oscillators** chapter of this book.

The second level of accuracy is achieved when the temperature around the crystal is stabilized, either by an "oven" or other nearby components. Crystals are usually designed to stabilize at temperatures far above any reached in normal operating environments. These oscillators are commonly good to 0.1 ppm per day and are widely used in the commercial two-way radio industry.

The third accuracy level uses a double oven with proportional heating. The two ovens compensate for each other automatically and provide excellent temperature stability. The ovens must be left on continuously, however, and warm-up requires several days to two weeks.

Crystal *aging* also affects frequency stability. Some crystals change frequency over time (age) so the circuit containing the crystal must contain components to compensate for this change. Other crystals become more stable over time and become excellent frequency standards. Many commercial laboratories go to the expense of buying and testing several examples of the same oscillator and select the best one for use. As a result, many surplus oscillators are surplus for a reason. Nevertheless, a good stable crystal oscillator can be accurate to 1×10^{-9} per day and very appropriate for amateur applications.

Time and Frequency Calibration

Many hams have digital frequency counters, which range from surplus lab equipment to new highly integrated instruments with nearly everything on one chip. Almost all of these are very precise and display nine or more digits. Many are even quite stable. Nonetheless, a 10-MHz oscillator accurate to 1 ppm per month can vary ±10 Hz in one month. This drift rate may be acceptable for many applications, but the question remains: How accurate is it?

This question can be answered by calibrating the oscillator. There are several ways to perform this calibration. The most accurate method compares the unit in question by leaving the oscillator operating, transporting it to an oscillator of known frequency and then making a comparison. A commonly used comparison method connects the output of the calibrated oscillator into the horizontal input of a high frequency oscilloscope, and the oscillator to be measured to the vertical input. It helps, but they need not be on the same frequency. By noting how long it takes the sine wave to travel one division at a given sweep speed, one can calculate the resulting drift in parts per million per minute (ppm/min).

Another technique of oscillator calibration uses a VLF phase comparator. This is a special direct-conversion receiver that picks up the signal from WWVB on 60 kHz. Phase comparison is used to compare WWVB with the divided frequency of the oscillator being tested. Many commercial units have a small strip chart printer attached and switches to determine the receiver frequency. Since these 60-kHz VLF Comparator receivers have been largely replaced by units that use Loran signals or rubidium standards, they can be found at very reasonable prices. A very effective 60-kHz antenna can be made by attaching an audio transformer with the low-impedance winding connected to the receiver antenna terminals by way of a series dc blocking capacitor. The high-impedance winding is then connected between ground and a random length of wire. A typical VLF Comparator can track an oscillator well into a few parts in 10^{-10}. This technique directly compares the oscillator with an NIST standard and can even characterize oscillator drift characteristics in ppm per day or week.

Another fairly direct method compares an oscillator with one of the WWV HF signals. The received signal is not immensely accurate, but if the oscillator of a modern HF transceiver is carefully compared, it will be accurate enough for all but the most demanding work.

The last and least accurate way to calibrate an oscillator is to compare it with another oscillator or counter owned by you or another local ham. Unless the calibration of the other oscillator or counter is known, this comparison could be very misleading. True accuracy is not determined by the label of a famous company or impressive looks. Metrologists (people who calibrate and measure equipment) spend more time calibrating oscillators than any other piece of equipment.

DC Instruments and Circuits

This section discusses the basics of analog and digital dc meters. It covers the design of range extenders for current, voltage and resistance; construction of a simple meter; functions of a digital volt-meter (DVM) and procedures for accurate measurements.

Basic Meters

In measuring instruments and test equipment suitable for amateur purposes, the ultimate readout is generally based on a measurement of direct current. There are two basic styles of meters: analog meters that use a moving needle display, and digital meters that display the measured values in digital form. The analog meter for measuring dc current and voltage uses a magnet and a coil to move a pointer over a calibrated scale in proportion to the current flowing through the meter.

The most common dc analog meter is the D'Arsonval type, consisting of a coil of wire to which the pointer is attached so that the coil moves (rotates) between the poles of a permanent magnet. When current flows through the coil, it sets up a magnetic field that interacts with the field of the magnet to cause the coil to turn. The design of the instrument normally makes the pointer move in direct proportion to the current.

Digital Multimeters

In recent years there has been a flood of inexpensive digital multimeters (DMMs) ranging from those built into probes to others housed in large enclosures. They are more commonly referred to as digital voltmeters (DVMs) even though they are multimeters; they usually measure voltage, current and resistance. After some years of refining circuits such as the "successive approximation" and "dual slope" methods, most meters now use the dual-slope method to convert analog voltages to a digital reading. DVMs have basically three main sections as shown in **Fig 26.1**.

The first section scales the voltage or current to be measured. It has four main circuits:

- a chain of multiplier resistors that reduce the input voltage to 0-1 V,
- a converter that changes 0-1 V ac to dc,
- an amplifier that raises signals in the 0-100 mV range to 0-1 V and
- a current driver that provides a constant current to the multiplier chain for resistance measurements.

The second section is an integrator. It is usually based on an operational amplifier that is switched by a timing signal. The timing signal initially shorts the input of the integrator to provide a zero reference. Next a reference voltage is connected to charge the capacitor for a determined amount of time. Finally the last part of the timing cycle allows the capacitor to discharge. The time it takes the capacitor to discharge is proportional to either the input voltage (V_{in}, after it was scaled into the range of 0 to 1 V) or 1 minus V_{in}, depending on the meter design. This discharge time is measured by the next section of the DVM, which is actually a frequency counter. Finally, the output of the frequency counter is scaled to the selected range of voltage or current and sent to the final section of the DVM — the digital display.

Since the timing is quite fast and the capacitor is not used long enough to drift much in value, the components that most determine accuracy are the reference voltage source and the range multiplier resistors. With the availability of integrated resistor networks that are deposited or diffused onto the same substrate, drift is automatically compensated because all branches of a divider drift in the same direction simultaneously. The voltage sources are generally Zener diodes on substrates with accompanying series resistors. Often the resistor and Zener have opposite temperature characteristics that cancel each other. In more complex DVMs, extensive digital circuitry can insert values to compensate for changes in the circuit and can even be automatically calibrated remotely in a few moments.

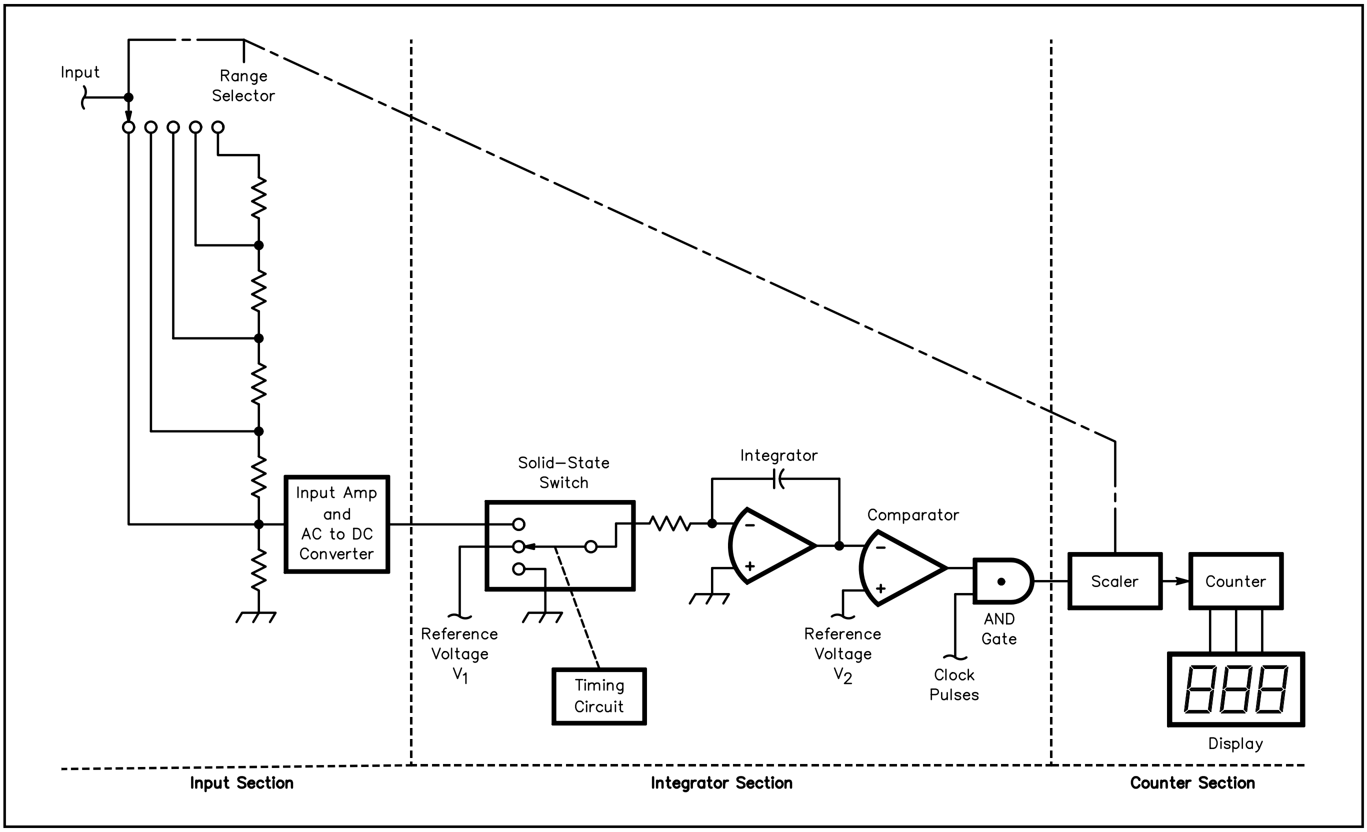


Fig 26.1 — A typical digital voltmeter consists of three parts: the input section for scaling, an integrator to convert voltage to pulse count, and a counter to display the pulse count representing the measured quantity.

Liquid crystal displays (LCD) are commonly used for commercial DVMs. As a practical matter they draw little current and are best for portable and battery-operated use. The usual alternative, light emitting diode (LED) displays, draw much more current but are better in low-light environments. Some older surplus units use gas plasma displays (orange-colored digits). You may have seen plasma displays on gas-station pumps. They are not as bright as LEDs, but are easier to read. On the down side, plasma displays require high-voltage power supplies, draw considerable current and often fail after 10 years or so.

The advantages of DVMs are high input resistance (10 MΩ on most ranges), accurate and precise readings, portability, a wide variety of ranges and low price. There is one disadvantage, however: Digital displays update rather slowly, often only one to two times per second. This makes it very difficult to adjust a circuit for a peak (maximum) or null (minimum) response using only a digital display. The changing digits do not give any clue of the measurement trend and it is easy to tune through the peak or null between display updates. In answer, many new DVMs are built with an auxiliary bar-graph display that is updated constantly, thus providing instantaneous readings of relative value and direction of changes.

Current Ranges

The sensitivity of an analog meter is usually expressed in terms of the current required for full-scale deflection of the pointer. Although a very wide variety of ranges is available, the meters of interest in amateur work give maximum deflection with currents measured in microamperes or milliamperes. They are called microammeters and milliammeters, respectively.

Thanks to the relationships between current, voltage and resistance expressed by Ohm's Law, it is possible to use a single low-range instrument (for example, 1 mA or less for full-scale pointer deflection) for a variety of direct-current measurements. Through its ability to measure current, the instrument can also be used indirectly to measure voltage. In the same way, a measurement of both current and voltage will obviously yield a value of resistance. These measurement functions are often combined in a single instrument: the volt-ohm-milliammeter or VOM, a multirange meter that is one of the most useful pieces of test equipment an amateur can possess.

Accuracy

The accuracy of a D'Arsonval-movement dc meter is specified by the manufacturer. A common specification is ±2% of full scale, meaning that a 0-100 μA meter, for example, will be correct to within 2 μA at any part of the scale. There are very few cases in amateur work where accuracy greater than this is needed. When the instrument is part of a more complex measuring circuit, however, the design and components can each cause error that accumulates to reduce the overall accuracy.

Extending Current Range

Because of the way current divides between two resistances in parallel, it is possible to increase the range (more specifically to decrease the sensitivity) of a dc current meter. The meter itself has an inherent resistance (its internal resistance) which determines the full-scale current passing through it when its rated voltage is applied. (This rated voltage is on the order of a few millivolts.) When an external resistance is connected in parallel with the meter, the current will divide between the two and the meter will respond only to that part of the current that flows through its movement. Thus, it reads only part of the total current; the effect makes more total current necessary for a full-scale meter reading. The added resistance is called a "shunt."

We must know the meter's internal resistance before we can calculate the value for a shunt resistor. Internal resistance may vary from a fraction of an ohm to a few thousand ohms, with greater resistance values associated with greater sensitivity. When this resistance is known, it can be used in the formula below to determine the required shunt for a given multiplication:

$$R = \frac{R_m}{n - 1} \quad (1)$$

where

R = shunt resistance, ohms

R_m = meter internal resistance, ohms

n = the factor by which the original meter scale is to be multiplied.

Often the internal resistance of a particular meter is unknown (when the meter is purchased at a flea market or is taken from a commercial piece of equipment, for example). Unfortunately, the internal resistance of a meter cannot be measured directly with an ohmmeter without risk of damage to the meter movement.

Fig 26.2 shows a method to safely measure the internal resistance of a linearly calibrated meter. It requires a calibrated meter that can measure the same current as the unknown meter. The system works as follows: S1 is switched off and R2 is set for maximum resistance. A supply of constant voltage is connected to the supply terminals (a battery will work fine) and R2 is adjusted so that the unknown meter reads exactly full scale. Note the current shown on M2. Close S1 and alternately adjust R1 and R2 so

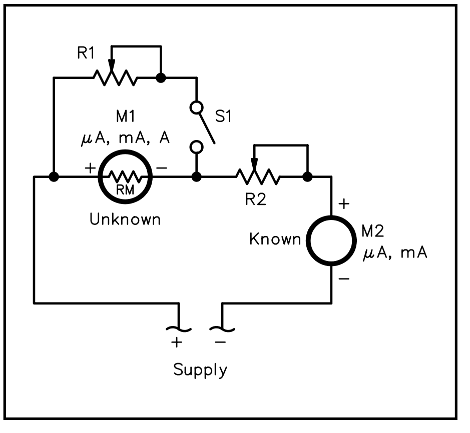


Fig 26.2 — This test setup allows safe measurement of a meter's internal resistance. See text for the procedure and part values.

that the unknown meter (M1) reads exactly half scale and the known meter (M2) reads the same value as in the step above. At this point, half of the current in the circuit flows through M1 and half through R1. To determine the internal resistance of the meter, simply open S1 and read the resistance of R1 with an ohm-meter.

The values of R1 and R2 will depend on the meter sensitivity and the supply voltage. The maximum resistance value for R1 should be approximately twice the expected internal resistance of the meter. For highly sensitive meters (10 µA and less), 1 kΩ should be adequate. For less-sensitive meters, 100 Ω should suffice. Use no more supply voltage than necessary.

The value for minimum resistance at R2 can be calculated using Ohm's Law. For example, if the meter reads 0 to 1 mA and the supply is a 1.5-V battery, the minimum resistance required at R2 will be:

$$R2 = \frac{1.5}{0.001}$$

$$R2(\text{min}) = 1500\,\Omega$$

In practice a 2- or 2.5-kΩ potentiometer would be used.

Making Shunts

Homemade shunts can be constructed from several kinds of resistance wire or from ordinary copper wire if no resistance wire is available. The copper wire table in the **Component Data** chapter of this *Handbook* gives the resistance per 1000 ft for various sizes of copper wire. After computing the resistance required, determine the smallest wire size that will carry the full-scale current, again from the wire table. Measure off enough wire to give the required resistance. A high-resistance 1- or 2-W carbon-composition resistor makes an excellent form on which to wind the wire, as the high resistance does not affect the value of the shunt. If the shunt gets too hot, go to a larger diameter wire of a greater length.

VOLTMETERS

If a large resistance is connected in series with a meter that measures current, as shown in **Fig 26.3**, the current multiplied by the resistance will be the voltage drop across the resistance. This is known as a multiplier. An instrument used in this way is calibrated in terms of the voltage drop across the multiplier resistor and is called a voltmeter.

Sensitivity

Voltmeter sensitivity is usually expressed in ohms per volt (Ω/V), meaning that the meter full-scale reading multiplied by the sensitivity will give the total resistance of the voltmeter. For example, the resistance of a 1 kΩ/V voltmeter is 1000 times the full-scale calibration voltage. Then by Ohm's Law the current required for full-scale deflection is l milliampere. A sensitivity of 20 kΩ/V, a commonly used value, means that the instrument is a 50-µA meter.

As voltmeter sensitivity (resistance) increases, so does accuracy. Greater meter resistance means that less current is drawn from the circuit and thus the circuit under test is less affected by connection of the meter. Although a 1000-Ω/V meter can be used for some applications, most good meters are 20 kΩ/V or more. Vacuum-tube voltmeters (VTVMs) and their modern equivalent FET voltmeters (FETVOMs) are usually 10-100 MΩ/V and DVMs can go even higher.

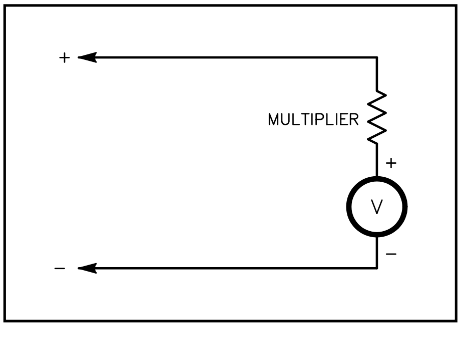


Fig 26.3 — A voltmeter is constructed by placing a current-indicating instrument in series with a high resistance, the "multiplier."

Multipliers

The required multiplier resistance is found by dividing the desired full-scale voltage by the current, in amperes, required for full-scale deflection of the meter alone. To be mathematically correct, the internal resistance of the meter should be subtracted from the calculated value. This is seldom necessary (except perhaps for very low ranges) because the meter resistance is usually very low compared with the multiplier resistance. When the instrument is already a voltmeter with an internal multiplier, however, the meter resistance is significant. The resistance required to extend the range is then:

$$R = R_m(n - 1) \qquad (2)$$

where
R_m = total resistance of the instrument
n = factor by which the scale is to be multiplied

For example, if a 1-kΩ/V voltmeter having a calibrated range of 0 to 10 V is to be extended to 1000 V, R_m is 1000 × 10 = 10 kΩ, n is 1000/10 = 100 and R = 10,000 × (100 – 1) = 990 kΩ.

When extending the range of a volt- meter or converting a low-range meter into a voltmeter, the rated accuracy of the instrument is retained only when the multiplier resistance is precise. High-precision, hand-made and aged wire-wound resistors are used as multipliers of high-quality instruments. These are relatively expensive, but the home constructor can do well with 1% tolerance metal-film resistors. They should be derated when used for this purpose. That is, the actual power dissipated in the resistor should not be more than $^1/_{10}$ to $^1/_4$ the rated dissipation. Also, use care to avoid overheating the resistor body when soldering. These precautions will help prevent permanent change in the resistance of the unit.

Many DVMs use special resistor groups that have been etched on quartz or sapphire and laser trimmed to value. These resistors are very stable and often quite accurate. They can be bought new from various suppliers. It is also possible to “rescue” the divider/multiplier resistors from an older DVM that no longer functions and use them as multipliers. Look for a series of four or five resistors that add up to 10 MΩ: .9,9,90,900,9,000,90,000 and 900,000 Ω. There is usually another 1-MΩ resistor in series to isolate the meter from the circuit under test. A few of these high-accuracy resistors in “odd” values can help calibrate less-expensive instruments.

DC Voltage Standards

For a long time NIST has statistically compared a bank of special Weston Cell or cadmium sulfate batteries to arrive at the standard volt. By using a special tapped resistor, a 1.08-V battery can be compared to other voltages and instruments compared. However, these are very high-impedance batteries that deliver almost no current and are relatively temperature sensitive. They are made up of a solution of cadmium and mercury in opposite legs of an “H” shaped glass container. You can read much more about them in *Calibration—Philosophy and Practice*, published by the John Fluke Co of Mount Lake Terrace, Washington.

Hams often use an ordinary flashlight battery as a convenient voltage reference. A fresh *D cell* usually provides 1.56 V under no load, as would be measured by a DVM. The Heath Company, which supplied thousands of kits to the ham community for many years, used such batteries as the calibration references for many of their kits.

Recently, NIST has been able to use a microwave to voltage converter called a “Josephson Junction” to determine the value of the volt. The converter transfers the accuracy of a frequency standard to the accuracy of the voltage that comes out of it. The converter generates only 5 mV, however, which then must be scaled to the standard 1-V level. One problem with high-accuracy measurements is stray noise (low-level voltages) that creates a floor below which measurements are meaningless. For that reason, meters with five or more digits must be very quiet and any comparisons must be made at a voltage high enough to be above the noise.

DC MEASUREMENT CIRCUITS

Current Measurement with a Voltmeter

A current-measuring instrument should have very low resistance compared with the resistance of the circuit being measured; otherwise, inserting the instrument will alter the current from its value when the instrument is removed. The resistance of many circuits in radio equipment is high and the circuit operation is affected little, if at all, by adding as much as a few hundred ohms in series. [Even better, use a resistor that is part of the working circuit if one exists. Unsolder one end of the resistor, measure its resistance, reinstall it and then make the measurement.—*Ed.*] In such cases the voltmeter method of measuring current in place of an ammeter, shown in **Fig 26.4**, is frequently convenient. A voltmeter (or low-range milliammeter provided with a multiplier and operating as a voltmeter) having a full-scale voltage range of a few volts is used to measure the voltage drop across a suitable value of resistance acting as a shunt.

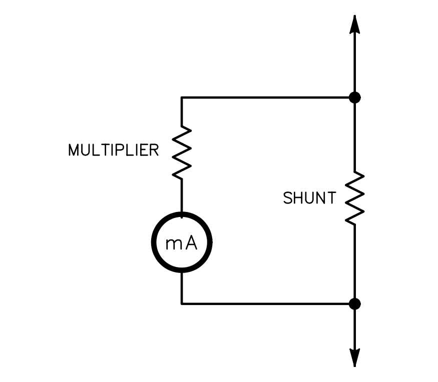


Fig 26.4 — A voltmeter can be used to measure current as shown. For reasonable accuracy, the shunt should be 5% of the circuit impedance or less, and the meter resistance should be 20 times the circuit impedance or more.

The value of shunt resistance must be calculated from the known or estimated maximum current expected in the circuit (allowing a safe margin) and the voltage required for full-scale movement of the meter with its multiplier. For example, to measure a current estimated at 15 A on the 2-V range of a DVM, we need to solve Ohm's Law for the value of R:

$$R_{shunt} = \frac{2\ V}{15\ A} = 0.133\,\Omega$$

This resistor would dissipate $15^2 \times 0.133 = 29.92$ W. For a short-duration measurement, 30 1.0-Ω, 1-W resistors could be parallel connected in two groups of 15 (0.067 Ω per group) that are series connected to yield 0.133 Ω. For long-duration measurements, 2- to 5-W resistors would be better.

Power

Power in direct-current circuits is usually determined by measuring the current and voltage. When these are known, the power can be calculated by multiplying voltage, in volts, by the current, in amperes. If the current is measured with a milliammeter, the reading of the instrument must be divided by 1000 to convert it to amperes.

The setup for measuring power is shown in **Fig 26.5A**, where R is any dc load, not necessarily an actual resistor. In this measurement it is always best to use the lowest voltmeter or ammeter scale that allows reading the measured quantity. This results in the percentage error being less than if the meter was reading in the very lowest part of the selected scale.

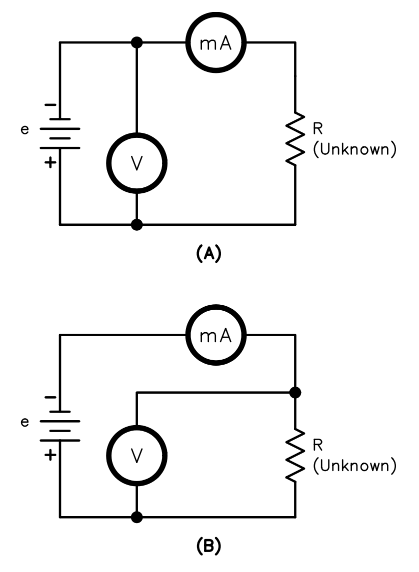


Fig 26.5 — Power or resistance can be calculated from voltage and current measurements. At A, error introduced by the ammeter is dominant. At B, error introduced by the voltmeter is dominant. The text gives an example.

Resistance

If both voltage and current are measured in a circuit such as that in Fig 26.5, the value of resistance R (in case it is unknown) can be calculated from Ohm's Law. For reasonable results, two conditions should be met:

1. The internal resistance of the current meter should be less than 5% of the circuit resistance.
2. The input impedance of the voltmeter should be greater than 20 times the circuit resistance.

These conditions are important because both meters tend to load the circuit under test. The current meter resistance adds to the unknown resistance, while the voltmeter resistance decreases the unknown resistance as a result of their parallel connection.

Ohmmeters

Although Fig 26.5B suffices for occasional resistance measurements, it is inconvenient when we need to make frequent measurements over a wide range of resistance. The device generally used for this purpose is the ohmmeter. Its simplest form is a voltmeter (or milliammeter, depending on the circuit used) and a small battery. The meter is calibrated so that the value of an unknown resistance can be read directly from the scale. **Fig 26.6** shows some typical ohmmeter circuits. In the simplest circuit, Fig 26.6A, the meter and battery are connected in series with the unknown resistance. If a given movement of the meter's needle is obtained with terminals A-B shorted, inserting the resistance to be measured will cause the meter reading to decrease. When the resistance of the voltmeter is known, the following formula can be applied.

$$R = \frac{eR_m}{E} - R_m \tag{3}$$

where
R = unknown resistance, ohms
e = voltage applied (A-B shorted)
E = voltmeter reading with R connected
R_m = resistance of the voltmeter.

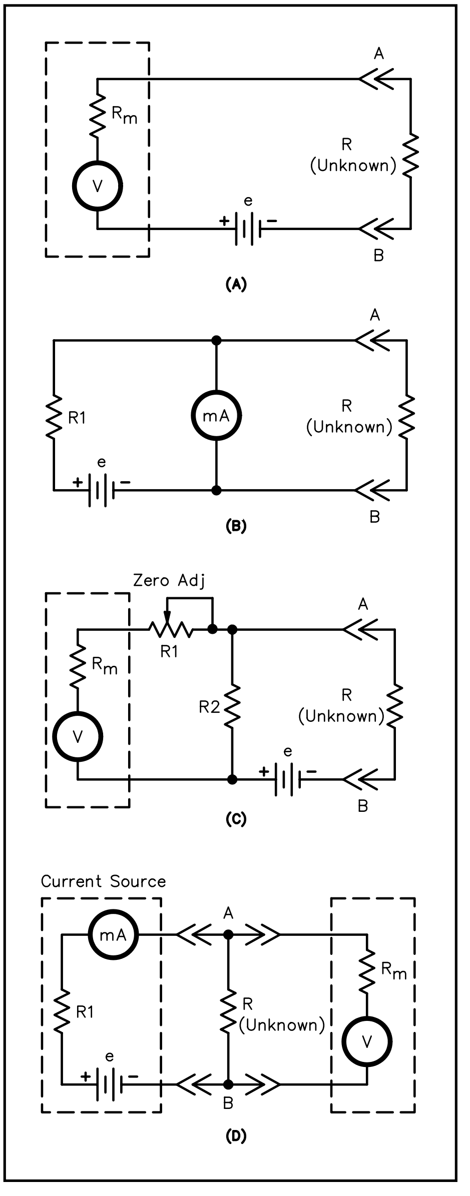


Fig 26.6 — Here are several kinds of ohmmeters. Each is explained in the text.

The circuit of Fig 26.6A is not suited to measuring low values of resistance (less than 100 Ω or so) with a high-resistance voltmeter. For such measurements the circuit of Fig 26.6B is better. The unknown resistance is

$$R = \frac{I_2 R_m}{I_1 - I_2} \tag{4}$$

where
R = unknown resistance, ohms
R_m = the internal resistance of the milliammeter, ohms
I_1 = current with R disconnected from terminals A-B, amps
I_2 = current with R connected, amps.

This formula is based on the assumption that the current in the complete circuit will be essentially constant whether or not the unknown terminals are short circuited. This requires that R1 be much greater than R_m. For example, 3000 Ω for a 1-mA meter with an internal resistance of perhaps 50 Ω. In this case, a 3-V battery would be necessary in order to obtain a full-scale deflection with the unknown terminals open. R1 can be an adjustable resistor, to permit setting the open-terminal current to exact full scale.

A third circuit for measuring resistance is shown in Fig 26.6C. In this case a high-resistance voltmeter is used to measure the voltage drop across a reference resistor, R2, when the unknown resistor is connected so that current flows through it, R2 and the battery in series. With suitable R2s (low values for low-resistance, high values for high-resistance unknowns), this circuit gives equally good results for resistance values in the range from one ohm to several megohms. The voltmeter resistance, R_m, must be much greater (50 times or more) than that of R2. A 20-kΩ/V instrument (50-µA movement) is generally used. If the current through the voltmeter is negligible compared with the current through R2, the formula for the unknown is

$$R = \frac{eR2}{E} - R2 \qquad (5)$$

where
R and R2 are in ohms
e = voltmeter reading with R removed and A shorted to B.
E = voltmeter reading with R connected.

R1 sets the voltmeter reading exactly to full scale when the meter is calibrated in ohms. A 10-kΩ pot is suitable with a 20-kΩ/V meter. The battery voltage is usually 3 V for ranges to 100 kΩ and 6 V for higher ranges.

Four-Wire Resistance Measurements

In situations where a very low resistance, like a 50-Ω dummy load, is to be measured, the resistance of the test leads can be significant. The average lead resistance is about 0.9 Ω through both leads, which would make a 50.5-Ω dummy load appear to be 51.4 Ω. To compensate for lead resistance, some meters allow for four-wire measurements. Briefly, two wires from the current source and two wires from the measuring circuit exit the meter case separately and connect directly to the unknown resistance (see Fig 26.6D). This eliminates the voltage drop in the current-source leads from the measurement. In practice, four-wire systems use special test clips that are similar to alligator clips, except that the jaws are insulated from each other and a meter lead is attached to each jaw. In some meters, an additional control allows the operator to short the test leads together and adjust the meter for a zero reading before making low-resistance measurements.

Bridge Circuits

Bridges are an important class of measurement circuits. They perform measurement by comparison with some known component or quantity, rather than by direct reading. VOMs, DVMs and other meters are convenient, but their accuracy is limited. The accuracy of manufactured analog meters is determined at the factory, while digital meters are accurate only to some percentage ±1 in the least-significant digit. The accuracy of comparison measurements, however, is determined only by the comparison standard and bridge sensitivity.

Bridge circuits are useful across most of the frequency spectrum. Most amateur applications are at RF, as shown later in this chapter. The principles of bridge operation are easier to understand at dc, however, where bridge operation is simple.

The Wheatstone Bridge

A simple resistance bridge, known as the Wheatstone bridge, is shown in **Fig 26.7**. All other bridge

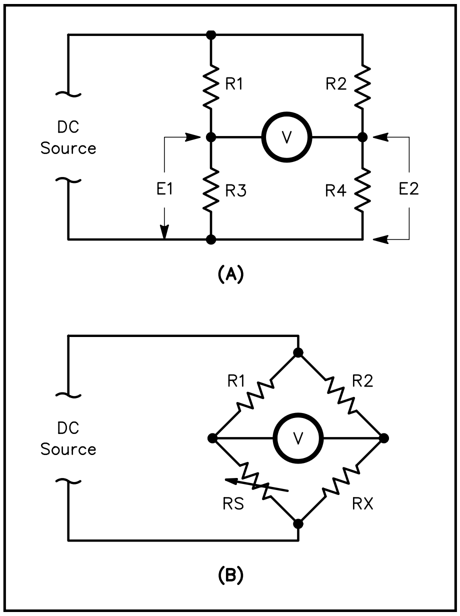


Fig 26.7 — A Wheatstone bridge circuit. A bridge circuit is actually a pair of voltage dividers (A). B shows how bridges are normally drawn.

circuits are based on this design. The four resistors, R1, R2, R3 and R4 in Fig 26.7A, are known as the bridge arms. For the voltmeter reading to be zero (null) the voltage dividers consisting of the pairs R1-R3 and R2-R4 must be equal. This means

$$\frac{R1}{R3} = \frac{R2}{R4} \tag{6}$$

When this occurs the bridge is said to be *balanced*.

The circuit is usually drawn as shown at Fig 26.7B when used for resistance measurement. Equation 6 can be rewritten

$$RX = RS\left(\frac{R2}{R1}\right) \tag{7}$$

RX is the unknown resistor. R1 and R2 are usually made equal; then the calibrated adjustable resistance (the standard), RS, will have the same value as RX when RS is set to show a null on the voltmeter.

Note that the resistance ratios, rather than the actual resistance values, determine the voltage balance. The values do have important practical effects on the sensitivity and power consumption, however. Bridge sensitivity is the ability of the meter to respond to slight unbalance near the null point; a sharper null means a more accurate setting of RS at balance.

The Wheatstone bridge is rarely used by amateurs for resistance measurement, since it is easier to measure resistances with VOMs and DVMs. Nonetheless, it is worthwhile to understand its operation as the basis of more complex bridges.

ELECTRONIC VOLTMETERS

We have seen that the resistance of a simple voltmeter (as in Fig 26.3) must be extremely high in order to avoid "loading" errors caused by the current that necessarily flows through the meter. The use of high-resistance meters tends to cause difficulty in measuring relatively low voltages because multiplier resistance progressively lessens as the voltage range is lowered.

Voltmeter resistance can be made independent of voltage range by using vacuum tubes, FETs or op amps as dc amplifiers between the circuit under test (CUT) and the indicator, which may be a conventional meter movement or a digital display. Because the input resistance of the electronic devices mentioned is extremely high (hundreds of megohms) they have negligible loading effect on the CUT. They do, however, require a closed dc path in their input circuits (although this path can have very high resistance). They are also limited in the voltage level that their input circuits can handle. Because of this, the device actually measures a small voltage across a portion of a high-resistance voltage divider connected to the CUT. Various voltage ranges are obtained from appropriate taps on the voltage divider.

In the design of electronic voltmeters it has become standard practice to use a voltage divider with a total resistance of 10 MΩ, tapped as required, in series with a 1-MΩ resistor incorporated in the meter. The total voltmeter resistance, including probe, is therefore 11 MΩ. The 1-MΩ resistor serves to isolate the voltmeter circuit from the CUT.

AC Instruments and Circuits

Most ac measurements differ from dc measurements in that the accuracy of the measurement depends on the purity of the sine wave. It is fairly easy to measure an ac voltage to between 1% and 5%, but getting down to 0.01% is difficult. Measurements to less than 0.01% must be left to precision laboratories. In general, amateurs measure ac voltages in household circuits, audio stages and RF power measurements, and 1% to 5% accuracy is usually close enough.

This section covers basic measurements, the nature of sine waves and meters. There are four common ways to measure ac voltage:

- Use a rectifier to change the ac to dc and then measure the dc.
- Heat a resistor in a Wheatstone bridge with the ac and measure the bridge unbalance.
- Heat a resistor surrounded by oil and measure the temperature rise.
- Use electronic circuits (such as multipliers and logarithmic amplifiers) with mathematical ac-to-dc conversion formulas. This method is not common, but it's interesting.

Calorimetric Meters

In a calorimetric meter, power is applied to a resistor that is immersed in the flow path of a special oil. This oil transfers the heat to another resistor that is part of a bridge. As the resistor heats, its resistance changes and the bridge becomes unbalanced. An attached meter registers the unbalance of the bridge as ac power. This type of meter is accurate for both dc and ac. They frequently operate from dc well into the GHz range. For calibration, an accurate dc voltage is applied and the reading is noted; a similar ac voltage is applied and the readings are compared. Some calorimetric meters are complicated, but others are simple.

Thermocouple Meters and RF Ammeters

In a thermocouple meter, alternating current flows through a low-resistance heating element. The power lost in the resistance generates heat that warms a thermocouple, which consists of a pair of junctions of two different metals. When one junction is heated a small dc voltage is generated in response to the difference in temperature of the two junctions. This voltage is applied to a dc milliammeter that is calibrated in suitable ac units. The heater-thermocouple/dc-meter combination is usually housed in a regular meter case.

Thermocouple meters are available in ranges from about 100 mA to many amperes. Their useful upper frequency limit is in the neighborhood of 100 MHz. Amateurs use these meters mostly to measure current through a known load resistance and calculate the RF power delivered to the load.

RECTIFIER INSTRUMENTS

The response of a rectifier RF ammeter is proportional (depending on the design) to either the peak or average value of the rectified ac wave, but never directly to the RMS value. These meters cannot be calibrated in RMS without knowing the relationship that exists between the real reading and the RMS value. This relationship may not be known for the circuit under test.

Average-reading ac meters work best with pure sine waves and RMS meters work best with complicated wave forms. Since many practical measurements involve nonsinusoidal forms, it is necessary to know what your instrument is actually reading, in order to make measurements intelligently. Most VOMs and VTVMs use averaging techniques, while DVMs may use either one. In all cases, check the meter instruction manual to be sure what it reads.

Peak and Average with Sine-Wave Rectification

Peak, average and RMS values of ac waveforms are discussed in the **AC Theory** chapter. Because the positive and negative half cycles of the sine wave have the same shape, half-wave rectification of either the positive half or the negative half gives exactly the same result. With full-wave rectification , the peak reading is the same, but the average reading is doubled, because there are twice as many half cycles per unit of time.

Asymmetrical Wave Forms

A nonsinusoidal waveform is shown in **Fig 26.8A**. When the positive half cycles of this wave are rectified, the peak and average values are shown at B. If the polarity is reversed and the negative half cycles are rectified, the result is shown in Fig 26.8C. Full-wave rectification of such a lopsided wave changes the average value, but the peak reading is always the same as that of the half cycle that produces the highest peak in half-wave rectification.

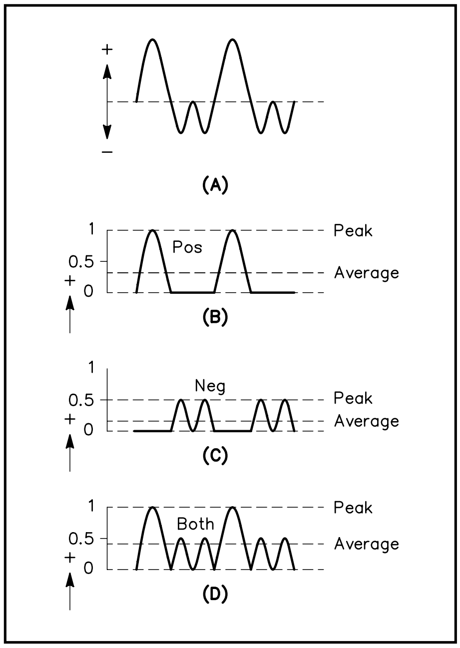


Fig 26.8— Peak vs average ac values for an asymmetrical wave form. Note that the peak values are different with positive or negative half-cycle rectification.

Effective-Value Calibration

The actual scale calibration of commercially made rectifier voltmeters is very often (almost always, in fact) in terms of RMS values. For sine waves, this is satisfactory and useful because RMS is the standard measurement at power-line frequencies. It is also useful for many RF applications when the waveform is close to sinusoidal. In other cases, particularly in the AF range, the error may be considerable when the waveform is not pure.

Turn-Over

From Fig 26.8 it is apparent that the calibration of an average-reading meter will be the same whether the positive or negative sides are rectified. A half-wave peak-reading instrument, however, will indicate different values when its connections to the circuit are reversed (turn-over effect). Very often readings are taken both ways, in which case the sum of the two is the peak-to-peak (P-P) value, a useful figure in much audio and video work.

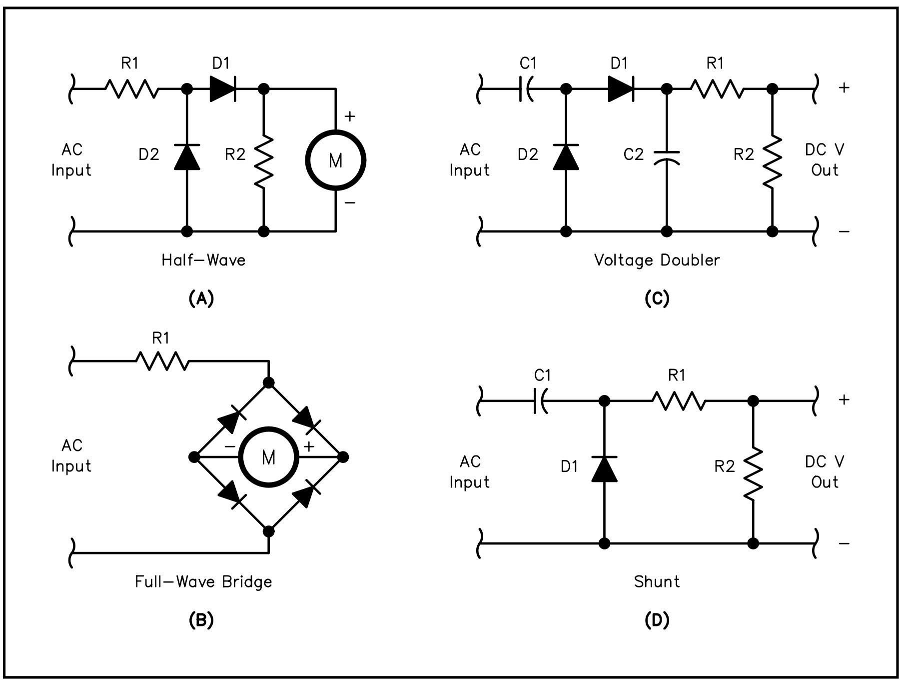


Fig 26.9 — Half (A) and full-wave (B) bridge rectifiers for average-reading, analog-display meters. Peak-reading circuits: a voltage doubler (C) and a shunt circuit (D). All circuits are discussed in the text.

Average- vs Peak-Reading Circuits

For traditional analog displays, the basic difference between average- and peak-reading rectifier circuits is that the output is not filtered for averaged readings, while a filter capacitor is charged to the peak value of the output voltage in order to measure peaks. **Fig 26.9A** and **B** show typical average-reading circuits, one half-wave and the other full-wave. In the absence of dc filtering, the meter responds to wave forms such as those shown at B, C and D in Fig 26.8; and since the inertia of the pointer system makes it unable to follow the rapid variations in current, it averages them out mechanically.

In Fig 26.9A, D1 actuates the meter; D2 provides a low-resistance dc return in the meter circuit on the negative half cycles. R1 is the voltmeter multiplier resistance. R2 forms a voltage divider with R1 (through D1) that prevents more than a few ac volts from appearing across the rectifier-meter combination. A corresponding resistor can be used across the full-wave bridge circuit.

In these two circuits there is no provision to isolate the meter from any dc voltage in the circuit under test. The resulting errors can be avoided by connecting a large nonpolarized capacitor in series with the hot lead. The reactance must be low compared with the meter impedance (at the lowest frequency of interest, more on this later) in order for the full ac voltage to be applied to the meter circuit. Some meters may require as much as 1 μF at line (60 Hz) frequencies. Such capacitors are usually not included in VOMs.

Voltage doubler and shunt peak-reading circuits are shown in Fig 26.9C and D. In both circuits, C1 isolates the rectifier from dc voltage in the circuit under test. In the voltage-doubler circuit, the time constant of the C2-R1-R2 combination must be very large compared with the period of the lowest ac frequency to be measured; similarly with C1-R1-R2 in the shunt circuit. This is so because the capacitor is charged to the peak value (V_{P-P} in C, V_P in D) when the ac wave reaches its maximum and then must hold the charge (so it can register on a dc meter) until the next maximum of the same polarity. If the time constant is 20 times the ac period, the charge will have decreased by about 5% when the next charge occurs. The average voltage drop will be smaller, so the error is appreciably less. The error will decrease rapidly with increasing frequency (if there is no change in the circuit values), but it will increase at lower frequencies.

In Fig 26.9C and D, R1 and R2 form a voltage divider that reduces the voltage to some desired value. For example, if R1 is 0 Ω in the voltage doubler, the voltage across R2 is approximately V_{P-P}; if R1 = R2, the output is approximately V_P (as long as the waveform is symmetrical).

The most common application of the shunt circuit is an RF probe to read V_{RMS}. In that case, R2 is the input impedance of a VTVM or DVM: 11 MΩ. R1 is chosen so that 71% of the peak value appears across R2. This converts the peak reading to RMS for sine-wave ac. R1 is therefore approximately 4.7 MΩ, making the total resistance nearly 16 MΩ. A capacitance of 0.05 μF is sufficient for low audio frequencies under these conditions. Much smaller values of capacitance may be used at RF.

Voltmeter Impedance

The impedance of a voltmeter at the frequency being measured may have an effect on the accuracy similar to that caused by the resistance of a dc voltmeter, as discussed earlier. The ac meter is a resistance in parallel with a capacitance. Since the capacitive reactance decreases with increasing frequency, the impedance also decreases with frequency. The resistance does change with voltage level, particularly at very low voltages (10 V or less) depending on the sensitivity of the meter and the kind of rectifier used.

The ac load resistance represented by a diode rectifier is about one-half of its dc-load resistance. In Fig 26.9A the dc load is essentially the meter resistance, which is generally quite low compared with the multiplier resistance R1. Hence, the total resistance will be about the same as the multiplier resistance. The capacitance depends on the components and construction, test-lead length and location, and other such factors. In general, the capacitance has little or no effect at lower line and audio frequencies, but ordinary VOMs lose accuracy at high audio frequencies and are of little use at RF. Rectifiers with very low inherent capacitance are used at RF and they are usually located at the probe tip to reduce losses.

Similar limitations apply to peak-reading circuits. In the shunt circuit, the resistive part of the impedance is smaller than in the voltage-doubler circuit because the dc load resistance, R1/R2, is directly across the circuit under test and in parallel with the diode ac load resistance. In both peak-reading circuits

Sources for RF Ammeters

When it comes to getting your own RF ammeter, there's good news and bad news. First, the bad news. New RF ammeters are expensive: about $70 to $200 (in 1994). AM radio stations are the main users of these today. 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 (see the Address List in the **References** chapter) has been a consistent RF-ammeter source. Ham flea markets are also worth trying. Some grubbing around in your nearest surplus store or some older ham's junk box may provide just the RF ammeter you need.

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.†

Another approach is to use a small low-voltage lamp as the heat/light element and use a photo detector 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.

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.††

Copper-Top Battery Testers as RF Ammeters

Finally, there are the *free* RF ammeters that come as the testers with Duracell batteries! For 1.5-V cells, these are actually 3 to 5-Ω resistors with built-in liquid-crystal displays. The resistor heats the liquid-crystal strip; the length of the "lighted" portion (heat turns the strip clear, exposing the fluorescent ink beneath) indicates the magnitude of the current.

Despite their "+" and "–" markings, these indicators are not polarized. Their resistance is low enough to have relatively little effect on a 50-Ω system. (For example, putting one in series with a 50-Ω dummy load would increase the system SWR from 1 to 1.1:1. These testers can measure about 200 to 400 mA. (You can achieve higher ranges by means of a shunt.) Best of all, if you burn out one of these "meters" during your tests, you can replace it at any drugstore, hardware store or supermarket for a few dollars, with some batteries thrown in free. *—John Stanley, K4ERO*

* F. Sutter, "What, No Meters?" *QST*, Oct 1938, p 49.
† C. Wright, "The Twin-Lamp," *QST*, Oct 1947, pp 22-23, 110 and 112.
†† Z. Lau, "A Relative RF Ammeter for Open-Wire Lines," *QST*, Oct 1988, pp 15-17.

the effective capacitance may range from 1 or 2 to a few hundred pF, with 100 pF typical in most instruments.

Scale Linearity

Fig 26.10 shows a typical current/voltage chart for a small semiconductor diode, which shows that the forward dynamic resistance of the diode is not constant, but rapidly decreases as the forward voltage increases from zero. The change from high to low resistance happens at much less than 1 V, but is in the range of voltage needed for a dc meter. With an average-reading circuit the current tends to be proportional to the square of the applied voltage. This makes the readings at the low end of the meter scale very crowded. For most measurement purposes, however, it is far more desirable for the output to be linear (that is, for the reading to be directly proportional to the applied voltage), which means that the markings on the meter are more evenly spaced.

To obtain that kind of linearity it is necessary to use a relatively large load resistance for the diode: Large enough that this resistance, rather than the diode resistance, will determine how much current flows. With this technique you can have a linear reading meter, but at the expense of sensitivity. The resistance needed depends on the type of diode; 5 kΩ to 50 kΩ is usually enough for a germanium rectifier, depending on the dc meter sensitivity, but several times as much may be needed for silicon diodes. Higher resistances require greater meter sensitivity; that is, the basic meter must be a microammeter rather than a low-range milliammeter.

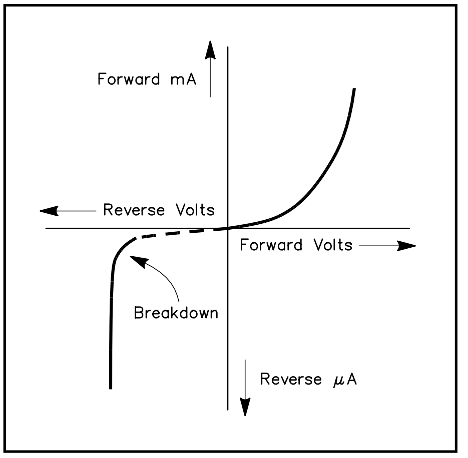


Fig 26.10 — Voltage vs current characteristics for a typical semiconductor diode. Actual values vary with different part numbers, but the forward current will always be increasing steeply with 1 V applied. Note that the forward current scale (mA) is 1000 times larger than that for reverse leakage current (μA). Breakdown voltage varies from 15 V to several hundred volts.

Reverse Current

When semiconductor diodes are reverse biased, a small leakage current flows. This reverse current flows during the half cycle when the diode should appear open, and the current causes an error in the dc meter reading. The quantity of reverse current is indicated by a diode's *back resistance* specification. This back resistance is so high that reverse current is negligible with silicon diodes, but back resistance may be less than 100 kΩ for germanium diodes.

The practical effect of semiconductor back resistance is to limit the amount of resistance that can be used in the dc load. This in turn affects the linearity of the meter scale. For practical purposes, the back resistance of vacuum-tube diodes is infinite.

RF Voltage

Special precautions must be taken to minimize the capacitive component of the voltmeter impedance at RF. If possible, the rectifier circuit should be installed permanently at the point where the RF voltage is to be measured, using the shortest possible RF connections. The dc meter can be remotely located, however.

For general RF measurements an RF probe is used in conjunction with a 10 MΩ electronic voltmeter. The circuit of **Fig 26.11**, which is basically the shunt peak-reading circuit of Fig 26.9D, is generally used. The series resistor, which is installed in the probe close to the rectifier, prevents RF from being fed through the probe cable to the electronic voltmeter. (In addition, the capacitance of the coaxial cable serves as a bypass for any RF on the lead.) This resistor, in conjunction with the 10-MΩ divider resistance of the electronic voltmeter, reduces the peak rectified voltage to a dc value equivalent to the RMS of the RF signal. Therefore, the RF readings are consistent with the regular dc calibration.

Of the diodes readily available to amateurs, the germanium point-contact or Schottky diode is preferred for RF-probe applications. It has low capacitance (on the order of 1 pF) and in high-back-resistance Schottky diodes, the reverse current is not serious. The principal limitation is that its safe reverse voltage is only about 50 to 75 V, which limits the applied voltage to 15 or 20 V RMS. Diodes can be series connected to raise the overall rating. At RF, however, it is more common to use capacitors or resistors as voltage dividers and apply the divider output to a single diode.

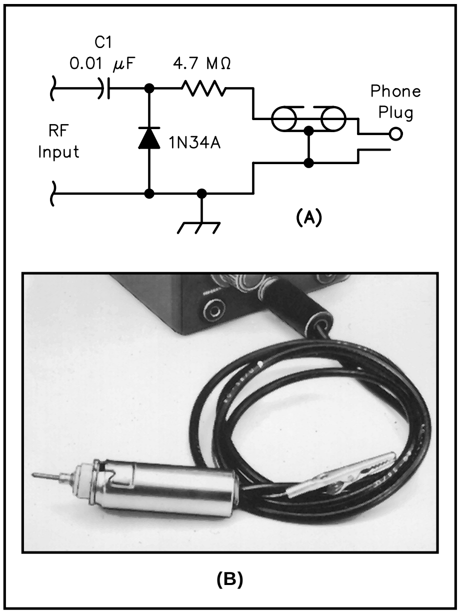


Fig 26.11 — At A is the schematic, at B a photo, of an RF probe for electronic voltmeters. The case of this probe is a seven-pin ceramic tube socket and a 2

RF Power

RF power can be measured by means of an accurately calibrated RF voltmeter connected across a dummy load in which the power is dissipated. If the load is a known pure resistance, the power, by Ohm's Law, is equal to E^2/R, where E is the RMS voltage.

The Hewlett-Packard 410B/C VTVM

The Hewlett-Packard 410B and 410C VTVMs have been standards of bench measurement for industry, and they are now available as industrial surplus. These units are not only excellent VTVMs, but they are also good wide-range RF power meters. Both models use a vacuum-tube detector mounted in a low-loss probe for ac measurements. With an adapter that allows the probe to contact the center conductor of a transmission line, it will give very good RF voltage measurements from 50 mV to 300 V and from 20 Hz to beyond 500 MHz. Very few other measuring instruments provide this range in a single sensor/meter. In addition to the 410B or C, you will also need a probe adapter HP model #11042A.

Do not take the probe apart for inspection because that can change the calibration. You can quickly check the probe by feeling it after it has been warmed up for about 15 minutes. If the body of it feels warm it is probably working. Inside, the 410B is quite different from the C model, with the B being simpler. The meter scales are also different; the 410C offers better resolution and perhaps better accuracy.

To make RF power measurements, remove the ac probe tip and twist lock the probe into the 11042A probe adapter. Attach the output of the adapter to a dummy load and use the formula

$$P = \frac{E^2}{R} \tag{8}$$

where

P = power, in watts

E = value given by the meter, in volts

R = resistance of the dummy load, in ohms.

The resistance of the dummy load should be accurately known to at least ±0.1 Ω, preferably measured with a four-wire arrangement as described in the section on ohmmeters. For frequent measurements, make a chart of voltage vs power at your most often used wattages.

THE MICROWATTER

This simple, easy to build terminating microwattmeter by Denton Bramwell, K7OWJ appeared in June 1997 *QST*.[1] It can measure power levels from below -50 dBm (10 nW) to -20 dBm (10 mW) with an accuracy of within 1 dB at frequencies from below the broadcast band up to 2 meters. Beyond that, it's still capable of making *relative* power measurements. The range of the meter can be extended by using an attenuator or by adding a broadband RF amplifier.

There are many uses for this meter. Combined with a signal generator, you can use it to characterize crystal and LC filters, for direct, on-air checks of field strength and measuring antenna patterns. If you are fortunate enough to have a 50-Ω oscilloscope probe (or a high-impedance probe), you can also use the Microwatter as an RF voltmeter for circuit testing.

Design

The circuit concept is quite simple (see **Fig 26.12**). Two nearly identical diodes (D1 and D2) are biased on by a small dc current. RF energy is coupled to one of the diodes, which then acts as a square-law detector. The difference between the voltages present on the two diodes is amplified by a differential amplifier (U1) and applied to an analog voltmeter equipped with hand-calibrated ranges. Getting adequate stability requires careful selection of parts.

D1 and D2 must have a low junction capacitance and a fast response in order to provide the desired bandwidth. Their temperature coefficients must also be extremely well matched. The solution is to use a very old part: the CA3039 diode array.[2] The diodes in this device are very fast silicon types, with exceptional thermal matching.[3] Any two diodes of the array—except the one tied to the substrate—can be used for D1 and D2. Unused IC pins can be cut off or bent out of the way. Similarly rated arrays of hot-carrier diodes may provide a better frequency response.

The Burr-Brown INA2128 provides two differential amplifiers with exceptional specifications: low thermal drift, superb common-mode rejection and low noise. In this circuit, one half of the INA2128 provides dc amplification of the detected voltage; the other half serves as a low-impedance offset-voltage source used for error nulling.

The RF-input terminating resistor is composed of two 100-Ω resistors (R3 and R4), providing a cleaner 50-Ω termination than a single resistor does. If you have them, use chip resistors. If not, clip the leads off two standard 100-Ω resistors and carefully scrape away any paint at the ends. After installing the resistors, verify the input resistance with an ohmmeter.

A stable power source is required. Operating the circuit from an unregulated battery supply allows slight relative changes in the local ground as the terminal voltage drops, creating drift in the reading. The Zener diodes (D3 and D4) following regulator U2 provide a simple, dynamic means of splitting the 12-V supply into the three voltages required to power the circuit.

C1 and C11 should be as close as physically possible to U1 pins 1 and 2. In combination with R1 and R5, these capacitors provide good immunity to unwanted RF. R14, the OFFSET NULL potentiometer, allows nulling the dc amplifier offset and any offset from slight mismatches in the CA3039 diodes. Use a multiturn potentiometer for R14—10 turns or more. Whether you use a panel-mount pot, or a PC-board-mounted trimmer as I did, R14 must be accessible. On the most-sensitive range, the instrument must be zeroed before each use.

The forward voltage drop difference between any two diodes in the CA3039 is specified as "typically less than 0.5 mV." If your device is typical, the nominal values of 5.1 kΩ for R10 and 3.9 kΩ for R9 will do just fine. The manufacturer does not reject parts until the forward voltage drop difference is *5 mV*—so the 0.5 mV figure isn't totally dependable. You can always rotate the diode array so that pins 1, 2, 3 and 4 occupy the spots designated for pins 5, 6, 7 and 8. That gives you a second set of diodes from which to choose. The simplest general solution for a homebrew project is to use nonidentical values for R9 and R10.

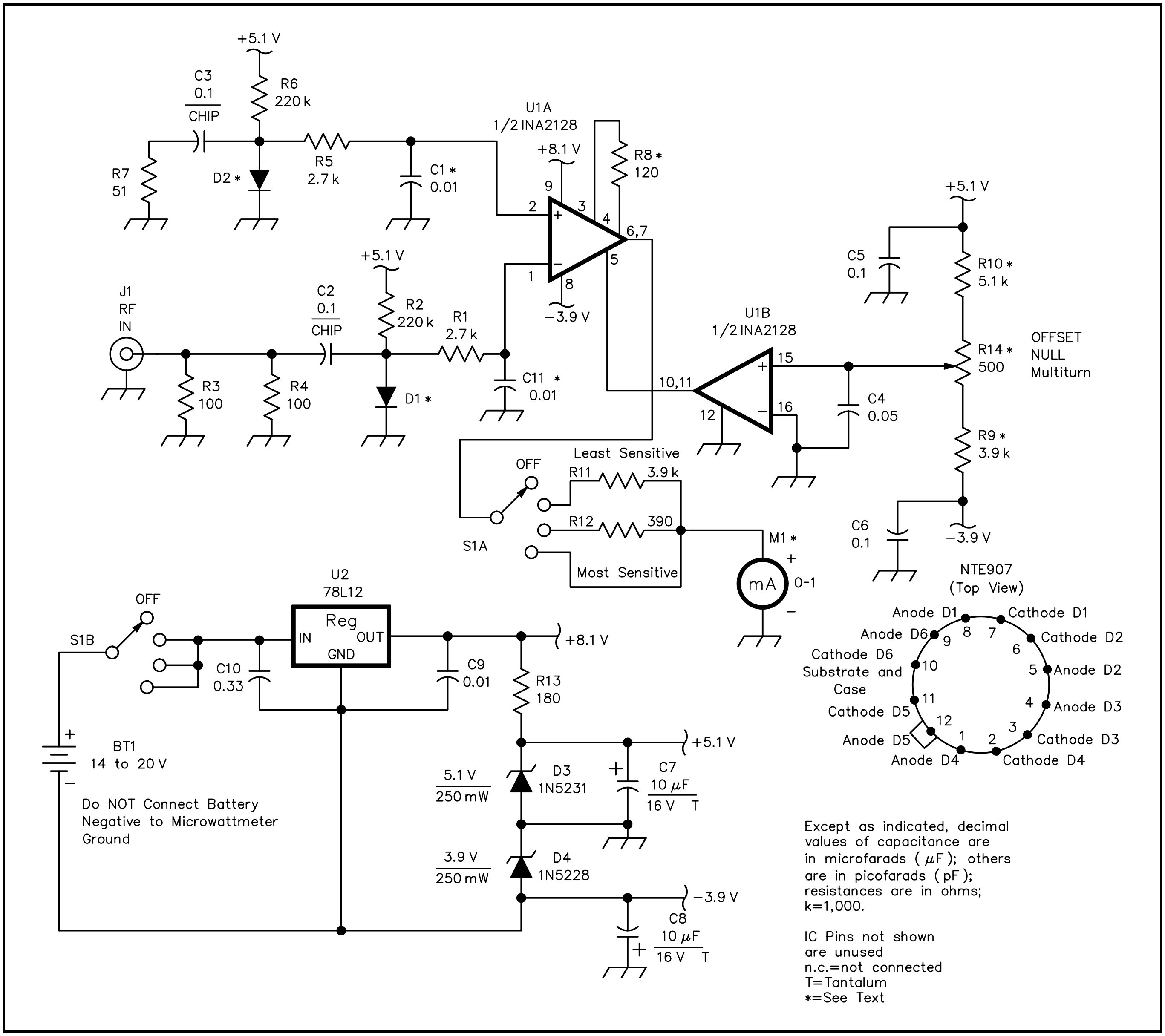


Fig 26.12—Schematic of the Microwatter circuit. Equivalent parts can be substituted. Unless otherwise specified, resistors are $^1/_4$-W, 5%-tolerance carbon-composition or film units.

D1, D2—Part of CA3039 or NTE907 diode array; see text and Notes 1 and 2. The NTE907 is available from Mouser Electronics (see References chapter Address List for contact information).

M1—1-mA meter movement, 50 Ω internal resistance; see text.

R14—500 Ω, 10-turn (or more) potentiometer; see text.

S1—2-pole, 4-position rotary switch.

U1—Burr Brown amplifier INA2128P; Digi-Key INA2128P; available from Digi-Key Corp (see References chapter Address List for contact information).

U2—78L12, positive 12-V, 100-mA voltage regulator; Digi-Key LM78L12ACZ.

If your diode array isn't typical, your Microwatter won't zero properly. The solution is simple. Remove R14 from the circuit. Temporarily connect one end of a 10-kΩ potentiometer to the +8.1 V supply and attach the other end to the –3.9 V supply. Connect the pot arm to the PC-board point of R14's arm. With this arrangement, you'll be able to zero your Microwatter, although the setting will be sensitive and might not produce an exact zero. Once you have the zero point, turn off the Microwatter and remove the 10-kΩ pot carefully without disturbing its setting. Measure the resistance between the pot arm and end that was connected to the +8.1 V point—this is the value of R10. R9's value is that measured between the pot arm and the end previously connected to the –3.9 V point. The combined value of these two resistors and that of R14 provides a total resistance of about 10.5 kΩ, and the resistance ratios will allow easy meter zeroing.

The Meter Movement

The specified meter movement has a 50-Ω internal resistance. Hence, on the most sensitive scale, a 50 mV output from U1 provides full-scale deflection. The middle and upper scales require 440 mV and 3.95 V, respectively, for full-scale deflection. Here's how you can use a meter that has a different internal resistance or current sensitivity.

Take, for example, a 25-µA meter movement with an internal resistance of 1910 Ω. A signal level of 47.8 mV (25 µA × 1910 Ω) provides full-scale deflection, so this meter can be substituted directly for the 1-mA, 50-Ω meter on the most sensitive scale. On the middle and top scales, the values of R12 and R11 should be 15.69 kΩ and 156.1 kΩ, or values close to those.

If the most sensitive meter you can find requires 80 to 100 mV to drive it to full scale deflection, don't worry. The Microwatter's most-sensitive scale will be compressed, but quite usable. The middle and top scales will still run full scale, probably not even requiring a change in the values of R11 and R12, if you use a 1-mA movement.

Construction and Calibration

For the prototype, components are mounted on double-sided PC board (see Note 1) with one side acting as a ground-plane. The cabinet is a 5 × 7 × 3-inch aluminum box, primed and painted gray (see **Fig 26.13**). Power is supplied by a set of NiCd batteries glued to the inside rear of the cabinet. The PC board is small enough to be mounted on the back of the input BNC connector without other support. Once you have completed PC-board assembly, remove the solder flux from the board.

To calibrate the Microwatter, you'll need a 50-Ω RF source with a known output level. A suitable signal source can be made from a crystal oscillator and an attenuator, using at least 6 dB of attenuation between the oscillator and the Microwatter at all times. The output of such a system can be calibrated with an oscilloscope, or with a simple RF meter. Any frequency in the mid-HF range is suitable for calibration—K7OWJ used 10 MHz.

Remove the meter-face cover and prepare the face for new markings. You can paint it white and use India ink to make the meter scale arcs and incremental marks. Before calibrating the Microwatter, let it stabilize at room temperature for an hour, and turn it on at least 15 minutes before use. (At these low power levels, even the heat generated by the small dc bias on the diodes needs time to stabilize.) Set the Microwatter to its most sensitive scale. Use R14 to set the meter needle to your chosen meter zero point and mark that point with a pencil. Then apply a –40 dBm signal and mark the top end of the scale. Decrease the input signal level in 1-dB steps, marking each step. Recheck your meter zeroing, switch the Microwatter to the middle scale, apply a –30 dBm signal and

Fig 26.13—Photo of the Microwatter.

mark this point. Again, decrease drive in 1-dB steps, marking each. Repeat with the least-sensitive scale, starting with –20 dBm. Once this is done, replace the meter-face cover and your Microwatter is ready to use.

DIRECTIONAL WATTMETERS

Directional wattmeters of varying quality are commonly used by the amateur community. The high quality standard is made by the Bird Electronic Corporation, who call their proprietary line THRULINE. The units are based on a sampling system built into a short piece of 50-Ω transmission line with plug-in elements for various power and frequency ranges.

AC BRIDGES

In its simplest form, the ac bridge is exactly the same as the Wheatstone bridge discussed earlier in the dc measurement section of this chapter. However, complex impedances can be substituted for resistances, as suggested by **Fig 26.14A**. The same bridge equation holds if Z (complex impedance) is substituted for R in each arm. For the equation to be true, however, both phase angles and magnitudes of the impedances must balance; otherwise, a true null voltage is impossible to obtain. This means that a bridge with all "pure" arms (pure resistance or reactance) cannot measure complex impedances; a combination of R and X must be present in at least one arm aside from the unknown.

The actual circuits of ac bridges take many forms, depending on the intended measurement and the frequency range to be covered. As the frequency increases, stray effects (unwanted capacitances and inductances) become more pronounced. At RF, it takes special attention to minimize them.

Most amateur built bridges are used for RF measurements, especially SWR measurements on transmission lines. The circuits at **Fig 26.14B** and C are favorites for this purpose.

Fig 26.14B is useful for measuring both transmission lines and lumped constant components. Combinations of resistance and capacitance are often used in one or more arms; this may be required for eliminating the effects of stray capacitance. The bridge shown in Fig 26.14C is used only on transmission lines and only on those lines having the characteristic impedance for which the bridge is designed.

SWR Measurement

The theory behind SWR measurement is covered in the **Transmission Lines** chapter and more fully in *The ARRL Antenna Book*. Projects to measure SWR appear in the **Station Setup and Accessory Projects** chapter of this *Handbook*.

Notes

[1] A PC board and some components for this project is available from FAR Circuits. A PC-board template package is not available.

[2] The NTE907 replacement is available from Mouser Electronics.

[3] The absolute value of the difference in forward drop between any two diodes is 1 mV/°C.

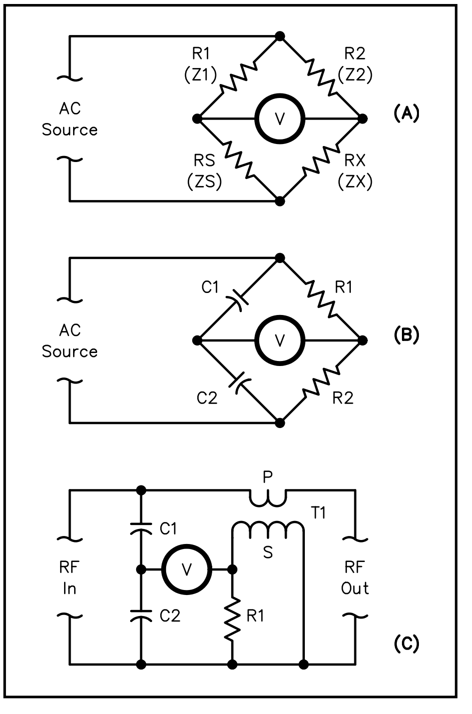


Fig 26.14—A shows a bridge circuit generalized for ac or dc use. B is a form of ac bridge for RF applications. C is an SWR bridge for use in transmission lines.

Frequency Measurement

The FCC Rules for Amateur Radio require that transmitted signals stay inside the frequency limits of bands consistent with the operator's license privileges. The exact frequency need not be known, as long as it is within the limits. On these limits there are no tolerances: Individual amateurs must be sure that their signal stays safely inside. The current limits for each license class can be found in the **References** chapter and in the current edition of *The FCC Rule Book*, published by the ARRL.

Staying within these limits is not difficult; many modern transceivers do so automatically, within limits. If your radio uses a PLL synthesized frequency source, just tune in WWV or another frequency standard occasionally.

Checks on older equipment require some simple equipment and careful adjusting. The equipment commonly used is the frequency marker generator and the method involves use of the station receiver, as shown in **Fig 26.15**.

FREQUENCY MARKER GENERATORS

A marker generator, in its simplest form, is a high-stability oscillator that generates a series of harmonic signals. When an appropriate fundamental is chosen, harmonics fall near the edges of the amateur frequency allocations.

Most US amateur band and subband limits are exact multiples of 25 kHz. A 25-kHz fundamental frequency will therefore produce the right marker signals if its harmonics are strong enough. But since harmonics appear at 25-kHz intervals throughout the spectrum, there is still a problem of identifying particular markers. This is easily solved if the receiver has reasonably good calibration. If not, most marker circuits provide a choice of fundamental outputs, say 100 and 50 kHz as well as 25 kHz. Then the receiver can be first set to a 100-kHz interval. From there, the desired 25-kHz (or 50-kHz) points can be counted. Greater frequency intervals are rarely required. Instead, tune in a signal from a station of known frequency and count off the 100-kHz points from there.

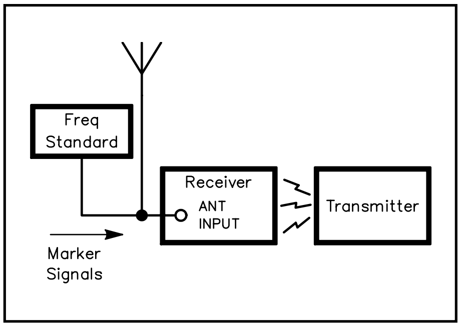


Fig 26.15 — A setup for checking transmitter frequency. Use care to ensure that the transmitter does not overload the receiver. False signals would result; see text.

Transmitter Checking

To check transmitter frequency, tune in the transmitter signal on a calibrated receiver and note the dial setting at which it is heard. To start, reduce the transmitter to its lowest possible level to avoid receiver overload. Also, place a direct short across the receiver antenna terminals, reduce the RF gain to minimum and switch in any available receiver attenuators to help prevent receiver IMD, overload and possible false readings.

Place the transmitter on standby (not transmitting) and use the marker generator as a signal source. Tune in and identify the nearest marker frequencies above and below the transmitter signal. The transmitter frequency is between these two known frequencies. If the marker frequencies are accurate, this is all you need to know, except that the transmitter frequency must not be so close to a band (or subband) edge that sidebands extend past the edge.

If the transmitter signal is inside a marker at the edge of an assignment, to the extent that there is an audible beat note with the receiver BFO turned off, normal CW sidebands are safely inside the edge. (So long as there are no abnormal sidebands such as those caused by clicks and chirps.) For 'phone the safety allowance is usually taken to be about 3 kHz, the usual width of one sideband. A frequency difference of this much can be estimated by noting the receiver dial settings for the two 25-kHz markers that are either side of the signal and dividing 25 by the number of dial divisions between them. This will give the number of kilohertz per dial division. It is a prudent practice to allow an extra kHz margin when setting the transmitter close to a band or subband edge (5-kHz is a safe HF margin for most modes on modern transmitters).

Transceivers

The method described above is good when the receiver and transmitter are separate pieces of equipment. When a transceiver is used and the transmitting frequency is automatically the same as that to which the receiver is tuned, setting the tuning dial to a spot between two known marker frequencies is all that is required. The receiver incremental tuning control (RIT) must be turned off.

The proper dial settings for the markers are those at which, with the BFO on, the signal is tuned to zero beat (the spot where the beat note disappears as tuning makes its pitch progressively lower). Exact zero beat can be determined by a very slow rise and fall of background noise, caused by a beat of a cycle or less per second. In receivers with high selectivity it may not be possible to detect an exact zero beat, because low audio frequencies from beat notes may be prevented from reaching the speaker or headphones.

Most commercial equipment has some way to match either the equipment's internal oscillator or marker generator with the signal received from WWV on one of its short-wave frequencies. It is a good idea to do this check on a new piece of gear. A recheck about a month later will show if anything has changed. Normal commercial equipment drifts less than 1 kHz after warm up.

Also check the dial linearity of equipment that has an analog dial or subdial. Often analog dials do not track frequency accurately across an entire band. Such radios usually provide for pointer adjustment so that dial error can be minimized at the most often used part of a band.

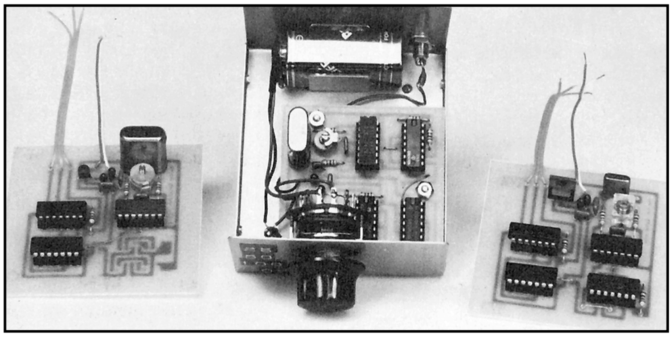
Frequency-Marker Circuits

The frequency in most amateur frequency markers is determined by a 100-kHz or 1-MHz crystal. Although the marker generator should produce harmonics every 25-kHz and 50-kHz, crystals (or other high-stability resonators) for frequencies lower than 100 kHz are expensive and rare. There is really no need for them, however, since it is easy to divide the basic frequency down to the desired frequency; 50- and 25-kHz steps require only two successive divisions by two (from 100 kHz). In the division process, the harmonics of the generator are strengthened so they are useful up to the VHF range. Even so, as frequency increases the harmonics weaken.

Current marker generators are based on readily available crystals. A 1 MHz basic oscillator would first be divided by 10 to produce 100 kHz and then followed by two successive divide-by-two stages to produce 50 kHz and 25 kHz.

A MARKER GENERATOR WITH SELECTABLE OUTPUT

Fig 26.16 shows a marker generator with selectable output for 100, 50 or 25-kHz intervals. It provides marker signals well up into the 2-m band. The project was first built by Bruce Hale, KB1MW, in the ARRL Lab. A more detailed presentation appeared in the **Station Accessories** chapter of *Handbooks* from 1987 through 1994. An etching pattern and parts-placement diagram are available for this project. See Chapter 30, **References**, for the Marker Generator template.



A 1, 2 or 4-MHz computer-surplus crystal is suitable for Y1. Several prototypes were built with such crystals and all could be tuned within 50 Hz

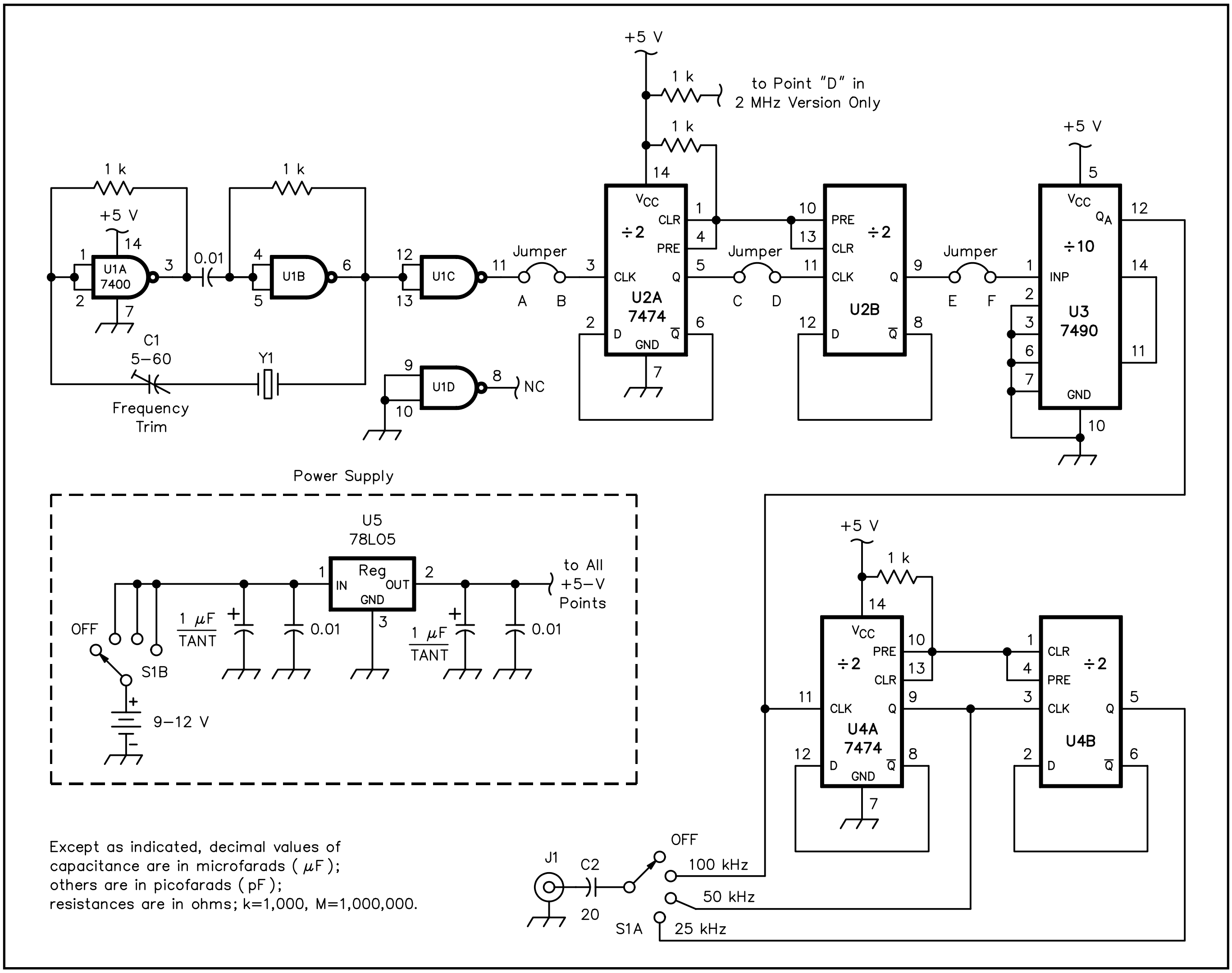


Fig 26.16 — Schematic diagram of the marker generator.

C1 — 5-60 pF miniature trimmer capacitor.
C2 — 20-pF disc ceramic.
U1 — 7400 or 74LS00 quad NAND gate.
U2, U4 — 7474 or 74LS74 dual D flip-flop.
U3 — 7490 or 74LS90 decade counter.
U5 — 78L05, 7805 or LM340-T5 5-V voltage regulator.
Y1 — 1, 2 or 4-MHz crystal in HC-25, HC-33 or HC-6 holder.

at 100 kHz. The marker division ratio must be chosen once the crystal frequency is selected. This is accomplished by means of several jumpers that disable part or all of U2A, the dual flip-flop IC. When Y1 is a 1-MHz crystal, U2 may be omitted entirely. **Table 26.2** gives the jumper placement for each crystal frequency.

The prototypes were built with TTL logic ICs. Unused TTL gate inputs should always be connected either to ground or to VCC through a pull-up resistor. If low-power Schottky (they have "LS" as part of their numerical designator) parts are available for U1 through U4, use them. They draw much less current than plain TTL, and will greatly extend battery life.

Table 26.2
Marker Generator Jumper Placement

Crystal Frequency	*Jumper Placement*
1 MHz	A to F (U2 not used)
2 MHz	A to B
	C to F
	D to +5 V, via 1-kW resistor
4 MHz	A to B
	C to D
	E to F

DIP METERS

This device is often called a transistor dip meter or a grid-dip oscillator (from vacuum-tube days). Most dip meters can also serve as absorption frequency meters (in this mode measurements are read at the current peak, rather than the dip). Further, some dip meters have a connection for headphones. The operator can usually hear signals that do not register on the meter. Because the dip meter is an oscillator, it can be used as a signal generator in certain cases where high accuracy or stability are not required.

A dip meter may be coupled to a circuit either inductively or capacitively. Inductive coupling results from the magnetic field generated by current flow. Therefore, inductive coupling should be used when a conductor with relatively high current is convenient. Maximum inductive coupling results when the axis of the pick-up coil is placed perpendicular to a nearby current path (see **Fig 26.17**).

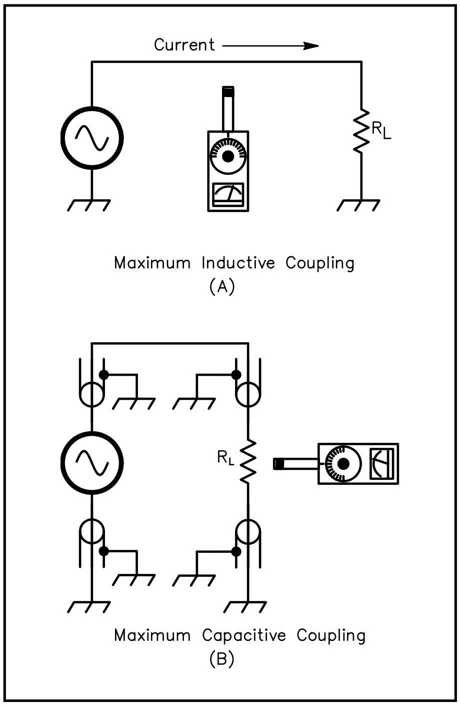


Fig 26.17 — Dip meter coupling. (A) uses inductive coupling and (B) uses capacitive coupling.

Capacitive coupling is required when current paths are magnetically confined or shielded. (Toroidal inductors and coaxial cables are common examples of magnetic self shielding.) Capacitive coupling depends on the electric field produced by voltage. Use capacitive coupling when a point of relatively high voltage is convenient. (An example might be the output of a 12-V powered RF amplifier. *Do not* attempt dip-meter measurements on true high-voltage equipment such as vacuum-tube amplifiers or switching power supplies while they are energized.) Capacitive coupling is maximum when the end of the pick-up coil is near a point of high voltage (see Fig 26.17). In either case, the circuit under test is affected by the presence of the dip meter. Always use the minimum coupling that yields a noticeable indication.

Use the following procedure to make reliable measurements. First, bring the dip meter gradually closer to the circuit while slowly varying the dip-meter frequency. When a current dip occurs, hold the meter steady and tune for minimum current. Once the dip is found, move the meter away from the circuit and confirm that the dip comes from the circuit under test (the current reading should increase with distance from the circuit until the dip is gone). Finally, move the meter back toward the circuit until the dip is just noticed. Retune the meter for minimum current and read the dip-meter frequency with a calibrated receiver or frequency meter.

The current dip of a good measurement is smooth and symmetrical. An asymmetrical dip indicates that the dip-meter oscillator frequency is being significantly influenced by the test circuit. Such conditions do not yield usable readings.

A measurement of effective unloaded inductor Q can be made with a dip meter and an RF voltmeter (or a dc voltmeter with an RF probe). Make a parallel resonant circuit using the inductor and a capacitance equal to that of the application circuit. Connect the RF voltmeter across this parallel combination and measure the resonant frequency. Adjust the dip-meter/circuit coupling for a convenient reading on the voltmeter, then maintain this dip-meter/circuit relationship for the remainder of the test. Vary the dip meter frequency until the voltmeter reading drops to 0.707 times that at resonance. Note the frequency of the dip meter and repeat the process, this time varying the frequency on the opposite side of resonance. The difference between the two dip meter readings is the test-circuit bandwidth. This can be used to calculate the circuit Q:

$$Q = \frac{f_0}{BW} \tag{9}$$

where

f_0 = operating frequency,

BW = measured bandwidth in the same units as the operating frequency.

When purchasing a dip meter, look for one that is mechanically and electrically stable. The coils should be in good condition. A headphone connection is helpful. Battery-operated models are convenient for antenna measurements.

A DIP METER WITH DIGITAL DISPLAY

An up-to-date dip meter was described by Larry Cicchinelli in the October 1993 issue of *QEX*. It consists of a dip meter with a three-digit frequency display. The analog portion of the circuit consists of an FET oscillator, voltage-doubler detector, dc-offset circuit and amplifier. The digital portion of the circuit consists of a high-impedance buffer, prescaler, counter, display driver, LED display and control circuit.

Circuit Description

The dip meter shown in **Fig 26.18** has four distinct functional blocks. The RF oscillator is a standard Colpitts using a common junction FET, Q1, as the active element. Its range is about 1.7 to 45 MHz. The 200-pF tuning capacitor gives a 2:1 tuning range. A 2:1 frequency range requires a capacitor with a 4:1 range. The sum of the minimum capacitance of the variable, the capacitors across the inductor and the strays must therefore be in the order of 70 pF. The values of L1 were determined experimentally by winding the coils and observing the lower and upper frequency values. [**Table 26.3** shows winding data calculated from the author's schematic. —*Ed.*] The coil sizes were experimentally determined, and the coils are constructed on $1^1/_4$-inch-diameter plug-in coil forms with #20 enameled wire. They are close wound. The number of turns shown are only a starting point; you may need to change them slightly in order to cover the desired frequency range.

The tapped capacitors are mounted inside the coil forms so that their values could be different for each band if required. The frequency spread of the lowest band is less than 2:1 because the tapped capacitor values are larger than those for the other bands. The frequency spread of the highest band is greater than 2:1 because its capacitors are smaller.

The analog display circuit begins with a voltage-doubler detector in order to get higher sensitivity. It drives a dc-offset circuit, U1A. R1 inserts a variable offset that is subtracted from the detector voltage. This allows the variable gain stage, U1B, to be more sensitive to variations in the detector output voltage. Q9 follows U1C to get an output gating voltage closer to ground. The resistor in series with the meter is chosen to limit the meter current to a safe value. For example, if a 1-mA meter is used, the resistor should be 8.2 kΩ.

The prescaler begins with a high-impedance buffer and amplifier, Q2 and Q3. If you are going to use the meter for the entire frequency range described, take care in the layout of both the oscillator and buffer/ amplifier circuits. The digital portion of the prescaler is a divide-by-100 circuit consisting of two divide-by-10 devices, U2 and U3. The devices used were selected because they were available. Any similar devices may be used as long as the reset circuit is compatible. Q5 is a level translator that shifts the 5-V signal to 9 V.

The first part of the digital display block is the oscillator circuit of U1C, which creates the gate time

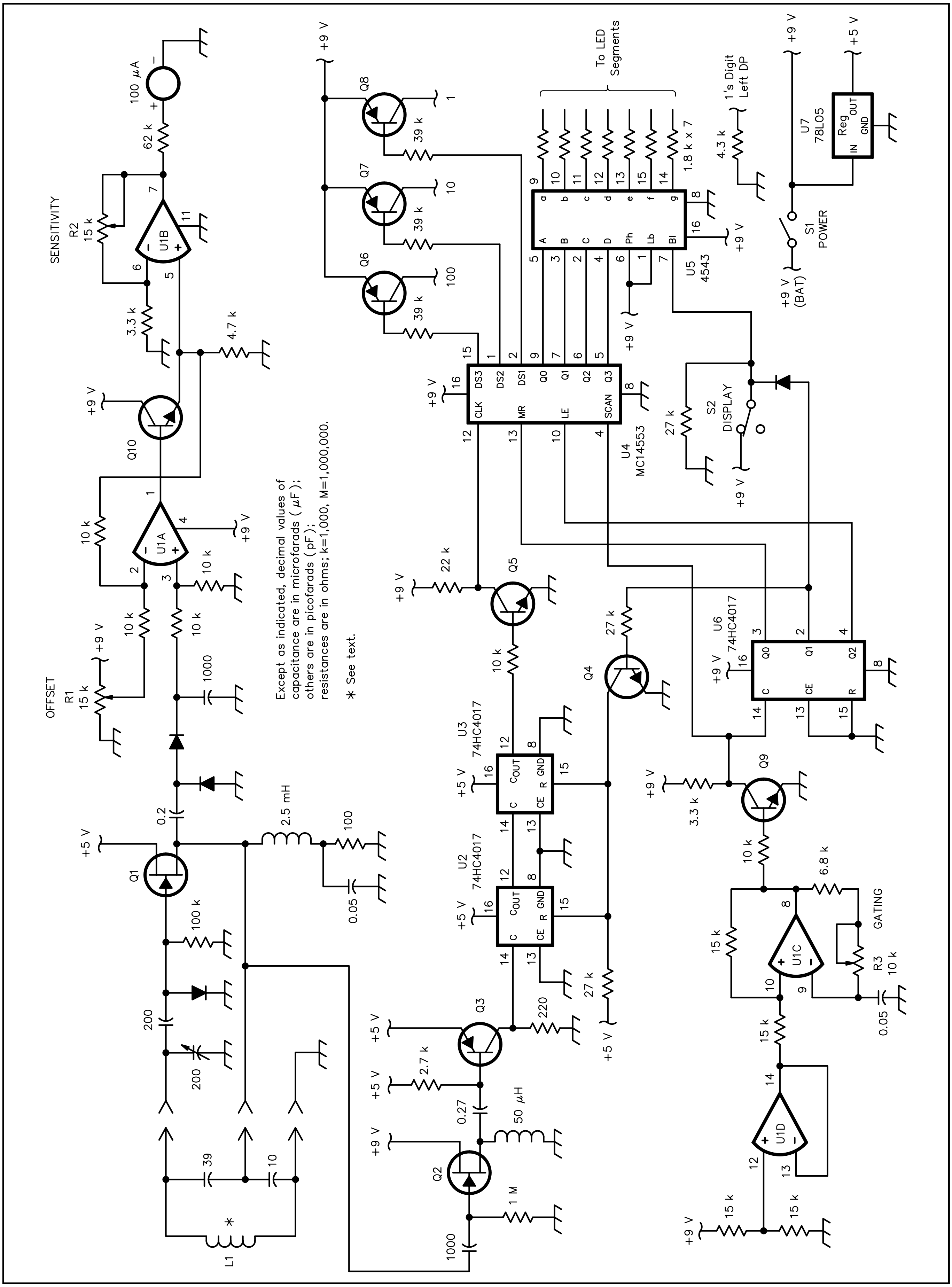
Table 26.3
Calculated Coil Data for the Dip Meter

Frequency (MHz)	*L µH*	*Turns*
1.7 to 3.1	48.6	52
2.8 to 5.9	16.3	23
5.6 to 11.9	4.0	9
9.7 to 20.7	1.3	5
19.0 to 45.0	0.3	2

Fig 26.18 — A schematic diagram of the dip meter. All diodes are 1N914 or similar. All resistors are $^1/_4$ W, 5%.

Q1, Q2 — MPF102 JFET transistor.
Q3, Q6, Q7, Q8 — 2N3906 PNP transistor.
Q4, Q5, Q9, Q10 — 2N3563 (or any general purpose NPN).
U1 — LM324.
U2, U3 — 74HC4017.
U4 — MC14553.
U5 — 4543.
U6 — 4017.
U7 — 78L05.

[Schematic on next page.]


Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms; k=1,000, M=1,000,000.
* See text.
SENSITIVITY
R2
15 k
OFFSET
R1
15 k
100 μA
62 k
3.3 k
4.7 k
U1A
U1B
U1C
U1D
Q1
Q2
Q3
Q4
Q5
Q6
Q7
Q8
Q9
Q10
L1
2.5 mH
50 μH
U2
74HC4017
U3
74HC4017
U4
MC14553
U5
4543
U6
74HC4017
U7
78L05
To LED
Segments
1.8 k x 7
1's Digit
Left DP
4.3 k
S1
POWER
S2
DISPLAY
GATING
R3
10 k
6.8 k
+9 V (BAT)
+9 V
+5 V
Reg
IN
OUT
GND

for the frequency counter. R3 adjusts the oscillator to a frequency of 500 Hz, yielding a 1-ms gate. The best way to set this frequency is to listen for the dip-meter output on a communications receiver and adjust R3 until the display agrees with the receiver. Once this calibration has been made for one of the bands, all the bands are calibrated. U1D gives a low-impedance voltage reference for U1C. Q9 was added to the output of the oscillator to remove a small glitch, which can cause the counter to trigger incorrectly. This type of oscillator has the advantage of simplicity. This circuit is fairly stable, easy to adjust and has a low parts count.

The digital system controller, U6, is a divide-by-10 counter. It has 10 decoded outputs, each of which goes high for one period of the input clock. The Q0 output is used to reset the frequency counter, U4. The Q1 output is used to enable the prescaler and disable the display and the Q2 output latches the count value into the frequency counter. Since the prescaler can only count while Q1 is high, it will be enabled for only 1 ms. Normally a 1-ms gate will yield 1-kHz resolution. Since the circuit uses a divide-by-100 prescaler, the resolution becomes 100 kHz.

The frequency counter, U4, is a three-digit counter with multiplexed BCD outputs. The clock input is driven from the prescaler, hence it is the RF oscillator frequency divided by 100. This signal is present for only 1 ms out of every 10 ms. The digit scanning is controlled by the 500-Hz oscillator of U1C.

U5 is a BCD-to-seven-segment decoder/driver. Its outputs are connected to each of the three common-anode, seven-segment displays in parallel. Only the currently active digit will be turned on by the digit strobe outputs of U5, via Q6, Q7 and Q8. The diode connected to the blanking input of U5 disables the display while U4 is counting. U7 is a 5-V regulator that allows the use of a single 9-V battery for both the circuit and the LEDs. S2 turns on the displays once the unit has been adjusted for a dip.

The circuit draws about 20 mA with the LEDs off and up to 35 mA with the LEDs on.

Many of the resistor values are not critical, and those used were chosen based upon availability; the op-amp circuits depend primarily on resistance ratios. The resistor at the collector of Q3 is critical and should not be varied. Use 0.27-µF monolithic capacitors. They have the required good high-frequency characteristics over the range of the meter.

Most of the parts can be purchased from Digi-Key. They did not have the ’4543 IC, which was purchased from Hosfelt Electronics. (See Address List in the **References** chapter.) The 74HC4017 may be substituted with a 74HCT4017. The circuit is built on a 4-inch-square perf board (with places for up to 12 ICs) and is housed in a 7 × 5 × 3-inch minibox.

Operation

To use the unit, set the gain, R2, fully clockwise for maximum sensitivity. With this setting, the output of the offset circuit (Q10 emitter) is at ground. As R1 is rotated, the voltage on the arm approaches and then becomes less than, the detector output. At this point the meter will start to deflect upward. Adjust R1 so that the meter reads about center scale. (Manual adjustment allows for variations in the output level of the RF oscillator.) As L1 is brought closer to the circuit under test, the meter will deflect downward as energy is absorbed by the circuit. For best results use the minimum possible coupling to the circuit being tested. If the dip meter is overcoupled to the test circuit, the oscillator frequency will be pulled.

FREQUENCY COUNTERS

One of the most accurate means of measuring frequency is a frequency counter. This instrument is capable of numerically displaying the frequency of the signal supplied to its input. For example, if an oscillator operating at 8.244 MHz is connected to a counter input, 8.244 would be displayed. At present, counters are usable well up into the gigahertz range. Most counters that are used at high frequencies make use of a prescaler ahead of a basic low-frequency counter. A prescaler divides the high-frequency signal by 10, 100, 1000 or some other fixed amount so that a low-frequency counter can display the operating frequency.

The accuracy of the counter depends on its internal crystal reference. A more accurate crystal reference yields more accurate readings. Crystals for frequency counters are manufactured to close tolerances. Most counters have a trimmer capacitor so that the crystal can be set exactly on frequency. Crystal frequencies of 1 MHz, 5 MHz or 10 MHz have become more or less standard. For calibration, harmonics of the crystal can be compared to a known reference station, such as those shown in Table 26.1, or other frequency standard and adjusted for zero beat.

Many frequency counters offer options to increase the accuracy of the counter timebase; this directly increases the counter accuracy. These options usually employ temperature-compensated crystal oscillators (TCXOs) or crystals mounted in constant temperature ovens that keep the crystal from being affected by changes in ambient (room) temperature. Counters with these options may be accurate to 0.1 ppm (part per million) or better. For example, a counter with a timebase accuracy of 5.0 ppm and a second counter with a TCXO accurate to 0.1 ppm are available to check a 436-MHz CW transmitter for satellite use. The counter with the 5-ppm timebase could have a frequency error of as much as 2.18 kHz, while the possible error of the counter with the 0.1 ppm timebase is only 0.0436 kHz.

Other Instruments and Measurements

This section covers a variety of test equipment that is useful in receiver and transmitter testing. It includes RF and audio generators, an inductance meter, oscilloscopes, spectrum analyzers, a calibrated noise source, a noise bridge, an advanced resonance indicator, combiners, attenuators and dummy loads. A number of applications of this equipment to basic transmitter and receiver testing is also included.

RF OSCILLATORS FOR CIRCUIT ALIGNMENT

Receiver testing and alignment uses equipment common to ordinary radio service work. Inexpensive RF signal generators are available, both complete and in kit form. However, any source of signal that is weak enough to avoid overloading the receiver usually will serve for alignment work. The frequency marker generator is a satisfactory signal source. In addition, its frequencies, although not continuously adjustable, are known far more precisely, since the usual signal-generator calibration is not highly accurate. An attenuator described later in this chapter can be added for relative dB measurements. When buying a used or inexpensive signal generator, look for these attributes: output level is calibrated, the output doesn't "ring" too badly when tapped, and doesn't drift too badly when warmed up. Many military surplus units are available that can work quite well. Commercial units such as the HP608 are big and stable, and they may be inexpensive.

AUDIO-FREQUENCY OSCILLATORS

An audio signal generator should provide a reasonably pure sine wave. The best oscillator circuits for this use are RC coupled, operating as close to a class-A amplifier as possible. Variable frequencies covering the entire audio range are needed for determining frequency response of audio amplifiers.

An oscillator generating one or two frequencies with good waveform is sufficient for most phone-transmitter testing and simple troubleshooting in AF amplifiers. A two-tone (dual) oscillator is very useful for testing and adjusting sideband transmitters.

A circuit of a simple RC oscillator that is useful for general testing is given in **Fig 26.19**. This Twin-T arrangement gives a waveform that is satisfactory for most purposes. The oscillator can be operated at any frequency in the audio range by varying the component values. R1, R2 and C1 form a low-pass network, while C2, C3 and R3 form a high-pass network. As the phase shifts are opposite, there is only one frequency at which the total phase shift from collector to base is 180°: Oscillation will occur at this frequency. When C1 is about twice the capacitance of C2 or C3 the best operation results. R3 should have a resistance about 0.1 that of R1 or R2 (C2 = C3 and R1 = R2). Output is taken across C1, where the harmonic distortion is least. Use a relatively high impedance load — 100 kΩ or more.

Most small-signal AF transistors can be used for Q1. Either NPN or PNP types are satisfactory if the supply polarity is set correctly. R4, the collector load resistor may be changed a little to adjust the oscillator for best output waveform.

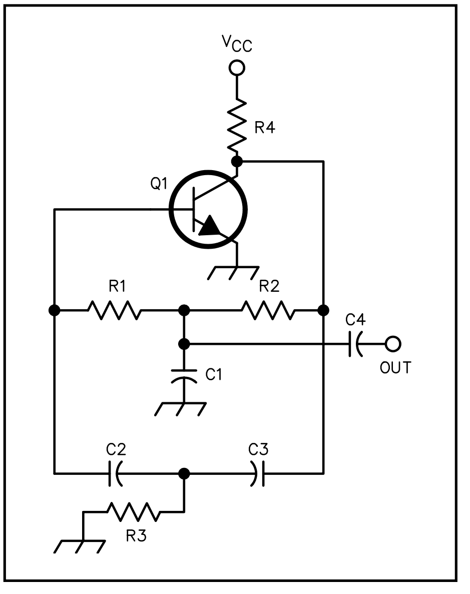


Fig 26.19 — Values for the twin-T audio oscillator circuit range from 18 kΩ for R1-R2 and 0.05 µF for C1 (750 Hz) to 15 kΩ and 0.02 µF for 1800 Hz. For the same frequency range, R3 and C2-C3 vary from 1800 Ω and 0.02 µF to 1500 Ω and 0.01 µF. R4 is 3300 Ω and C4, the output coupling capacitor, can be 0.05 µF for high-impedance loads.

A WIDE-RANGE AUDIO OSCILLATOR

A wide-range audio oscillator that will provide a moderate output level can be built from a single 741 operational amplifier (**Fig 26.20**). Power is supplied by two 9-V batteries from which the circuit draws 4 mA. The frequency range is selectable from 8 Hz to 150 kHz. Distortion is approximately 1%. The output level under a light load (10 kΩ) is 4 to 5 V. This can be increased by using higher battery voltages, up to a maximum of plus and minus 18 V, with a corresponding adjustment of R_F.

Pin connections shown are for the TO-5 case and the eight-pin DIP package. Variable resistor R_F is trimmed for an output level of about 5% below clipping as seen on an oscilloscope. This should be done for the temperature at which the oscillator will normally operate, as the lamp is sensitive to ambient temperature. This unit was originally described by Shultz in November 1974 *QST*; it was later modified by Neben as reported in June 1983 *QST*.

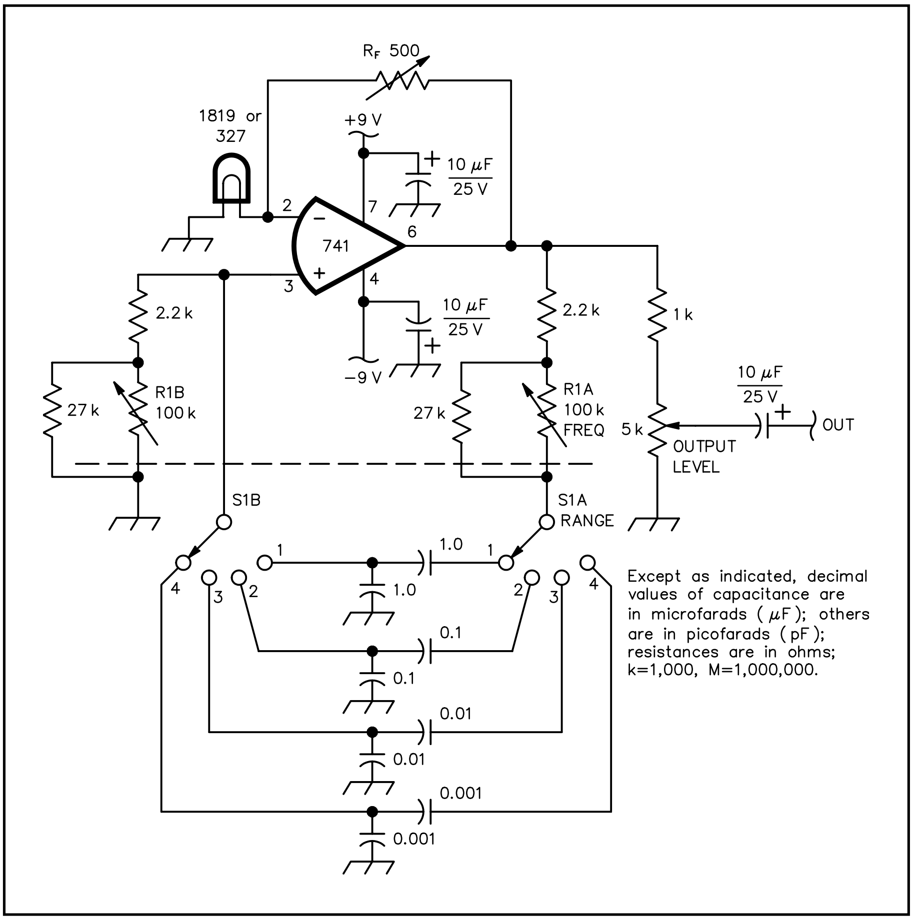


Fig 26.20 — A single IC (741 op amp) based audio oscillator. The frequency range is set by switch S1.

MEASURE INDUCTANCE AND CAPACITANCE WITH A DVM

Many of us have a DVM (Digital Voltmeter) or VOM (Volt Ohm Meter) in the shack, but few of us own an inductance or capacitance meter. If you have ever looked into your junk box and wanted to know the value of the unmarked parts, these simple circuits will give you the answer. They may be built in one evening (**Fig 26.23** and **Fig 26.24**), and will adapt your DVM or VOM to measure inductance or capacitance. The units are calibrated against a known part. Therefore, the overall accuracy depends only on the calibration values and not on the components used to build the circuits. If it is carefully calibrated, an overall accuracy of 10% may be expected if used with a DVM and slightly less with a VOM.

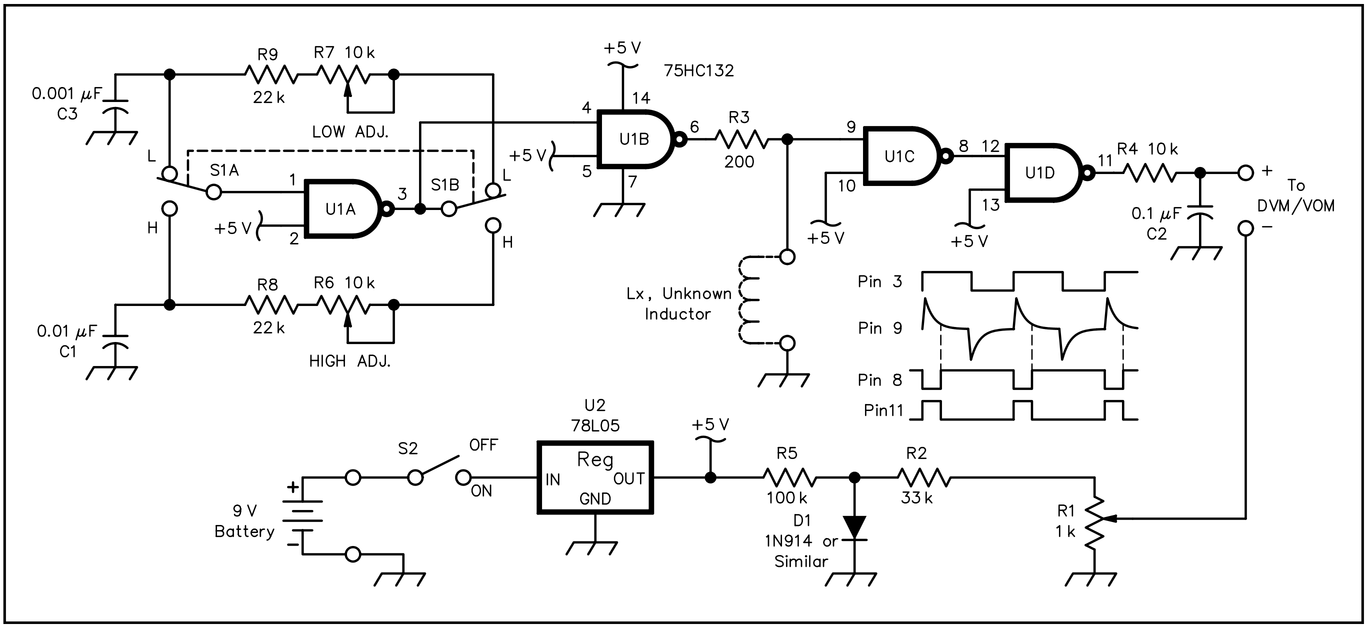


Fig 26.23—All components are 10% tolerance. 1N4148 or equivalent may be substituted for D1. An LM7805 may be substituted for the 78L05. All fixed resistors are 1/4-W carbon composition. Capacitors are in µF.

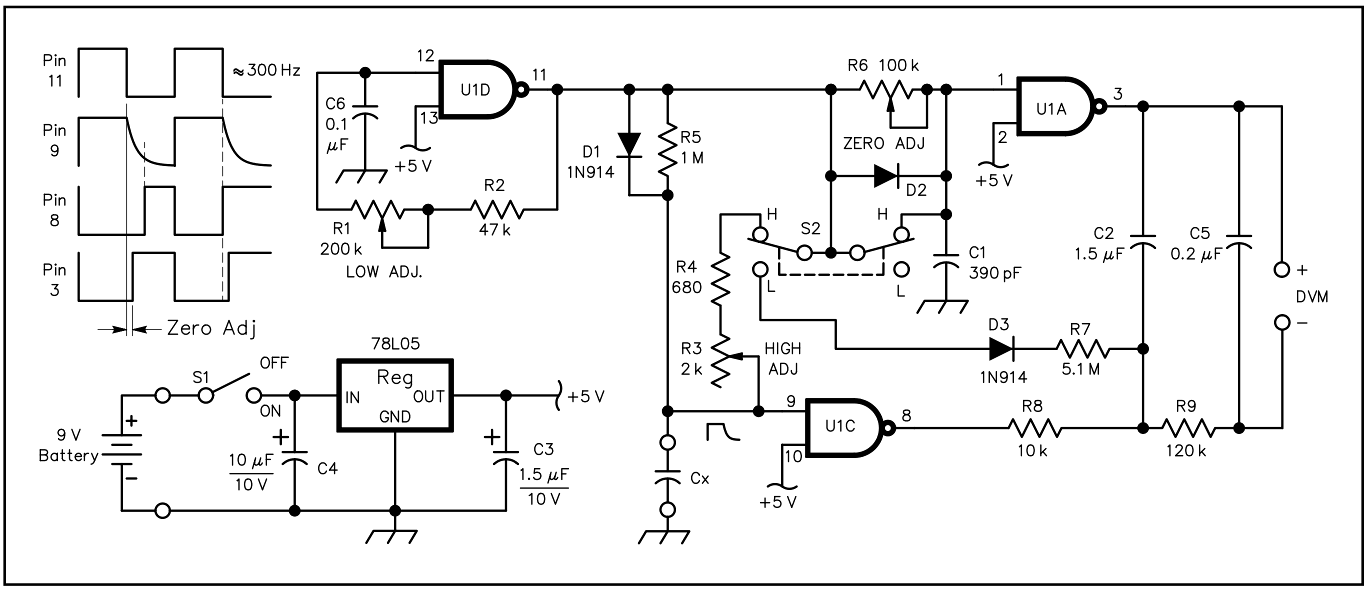


Fig 26.24—All components are 10% tolerance. An LM7805 may be substituted for the 78L05. All fixed resistors are 1/4-W carbon composition. Capacitors are in µF unless otherwise indicated.

CONSTRUCTION

The circuits may be constructed on a small perf board (Radio Shack dual mini board, # 276-168), or if you prefer, on a PC board. A template and parts layout may be obtained from the ARRL.[1] Layout is non-critical—almost any construction technique will suffice. Wire-wrapping or point-to-point soldering may be used.

The circuits are available in kit form from Electronic Rainbow Inc. (see Address List in the **References** chapter). The IA inductance adapter kit is $14.95. A separate cabinet is available for $8.95. The PC board is 1.75 × 2.5 inches. The CA-1 capacitance adapter kit is $12.95 and comes with a 1.80 × 2.0-inch PC board.

INDUCTANCE ADAPTER FOR DVM/VOM

Description

The circuit shown in Fig 26.23 converts an unknown inductance into a voltage that can be displayed on a DVM or VOM. Values between 3 and 500 µH are measured on the low range and from 100 µH to 7 mH on the high range. NAND gate U1A is a two frequency RC square-wave oscillator. The output frequency (pin 3) is approximately 60 kHz in the low range and 6 kHz in the high range. The square-wave output is buffered by U1B and applied to a differentiator formed by R3 and the unknown inductor, LX. The stream of spikes produced at pin 9 decay at a rate proportional to the time constant of R3-LX. Because R3 is a constant, the decay time is directly proportional to the value of LX. U1C squares up the positive going spikes, producing a stream of negative going pulses at pin 8 whose width is proportional to the value of LX.

They are inverted by U1D (pin 11) and integrated by R4-C2 to produce a steady dc voltage at the + output terminal. The resulting dc voltage is proportional to LX and the repetition rate of the oscillator. R6 and R7 are used to calibrate the unit by setting a repetition rate that produces a dc voltage corresponding to the unknown inductance. D1 provides a 0.7 volt constant voltage source that is scaled by R1 to produce a small offset reference voltage for zeroing the meter on the low inductance range.

When S1 is low, *mV* corresponds to *µH* and when high, *mV* corresponds to *mH*. A sensitive VOM may be substituted for the DVM with a sacrifice in resolution.

Test and Calibration

Short the LX terminals with a piece of wire and connect a DVM set to the 200-mV range to the output. Adjust R1 for a zero reading. Remove the short and substitute a known inductor of approximately 400 µH. Set S1 to the low (in) position and adjust R7 for a reading equal to the known inductance. Switch S1 to the high position and connect a known inductor of about 5 mH. Adjust R6 for the corresponding value. For instance, if the actual value of the calibration inductor is 4.76 mH, adjust R7 so the DVM reads 476 mV.

CAPACITANCE ADAPTER FOR DVM/VOM

Description

The circuit shown in Fig 26.24 measures capacitance from 2.2 to 1000 pF in the low range, from 1000 pF to 2.2 µF in the high range. U1D of the 74HC132 (pin 11) produces a 300 Hz square-wave clock. On the rising edge CX rapidly charges through D1. On the falling edge CX slowly discharges through R5 on the low range and through R3-R4 on the high range. This produces an asymmetrical waveform at pin 8 of U1C with a duty cycle proportional to the unknown capacitance, CX. This signal is integrated by R8-R9-C2 producing a dc voltage at the negative meter terminal proportional to the unknown capacitance. A constant reference voltage is produced at the positive meter terminal by integrating the square-wave at U1A, pin 3. R6 alters the symmetry of this square-wave producing a small change in the reference voltage at the positive meter terminal. This feature provides a zero adjustment on the low range. The DVM measures the difference between the positive and negative meter terminals. This difference is proportional to the unknown capacitance.

Test and Calibration

Without a capacitor connected to the input terminals, set SW2 to the low range (out) and attach a DVM to the output terminals. Set the DVM to the 2-volt range and adjust R6 for a zero meter reading. Now connect a 1000 pF calibration capacitor to the input and adjust R1 for a reading of 1.00 volt. Switch to the high range and connect a 1.00 µF calibration capacitor to the input. Adjust R3 for a meter reading of 1.00 volts. The calibration capacitors do not have to be exactly 1000 pF or 1.00 µF, as long as you know their exact value. For instance, if the calibration capacitor is known to be 0.940 µF, adjust the output for a reading of 940 mV.

Note

[1] See Chapter 30, **References**, for the template.

A SIX DIGIT PROGRAMMABLE FREQUENCY COUNTER AND DIGITAL DIAL

This six digit programmable frequency counter has multiple uses around the shack. It can be used as a general purpose frequency counter, a digital dial for home-built rigs, an add-on digital dial for commercial rigs, or it can be added to vintage tube-type receivers for precise frequency spotting. The counter displays frequencies greater than 50 MHz with 100-Hz resolution. It updates 40 times per second, providing near real time tuning response, and is programmable to account for the offset between a receiver local oscillator and the operating frequency. It also will display the correct operating frequency with reverse tuning local oscillators.

The counter is based on the PIC16C57, an eight bit CMOS Microcontroller by Microchip Technology Inc. This part is custom programmed by Radio Adventures Corp. (see Address List in the **References** chapter), and is supplied by them as the C5 frequency counter chip. The easiest way to build this project is to order the complete kit (model A-2) including a silk-screened punched chassis from Radio Adventures for $79.95. The PC boards may be ordered separately, the counter board (BK-172) is $37.95 and the display board (BK-171) is $24.95. The programmed C5 chip alone is also available for $14.95. The kit comes with a metal case, instructions, and all parts including two high quality PC boards. If you choose to build the counter from scratch, a PC board layout and a parts location drawing are available from ARRL.[1]

Some of the advanced features of the counter include:

- an anti-jitter code that reduces last digit jitter
- five push-button selectable, non-volatile programmable offsets (total of 16)
- programmable reverse counting for reverse tuning VFOs
- selectable direct frequency readout
- programmable 100-Hz digit blanking
- programmable automatic display blanking for power conservation
- automatic display enable when the frequency changes
- leading zero blanking of megahertz digits.

The frequency counter requires 9 to 13-volts dc. Current requirements vary with operating frequency and the status of the display. At 5 MHz the operating current is approximately 100 mA with the display active, and approximately 40 mA with the display blanked. The main board is about 2.2 × 3.9-inches. The display board is a separate 1.4 × 3.9-inch assembly.

CIRCUIT DESCRIPTION

The block diagram (**Fig 26.25**) shows the crystal time base, counter control logic, six digit counter logic, multiplexed display driver and memory sections. The crystal is 4.032 MHz, which puts the fundamental and most lower order harmonics outside the ham bands.

The unit is built on two PC boards, the C5 frequency counter chip on one and the display on the other. A complete schematic is shown in **Fig 26.26**. Capacitor

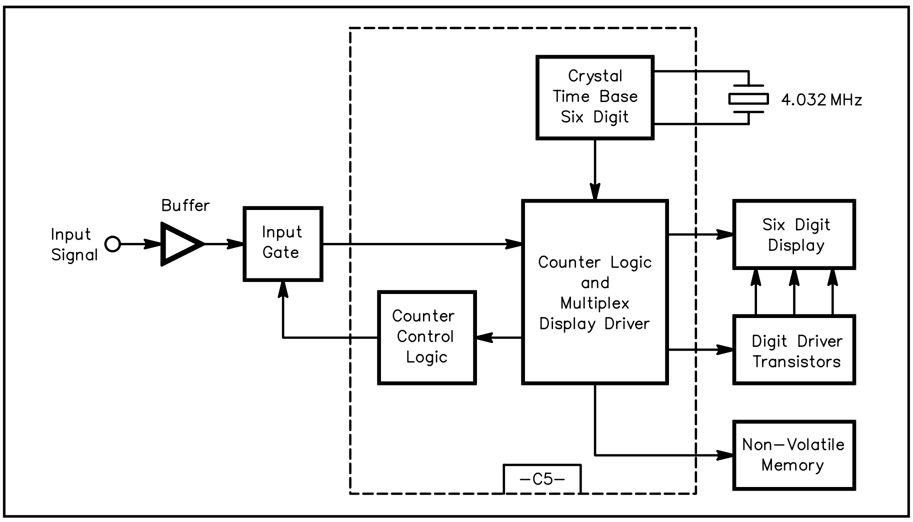


Fig 26.25—Block diagram showing the crystal time base, counter control logic, six digit counter control logic, multiplexed display driver and memory control for the Six Digit Programmable Frequency Counter.

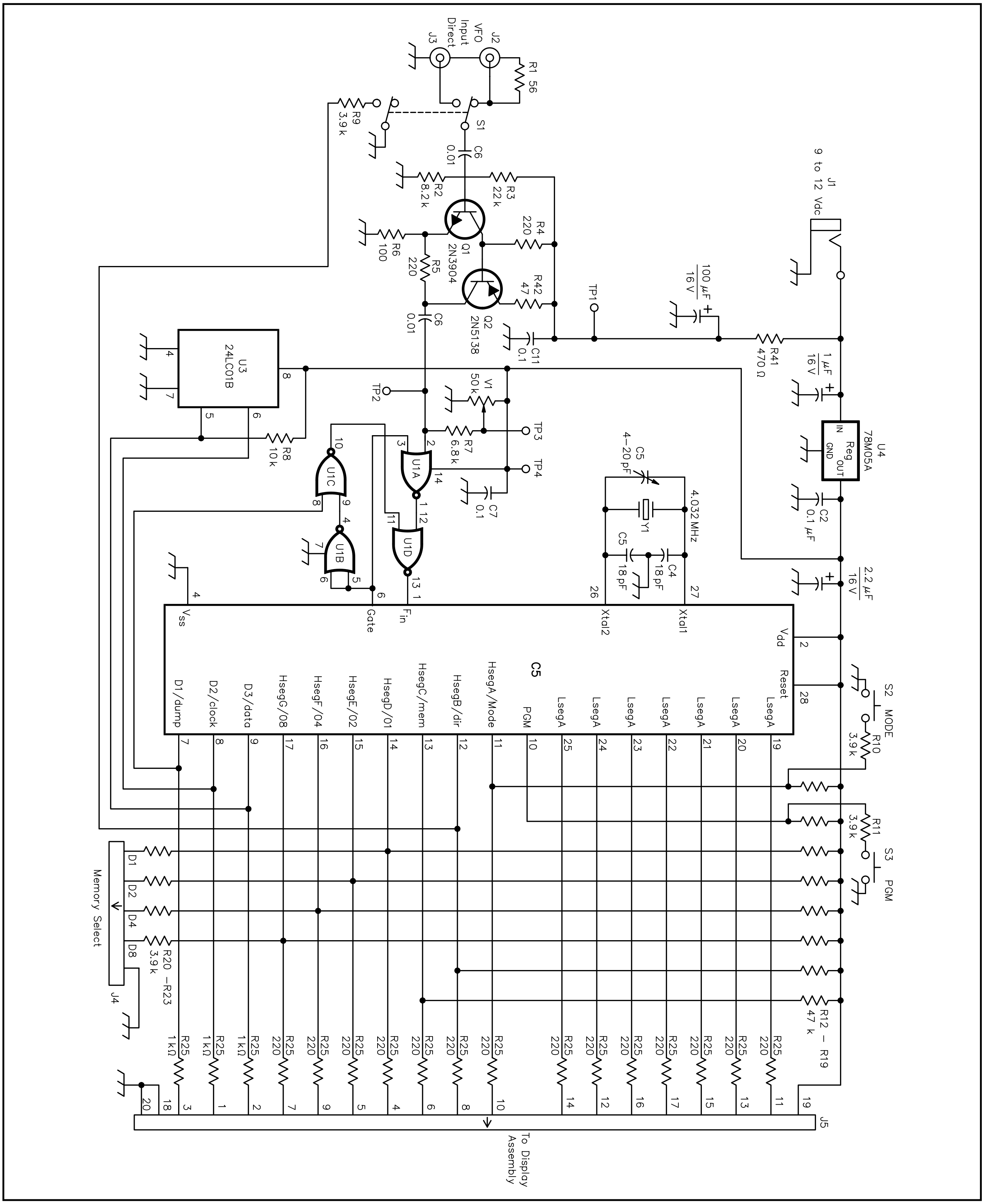


Fig 26.26A—Schematic of the C5 frequency counter chip and associated circuitry. All fixed resistors are $^{1}/_{4}$-W, 5%-tolerance carbon film. Capacitance values are in microfarads (μF).

U1—74HC02 quad 2 input NOR gate (Digi-Key MM74HC02N-ND)

U2—C5 frequency counter IC (8 bit microcontroller) see text

U3—24LC01B/P 128×8 EEPROM (Digi-Key 24LC01B/P-ND)

Y1—4.032 MHz 20 pF HC-49/U crystal (Mouser 520-HCA403-20)

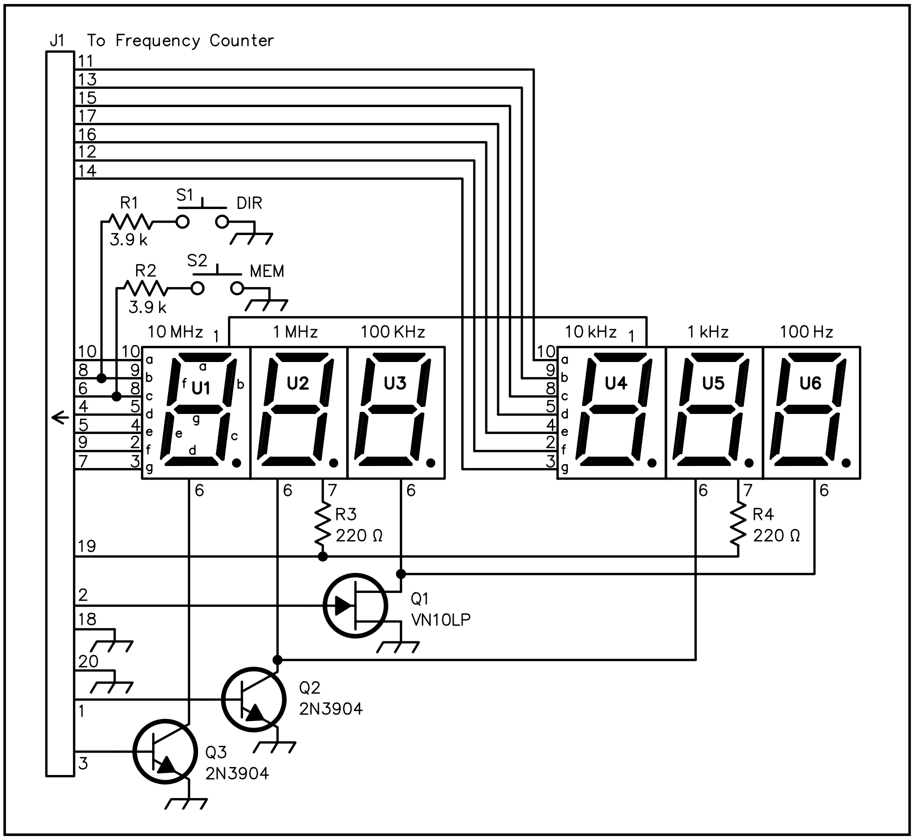


Fig 26.26B—Display circuitry schematic. J1 is connected to J5 through a 20-conductor ribbon cable. All fixed resistors are $^1/_4$-W, 5%-tolerance carbon film. Capacitance values are in microfarads (µF).

Q1—VN10LP VMOS power FET (Digi-Key VN10LP-ND)
Q2, Q3—2N3904 NP transistor
U1, U2, U6—9 mm green LED 7-segment display (Kingbright SC36-11GWA)
U3, U4, U5—9 mm yellow LED 7-segment display (Kingbright SC36-11YWA)

C5 is used to calibrate the counter. A two transistor buffer precedes the count gate and provides some amplification and buffering between the counter and the circuit being measured. Two switch selected input connections are provided—a 50-Ω termination for the digital dial functions and a high impedance termination for direct frequency measurement. The signal passes through U1A to the counter chip.

Using the pulse count stored in the C5 count registers, the chip makes necessary calculations and drives a common cathode display consisting of two groups of three digits each. The display and front panel switches are mounted on a separate PC board for ease of mounting and are connected to the main board by a ribbon cable. Current limiting resistors mounted on the main board are used to provide segment current limiting and provide protection to the C5 in case wiring to the display is accidentally shorted.

Programmed parameters such as offset, slope and auto-blanking are stored in U3, a non-volatile EEPROM memory chip. The programming and operating switches, which share pins with the display, are isolated by 3.9-kΩ resistors.

The C5 counter chip provides 16 channels of program storage. Each program stores frequency offset, normal/inverted counting, continuous/automatic blanking display and 100/1000 Hz resolution selection. The first five channels, 00 through 04, are selected by the front panel MEM switch. All sixteen channels may be selected through the binary jumper connected to J4 (MEMORY SELECT), on the main PC board. The terminals of J4 are pulled low to select a certain channel. For instance, to select memory channel five, terminals 1 and 4 would be connected to the ground pin.

Since the PGM and MODE switches are only used when the counter is being set up, they are located on the rear panel. The MEM and DIR frequency functions are used more often, therefore, they are accessible from the front panel.

The power supply for the counter should provide between 9 to 13 V of ripple-free dc. The counter main board contains a 5-V regulator (U4) that provides power to the counter, logic and memory chips. The 9 to 13-V power is passed through an RC filter to the two-transistor buffer. If voltages below 9 V or above 13 V are used, the voltage dropping resistor, R41, can be changed such that the voltage at test point TP1 is about 6.5 V.

CONSTRUCTION

Construction is very straightforward if the circuit board kit is used. If you choose to hand wire your unit on perfboard or lay out your own PC board, use short connections in the front end of the counter (around J1, J2, Q1, Q2 and U1). The wiring to the display, memory and switches is less critical. The use of sockets for the counter, gate and memory ICs is strongly recommended. A good quality magnifying glass can be very helpful in identifying the various small components and checking the solder joints after assembly. A small 20 to 25 W soldering iron with a small tip will be necessary for making reliable solder joints without bridging. Use only electronic grade rosin core solder.

TEST AND CALIBRATION

After assembly, but before inserting the ICs into their sockets, apply dc power and check for +5 V ± 10% at the following pins:

U1, pin 14

U2, pins 2, 9, 10, 28

U3, pins 5, 8

The voltage at U1, pin 2, should vary between 0 and 5 V as potentiometer V1 is rotated. Set this voltage to 2.5 V. U2 should have 1.65 V ± 10% on pins 11, 12, 13, 14, 15, 16, 17.

All other pins on U1, U2 and U3 should measure 0 V. The voltage at TP1 should be 5-8 V and should be several volts less than the voltage powering the unit. If the unit fails one or more of these tests, there is a wiring error of some sort. Correct the error before proceeding.

When all is well with the above tests, remove the dc power and insert the ICs into their appropriate sockets. Re-apply power. The display should come alive. Position the input selector switch to the direct frequency position with no signal applied. The display should indicate 000.X where X may be a 0 or 1. If the display is indicating random numbers, adjust the trimmer pot V1 slightly to obtain the correct display. Adjusting V1 beyond the point where the display stabilizes will reduce the sensitivity of the counter. Attach a signal source with a known frequency to the direct frequency input. The signal should be in the range of 75 mV to 1 V rms. The higher the frequency, up to 50 MHz, the better. Adjust the trimmer capacitor C5 until the frequency displayed by the counter matches the applied frequency. The unit is now calibrated and ready for use.

OPERATION

The finished boards may be mounted in a stand alone enclosure or may be built into another piece of equipment. Since the counter is a digital device, it is possible that some noise may be heard in sensitive receivers on a quiet band. Very good results have been obtained when the counter is mounted in its own enclosure. If noise is present in built-in installations, some shielding may be necessary. Additional filtering of the supply voltage line may also help. Minimize the length of the ribbon cable connecting the main board and the display board.

In the digital dial mode, the counter requires about +10 dBm of signal over most of its operating range. Input to the counter is taken from the tunable oscillator or from the output of the pre-mix system. The general rule is to sample the signal at a low impedance point such as the emitter or source for transistors or the cathode in tube type equipment. Use small diameter coaxial cable, such as RG-174, to connect the digital dial to the equipment. A small capacitor, usually in the range of 10-100 pF, connected in series with the center conductor of the cable can be used to establish the proper signal level. Use the smallest value that gives reliable counter operation over the frequency range of interest. In broadband tube type equipment the oscillator signal amplitude may vary too much for reliable counting. In such cases, back-to-back parallel diodes may be connected to limit the signal level. If a broadband oscilloscope is available, observe the signal level at TP2 and adjust the value of the coupling capacitor to obtain between 0.6 and 2.5 V peak-to-peak.

Once a stable reading is obtained over the operating frequency range, the counter can be programmed with desired features and offsets. Programming is carried out in the following order. First, select the desired memory channel to be programmed, second, program the desired display mode, third, program the MHz offset and finally, program the kHz offset. There is no need to program unused memory channels. A detailed description of each programming step follows:

- Select the desired memory channel by closing the MEM switch and holding it until the desired channel number is displayed. Alternatively, switch the MEMORY SELECT lines to the binary value desired. Remember that the lines are active low. For instance, if you wanted to select memory channel 7, you would pull lines 01, 02 and 04 low.
- Program the mode by closing and holding the PGM switch until the display quits blinking *PROG*. Release the PGM switch. Close and hold the MODE switch until the display indicates the desired mode. Release the MODE switch. See **Table 26.5** to select desired mode. Store the selected mode by momentarily closing the PGM switch. The counter will now display frequency, indicating that the program mode has been exited.
- Program the MHz offset by closing and holding the PGM switch until the display quits blinking *PROG*. While the display is still blinking, momentarily close the DIR switch on the front panel. This action places the counter into the MHz program mode. When the display quits blinking release the PGM switch and use the DIR and MEM front panel switches to retard or advance the MHz display until it displays the frequency being monitored. Momentarily close the PGM switch to store the MHz offset.
- Program the kHz offset by closing and holding the PGM switch until the display quits blinking *PROG*. Release the PGM switch and use the DIR and MEM front panel switch to retard or advance the kHz display until it displays the frequency being monitored. Momentarily close the PGM switch to store the kHz offset.
  This same procedure is used to program each memory channel.

Table 26.5
Counter Display Modes

Mode	*Invert*	*Blank*	*100 Hz*
00			X
01	X		
20		X	X
30	X	X	X
40			
50	X		
60		X	
70	X	X	

Note

[1] A template for the six digit programmable frequency counter is available in the **References** chapter.

OSCILLOSCOPES

Most engineers and technicians will tell you that the most useful single piece of test and design equipment is the triggered-sweep oscilloscope (commonly called just a "scope"). This section was written by Dom Mallozzi, N1DM.

Oscilloscopes can measure and display voltage relative to time, showing the waveforms seen in electronics textbooks. Scopes are broken down into two major classifications: analog and digital. This does not refer to the signals they measure, but rather to the methods used inside the scope to process signals for display.

ANALOG OSCILLOSCOPES

Fig 26.27 shows a simplified diagram of a triggered-sweep oscilloscope. At the heart of nearly all scopes is a cathode-ray tube (CRT) display. The CRT allows the visual display of an electronic signal by taking two electric signals and using them to move (deflect) a beam of electrons that strikes the screen. Unlike a television CRT, an oscilloscope uses electrostatic deflection rather than magnetic deflection. Wherever the beam strikes the phosphorescent screen of the CRT it causes a small spot to glow. The exact location of the spot is a result of the voltage applied to the vertical and horizontal inputs.

All of the other circuits in the scope are used to take the real-world signal and convert it to a form usable by the CRT. To trace how a signal travels through the oscilloscope circuitry start by assuming that the trigger select switch is in the INTERNAL position.

The input signal is connected to the input COUPLING switch. The switch allows selection of either the ac part of an ac/dc signal or the total signal. If you wanted to measure, for example, the RF swing at the collector of an output stage (referenced to the dc level), you would use the dc-coupling mode. In the ac position, dc is blocked from reaching the vertical amplifier chain so that you can measure a small ac signal superimposed on a much larger dc level. For example, you might want to measure a 25 mV 120-Hz ripple on a 13 Vdc power supply. Note that you should not use ac coupling at frequencies below 30 Hz, because the value of the blocking capacitor represents a considerable series impedance to very low-frequency signals.

After the coupling switch, the signal is connected to a calibrated attenuator. This is used to reduce the signal to a level that can be tolerated by the scope's vertical amplifier. The vertical amplifier boosts the signal to a level that can drive the CRT and also adds a bias component to locate the waveform on the screen.

A small sample of the signal from the vertical amplifier is sent to the trigger circuitry. The trigger circuit feeds a start pulse to the sweep generator when the input signal reaches a certain level. The sweep generator gives a precisely timed signal that looks like a triangle (see **Fig 26.28**). This triangular signal

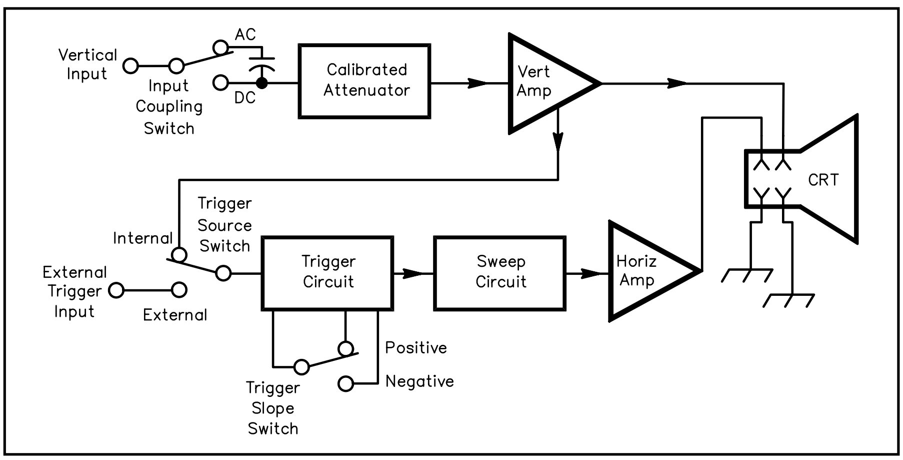


Fig 26.27 — Typical block diagram of a simple triggered-sweep oscilloscope.

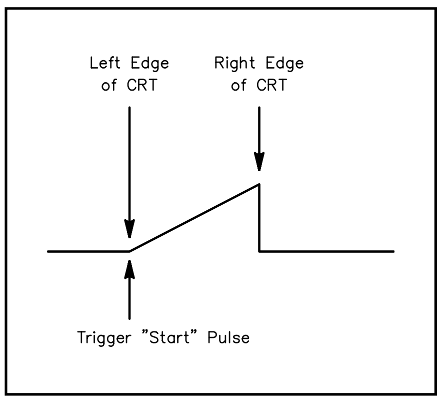


Fig 26.28 — The sweep trigger starts the ramp waveform that sweeps the CRT electron beam from side to side.

causes the scope trace to sweep from left to right, with the zero-voltage point representing the left side of the screen and the maximum voltage representing the right side of the screen.

The sweep circuit feeds the horizontal amplifier that, in turn, drives the CRT. It is also possible to trigger the sweep system from an external source (such as the system clock in a digital system). This is done by using an external input jack with the trigger select switch in the EXTERNAL position.

The trigger system controls the horizontal sweep. It looks at the trigger source (internal or external) to find out if it is positive- or negative-going and to see if the signal has passed a particular level. **Fig 26.29A** shows a typical signal and the dotted line on the figure represents the trigger level. It is important to note that once a trigger circuit is "fired" it cannot fire again until the sweep has moved all the way across the screen from left to right. In normal operation. the TRIGGER LEVEL control is manually adjusted until a stable display is seen. Some scopes have an AUTOMATIC position that chooses a level to lock the display in place without manual adjustment.

Fig 26.29B shows what happens when the level has not been properly selected. Because there are two points during a single cycle of the waveform that meet the triggering requirements, the trigger circuit will have a tendency to jump from one trigger point to another. This will make the waveform jitter from left to right. Adjustment of the TRIGGER control will fix this problem.

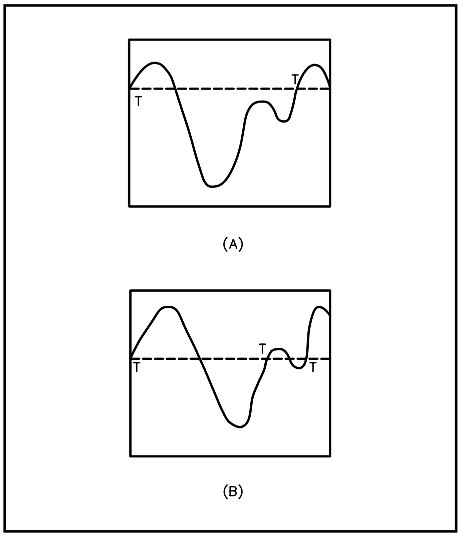


Fig 26.29 — In order to produce a stable display the selection of the trigger point is very important. Selecting the trigger point in A produces a stable display, but the trigger shown at B will produce a display that "jitters" from side to side.

The horizontal travel of the trace is calibrated in units of time. If the time of one cycle is known, we can calculate the frequency of the waveform. In **Fig 26.30**, for example, if the SWEEP speed selector is set at 10 µs/division and we count the number of divisions (vertical bars) between peaks of the waveform (or any similar well defined points that occur once per cycle) we can find the period of one cycle. In this case it is 80 µs. This means that the frequency of the waveform is 12,500 Hz (1/80 µs). The accuracy of the measured frequency depends on the accuracy of the scope's sweep oscillator (usually approximately 5%) and the linearity of the ramp generator. This accuracy cannot compete with even the least-expensive frequency counter, but the scope can still be used to determine whether a circuit is functioning properly.

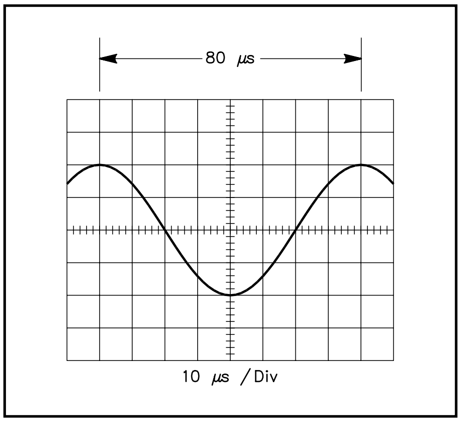


Fig 26.30 — Oscilloscopes with a calibrated sweep rate can be used to measure frequency. Here the waveform shown has a period of 80 microseconds (8 divisions × 10 µs per division) and therefore a period of 1/80 µs or 12.5 kHz.

Dual-Trace Oscilloscopes

Dual-trace oscilloscopes can display two waveforms at once. This type of scope has two vertical input channels that can be displayed either alone, together or one after the other. Fig 26.31 shows a simplified block diagram of a dual-trace oscilloscope. The only differences between this scope and the previous example are the additional vertical amplifier and the "channel switching circuit." This block determines whether we display channel A, channel B or both (simultaneously). The dual display is not a true dual display (there is only one electron gun in the CRT) but the dual traces are synthesized in the scope.

There are two methods of synthesizing a dual-trace display from a single-beam scope. These two methods are referred to as

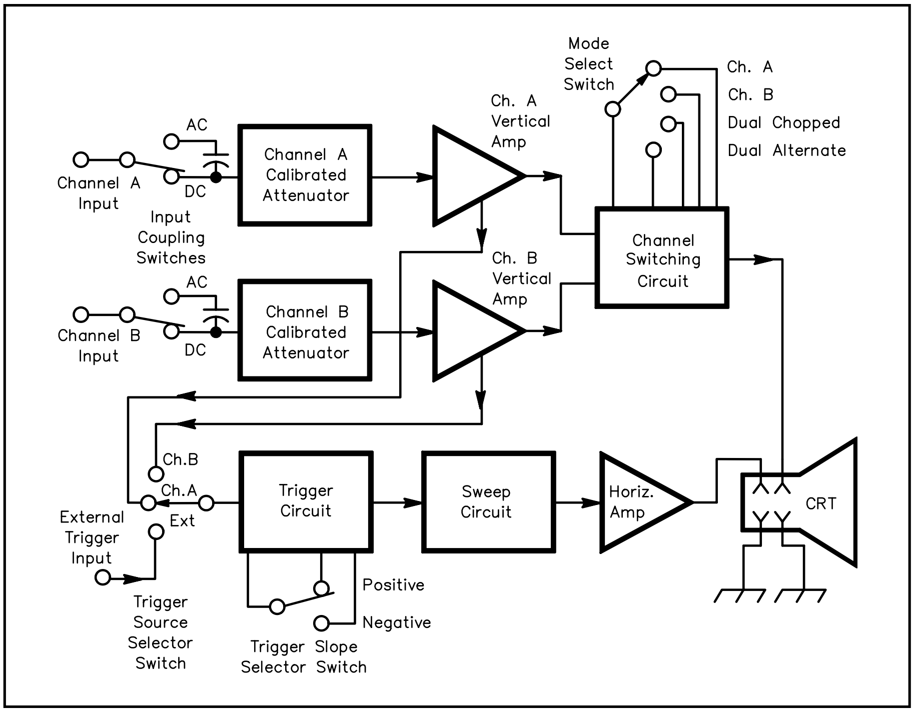


Fig 26.31 — Simplified Dual-trace oscilloscope block diagram. Note the two identical input channels and amplifiers.

chopped mode and *alternate mode*. In the chopped mode a small portion of the channel A waveform is written to the CRT, then a corresponding portion of the channel B waveform is written to the CRT. This procedure is continued until both waveforms are completely written on the CRT. The chopped mode is especially useful where an actual measure of the phase difference between the two waveforms is required. The chopped mode is usually most useful on slow sweep speeds (times greater than a few microseconds per division).

In the alternate mode, the complete channel A waveform is written to the CRT followed immediately by the complete channel B waveform. This happens so quickly that it appears that the waveforms are displayed at the same time. This mode of operation is not useful at very slow sweep speeds, but is good at most other sweep speeds.

Most dual-trace oscilloscopes also have a feature called "X-Y" operation. This feature allows one channel to drive the horizontal amplifier of the scope (called the X channel) while the other channel (called Y in this mode of operation) drives the vertical amplifier. Some oscilloscopes also have an external Y input. X-Y operation allows the scope to display Lissajous patterns for frequency and phase comparison and to use specialized test adapters such as curve tracers or spectrum analyzer front ends. Because of frequency limitations of most scope horizontal amplifiers the X channel is usually limited to a 5 or 10-MHz bandwidth.

DIGITAL OSCILLOSCOPES

The classic analog oscilloscope just discussed has existed for over 50 years. In the last 15 years, the digital oscilloscope has advanced from a specialized laboratory device to a very useful general-purpose tool, with a price attractive to an active experimenter. It uses digital circuitry and microprocessors to enhance the processing and display of signals. These result in dramatically improved accuracy for both amplitude and time measurements. When configured as a digital storage oscilloscope (DSO) it can read a stored waveform for as long as you wish without time limitations incurred by an analog type of storage scope.

Examine the simplified block diagram shown in Fig 26.32. After the signal goes through the vertical input attenuators and amplifiers, it arrives at the analog-to-digital converter (ADC). The ADC assigns a digital value to the level of the analog input signal and puts this in a memory similar to computer RAM. This value is stored with an assigned time, determined by the trigger circuits and the crystal timebase. The digital oscilloscope takes discrete amplitude samples at regular time intervals. If you were to take this data directly from memory, it would put a series of dots on the screen. You would then have to connect the dots to reconstruct the original waveform. The digital scope's microprocessor does this for

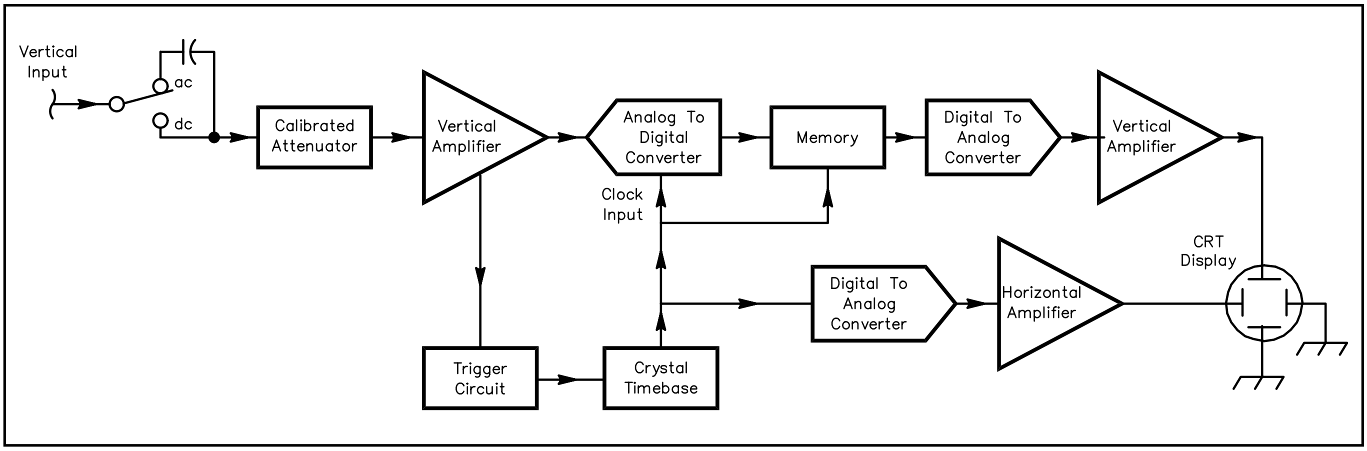


Fig 26.32 — Simplified block diagram of a digital oscilloscope. Note: The microprocessors are not shown for clarity after *ABC's of Oscilloscopes* copyright Fluke Corporation (reproduced with Permission).

you by mathematically processing the signal while reading it back from the memory and driving a digital-to-analog converter (DAC), which then drives the vertical deflection amplifier. A DAC also takes the digital stored time data and uses it to drive the horizontal deflection amplifier.

For the vertical signals you will see manufacturers refer to "8-bit digitizing," or perhaps "10-bit resolution." This is a measure of how many digital levels that are shown along the vertical (voltage) axis. More bits give you better resolution and accuracy of measurement. An 8-bit vertical resolution means each vertical screen has 2^8 (or 256) discrete values; similarly, 10 bits resolution yields 2^{10} (or 1024) discrete values.

It is important to understand some of the limitations resulting from sampling the signal rather than taking a continuous, analog measurement. When you try to reconstruct a signal from individual discrete samples, you must take samples at least twice as fast as the highest frequency signal being measured. If you digitize a 100-MHz sine wave, you should take samples at a rate of 200 million samples a second (referred to as 200 Megasamples/second). Actually, you really would like to take samples even more often, usually at a rate at least five times higher than the input signal.

If the sample rate is not high enough, very fast signal changes between sampling points will not appear on the display. For example, **Fig 26.33** shows one signal measured using both analog and digital scopes. The large spikes seen in the analog-scope display are not visible on the digital scope. The sampling frequency of the digital scope is not fast enough to store the higher frequency components of the waveform. If you take samples at a rate less than twice the input frequency, the reconstructed signal has a wrong apparent frequency; this is referred to as *aliasing*. In Fig 26.33 you can see that there is about one sample taken per cycle of the input waveform. This does not meet the 2:1 criteria established above. The result is that the scope reconstructs a waveform with a different apparent frequency.

Many older digital scopes had potential problems with aliasing.

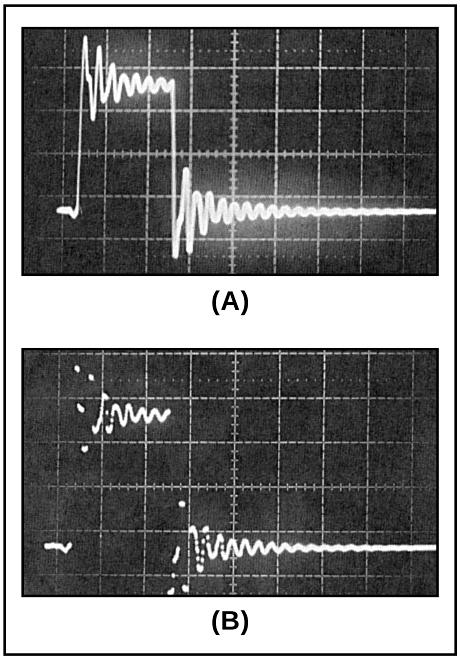


Fig 26.33 — Comparison of an analog scope waveform (A) and that produced by a digital oscillo-scope (B). Notice that the digital samples in B are not continuous, which may leave the actual shape of the waveform in doubt for the fastest rise time displays the scope is capable of producing.

Newer scopes use advanced techniques to check themselves. A simple manual check for aliasing is to use the highest practical sweep speed (shortest time per division) and then to change to other sweep speeds to verify that the apparent frequency doesn't change.

LIMITATIONS

Oscilloscopes have fundamental limits, primarily in frequency of operation and range of input voltages. For most purposes the voltage range of a scope can be expanded by the use of appropriate probes. The frequency response (also called the bandwidth) of a scope is usually the most important limiting factor. At the specified maximum response frequency, the response will be down 3 dB (0.707 voltage). For example, a 100-MHz 1-V sine wave fed into a 100-MHz bandwidth scope will read approximately 0.707 V on the scope display. The same scope at frequencies below 30 MHz (down to dc) should be accurate to about 5%.

A parameter called *rise time* is directly related to bandwidth. This term describes a scope's ability to accurately display voltages that rise very quickly. For example, a very sharp and square waveform may appear to take some time in order to reach a specified fraction of the input voltage level. The rise time is usually defined as the time required for the display to show a change from the 10% to 90% points of the input waveform, as shown in **Fig 26.34**. The mathematical definition of rise time is given by:

$$t_r = \frac{0.35}{BW} \tag{10}$$

where

t_r = rise time, µs

BW = bandwidth, MHz.

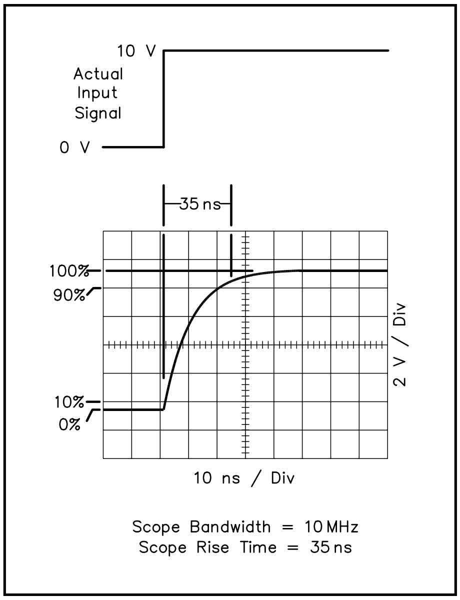


Fig 26.34 — The bandwidth of the oscilloscope vertical channel limits the rise time of the signals displayed on the scope.

It is also important to note that all but the most modern (and expensive) scopes are not designed for precise measurement of either time or frequency. At best, they will not have better than 5% accuracy in these applications. This does not change the usefulness of even a moderately priced oscilloscope, however. The most important value of an oscilloscope is that it presents an image of what is going on in a circuit and quickly shows which component or stage is at fault. It can show modulation levels, relative gain between stages and oscillator output.

OSCILLOSCOPE PROBES

Oscilloscopes are usually connected to a circuit under test with a short length of shielded cable and a probe. At low frequencies, a piece of small-diameter coax cable and some sort of insulated test probe might do. Unfortunately, at higher frequencies the capacitance of the cable would produce a capacitive reactance much less than the one-megohm input impedance of the oscilloscope. In addition each scope has a certain built-in capacitance at its input terminals (usually between 5 and 35 pF). These two capacitances cause problems when probing an RF circuit with a relatively high impedance.

The simplest method of connecting a signal to a scope is to use a specially designed probe. The most common scope probe is a *×10 probe* (called a times ten probe). This probe forms a 10:1 voltage divider using the built-in resistance of the probe and the input resistance of the scope. When using a ×10 probe, all voltage readings must be multiplied by 10. For example, if the scope is on the 1 V/division range and a ×10 probe was in use, the signals would be displayed on the scope face at 10 V/division.

Unfortunately a resistor alone in series with the scope input seriously degrades the scope's rise-time performance and therefore its bandwidth. Since the scope input looks like a parallel RC circuit, the series resistor feeding it causes a significant reduction in available charging current from the source. This may be corrected by using a compensating capacitor in parallel with the series resistor. Thus two dividers are formed: one resistive voltage divider and one capacitive voltage divider. With these two dividers connected in parallel and the RC relationships shown in **Fig 26.35**, the probe and scope should have a flat response curve through the whole bandwidth of the scope.

To account for manufacturing tolerances in the scope and probe the compensating capacitor is made variable. Most scopes provide a "calibrator" output that produces a known-frequency square wave for the purpose of adjusting the compensating capacitor in a probe. **Fig 26.36** shows possible responses when the probe is connected to the oscilloscope's calibrator jack.

If a probe cable is too short, do not attempt to extend the length of the cable by adding a piece of common coaxial cable. The cable usually used for probes is much different than common 50 or 75-Ω coax. In addition the compensating capacitor in the probe is chosen to compensate for the provided length of cable. It usually will not have enough range to compensate for extra lengths.

The shortest ground lead possible should be used from the probe to the circuit ground. Long ground leads are inductors at high frequencies. In these circuits they cause ringing and other undesirable effects.

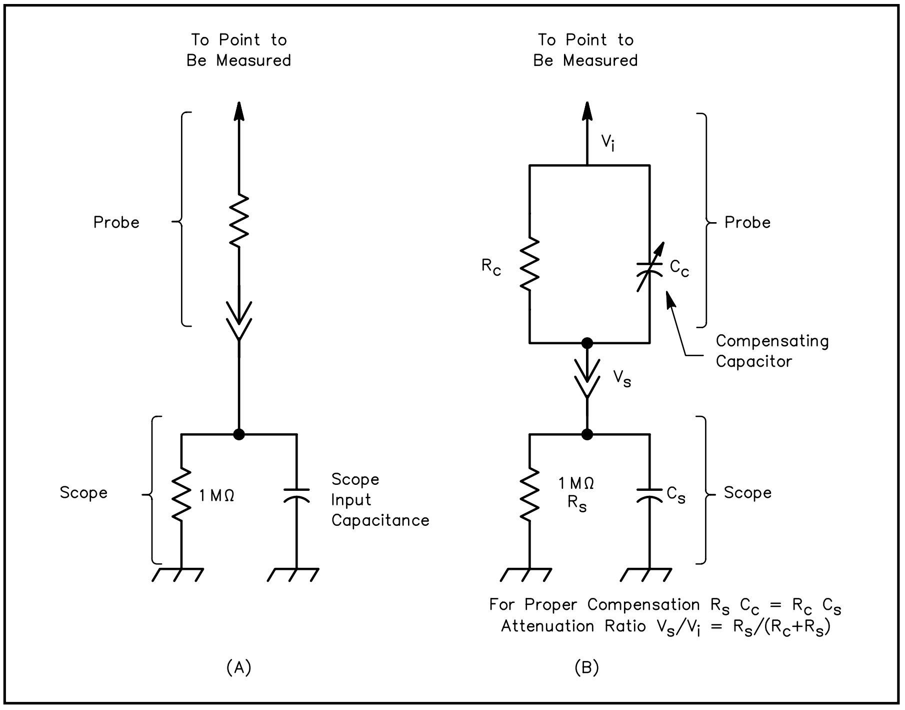


Fig 26.35 — Uncompensated probes such as the one at A are sufficient for low-frequency and slow-rise-time measurements. However, for accurate display of fast rise times with high-frequency components the compensated probe at B must be used. The variable capacitor is adjusted for proper compensation (see text for details).

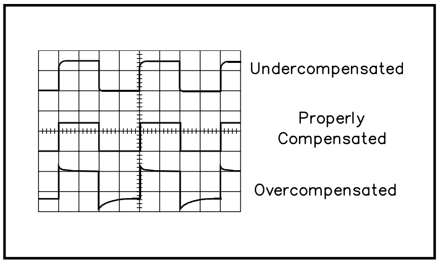


Fig 26.36 — Displays of a square-wave input illustrating undercompensated, properly compensated and overcompensated probes.

THE MODERN SCOPE

For many years a scope (even a so-called portable) was big and heavy. Computers and modern ICs have reduced the size and weight. Modern scopes can take other forms than the traditional large cabinet with built-in CRT. Some modern scopes use an LCD display for true portability. Some scopes take the form of a card plugged into a PC, where they use the PC and monitor for display. Even if they don't use a PC for a display, many scopes can attach to a PC and download their data for storage and analysis using advanced mathematical techniques. Many high-end scopes now incorporate nontraditional functions, such as Fast Fourier Transforms (FFT). This allows limited spectrum analysis or other advanced mathematical techniques to be applied to the displayed waveform.

BUYING A USED SCOPE

Many hams will end up buying a used scope due to price. If you buy a scope and intend to service it yourself, be aware all scopes that use tubes or a CRT contain lethal voltages. Treat an oscilloscope with the same care you would use with a tube-type high-power amplifier. The CRT should be handled carefully because if dropped it will crack and implode, resulting in pieces of glass and other materials being sprayed everywhere in the immediate vicinity. You should wear a full-face safety shield and other appropriate safety equipment to protect yourself.

Another concern when servicing an older scope is the availability of parts. Many scopes made since about 1985 have used special ICs, LCDs and microprocessors. Some of these may not be available or may be prohibitive in cost. You should buy a used scope from a reputable vendor— even better yet, try it out before you buy it. Make sure you get the operators manual also.

AN HF ADAPTER FOR NARROW-BANDWIDTH OSCILLOSCOPES

Fig 26.37 shows the circuit of a simple piece of test equipment that will allow you to display signals that are beyond the normal bandwidth of an inexpensive oscilloscope. This circuit was built to monitor modulation of a 10-m signal on a scope that has a 5-MHz upper-frequency limit. This design features a Mini-Circuits Laboratory SRA-1 mixer. (See the Address List in the **References** chapter for contact information.) Any stable oscillator or VFO with an output of 10 dBm can be used for the local oscillator (LO), which mixes with the HF signal to produce an IF in the bandwidth of the oscilloscope.

The mixer can handle RF signal levels up to –3 dBm without clipping, so this was set as an upper limit for the RF input. A toroidal transformer coupler is constructed by winding a 31-turn secondary of #28-AWG wire on a 3E2A core, which has a 0.038-inch diameter. An FT-37-75 is suitable. The primary is a piece of coaxial cable passed through the core center. The coupler gives 30 dB of attenuation and has a flat response from 0.5 to 100 MHz. An additional 20-dB of attenuation was added for a total of 50 dB before the mixer. One-watt resistors will do fine for the attenuator. The completed adapter should be built into a shielded box.

This circuit, with a 25-MHz LO frequency, is useful on frequencies in the 20 to 30-MHz range with transmitters of up to 50-W power output. By changing the frequency of the LO, any frequency in the range of the coupler can be displayed on a 5-MHz-bandwidth oscilloscope. The frequency displayed will be the difference between the LO and the input signal. As an example, a 28.1-MHz input and a 25-MHz LO will be seen as a 3.1-MHz signal on the oscilloscope.

More attenuation will be required for higher-power transmitters. This circuit was described by Kenneth Stringham Jr., AE1X, in the Hints and Kinks column of February 1982 *QST*.

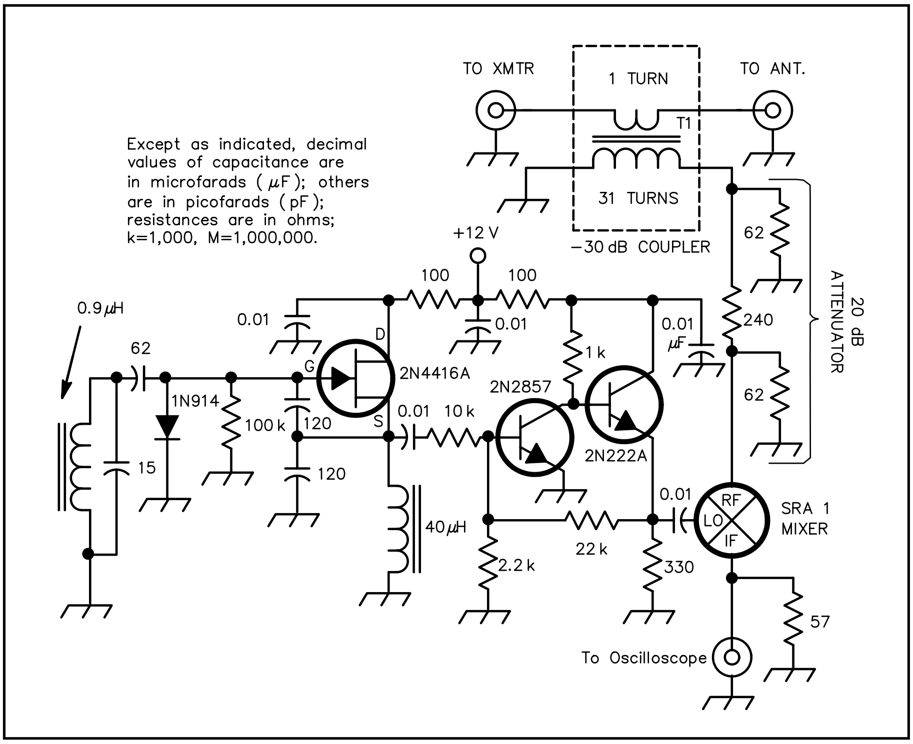


Fig 26.37 — This adapter displays HF signals on a narrow-bandwidth oscilloscope. It uses a 10-dBm 25-MHz LO, –30-dB coupler, 20-dB attenuator and diode-ring mixer. See text for further information.

A CALIBRATED NOISE SOURCE

NOISE FIGURE MEASUREMENT

One of the most important measurements in communications is the noise figure of a receiving setup. Relative measurements are often easy, while accurate ones are more difficult and expensive. One EME (moon bounce) station checks noise and system performance by measuring the noise of the sun reflected off the moon. While the measurement source (use of the sun and moon) is not expensive, the measuring equipment on 2 m consists of 48 antennas (each over 30 ft long). This measurement equipment is not for everyone!

The rest of us use more conventional noise sources and measuring techniques. Coverage of noise figure and its measurement appear in the **Transceivers** chapter of this *Handbook*.

Most calibrated and stable noise sources are expensive, but not this unit developed by Bill Sabin, W0IYH. It first appeared in May 1994 *QST*. When hams use a noise source, it is usually included in an RF bridge used to measure impedances and adjust antenna tuners. A somewhat different device (an *accurately calibrated and stable* noise source) is also useful. Combining a broadband RF noise source of known power output and a known output impedance with a true-RMS voltmeter, results in an excellent instrument for making interesting and revealing measurements on a variety of circuits hams commonly use. (Later on, some examples will be described.) The true-RMS voltmeter can be an RF voltmeter, a spectrum analyzer or an AF voltmeter at the output of a linear receiver.[1]

Calibrated noise generators and noise-figure meters are available at medium to astronomical prices. Here is a low-cost approach which can be used with reasonable confidence for many amateur applications where accuracy to tenths of a decibel is not needed, but where precision (repeatability) and comparative measurements are much more important. PC boards are available for this project.[2]

Semiconductor Noise Diodes

Any Zener diode can be used as a source of noise. If, however, the source is to be calibrated and used for reliable measurements, avalanche diodes specially designed for this purpose are preferable by far.[3] A good noise diode generates its noise through a carefully controlled *bulk avalanche* mechanism which exists *throughout* the PN junction, not merely at the junction surfaces where unstable and unreliable surface effects predominate due to local breakdown and impurity.[4] A true noise diode has a very low *flicker noise* (1/f) effect and tends to create a uniform level of truly *Gaussian noise* over a wide band of frequencies.[5] In order to maximize its bandwidth, the diode also has very low junction capacitance and lead inductance.

[1] W. Sabin, "Measuring SSB/CW Receiver Sensitivity," *QST*, Oct 1992, pp 30-34. See also Technical Correspondence, *QST*, Apr 1993, pp 73-75.

[2] PC boards are available from FAR Circuits, See **References** chapter Address List; price, $3.75 plus $1.50 shipping. A PC board template for the Sabin noise source is available at: **http://www.arrl.org/notes**.

[3] The term *Zener diode* is commonly used to denote a diode that takes advantage of avalanche effect, even though the Zener effect and the avalanche effect are not exactly the same thing at the device-physics level.

[4] The term *bulk avalanche* refers to the avalanche multiplication effect in a PN junction. A carrier (electron or hole) with sufficient energy collides with atoms and causes more carriers to be knocked loose. This effect "avalanches" and it occurs throughout the volume of the PN junction. This mechanism is responsible for the high-quality noise generation in a true noise diode.

[5] *Gaussian noise* refers to the instantaneous values of a noise voltage. These values conform to the Gaussian probability density function of statistics.

This project uses the NOISE/COM NC302L diode. It consists of a glass, axial-lead DO-35 package and is rated for use from 10 Hz to 3 GHz, if appropriate construction methods are followed. Prior to sale, the diodes are factory aged for 168 hours and are well stabilized. NOISE/COM has kindly agreed to make these diodes available to amateur experimenters for the special price of $10 each; the usual low-quantity price is about $25.[6]

Noise Source Design

The noise source presents two kinds of available output power. One is the thermal noise (–174 dBm/Hz at room temperature) when the diode is turned off. This is called N_{OFF}. The other is the sum of this same thermal noise and an "excess" noise, N_E, which is created by the diode when turned on, called N_{ON} (equivalent to $N_{OFF} + N_E$). For accurate measurements, the output impedance of the test apparatus must be the same (on or off) so that the device under test (DUT) always sees the same generator impedance. In Amateur Radio work, this impedance is usually 50 Ω, resistive. The circuit design must guarantee this condition.

For maximum frequency coverage, a PC-board layout and coax connector suitable for use at microwaves are needed. For lower frequency usage, a less stringent approach can be employed. Two noise sources are presented here. One is for the 0.5 to 500-MHz region and uses conventional components that many amateurs already have. The other is for the 1-MHz to 2.5-GHz range; it uses chip components and an SMA connector.

Circuit Diagram and Construction

Figs 26.38 and 26.39 show the simple schematics of the two noise sources. In series with the diode is a 46.4-Ω resistor that combines with the dynamic resistance of the diode in the avalanche, noise generator mode (about 4 Ω) to total about 50 Ω. When the applied voltage polarity is reversed, the diode is forward conducting and its dynamic resistance is still about 4 Ω, but the avalanche noise is now turned off. As a result, the noise source output impedance is always about 50 Ω. The 5-dB pad reduces the effect of any small impedance differences, so that the output impedance is nearly constant from the *on* to the *off* condition, and the SWR is less than 2:1.

Consider the noise situation of the noise diode when it is forward conducting. The resistance of the forward biased PN junction is a *dynamic* resistance. This dynamic resistance is *not* a source of ther-

6 NOISE/COM Co, for contact information see the References chapter Address List.

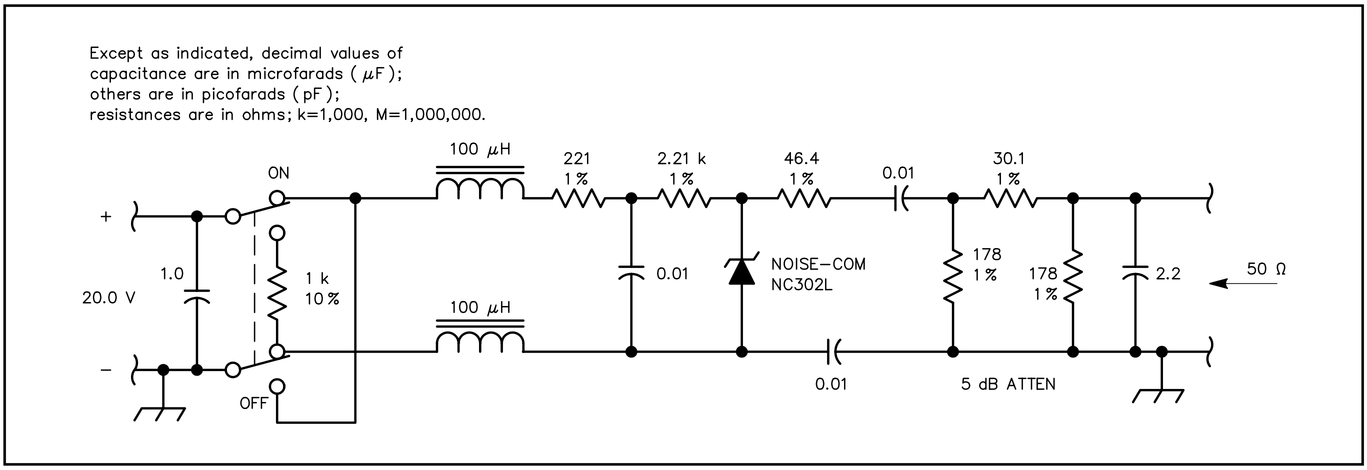


Fig 26.38 — Schematic of the 0.5 to 500-MHz calibrated noise source. Resistors are $^1/_8$-W, 1%-tolerance metal-film units. 1% resistors are available from Digi-Key. See the References chapter for the address.

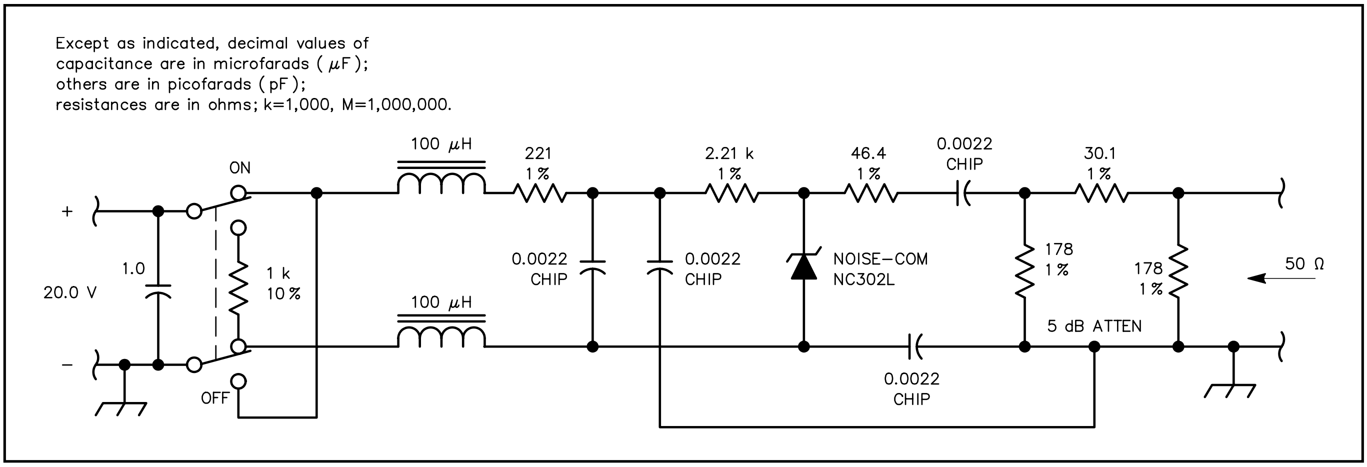


Fig 26.39 — The 1-MHz to 2.5-GHz calibrated noise source uses 1%-tolerance, 0.1-W chip resistors and chip capacitors. 1% resistors are available from Digi-Key. See the References chapter for the address.

mal noise, since it is not an actual physical resistance such as in a resistor or lossy network. However, the 0.6-V forward drop across the PN junction does produce a shot noise effect. The mathematics of this shot noise shows that the noise power associated with this effect is only about 50% of the thermal noise power that would be available from a physical resistor having the same value as the dynamic resistance. Therefore, the forward biased junction does *not* add excess noise to the system.[7] There is an 1/f noise effect associated with this shot noise in the diode, but its corner frequency is at about 100 kHz and of no importance at higher frequencies. Also, the small amount of bulk resistance contributes a little thermal noise.

In order to maximize the unit's flatness and frequency response bandwidth, noise-source construction methods should aim for RF circuit lead lengths as close to zero as possible as well as minimum inductance in the ground path and the coupling capacitors. The power-supply voltage must be clean, well bypassed and set accurately. **Fig 25.40** shows a 0.5 to 500-MHz unit. This construction method satisfies quite well the electrical requirements wanted for this model. At 500 MHz, the return loss with respect to 50 Ω at the output jack decreased to 10 dB. A calibration chart (**Fig 26.41**) is attached to the unit's top for easy reference. **Fig 26.42** shows the inside of the 1-MHz to 2.5-GHz noise source.

[7] Motchenbacher and Fitchen, *Low Noise Electronic Design* (New York: Wiley & Sons, 1973), p 22.

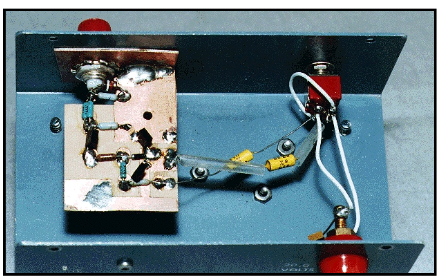


Fig 26.40 — An inside view of the 0.5 to 500-MHz noise source.

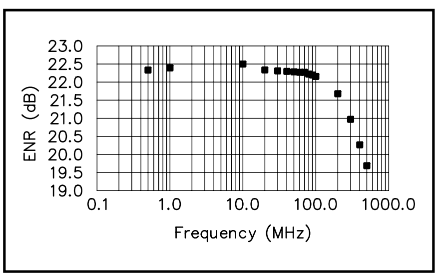


Fig 26.41 — Sample calibration chart for the 0.5 to 500-MHz noise source.

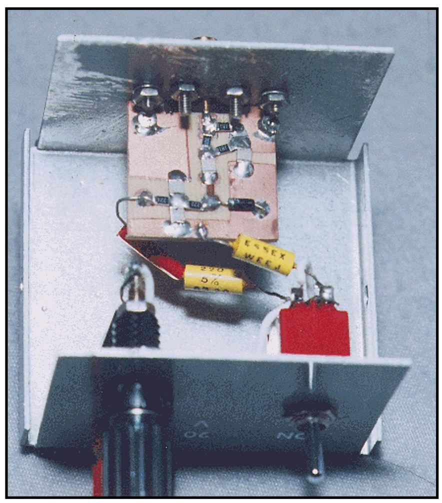


Fig 26.42 — A view inside of the 1-MHz to 2.5-GHz noise source.

Calibrating the Noise Source

If the construction is solid, the calibration should last for a long time. There are two ways to calibrate the noise source. If the unit *has been carefully constructed and its correct operation verified,* NOISE/COM will calibrate home-built units over the desired frequency range for $25 plus return shipping charges. Note that one factory calibrated unit can be used as a reference for many home calibrated units. **Fig 26.43** shows the NOISE/COM calibration data for both models of prototype noise sources, including SWR data. The noise data is strictly valid only at room temperature, so it's necessary to avoid extreme temperature environments.

Frequency (MHz)	***0.5 to 500 MHz Unit***		***1.0 to 2500 MHz Unit***	
	ENR (dB)	*SWR*	*NR (dB)*	*SWR*
0.5	22.33	1.03		
1	22.38	1.03	21.38	1.03
10	22.45	1.04	21.46	1.03
20	22.35	1.06		
30	22.32	1.06		
40	22.32	1.09		
50	22.30	1.11		
60	22.29	1.12		
70	22.25	1.15		
80	22.22	1.17		
90	22.20	1.20		
100	22.15	1.23	21.80	1.07
200	21.65	1.42		
300	20.96	1.62		
400	20.25	1.70		
500	19.60	1.90	20.71	1.44
1000			20.12	1.86
1500			20.00	2.06
2000			20.70	2.14
2500			21.51	1.88

Fig 26.43 — NOISE/COM calibration data for both prototype noise sources. The data is not universal; it varies from unit to unit.

The second calibration method requires a signal generator with known output levels at the various desired calibration frequencies. One approach is to build a tunable weak-signal oscillator that can be compared to some accessible high-quality signal generator, using a sensitive receiver as a detector.[8] The level of the signal source in dBm is needed.

Access to a multistage attenuator is also desirable. Build the attenuator using the nearest 1% values of metal-film resistors, so that systematic errors are minimized. A total attenuation of 25 dB in 0.1-dB steps is desirable. Attenuator construction must be appropriate for use at the intended frequency range. In some cases, a high-frequency correction chart may be needed.

With the calibrated signal source and the attenuator feeding the receiver in an SSB or CW mode the techniques discussed in the reference of Note 1 should be used to determine the excess noise (N_E) of the noise source and the noise bandwidth (B_N) of the receiver.

Excess Noise Ratio

A few words about excess noise ratio (ENR) are needed. It is defined as the ratio of excess noise to thermal noise. That is,

$$\mathrm{ENR} = \frac{N_{ON} - N_{OFF}}{N_{OFF}} = \frac{N_E}{N_{OFF}} \tag{12}$$

When the noise source is turned on, its output is $N_{OFF} + N_E$. The ratio of N_{ON} to N_{OFF} is then

$$\frac{N_{ON}}{N_{OFF}} = \frac{N_{OFF} + N_E}{N_{OFF}} = 1 + \frac{N_E}{N_{OFF}} = 1 + \mathrm{ENR} \tag{13}$$

[8] W. Hayward and D. DeMaw, *Solid State Design for the Radio Amateur* (Newington: ARRL, 1986).

Therefore, ENR is a measure of how much the noise increases and the noise generator can be calibrated in terms of its ENR.

Normalizing ENR to a 1-Hz bandwidth and converting to decibels, this is

$$\mathrm{ENR}\,(\mathrm{dB}) = 174\,(\mathrm{dBm/Hz}) + \frac{N_E\,(\mathrm{dBm})}{B_N\,(\mathrm{Hz})} \qquad (14)$$

Prepare a calibration chart and attach it to the top of the unit (see Fig 26.41). If the unit is to be factory calibrated, first perform the calibration procedure to ensure everything is working properly. Remember, a factory calibrated unit can be used as a reference for other home calibrated units, once the calibration-transfer procedures have been worked out. This requires some careful thinking and proper techniques. Generally speaking, a NOISE/COM calibration is the best choice.

Noise-Figure Measurement

The thermal noise power available from the attenuator remains constant for any value of attenuator setting. But the excess noise and therefore the ENR (in dB) due to the noise diode is equal to the calibration point of the source minus the setting (in dB) of the attenuator.

The noise-figure measurement of a device under test (DUT) uses the Y method and the setup in **Fig 26.44**. If the DUT has a noise-generator input and a true-RMS noise-measuring instrument at the output, then the total output noise (including the contribution of the measuring instrument) with the noise generator turned off is

$$N_{OFF(TOT)} = kTB_N\, F_{TOT}\, G_{DUT}\, G_{NMI} \qquad (15)$$

where

kTB_N = thermal noise,
G_{DUT} = gain of the DUT,
G_{NMI} = gain of the noise-measuring instrument, and
F_{TOT} = noise factor of the combination of the DUT and the noise-measuring instrument.

When the noise generator is turned on, the output noise is

$$N_{ON(TOT)} = kTB_N\, F_{TOT}\, G_{DUT}\, G_{NMI} + (ENR)kTB_N\, G_{DUT}\, G_{NMI} \qquad (16)$$

Where the last term is the contribution of excess noise by the noise generator. Note that none of these values is in dB or dBm.

If we divide equation 16 by equation 15 and say that the ratio

$$\frac{N_{ON(TOT)}}{N_{OFF(TOT)}} = Y \qquad (17)$$

then,

$$\frac{F_{TOT} + ENR}{F_{TOT}} = 1 + \frac{ENR}{F_{TOT}}$$

Note that kTBN, G_{DUT} and G_{NMI} disappear, so that these quantities need not be known to measure noise factor. If we solve equation 17 for F_{TOT}, we get the noise factor

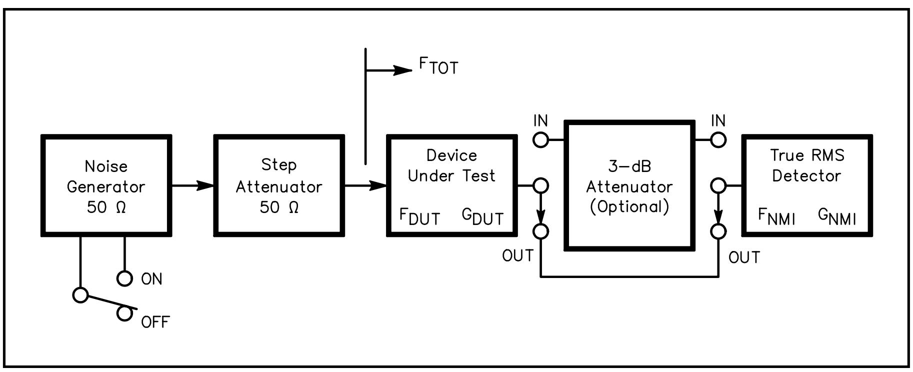


Fig 26.44 — Setup for measuring noise figure of a device under test (DUT).

$$F_{TOT} = \frac{ENR}{Y-1} \tag{18}$$

If the noise output doubles (increases by 3 dB) when we turn on the noise source, then Y = 2 and the noise factor is numerically equal to the excess noise ratio (ENR). If the attenuator steps are not fine enough or if the attenuator is not reliable over the entire frequency range, use equation 18 to get a better answer. (It's much simpler to use a good fine-step attenuator.) The value of F_{TOT} is that of the DUT in cascade with the noise-measuring instrument. To find F_{DUT}, we must know the noise factor F_{NMI} of the noise-measuring instrument and G_{DUT} and then use the Friis formula, unless G_{DUT} is very large (as it would be if the DUT were a high-gain receiver (see footnote 1).

$$F_{DUT} = F_{TOT} - \frac{F_{NMI}-1}{G_{DUT}} \tag{19}$$

The validity of equation 19 (if we need to use it) requires that the noise bandwidth of the noise-measuring instrument be less than the noise bandwidth of the DUT (see the reference of Note 1). Verify this before proceeding.

There's another advantage to using the power-doubling method. If the 3-dB attenuator of Fig 26.44 is used to maintain a constant noise level into the following stages and the RMS meter, this means that the noise factor, using the calibration scale and the input attenuator (without using equation 18), is

$$F_{DUT} = ENR + \frac{1}{G_{DUT}} \tag{20}$$

If G_{DUT} is large, then the last term can be neglected. If G_{DUT} is small, we need to know its value. However, we do not need to know the noise factor F_{NMI} of the circuitry after the DUT, as we did in the previous discussion.

The 3-dB attenuator method also removes all restrictions regarding the type of noise measuring instrument, since the meter reading is now used only as a reference point. This last statement applies only when two noise (or two signal generator) inputs are being compared.

Frequency Response Measurements

The noise generator, in conjunction with a spectrum analyzer, is an excellent tool for measuring the frequency response of a DUT, if the noise source is much stronger than the internal noise of the DUT and that of the spectrum analyzer. Many spectrum analyzers are not equipped with tracking generators, which can be quite expensive for an amateur's budget.

The spectrum analyzer needs to be calibrated for a noise input, if accurate amplitude measurements are needed, because it responds differently to noise signals than to sine-wave signals. The envelope detection of noise, combined with the logarithmic amplification of the spectrum analyzer, creates an error of about 2.5 dB for a noise signal (the noise is that much greater than the instrument indicates). Also, the noise bandwidth of the IF filter is different from its resolution bandwidth. Some modern spectrum analyzers have internal DSP algorithms that make the corrections so that external noise sources and also carrier-to-noise ratios, normalized to some noise bandwidth like 1.0 Hz, can be measured with fair accuracy if the input noise is a few decibels above thermal. One example is the Tektronix Model 2712. If only relative response readings are needed, then these corrections are not needed.

Also, the noise source itself can be used to establish an accurate reference level (in dBm) on the screen. An accurate, absolute measurement with the DUT in place will then be this reference level (in dBm), plus the increment in decibels produced by the DUT.

The noise-generator output can be viewed as a collection of sine waves separated by, say, 1 Hz. Each separated frequency "bin" has its own Gaussian amplitude and random phase with respect to all the others. So the DUT is simultaneously looking at a collection or "ensemble," of input signals. As the spectrum analyzer frequency sweeps, it looks simultaneously at all of the DUT frequencies that fall within the spectrum analyzer's IF noise bandwidth. The spectrum display is thus the "convolution" of the IF filter frequency response and the DUT frequency response. If the DUT is a narrow filter, a very narrow resolution and a slow sweep are needed in the spectrum analyzer. In addition, the analyzer's video, or post-detection, filter has a narrow bandwidth and also requires some settling time to get an accurate reading. So, some experience and judgment are required to use a spectrum analyzer this way.

Using Your Station Receiver

Your station receiver can also be used as a spectrum analyzer. Place a variable attenuator between the DUT and the receiver. As you tune your receiver, in a narrow CW mode, adjust the attenuator for a constant reference level receiver output. The attenuator values are inversely related to the frequency response.

A calibrated noise source with an adjustable attenuator that can be easily switched into a receiver antenna jack is an excellent tool for measuring antenna noise level or incoming weak signal level (in dBm) or for establishing correct receiver operation.

The noise source can also be combined with a locally generated data-mode waveform of a known dBm value to get an approximate check on modem performance or to make adjustments that might assure correct operation of the system. The rigorous evaluation of system performance requires special equipment and techniques that may be unavailable at most amateur stations. Or, you could evaluate the intelligibility improvement of your SSB transmitter's speech processor in a noise background.

Summary

The calibrated, flat-spectrum noise generator described in this article is quite a useful instrument for amateur experimenters. Its simplicity and low cost make it especially attractive. Getting a good calibration is the main challenge, but once it is achieved, the calibration lasts a long time, if the right diode is used. The ENR of the units described here is in the range of 20 dB. Use of a high-quality, external, 10-dB attenuator barrel will get into the range of 10-dB ENR. If the unit is sent to NOISE/COM the attenuator should also be sent, with the request that it be included in the calibration. That attenuator then "belongs" to the noise source and should be so tagged. If the attenuator is of high quality, the output SWR will also be improved. NOISE/COM suggests periodic recalibration, at your discretion.

A NOISE BRIDGE FOR 1.8 THROUGH 30 MHz

The noise bridge, sometimes referred to as an antenna noise bridge or RX noise bridge, is an instrument that measures the impedances of antennas or other electrical circuits. The unit shown in **Fig 26.45** provides adequate accuracy for most measurements in the 1.8- through 30-MHz range. Battery operation and small physical size make this unit ideal for remote-location use. This classic bridge circuit was updated by Mark Shelhamer, WA3YNO. Additional information about using the noise bridge for transmission-line measurements appears in that article and in the Transmission Line and Antenna Measurements chapter of *The ARRL Antenna Book*. A detector, such as the station receiver, is required for operation. An etching pattern and parts placement diagram are in Chapter 30, **References**.

The noise bridge consists of two parts: the noise generator and the bridge circuitry. See **Fig 26.46**. A 6.8-V Zener diode serves as the noise source. The broadband noise signal is amplified by U1 and associated components to produce an approximate S9 signal in the receiver.

The bridge portion of the circuit consists of T1, C1 and R1. T1 is a ferrite core wound as shown in the schematic detail. This design eliminates phase shift and the ferrite core has sufficient permeability to eliminate low-frequency resistance shift. One winding of T1 couples noise energy into the bridge circuit. The remaining two windings are each in one arm of the bridge. C1 and R1 complete the known arm; the UNKNOWN circuit with C3 comprises the remainder of the bridge. The terminal labeled RCVR is for connection to the detector.

The reactance range of a noise bridge depends on several factors, including operating frequency, value of the series capacitor (C3 in the figure) and the range of the variable capacitor (C1 in the figure). The zero-reactance point occurs when C1 is either nearly fully meshed or fully unmeshed.

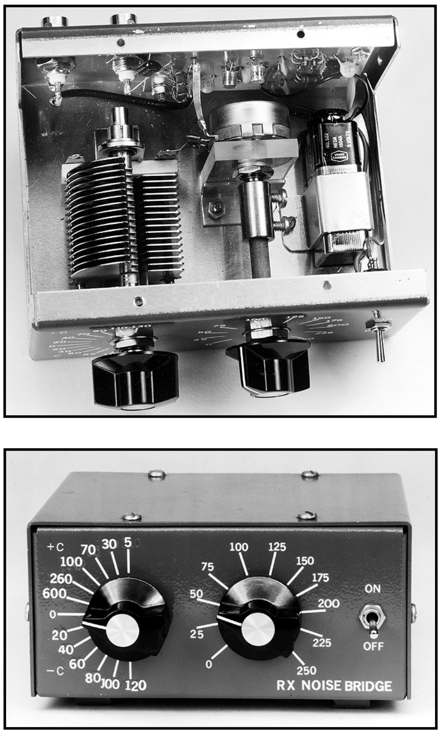


Fig 26.45 — Noise bridge construction details. Press-on lettering is used for the calibration marks. Note that the potentiometer must be isolated from ground.

Construction

The noise bridge can be put in a home-made aluminum enclosure that measures $5 \times 2^3/_8 \times 3^3/_4$ inches. Many of the circuit components are mounted on a circuit board that is fastened to the rear wall of the

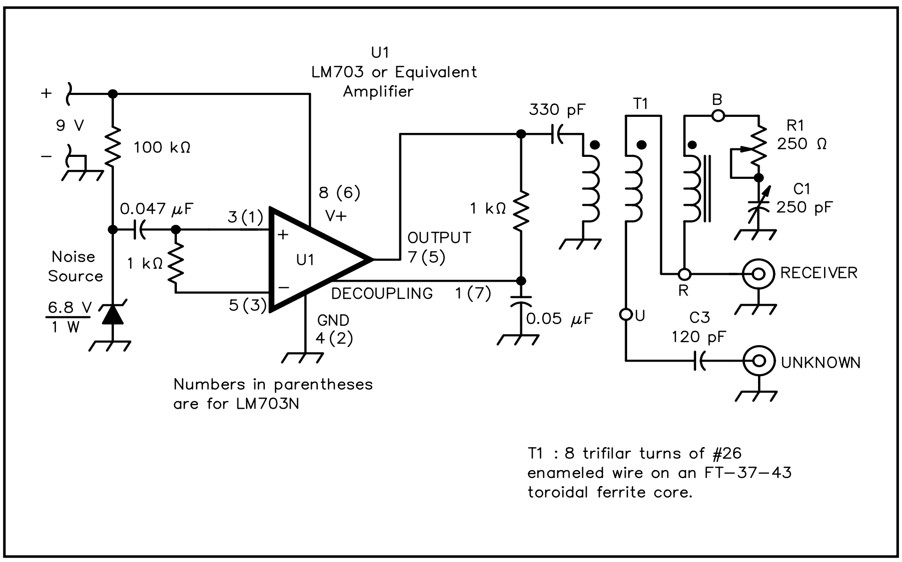


Fig 26.46 — Noise bridge schematic. The pin numbers on U1 refer to a metal can LM703 or the NTE/ECG replacement. Those in parentheses are for a National Semiconductor mini-DIP LM703N. See the text for placement of compensation C or L at points B, U or R.

cabinet. The circuit-board layout keeps the lead lengths to the board from the bridge and coaxial connectors to a minimum.

Potentiometer R1 must be mounted carefully. For accurate readings it must be very well insulated from ground. In the prototype the control is mounted on a piece of Plexiglas, which was fastened to the chassis with a piece of aluminum angle stock. Additionally, a $^1/_4$-inch control-shaft coupling and a length of phenolic rod were used to keep the control away from the front panel. Use a high-quality potentiometer to ensure good measurement results.

The variable capacitor is easier to mount because the rotor is grounded. It should be a high-quality unit. Two female RF fittings on the rear panel are connected to a detector (receiver) and to the UNKNOWN circuit. Plastic insulated phono connectors should *not* be used because they might influence bridge accuracy at higher frequencies. Miniature coaxial cable (RG-174) is used for the connection between the RCVR connector and circuit board. Attach one end of C3 to the circuit board and the other directly to the UNKNOWN circuit connector.

Bridge Compensation

Stray capacitance and inductance in the bridge circuit can affect impedance readings. If a very accurate bridge is required, use the next steps to make readings more accurate.

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

Construction details of the loads are shown in **Fig 26.47**. Each load is constructed inside a connector. The leads should be kept as short as possible. The resistors must be noninductive (not wire wound). Carbon-composition ($^1/_4$-W) resistors should work fine. The potentiometer in the variable-resistance load is a miniature PC-mount unit with a maximum resistance of 100 Ω, or less. The potentiometer wiper and one of the end leads are connected to the center pin of the connector; the other lead is connected to ground.

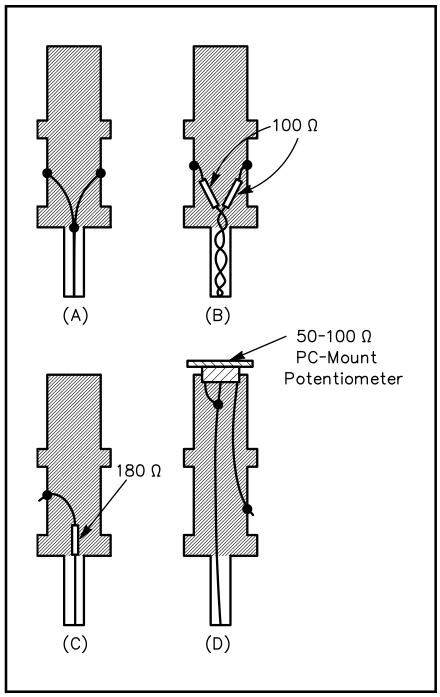


Fig 26.47 — Loads used to check and calibrate a noise bridge are built into a PL-259 shell. Leads should be kept as short as possible to minimize parasitic inductance. The connector shell is screwed in place after construction and is not shown in the figure. (A) is a short circuit; (B) depicts a 50-Ω load; (C) is a 180-Ω load; (D) shows a variable-resistance load used to determine the loss in a coaxial cable.

Stray Capacitance

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

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

$$C_P = C_S\left(\sqrt{\frac{R1}{R_L}} - 1\right) \quad (21)$$

where
R_L = load resistance (as measured), ohms
R1 = resistance of the variable resistor, ohms
C_S = series capacitance (either C3 or C4, whichever is selected), pF.

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

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

Compensate the bridge by setting the dc resistance of R1 equal to R_L. With the bridge at 1.8 MHz. alternately adjust C_c and C1 to obtain a null.

Stray Inductance

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

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

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

Calibration

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

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

$$X_{u(f)} = X_{u(10)} \frac{10}{f} \qquad (22)$$

where
f = frequency, MHz
$X_{u(10)}$ = reactance of the unknown load at 10 MHz.
$X_{u(f)}$ = reactance of the unknown load at f.

A shorted piece of coaxial cable serves as a reactance source. (The reactance of a shorted, low-loss coaxial cable is dependent only on the cable length, the measurement frequency and the cable characteristic impedance.) RadioShack RG-8M is used here because it is easy to get, has relatively low loss and has an almost purely resistive characteristic impedance. Prepare the calibration cable as follows:

1. Cut a length of coaxial cable that is slightly longer than λ/4 at 10 MHz (about 20 ft for RG-8M). Attach a suitable connector to one end of the cable; leave the other end open circuited.
2. Connect the 0-Ω load to the noise bridge UNKNOWN connector and set the receiver frequency to 10 MHz. Adjust the noise bridge for a null. Do not adjust the reactance control after the null is found.
3. Connect the calibration cable to the bridge UNKNOWN terminal. Null the bridge by adjusting only the variable resistor and the receiver frequency. The receiver frequency should be less than 10 MHz; if it is above 10 MHz, the cable is too short and you need to prepare a longer one.
4. Gradually cut short lengths from the end of the coaxial cable until you obtain a null at 10 MHz by adjusting only the resistance control. Then connect the cable center and shield conductors at the open end with a short length of braid. Verify that the bridge nulls with zero reactance at 20 MHz.
5. The reactance of the coaxial cable (normalized to 10 MHz) can be calculated from:

$$X_{i(10)} = R_0 \frac{f}{10} \tan\left(2\pi \frac{f}{40}\right) \quad (23)$$

where

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

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

f = frequency in MHz.

The results have less than 5% error for reactances less than 500 Ω, as long as the test-cable loss is less than 0.2 dB. This error becomes significantly less at lower reactances (2% error at 300 Ω for a 0.2-dB loss cable). The loss in 18 ft of RG-8M is 0.13 dB at 10 MHz. Reactance data for RadioShack RG-8M is given in **Table 26.6**.

With the prepared cable and calibration values on hand, go on to calibrate the reactance scale. Tune the receiver to the

Table 26.6

Noise Bridge Calibration with Coaxial Cable

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

Capacitance				***Reactance***			
C(pF)	*f(MHz)*	*C(pF)*	*f(MHz)*	*Xi*	*f(MHz)*	*Xi*	*f(MHz)*
10	9.798	–10	10.219	10	3.318	–10	19.376
20	9.612	–20	10.459	20	4.484	–20	18.722
30	9.440	–30	10.721	30	5.262	–30	18.048
40	9.280	–40	11.010	40	5.838	–40	17.368
50	9.130	–50	11.328	50	6.286	–50	16.701
60	8.990	–60	11.679	60	6.647	–60	16.063
70	8.859	–70	12.064	70	6.943	–70	15.472
80	8.735	–80	12.484	80	7.191	–80	14.938
90	8.618	–90	12.935	90	7.404	–90	14.459
100	8.508	–100	13.407	100	7.586	–100	14.045
110	8.403	–110	13.887	110	7.747	–110	13.683
120	8.304	–120	14.357	120	7.884	–120	13.370
130	8.209	–130	14.801	130	8.009	–130	13.097
140	8.119	–140	15.211	140	8.119	–140	12.861
				150	8.217	–150	12.654
				160	8.306	–160	12.473
				170	8.387	–170	12.313
				180	8.460	–180	12.172
				190	8.527	–190	12.045
				200	8.588	–200	11.932
				210	8.645	–210	11.831
				220	8.697	–220	11.739
				230	8.746	–230	11.655
				240	8.791	–240	11.579
				250	8.832	–250	11.510
				260	8.872	–260	11.446
				270	8.908	–270	11.387
				280	8.942	–280	11.333
				290	8.975	–290	11.283
				300	9.005	–300	11.236
				350	9.133	–350	11.045
				400	9.232	–400	10.905
				450	9.311	–450	10.798
				500	9.375	–500	10.713

appropriate frequency for the desired reactance (given in the table or found using the equation). Adjust the resistance and reactance controls to null the bridge. Mark the reactance reading on the front panel. Repeat this process until all desired reactance values have been marked. The resistance values needed to null the bridge during this calibration procedure may be quite high (more than 100 Ω) at the higher reactances.

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

Finding the Null

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

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

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

A SIGNAL GENERATOR FOR RECEIVER TESTING

The oscillator shown in **Fig 26.48** and **Fig 26.49** was designed for testing high-performance receivers. Parts cost for the oscillator has been kept to a minimum by careful design. While the stability is slightly less than that of a well-designed crystal oscillator, the stability of the unit should be good enough to measure most amateur receivers. In addition, the ability to shift frequency is important when

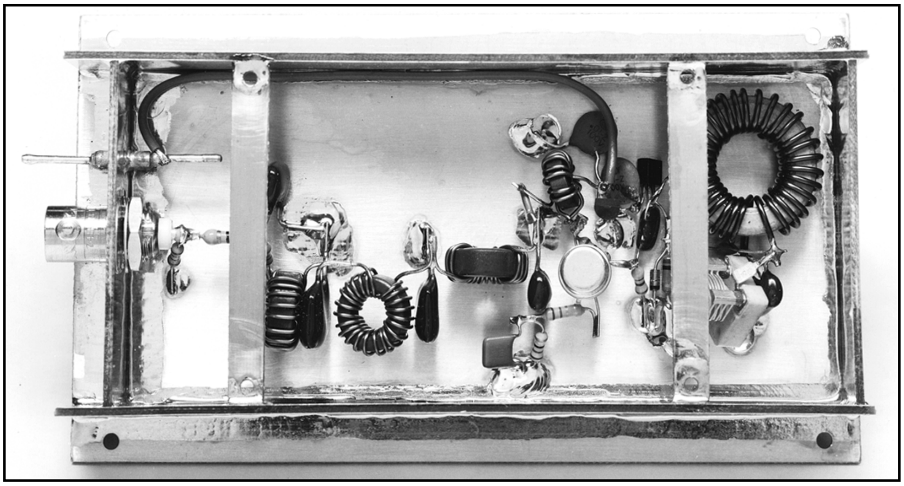


Fig 26.48 — A low-cost LC oscillator for receiver measurements. Toroidal cores are used for all of the inductances.

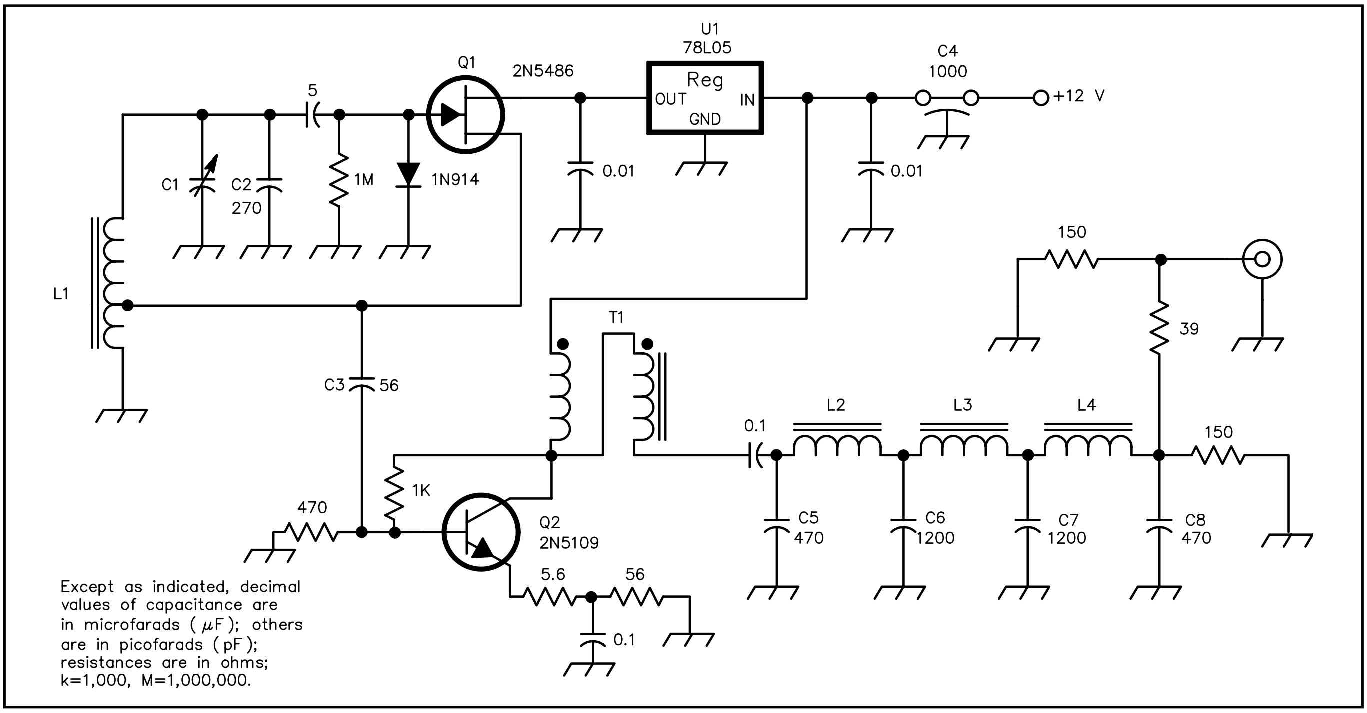


Fig 26.49 — Schematic diagram of the LC oscillator operating at 3.7 MHz. All resistors are $^1/_4$ W, 5% units.

C1 — 1.4 to 9.2-pF air trimmer (value and type not critical)

C2 — 270-pF silver-mica or NP0 capacitor. Value may be changed slightly to compensate for variations in L1.

C3 — 56-pF silver mica or NP0 capacitor. Value may be changed to adjust output power.

C4 — 1000 pF solder-in feedthrough capacitor. Available from Microwave Components of Michigan (see References chapter).

C5-C8 — Silver-mica, NP0 disc or polystyrene capacitor.

D1 — 1N914, 1N4148.

L1 — 31t #18 enameled wire on T-94-6 core. Tap 8 turns from ground end (7.5 µH).

L2, L4 — 21t #22 enameled wire on a T-50-2 core (2.5 µH, 2.43 µH ideal).

L3 — 23t #22 enameled wire on a T-50-2 core. (2.9 µH, 3.01 µH ideal).

T1 — 7t #22 enameled wire bifilar wound on an FT-37-43 core.

Q1 — 2N5486 JFET. MPF102 may give reduced output.

Q2 — 2N5109.

U1 — 78LO5 low-current 5-V regulator.

dealing with receivers that have spurious responses. More importantly, LC oscillators with high-Q components often have much better phase noise performance than crystal oscillators, because of power limitations in the crystal oscillators (crystals are easily damaged by excessive power).

The circuit is a Hartley oscillator followed by a class-A buffer amplifier. A 5-V regulator is used to keep the power supply output stable. The amplifier is cleaned up by a seven-element Chebyshev low-pass filter, which is terminated by a 6-dB attenuator. The attenuator keeps the filter working properly, even with a receiver that has an input impedance other than 50 Ω. A receiver designed to work with a 50-Ω system may not have a 50-Ω input impedance. The +4 dBm output is strong enough for most receiver measurements. It may even be too strong for some receivers. Note that sensitive components like crystal filters may require a step attenuator to lower the output level.

Construction

This unit is built in a box made of double-sided circuit board. Its inside dimensions are 1 × 2.2 × 5 inches (HWD). The copper foil of the circuit board makes an excellent shield, while the fiberglass helps temperature stability. Capacitor C2 should be soldered directly across L1 to ensure high Q. Since this is an RF circuit, leads should be kept short. While silver mica capacitors have slightly better Q, NP0 capacitors may offer better stability. Mounting the three inductors orthogonal (axis of the inductors 90° from each other) reduced the second-order harmonic by 2 dB when it was compared to the first unit that was made.

Alignment and Testing

The output of the regulator should be +5 V. The output of the oscillator should be +4 dBm (2.5 mW) into a 50-Ω load. Increasing the value of C3 will increase the power output to a maximum of about 10 mW. The frequency should be around 3.7 MHz. Additional capacitance across L1 (in parallel with C2) will lower the frequency if desired, while the trimmer capacitor (C1) specified will allow adjustment to a specific frequency. The drift of one of the first units made was 5 Hz over 25 minutes after a few minutes of warm up. If the warm-up drift is large, changing C2 may improve the situation somewhat. For most receivers, a drift of 100 Hz while you are doing measurements is not bad.

HYBRID COMBINERS FOR SIGNAL GENERATORS

Many receiver performance measurements require two signal generators to be attached to a receiver simultaneously. This, in turn, requires a combiner that isolates the two signal generators (to keep one generator from being frequency or phase modulated by the other). Commercially made hybrid combiners are available from Mini-Circuits Labs (see the Address List in the **References** chapter).

Alternatively, a hybrid combiner is not difficult to construct. The combiners described here (see **Fig 26.50**) provide 40 to 50 dB of isolation between ports (connections) while attenuating the desired signal paths (each input to output) by 6 dB. The 50-Ω impedance of the system is kept constant (very important if accurate measurements are to be made).

The combiners are constructed in small boxes made from double-sided circuit-board material. Each piece is soldered to the next one along the entire length of the seam. This makes a good RF-tight enclosure. BNC coaxial fittings are used on the units shown. However, any type of coaxial connector can be used. Leads must be kept as short as possible and precision resistors (or matched units from the junk box) should be used. The circuit diagram for the combiners is shown in **Fig 26.51**.

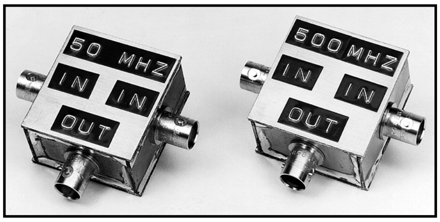


Fig 26.50 — The hybrid combiner on the left is designed to cover the 1 to 50-MHz range; the one on the right 50 to 500 MHz.

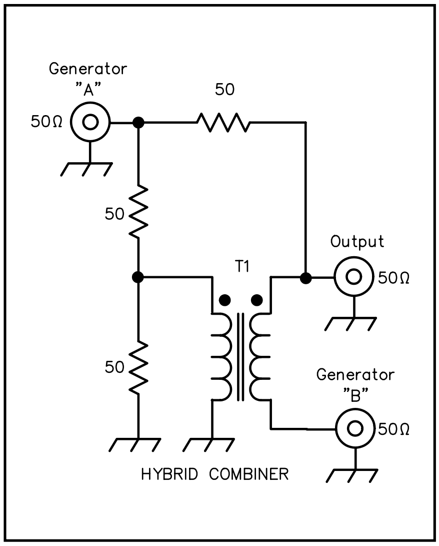


Fig 26.51 — A single bifilar wound transformer is used to make a hybrid combiner. For the 1 to 50-MHz model, T1 is 10 turns of #30 enameled wire bifilar wound on an FT-23-77 ferrite core. For the 50 to 500-MHz model, T1 consists of 10 turns of #30 enameled wire bifilar wound on an FT-23-63 ferrite core. Keep all leads as short as possible when constructing these units.

Return Loss Bridges

Return loss is a measure of how closely an impedance matches a reference impedance in phase angle and magnitude. If the reference impedance equals the measured impedance level with a 0° phase difference it has a return loss of infinity. **Fig A** shows basic return-loss measurement setups. Return-loss bridges are good for measuring filter response because return loss measurements are a more sensitive measure of passband response than insertion-loss measurements.

A 100 Hz to 100-kHz Return-Loss Bridge

Ed Wetherhold, W3NQN, has developed a low-frequency return-loss bridge (RLB) that can be adapted to different impedance levels. (See **Fig B**.) This bridge is used primarily for testing passive-LC filters that have been designed to work at a certain impedance level. Return-loss measurements require that the signal generator and RLB match the specific filter impedance level.

The characteristic impedance of this RLB is set by the values of four resistors (R1 = R2 = R3 = R4 = characteristic impedance). Ed mounted the four resistors on a plug-in module and placed their interconnections on the socket. Additional modules may be built for any impedance.

Table A provides computed values of return loss for several known loads and a 500-Ω RLB. If you build this RLB, use the values in the table to check its operation. For this frequency range, use an ac voltmeter in place of the power meter shown in Fig A. Choose one with good ac response well above 100 kHz.

Ed originally described this bridge circuit in an article published in 1993. That article provides complete construction details. Reprints are available from Ed Wetherhold, W3NQN (for $3). (For contact information, see the **References** chapter Address List.) The RLB shown is useful from 100 Hz to 100 kHz.

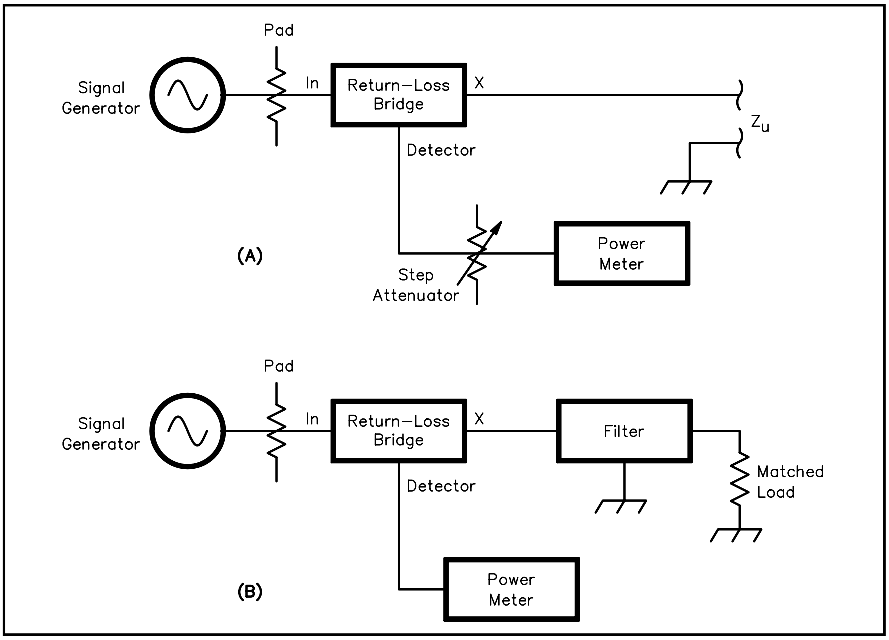


Fig A — A shows a setup to measure an unknown impedance with a return loss bridge. B is for measuring filter response.

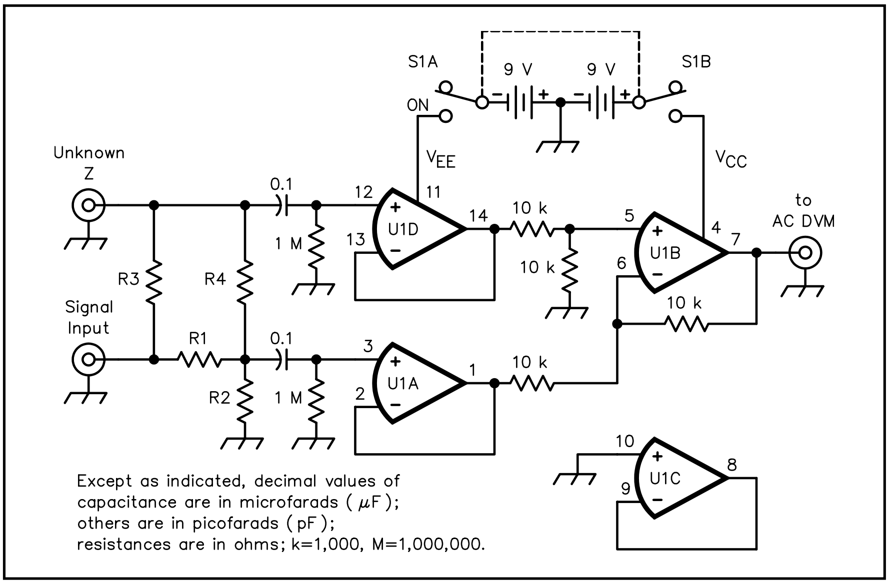


Fig B — An active low-frequency RLB. U1 is an LM324 quad op amp. R1, R2, R3 and R4 are $^{1}/_{4}$-W, 1% resistors with the same value as the output impedance of the signal generator and the input impedance of the test circuit.

Table A

Performance and Test Data for the 100-Hz to 100-kHz RLB

Bridge and signal-generator impedance = 500 Ω

Bridge Directivity

Frequency	*Return Loss (dB, Unknown = 500-Ω)*
100 Hz to 10 kHz	> 45
10 kHz to 80 kHz	40
80 kHz to 100 kHz	30

Return Loss of Known Resistive Loads

$R_{LOAD} = LF \times Z_{BRIDGE}$

where

R_{LOAD} = Load resistance, ohms

LF = Load factor

Z_{BRIDGE} = characteristic bridge impedance, ohms.

LF	*Return Loss (dB)*
5.848	3
3.009	6
1.925	10
1.222	20
1.065	30

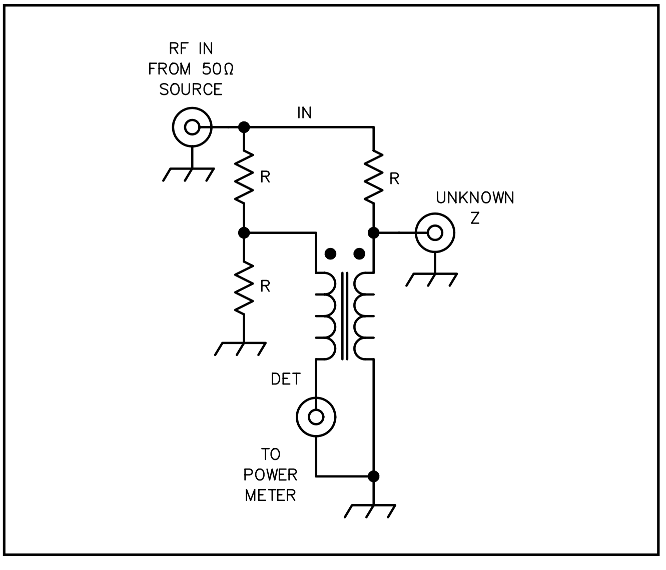


Fig C — An RLB for RF. Keep the lead lengths short. Wind the transformer on a high-permeability ferrite core. Use either 51-Ω carbon or 49.9-Ω 1% metal-film resistors.

An RF Return-Loss Bridge

At HF and higher frequencies, return-loss bridges are used as shown in Fig A for making measurements in RF circuits. The schematic of a simple bridge is shown in **Fig C**. (Notice that the circuit is identical to that of a hybrid combiner.) It is built in a small box with short leads to the coax connectors. Either 49.9-Ω 1% metal-film or 51-Ω $^{1}/_{4}$-W carbon resistors may be used. The transformer is wound with 10 bifilar turns of #30 enameled wire on a high permeability ferrite core such as an FT-23-43 or similar.

Apply the output of the signal generator to the RF INPUT port of the RLB. It may be necessary to attenuate the generator output to avoid overloading the amplifier under test. Connect the bridge DETECTOR port to a power meter through a step attenuator and leave the UNKNOWN port of the bridge open circuited. Set the step attenuator for a relatively high level of attenuation and note the power meter indication.

Now connect the unknown impedance, Z_u, to the bridge. The power meter reading will decrease. Adjust the step attenuator to produce the same reading obtained when the UNKNOWN port was open circuited. The difference between the two settings of the attenuator is the return loss, measured in dB.

The unknown impedance measured by this technique is not limited to amplifier inputs. Coax cable attached to an antenna, a filter, or any other fixed impedance device can be characterized by return loss. Return loss is measured in dB, and it is related to a quantity known as the voltage reflection coefficient, ρ:

$$RL = -20 \log |\rho|$$

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

where

RL = return loss, dB

ρ = voltage reflection coefficient.

The relationship of return loss to SWR is:

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

Receiver Performance Tests

Comparing the performance of one receiver to another is difficult at best. The features of one receiver may outweigh a second, even though its performance under some conditions is not as good as it could be. Although the final decision on which receiver to purchase will more than likely be based on personal preference and cost, there are ways to compare receiver performance characteristics. Some of the more important parameters are sensitivity, blocking dynamic range and two-tone IMD dynamic range.

Instruments for measuring receiver performance should be of suitable quality and calibration. Always remember that accuracy can never be better than the tools used to make the measurements. Common instruments used for receiver testing include:

- Signal generators
- Hybrid combiner
- Audio ac voltmeter
- Distortion meter (FM measurements only)
- Noise figure meter (only required for noise figure measurements)
- Step attenuators (10 dB and 1 dB steps are useful)

Signal generators must be calibrated accurately in dBm or µV. The generators should have extremely low leakage. That is, when the output of the generator is switched off, no signal should be detected at the operating frequency with a sensitive receiver. Ideally, at least one of the signal generators should be capable of amplitude modulation. A suitable lab-quality piece would be the HP-8640B.

While most signal generators are calibrated in terms of microvolts, the real concern is not with the voltage from the generator but with the power available. The unit that is used for most low-level RF work is the milliwatt, and power is often specified in decibels with respect to 1 mW (dBm). Hence, 0 dBm would be 1 mW. The dBm level, in a 50-Ω load, can be calculated with the aid of the following equation:

$$\text{dBm} = 10\log_{10}\left[20\left(V_{RMS}\right)^2\right] \qquad (24)$$

where

dBm = power with respect to 1 mW

V = RMS voltage available at the output of the signal generator.

The convenience of a logarithmic power unit such as the dBm becomes apparent when signals are amplified or attenuated. For example, a –107 dBm signal that is applied to an amplifier with a gain of 20 dB will result in an output increased by 20 dB. Therefore in this example (–107 dBm + 20 dB) = 87 dBm. Similarly, a –107 dBm signal applied to an attenuator with a loss of 10 dB will result in an output of (–107 dBm – 10 dB) or –117 dBm.

A hybrid combiner is a three-port device used to combine the signals from a pair of generators for all dynamic range measurements. It has the characteristic that signals applied at ports 1 or 2 appear at port 3 and are attenuated by 3 dB. However, a signal from port 1 is attenuated 30 or 40 dB when sampled at port 2. Similarly, signals applied at port 2 are isolated from port 1 some 30 to 40 dB. The isolating properties of the box prevent one signal generator from being frequency or phase modulated by the other. A second feature of a hybrid combiner is that a 50-Ω impedance level is maintained throughout the system.

Audio voltmeters should be calibrated in dB as well as volts. This facilitates easy measurements and eliminates the need for cumbersome calculations. Be sure that the step attenuators are in good working order and suitable for the frequencies involved. A distortion meter, such as the Hewlett- Packard 339A, is required for FM sensitivity measurements and a noise figure meter, such as the Hewlett-Packard 8970A, is excellent for certain kinds of sensitivity measurements.

Receiver Sensitivity

Several methods are used to determine receiver sensitivity. The mode under consideration often determines the best choice. One of the most common sensitivity measurements is minimum discernible signal (MDS) or noise floor. It is suitable for CW and SSB receivers.

This measurement indicates the minimum discernible signal that can be detected with the receiver. This level is defined as that which will produce the same audio-output power as the internally generated receiver noise. Hence, the term "noise floor."

To measure MDS, use a signal generator tuned to the same frequency as the receiver (see **Fig 26.56**). With the generator output at 0 or with maximum attenuation of its output note the voltmeter reading. Next increase the generator output level until the ac voltmeter at the receiver audio-output jack shows a 3-dB increase. The signal input at this point is the MDS. Be certain that the receiver is peaked on the generator signal. The filter bandwidth can affect the MDS. Always compare MDS readings taken with identical filter bandwidths. (A narrow bandwidth tends to improve MDS performance.) MDS can be expressed in μV or dBm.

In the hypothetical example of Fig 26.56, the output of the signal generator is –133 dBm and the step attenuator is set to 4 dB. Here is the calculation:

$$\text{Noise floor} = -133\text{ dBm} - 4\text{ dB} = -137\text{ dBm} \tag{25}$$

where the noise floor is the power available at the receiver antenna terminal and –4 dB is the loss through the attenuator.

Receiver sensitivity is also often expressed as 10 dB S+N/N (a 10-dB ratio of signal + noise to noise) or 10 dB S/N (signal to noise). The procedure and measurement are identical to MDS, except that the input signal is increased until the receiver output increases by 10 dB for 10 dB S+N/N and 9.5 dB for 10 dB S/N (often called "10 dB signal to noise ratio"). AM receiver sensitivity is usually expressed in this manner with a 30% modulated, 1-kHz test signal. (The modulation in this case is keyed on and off and the signal level is adjusted for the desired increase in the audio output.)

SINAD is a common sensitivity measurement normally associated with FM receivers. It is an acronym for "*si*gnal plus *n*oise *a*nd *d*istortion." SINAD is a measure of signal quality:

$$\text{SINAD} = \frac{\text{signal} + \text{noise} + \text{distortion}}{\text{noise} + \text{distortion}} \tag{26}$$

where SINAD is expressed in dB. In this example, all quantities to the right of the equal sign are expressed in volts, and the ratio is converted to dB by multiplying the log of the fraction by 20.

$$\text{SINAD}(\text{dB}) = 20\log\left(\frac{\text{Signal}(\text{V}) + \text{Noise}(\text{V}) + \text{Distortion}(\text{V})}{\text{Noise}(\text{V}) + \text{Distortion}(\text{V})}\right)$$

Let's look at this more closely. We can consider distortion to be a part of the receiver noise because distortion, like noise, is an unwanted signal added to the desired signal by the receiving system. Then, if we assume that the desired signal is much stronger than the noise, SINAD closely approximates the signal to noise ratio. The common 12-dB SINAD specification therefore corresponds

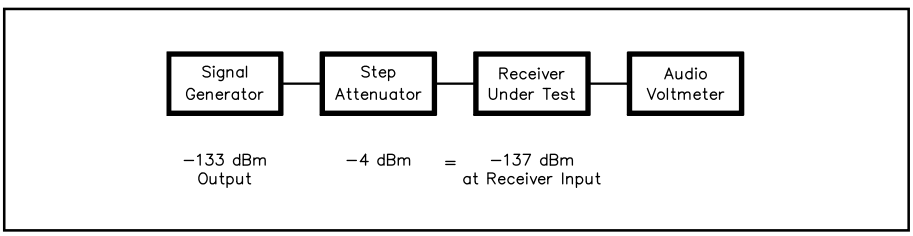


Fig 26.56 — A general test setup for measuring receiver MDS, or noise floor. Signal levels shown are for an example discussed in the text.

to a 4:1 S/N ratio (noise + distortion = 0.25 × signal).

The basic test setup for measuring SINAD is shown in **Fig 26.57**. The level of input signal is adjusted to provide 25% distortion (12 dB SINAD). Narrow-band FM signals, typical for amateur communications, usually have 3-kHz peak deviation when modulated at 1000 Hz.

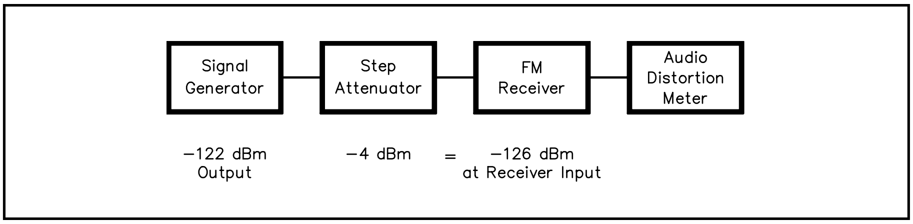


Fig 26.57 — FM SINAD test setup.

Noise figure is another measure of receiver sensitivity. It provides a sensitivity evaluation that is independent of the system bandwidth. Noise figure is discussed further in the **Transceivers** chapter.

Dynamic Range

Dynamic range is the ability of the receiver to tolerate strong signals outside of its band-pass range. Two kinds will be considered:

Blocking dynamic range (blocking DR) is the difference, in dB, between the noise floor and a signal that causes 1 dB of gain compression in the receiver. It indicates the signal level, above the noise floor, that begins to cause desensitization.

IMD dynamic range (IMD DR) measures the impact of two-tone IMD on a receiver. IMD is the production of spurious responses that results when two or more signals mix. IMD occurs in any receiver when signals of sufficient magnitude are present. IMD DR is the difference, in dB, between the noise floor and the strength of two equal incoming signals that produce a third-order product 3 dB above the noise floor.

What do these measurements mean? When the IMD DR is exceeded, false signals begin to appear along with the desired signal. When the blocking DR is exceeded, the receiver begins losing its ability to amplify weak signals. Typically, the IMD DR is 20 dB or more below the blocking DR, so false signals appear well before sensitivity is significantly decreased. IMD DR is one of the most significant parameters that can be specified for a receiver. It is generally a conservative evaluation for other effects, such as blocking, which will occur only for signals well outside the IMD dynamic range of the receiver.

Both dynamic range tests require two signal generators and a hybrid combiner. When testing blocking DR (see **Fig 26.58**), one generator is set for a weak signal of roughly –110 dBm. The receiver is tuned to this frequency and peaked for maximum response. (ARRL Lab procedures require this level to be about 10 dB below the 1-dB compression point, if the AGC can be disabled. Otherwise, the level is set to 20 dB above the MDS.)

The second generator is set to a frequency 20 kHz away from the first and its level is increased until the receiver output drops by 1 dB, as measured with the ac voltmeter.

In the example shown, the output of the generator is –7 dBm, the loss through the combiner is fixed at 3 dB and the

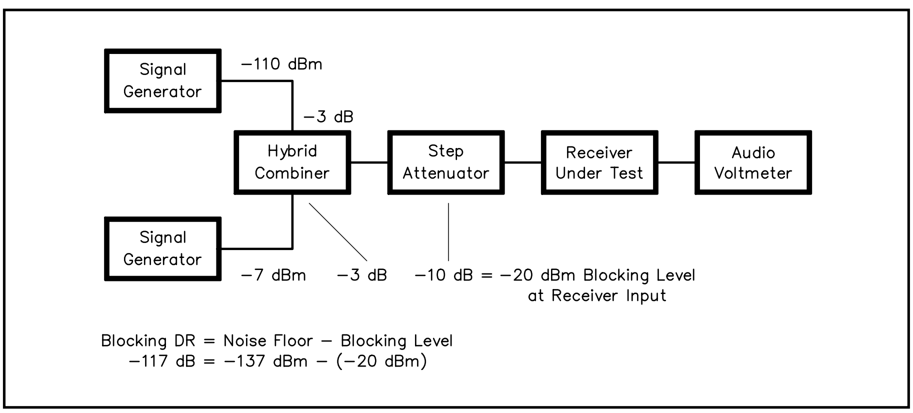


Fig 26.58 — Receiver Blocking DR is measured with this equipment and arrangement. Measurements shown are for the example discussed in the text.

step attenuator is set to 10 dB. The 1-dB compression level is calculated as follows:

$$\begin{aligned}\text{Blocking level} &= -7\text{ dBm} - 3\text{ dBm} - 10\text{ dBm}\\ &= -20\text{ dBm}\end{aligned} \tag{27}$$

To express this as a dynamic range, the blocking level is referenced to the receiver noise floor (calculated earlier). Calculate it as follows:

$$\begin{aligned}\text{Blocking DR} &= \text{noise floor} - \text{blocking level}\\ &= -137\text{ dBm} - (-20\text{ dBm})\\ &= -117\text{ dB}\end{aligned} \tag{28}$$

This value is usually expressed as an absolute value: 117 dB.

Two-Tone IMD Test

The setup for measuring IMD DR is shown in **Fig 26.59**. Two signals of equal level, spaced 20-kHz apart are injected into the receiver input. When we call these frequencies f1 and f2, the so-called third-order IMD products will appear at frequencies of (2f1 – f2) and (2f2 – f1). If the two input frequencies are 14.040 and 14.060 MHz, the third-order products will be at 14.020 and 14.080 MHz. Let's talk through a measurement with these frequencies.

First, set the generators for f1 and f2. Adjust each of them for an output of –10 dBm. Tune the receiver to either of the third-order IMD products. Adjust the step attenuator until the IMD product produces an output 3 dB above the noise level as read on the ac voltmeter.

For an example, say the output of the generator is –10 dBm, the loss through the combiner is 3 dB and the amount of attenuation used is 30 dB. The signal level at the receiver antenna terminal that just begins to cause IMD problems is calculated as:

$$\begin{aligned}\text{IMD level} &= -10\text{ dBm} - 3\text{ dB} - 30\text{ dB}\\ &= -43\text{ dBm}\end{aligned} \tag{29}$$

To express this as a dynamic range the IMD level is referenced to the noise floor as follows:

$$\begin{aligned}\text{IMD DR} &= \text{noise floor} - \text{IMD level}\\ &= -137\text{ dBm} - (-43\text{ dBm})\\ &= -94\text{ dB}\end{aligned} \tag{30}$$

Therefore, the IMD dynamic range of this receiver would be 94 dB.

Third-Order Intercept

Another parameter used to quantify receiver performance is the third-order input intercept (IP^3). This is the point at which the desired response and the third-order IMD response intersect, if extended beyond their linear regions (see **Fig 26.60**). Greater IP^3 indicates better receiver performance. Calculate IP^3 like this:

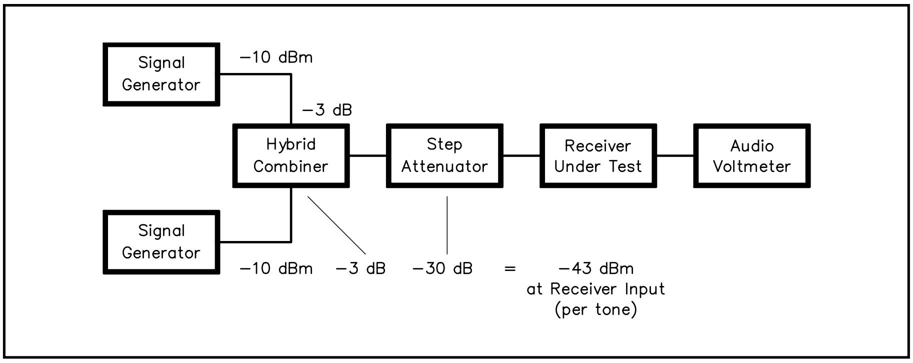


Fig 26.59 — Receiver IMD DR test setup. Signal levels shown are for the example discussed in the text.

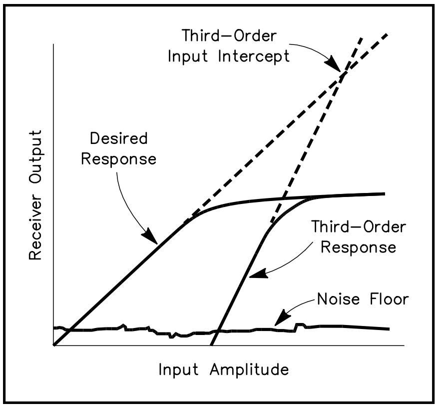


Fig 26.60 — A plot of the receiver characteristics that determine third-order input intercept, a measure of receiver performance.

$$IP^3 = 1.5 \text{ (IMD dynamic range in dB)} + \text{(MDS in dBm)} \qquad (31)$$

For our example receiver:

$$IP^3 = 1.5 \text{ (94 dB)} + (–137 \text{ dBm}) = +4 \text{ dBm}$$

The example receiver we have discussed here is purely imaginary. Nonetheless, its performance is typical of contemporary communications receivers.

Evaluating the Data

Thus far, a fair amount of data has been gathered with no mention of what the numbers really mean. It is somewhat easier to understand exactly what is happening by arranging the data as shown in **Fig 26.61**. The base line represents power levels with a very small level at the left and a higher level (0 dBm) at the right.

The noise floor of our hypothetical receiver is at –137 dBm, the IMD level (the level at which signals will begin to create spurious responses) at –43 dBm and the blocking level (the level at which signals will begin to desensitize the receiver) at –20 dBm. The IMD dynamic range is some 23 dB smaller than the blocking dynamic range. This means IMD products will be heard long before the receiver begins to desensitize, some 23 dB sooner.

SPECTRUM ANALYZERS

A spectrum analyzer is similar to an oscilloscope. Both visually present an electrical signal through graphic representation. The oscilloscope is used to observe electrical signals in the time domain (amplitude as a function of time). The time domain, however, gives little information about the frequencies that make up complex signals. Amplifiers, mixers, oscillators, detectors, modulators and filters are best characterized in terms of their frequency response. This information is obtained by viewing electrical signals in the *frequency domain* (amplitude as a function of frequency). One instrument that can display the frequency domain is the spectrum analyzer.

Time and Frequency Domain

To better understand the concepts of time and frequency domain, see **Fig 26.62**. The three-dimensional coordinates show time (as the

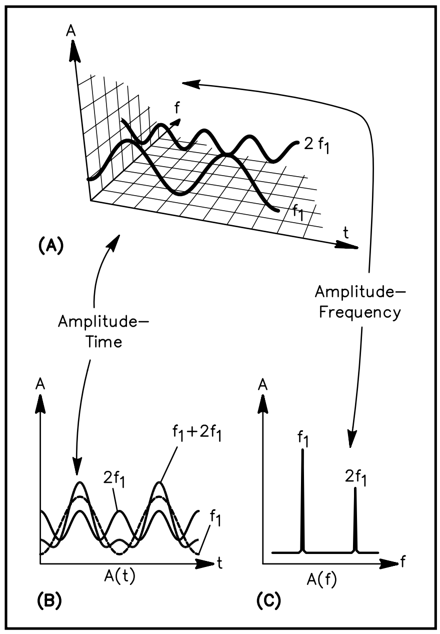


Fig 26.62 — A complex signal in the time and frequency domains. A is a three-dimensional display of amplitude, time and frequency. B is an oscilloscope display of time vs amplitude. C is spectrum analyzer display of the frequency domain and shows frequency vs amplitude.

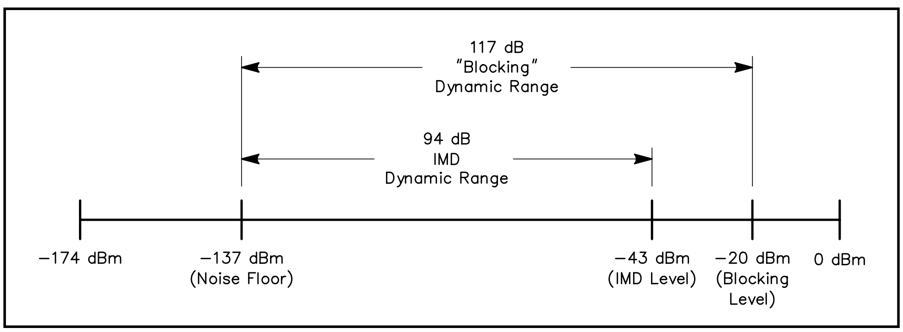


Fig 26.61 — Performance plot of the receiver discussed in the text. This is a good way to visualize the interaction of receiver-performance measurements.

line sloping toward the bottom right), frequency (as the line rising toward the top right) and amplitude (as the vertical axis). The two discrete frequencies shown are harmonically related, so we'll refer to them as f1 and 2f1.

In the representation of time domain at B, all frequency components of a signal are summed together. In fact, if the two discrete frequencies shown were applied to the input of an oscilloscope, we would see the solid line (which corresponds to f1 + 2f1) on the display.

In the frequency domain, complex signals (signals composed of more than one frequency) are separated into their individual frequency components. A spectrum analyzer measures and displays the power level at each discrete frequency; this display is shown at C.

The frequency domain contains information not apparent in the time domain and therefore the spectrum analyzer offers advantages over the oscilloscope for certain measurements. As might be expected, some measurements are best made in the time domain. In these cases, the oscilloscope is a valuable instrument.

Spectrum Analyzer Basics

There are several different types of spectrum analyzers, but by far the most common is nothing more than an electronically tuned superheterodyne receiver. The receiver is tuned by means of a ramp voltage. This ramp voltage performs two functions: First, it sweeps the frequency of the analyzer local oscillator; second, it deflects a beam across the horizontal axis of a CRT display, as shown in **Fig 26.63**. The vertical axis deflection of the CRT beam is determined by the strength of the received signal. In this way, the CRT displays frequency on the horizontal axis and signal strength on the vertical axis.

Most spectrum analyzers use an up-converting technique so that a fixed tuned input filter can remove the image. Only the first local oscillator need be tuned to tune the receiver. In the up-conversion design, a wide-band input is converted to an IF higher than the highest input frequency. As with most up-converting communications receivers, it is not easy to achieve the desired ultimate selectivity at the first IF, because of the high frequency. For this reason, multiple conversions are used to generate an IF low enough so that the desired selectivity is practical. In the example shown, dual conversion is used: The first IF is at 400 MHz; the second at 10.7 MHz.

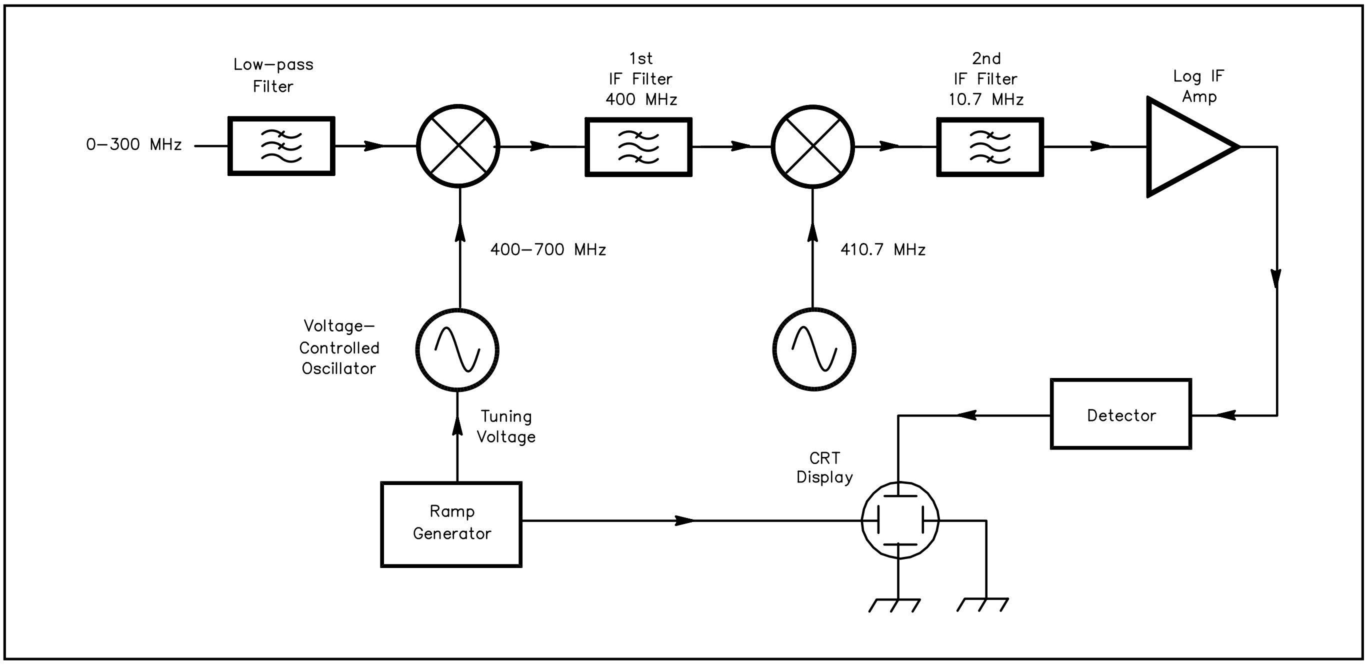


Fig 26.63 — A block diagram of a superheterodyne spectrum analyzer. Input frequencies of up to 300 MHz are up converted by the local oscillator and mixer to a fixed frequency of 400 MHz.

In the example spectrum analyzer, the first local oscillator is swept from 400 MHz to 700 MHz; this converts the input (from nearly 0 MHz to 300 MHz) to the first IF of 400 MHz. The usual rule of thumb for varactor tuned oscillators is that the maximum practical tuning ratio (the ratio of the highest frequency to the lowest frequency) is an octave, a 2:1 ratio. In our example spectrum analyzer, the tuning ratio of the first local oscillator is 1.75:1, which meets this specification.

The image frequency spans 800 MHz to 1100 MHz and is easily eliminated using a low-pass filter with a cut-off frequency around 300 MHz. The 400-MHz first IF is converted to 10.7 MHz where the ultimate selectivity of the analyzer is obtained. The image of the second conversion, (421.4 MHz), is eliminated by the first IF filter. The attenuation of the image should be great, on the order of 60 to 80 dB. This requires a first IF filter with a high Q; this is achieved by using helical resonators, SAW resonators or cavity filters. Another method of eliminating the image problem is to use triple conversion; converting first to an intermediate IF such as 50 MHz and then to 10.7 MHz. As with any receiver, an additional frequency conversion requires added circuitry and adds potential spurious responses.

Most of the signal amplification takes place at the lowest IF; in the case of the example analyzer this is 10.7 MHz. Here the communications receiver and the spectrum analyzer differ. A communications receiver demodulates the incoming signal so that the modulation can be heard or further demodulated for RTTY or packet or other mode of operation. In the spectrum analyzer, only the signal strength is needed.

In order for the spectrum analyzer to be most useful, it should display signals of widely different levels. As an example, signals differing by 60 dB, which is a thousand to one difference in voltage or a million to one in power, would be difficult to display. This would mean that if power were displayed, one signal would be one million times larger than the other (in the case of voltage one signal would be a thousand times larger). In either case it would be difficult to display both signals on a CRT. The solution to this problem is to use a logarithmic display that shows the relative signal levels in decibels. Using this technique, a 1000:1 ratio of voltage reduces to a 60-dB difference.

The conversion of the signal to a logarithm is usually performed in the IF amplifier or detector, resulting in an output voltage proportional to the logarithm of the input RF level. This output voltage is then used to drive the CRT display.

Spectrum Analyzer Performance Specifications

The performance parameters of a spectrum analyzer are specified in terms similar to those used for radio receivers, in spite of the fact that there are many differences between a receiver and a spectrum analyzer.

The sensitivity of a receiver is often specified as the minimum discernible signal, which means the smallest signal that can be heard. In the case of the spectrum analyzer, it is not the smallest signal that can be heard, but the smallest signal that can be seen. The dynamic range of the spectrum analyzer determines the largest and smallest signals that can be simultaneously viewed on the analyzer. As with a receiver, there are several factors that can affect dynamic range, such as IMD, second- and third-order distortion and blocking. IMD dynamic range is the maximum difference in signal level between the minimum detectable signal and the level of two signals of equal strength that generate an IMD product equal to the minimum detectable signal.

Although the communications receiver is an excellent example to introduce the spectrum analyzer, there are several differences such as the previously explained lack of a demodulator. Unlike the communications receiver, the spectrum analyzer is not a sensitive radio receiver. To preserve a wide dynamic range, the spectrum analyzer often uses passive mixers for the first and second mixers. Therefore, referring to Fig 26.63, the noise figure of the analyzer is no better than the losses of the input low-pass filter plus the first mixer, the first IF filter, the second mixer and the loss of the second IF filter. This often results in a combined noise figure of more than 20 dB. With that kind of noise figure the spectrum analyzer is obviously not a communications receiver for extracting very weak signals from the noise but a measuring instrument for the analysis of frequency spectrum.

The selectivity of the analyzer is called the resolution bandwidth. This term refers to the minimum frequency separation of two signals of equal level that can be resolved so there is a 3-dB dip between the two. The IF filters used in a spectrum analyzer differ from a communications receiver in that the filters in a spectrum analyzer have very gentle skirts and rounded passbands, rather than the flat passband and very steep skirts used on an IF filter in a high-quality communications receiver. This rounded passband is necessary because the signals pass into the filter passband as the spectrum analyzer scans the desired frequency range. If the signals suddenly pop into the passband (as they would if the filter had steep skirts), the filter tends to ring; a filter with gentle skirts is less likely to ring. This ringing, called scan loss, distorts the display and requires that the analyzer not sweep frequency too quickly. All this means that the scan rate must be checked periodically to be certain the signal amplitude is not affected by fast tuning.

Spectrum Analyzer Applications

Spectrum analyzers are used in situations where the signals to be analyzed are very complex and an oscilloscope display would be an indecipherable jumble. The spectrum analyzer is also used when the frequency of the signals to be analyzed is very high. Although high-performance oscilloscopes are capable of operation into the UHF region, moderately priced spectrum analyzers can be used well into the gigahertz region.

A spectrum analyzer can also be used to view very low-level signals. For an oscilloscope to display a VHF waveform, the bandwidth of the oscilloscope must extend from zero to the frequency of the waveform. If harmonic distortion and other higher-frequency distortions are to be seen the bandwidth of the oscilloscope must exceed the fundamental frequency of the waveform. This broad bandwidth can also admit a lot of noise power. The spectrum analyzer, on the other hand, analyzes the waveform using a narrow bandwidth; thus it is capable of reducing the noise power admitted.

Probably the most common application of the spectrum analyzer is the measurement of the harmonic content and other spurious signals in the output of a radio transmitter. **Fig 26.64** shows two ways to connect the transmitter and spectrum analyzer. The method shown at A should not be used for wide-band measurements since most line-sampling devices do not exhibit a constant-amplitude output over a broad frequency range. Using a line sampler is fine for narrow-band measurements, however. The method shown at B is used in the ARRL Lab. The attenuator must be capable of dissipating the transmitter power. It must also have sufficient attenuation to protect the spectrum analyzer input. Many spectrum analyzer mixers can be damaged by only a few milliwatts, so most analyzers have an adjustable input attenuator that will provide a reasonable amount of attenuation to protect the sensitive input mixer from damage. The power limitation of the attenuator it-

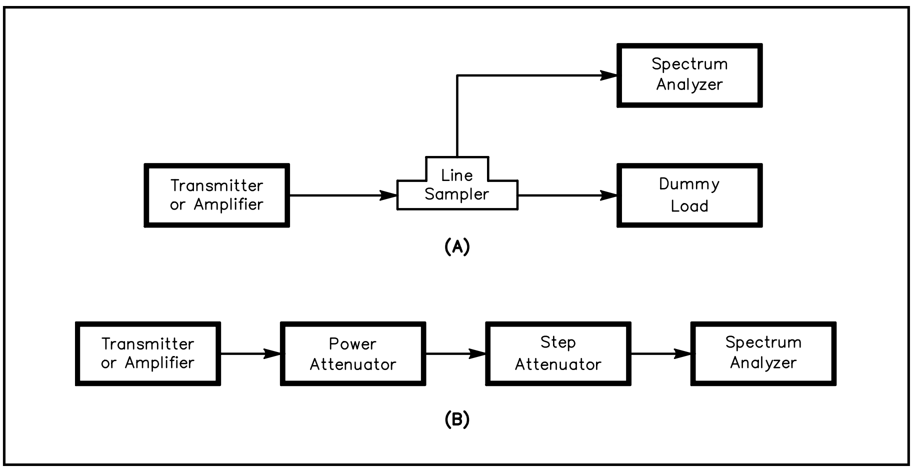


Fig 26.64 — Alternate bench setups for viewing the output of a high power transmitter or oscillator on a spectrum analyzer. A uses a line sampler to pick off a small amount of the transmitter or amplifier power. In B, most of the transmitter power is dissipated in the power attenuator.

self is usually on the order of a watt or so, however. This means that 20 dB of additional attenuation is required for a 100-W transmitter, 30 dB for a 1000-W transmitter and so on, to limit the input to the spectrum analyzer to 1 W. There are specialized attenuators that are made for transmitter testing; these attenuators provide the necessary power dissipation and attenuation in the 20 to 30-dB range.

When using a spectrum analyzer it is very important that the maximum amount of attenuation be applied before a measurement is made. In addition, it is a good practice to start with maximum attenuation and view the entire spectrum of a signal before the attenuator is adjusted. The signal being viewed could appear to be at a safe level, but another spectral component, which is not visible, could be above the damage limit. It is also very important to limit the input power to the analyzer when pulse power is being measured. The average power may be small enough so the input attenuator is not damaged, but the peak pulse power, which may not be readily visible on the analyzer display, can destroy a mixer, literally in microseconds.

When using a spectrum analyzer it is necessary to ensure that the analyzer does not generate additional spurious signals that are then attributed to the system under test. Some of the spurious signals that can be generated by a spectrum analyzer are harmonics and IMD. If it is desired to measure the harmonic levels of a transmitter at a level below the spurious level of the analyzer itself, a notch filter can be inserted between the attenuator and the spectrum analyzer as shown in **Fig 26.65**. This reduces the level of the fundamental signal and prevents that signal from generating harmonics within the analyzer, while still allowing the harmonics from the transmitter to pass through to the analyzer without attenuation. Use caution with this technique; detuning the notch filter or inadvertently changing the transmitter frequency will allow potentially high levels of power to enter the analyzer. In addition, use care when choosing filters; some filters (such as cavity filters) respond not only to the fundamental but notch out odd harmonics as well.

It is good practice to check for the generation of spurious signals within the spectrum analyzer. When a spurious signal is generated by a spectrum analyzer, adding attenuation at the analyzer input will cause the internally generated spurious signals to decrease by an amount greater than the added attenuation. If attenuation added ahead of the analyzer causes all of the visible signals to decrease by the same amount, this indicates a spurious-free display.

The input impedance for most RF spectrum analyzers is 50 Ω; not all circuits have convenient 50-Ω connections that can be accessed for testing purposes, however. Using a probe such as the one shown in **Fig 26.66** allows the analyzer to be used as a troubleshooting tool. The probe can be used to track

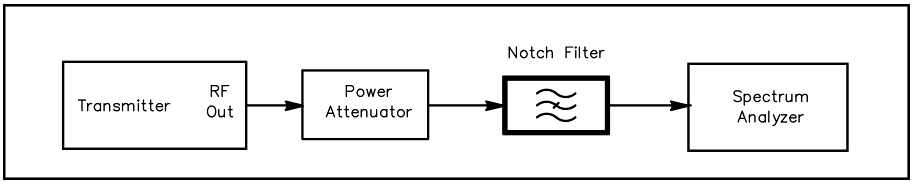


Fig 26.65 — A notch filter is another way to reduce the level of a transmitter's fundamental signal so that the fundamental does not generate harmonics within the analyzer. However, in order to know the amplitude relationship between the fundamental and the transmitter's actual harmonics and spurs, the attenuation of the fundamental in the notch filter must be known.

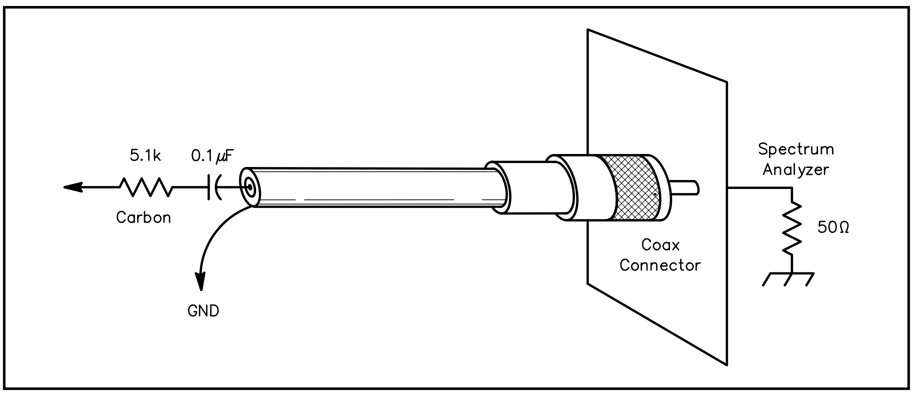


Fig 26.66 — A voltage probe designed for use with a spectrum analyzer. Keep the probe tip (resistor and capacitor) and ground leads as short as possible.

down signals within a transmitter or receiver, much like an oscilloscope is used. The probe shown offers a 100:1 voltage reduction and loads the circuit with 5000 Ω. A different type of probe is shown in **Fig 26.67**. This inductive pickup coil (sometimes called a "sniffer") is very handy for troubleshooting. The coil is used to couple signals from the radiated magnetic field of a circuit into the analyzer. A short length of miniature coax is wound into a pick-up loop and soldered to a larger piece of coax. The use of the coax shields the loop from coupling energy from the electric field component. The dimensions of the loop are not critical, but smaller loop dimensions make the loop more accurate in locating the source of radiated RF. The shield of the coax provides a complete electrostatic shield without introducing a shorted turn.

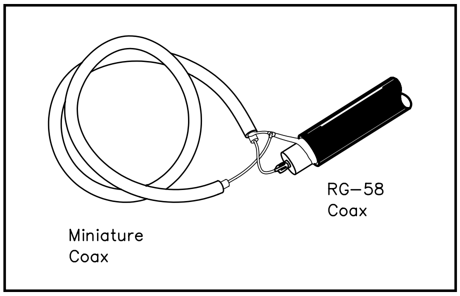


Fig 26.67 — A "sniffer" probe consisting of an inductive pick-up. It has an advantage of not loading the circuit under test. See text for details.

The sniffer allows the spectrum analyzer to sense RF energy without contacting the circuit being analyzed. If the loop is brought near an oscillator coil, the oscillator can be tuned without directly contacting (and thus disturbing) the circuit. The oscillator can then be checked for reliable starting and the generation of spurious sidebands. With the coil brought near the tuned circuits of amplifiers or frequency multipliers, those stages can be tuned using a similar technique.

Even though the sniffer does not contact the circuit being evaluated, it does extract some energy from the circuit. For this reason, the loop should be placed as far from the tuned circuit as is practical. If the loop is placed too far from the circuit, the signal will be too weak or the pick-up loop will pick up energy from other parts of the circuit and not give an accurate indication of the circuit under test.

The sniffer is very handy to locate sources of RF leakage. By probing the shields and cabinets of RF-generating equipment (such as transmitters) egress and ingress points of RF energy can be identified by increased indications on the analyzer display.

One very powerful characteristic of the spectrum analyzer is the instrument's capability to measure very low-level signals. This characteristic is very advantageous when very high levels of attenuation are measured. **Fig 26.68** shows the setup for tuning the notch and passband of a VHF duplexer. The spectrum analyzer, being capable of viewing signals well into the low microvolt region, is capable of measuring the insertion loss of the notch cavity more than 100 dB below the signal generator output. Making a measurement of this sort requires care in the interconnection of the equipment and a well designed spectrum analyzer and signal generator.

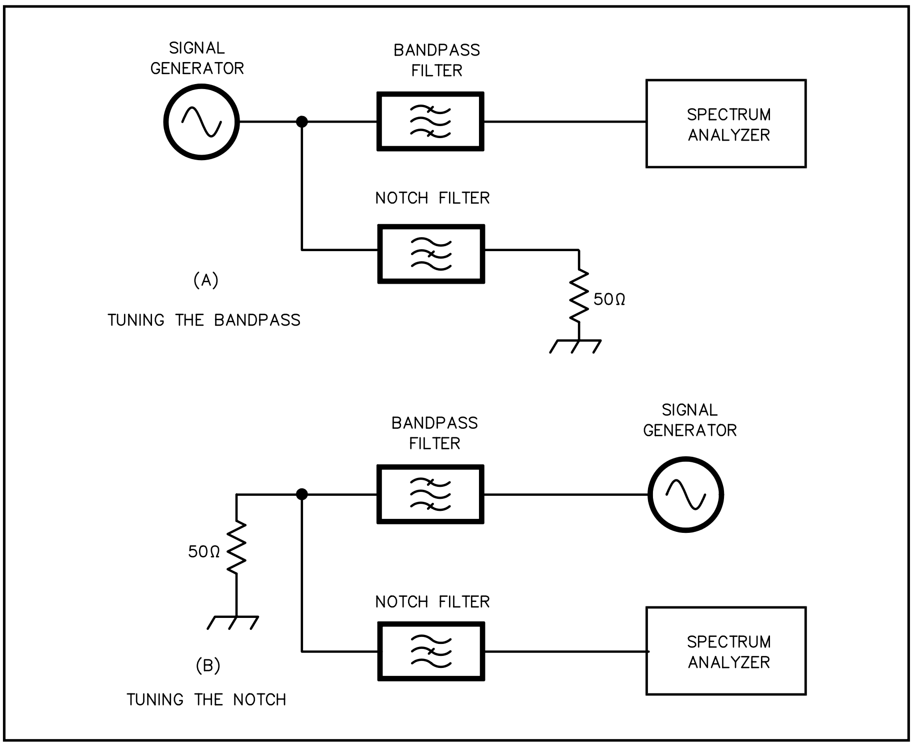


Fig 26.68 — Block diagram of a spectrum analyzer and signal generator being used to tune the band-pass and notch filters of a duplexer. All ports of the duplexer must be properly terminated and good quality coax with intact shielding used to reduce leakage.

RF energy leaking from the signal generator cabinet, line cord or even the coax itself, can get into the spectrum analyzer through similar paths and corrupt the measurement. This leakage can make the measurement look either better or worse than the actual attenuation, depending on the phase relationship of the leaked signal.

Extensions of Spectral Analysis

What if a signal generator is connected to a spectrum analyzer so that the signal generator output frequency is exactly the same as the receiving frequency of the spectrum analyzer? It would certainly appear to be a real convenience not to have to continually reset the signal generator to the desired frequency. It is, however, more than a convenience. A signal generator connected in this way is called a tracking generator because the output frequency tracks the spectrum analyzer input frequency. The tracking generator makes it possible to make swept frequency measurements of the attenuation characteristics of circuits, even when the attenuation involved is large.

Fig 26.69 shows the connection of a tracking generator to a circuit under test. In order for the tracking generator to create an output frequency exactly equal to the input frequency of the spectrum analyzer, the internal local oscillator frequencies of the spectrum analyzer must be known. This is the reason for the interconnections between the tracking generator and the spectrum analyzer. The test setup shown will measure the gain or loss of the circuit under test. Only the magnitude of the gain or loss is available; in some cases, the phase angle between the input and output would also be an important and necessary parameter.

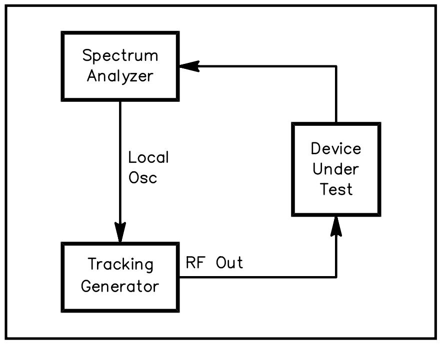


Fig 26.69 — A signal generator (shown in the figure as the "Tracking Generator") locked to the local oscillator of a spectrum analyzer can be used to determine filter response over a range of frequencies.

The spectrum analyzer is not sensitive to the phase angle of the tracking generator output. In the process of generating the tracking generator output, there are no guarantees that the phase of the tracking generator will be either known or constant. This is especially true of VHF spectrum analyzers/tracking generators where a few inches of coaxial cable represents a significant phase shift.

One effective way of measuring the phase angle between the input and output of a device under test is to sample the phase of the input and output of device under test and apply the samples to a phase detector. **Fig 26.70** shows a block diagram of this technique. An instrument that can measure both the magnitude and phase of a signal is called a vector network analyzer or simply a network analyzer. The magnitude and phase can be displayed either separately or together. When the magnitude and phase are displayed together the two can be presented as two separate traces, similar to the two traces on a dual-trace oscilloscope. A much more useful method of display is to present the magnitude and phase as a polar plot where the locus of the points of a vector having the length of the magnitude and the angle of the phase are displayed. Very sophisticated network analyzers can display all of the S parameters of a circuit in either a polar format or a Smith Chart format.

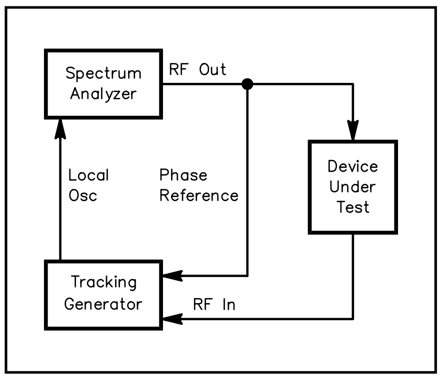


Fig 26.70 — A network analyzer is usually found in commercial communications development labs. It can measure both the phase and magnitude of the filter input and output signals. See text for details.

Transmitter Performance Tests

The test setup used in the ARRL Laboratory for measuring an HF transmitter or amplifier is shown in **Fig 26.71**. As can be seen, different power levels dictate different amounts of attenuation between the transmitter or amplifier and the spectrum analyzer.

Spurious Emissions

Fig 26.72 shows the broadband spectrum of a transmitter, showing the harmonics in the output. The horizontal (frequency) scale is 5 MHz per division; the main output of the transmitter at 7 MHz can be seen about 1.5 major divisions from the left of the trace. A very large apparent signal is seen at the extreme left of the trace. This occurs at what would be zero frequency and it is caused by the first local oscillator frequency being exactly the first IF. All up-converting superheterodyne spectrum analyzers have this IF feedthrough; in addition, this signal is occasionally accompanied by a smaller spurious signal, generated within the analyzer. To determine what part of the displayed signal is a spurious response caused by IF feedthrough and what is an actual input signal, simply remove the input signal and observe the trace. It is not necessary or desirable that the transmitter be modulated for this broadband test.

Other transmitter tests that can be performed with a spectrum analyzer include measurement of two-tone IMD and SSB carrier and unwanted sideband suppression.

Two-Tone IMD

Investigating the sidebands from a modulated transmitter requires a narrow-band spectrum analysis and produces displays similar to that shown in Fig 26.73. In this example, a two-tone test signal is used to modulate the transmitter. The display shows the two test tones plus some of the IMD produced by the SSB transmitter. The test setup used to produce this display is shown in Fig 26.74.

In this example, a two-tone test signal with frequencies of 700 and 1900 Hz is used to modulate the transmitter. Set the transmitter output and au-

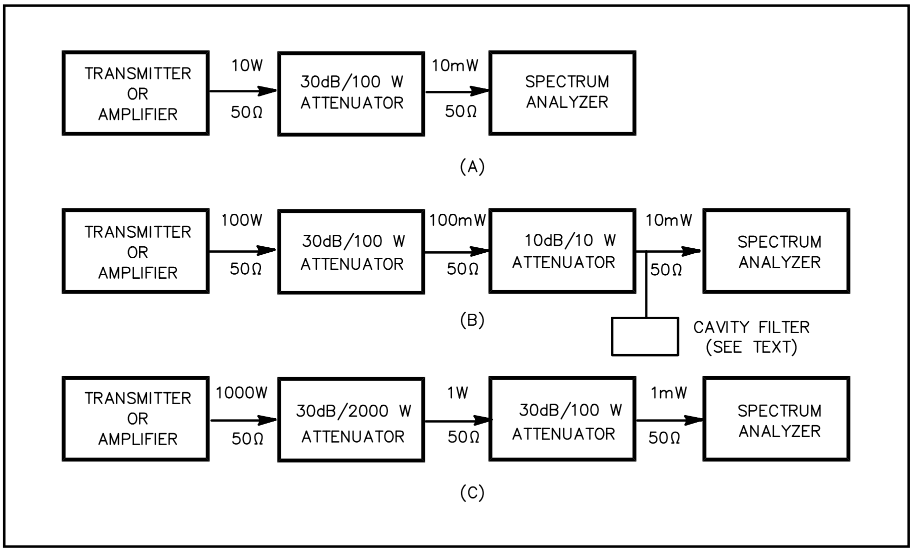
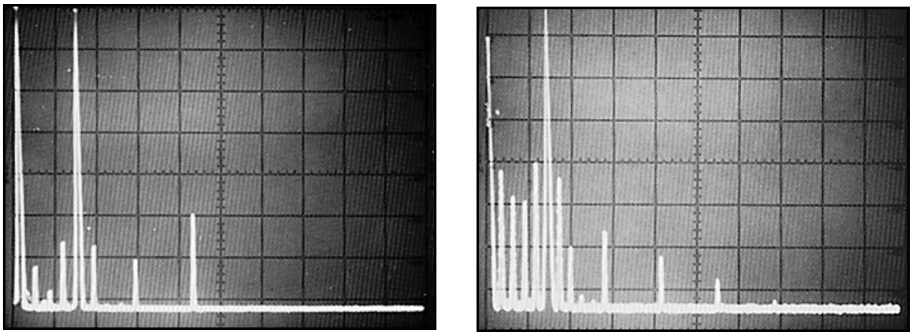


Fig 26.71 — These setups are used in the ARRL Laboratory for testing transmitters or amplifiers with several different power levels.



(A) **(B)**

Fig 26.72 — Comparison of two different transmitters on the 40-m band as seen on a spectrum analyzer display. The photograph at A shows a relatively clean transmitted signal but the transmitter at B shows more spurious signal content. Horizontal scale is 5 MHz per division; vertical is 10 dB per division. According to current FCC spectral purity requirements both transmitters are acceptable.

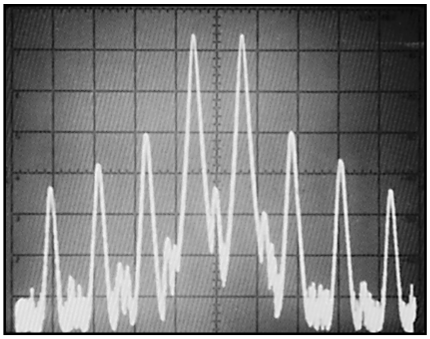


Fig 26.73 — An SSB transmitter two-tone test as seen on a spectrum analyzer. Each horizontal division represents 1 kHz and each vertical division is 10 dB. The third-order products are 30 dB below the PEP (top line), the fifth-order products are down 37 dB and seventh-order products are down 44 dB. This represents acceptable (but not ideal) performance.

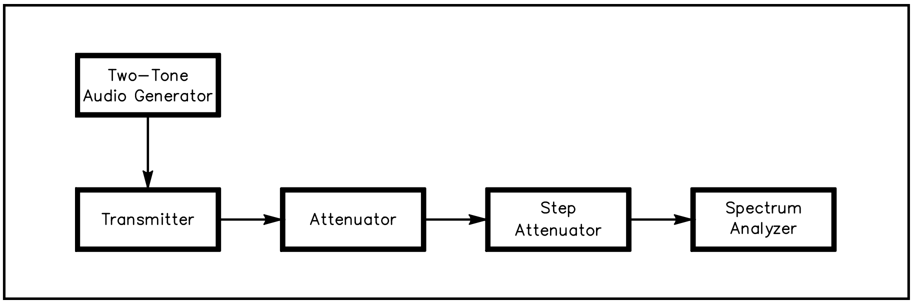


Fig 26.74 — The test setup used in the ARRL Laboratory to measure the IMD performance of transmitters and amplifiers.

dio input to the manufacturer's specifications. Each desired tone is adjusted to be equal in amplitude and centered on the display. The step attenuators and analyzer controls are then adjusted to set the two desired signals 6 dB below the 0-dB reference (top) line. The IMD products can then be read directly from the display in terms of "dB below Peak Envelope Power (PEP)." (In the example shown, the third-order products are 30 dB below PEP, the fifth-order products are 37 dB down, the seventh- order products are down 44 dB.)

Carrier and Unwanted Sideband Suppression

Single-tone audio input signals can be used with the same setup to measure unwanted sideband and carrier suppression of SSB signals. In this case, set the single tone to the 0-dB reference line. (Once the level is set, the audio can be disabled for carrier suppression measurements in order to eliminate IMD and other effects.)

Phase Noise

Phase/composite noise is also measured with spectrum analyzers in the ARRL Lab. This test requires specialized equipment and is included here for information purposes only.

The purpose of the Composite-Noise test is to observe and measure the phase and amplitude noise, as well as any close-in spurious signals generated by a transmitter. Since phase noise is the primary noise component in any well-designed transmitter, almost all of the noise observed during this test is phase noise.

This measurement is accomplished in the lab by converting the transmitter output down to a frequency band about 10 or 20 kHz above baseband. A mixer and a signal generator (used as a local oscillator) are used to perform this conversion. Filters remove the 0-Hz component as well as any unwanted heterodyne components. A spectrum analyzer (see **Fig 26.75**) displays the remaining noise and spurious signals from 2 to 20 kHz from the carrier frequency (in the CW mode).

Tests in the Time Domain

Oscilloscopes are used for transmitter testing in the time domain. Dual-trace instruments are best in most cases, providing easy to read time-delay measurements between keying input and RF- or audio-output signals. Common transmitter measurements performed with 'scopes include CW keying wave shape and time delay and SSB/FM transmit-to-audio turnaround tests (important for many digital modes).

A typical setup for measuring CW keying waveform and time delay is shown in **Fig 26.76**. A keying test generator is used to repeatedly key the transmitter at a controlled rate. The generator

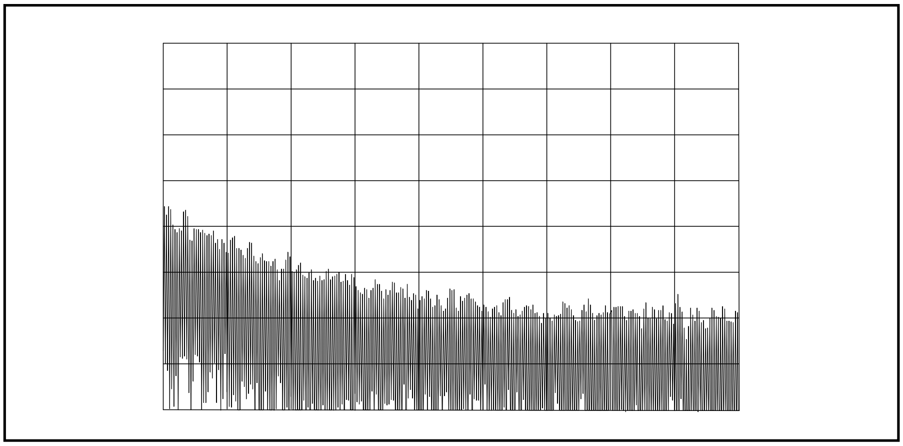


Fig 26.75 — The spectral-display results of a composite-noise test in the ARRL Lab. This display is for the ICOM IC-707 reviewed in April 1994 *QST*. Power output is 100 W at 14 MHz. Vertical divisions are 10 dB; horizontal divisions are 2 kHz. The log reference level (the top horizontal line on the scale) represents –60 dBc/Hz and the baseline is –140 dBc/Hz. The carrier, off the left edge of the plot, is not shown. This plot shows composite transmitted noise 2 to 20 kHz from the carrier.

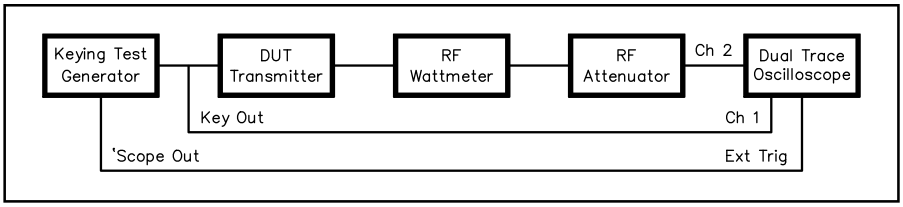


Fig 26.76 — CW keying waveform test setup.

can be set to any reasonable speed, but ARRL tests are usually conducted at 20-ms on and 20-ms off (25 Hz, 50% duty cycle). **Fig 26.77** shows a typical display. The rise and fall times of the RF output pulse are measured between the 10% and 90% points on the leading and trailing edges, respectively. The delay times are measured between the 50% points of the keying and RF output waveforms. Look at the **Transceivers** chapter for further discussion of CW keying issues.

For voice modes (SSB/FM), a PTT-to-RF output test is similar to CW keying tests. It measures rise and fall times, as well as the on- and off-delay times just as in the CW test. See **Fig 26.78** for the test setup.

"Turnaround time" is the time it takes for a transceiver to switch from the 50% fall time of a keying pulse to 50% rise of audio output. The test setup is shown in **Fig 26.79**. Turnaround time measurements require extreme care with respect to transmitter output power, attenuation, signal-generator output and the maximum input signal that can be tolerated by the generator. The generator's specifications must not be exceeded and the input to the receiver must be at the required level, usually S9. Receiver AGC is usually off for this test, but experimentation with AGC and signal input level can reveal surprising variations.

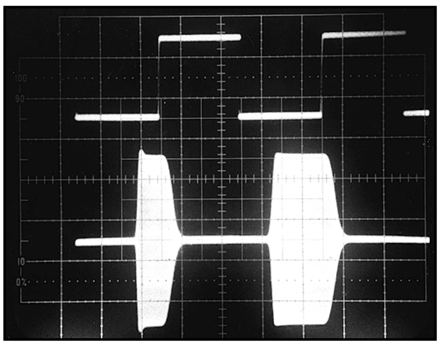


Fig 26.77 — Typical CW keying waveform test results. This display is for the ICOM IC-707 (semi-break-in mode) reviewed in April 94 *QST*. The upper trace is the actual key closure; the lower trace is the RF envelope. Horizontal divisions are 10 ms. The transceiver was being operated at 100 W output at 14 MHz.

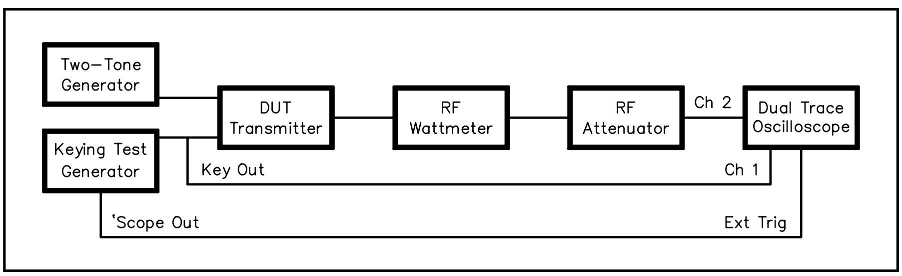


Fig 26.78 — PTT-to-RF-output test setup for voice-mode transmitters.

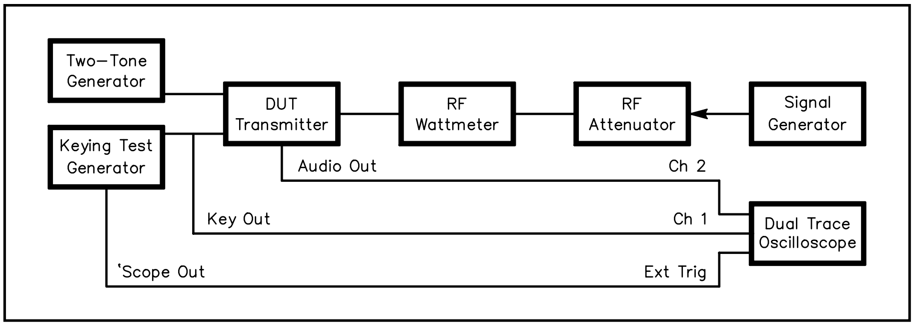


Fig 26.79 — Transmit-receive turn-around time test setup.

The keying rate must be considerably slower than the turn-around time; rates of 200-ms on/200-ms off or faster, have been used with success in Product Review tests at the ARRL Lab.

Turn-around time is an important consideration with some digital modes. AMTOR, for example, requires a turn-around time of 35 ms or less.

Oscilloscope Bibliography

R. vanErk, *Oscilloscopes, Functional Operation and Measuring Examples*, McGraw-Hill Book Co, New York, 1978.

V. Bunze, *Probing in Perspective—Application Note 152*, Hewlett-Packard Co, Colorado Springs, CO, 1972 (Pub No. 5952-1892).

The XYZs of Using a Scope, Tektronix, Inc, Portland, OR, 1981 (Pub No. 41AX-4758).

Basic Techniques of Waveform Measurement (Parts 1 and 2), Hewlett-Packard Co, Colorado Springs, CO, 1980 (Pub No. 5953-3873).

J. Millman, and H. Taub, *Pulse Digital and Switching Waveforms*, McGraw-Hill Book Co, New York, 1965, pp 50-54.

V. Martin, *ABCs of DMMs*, Fluke Corp, PO Box 9090, Everett, WA 98206.

Glossary

AC (Alternating current) — The polarity constantly reverses, as contrasted to dc (direct current) where polarity is fixed.

Analog — Signals which have a full set of values. If the signal varies between 0 and 10 V all values in this range can be found. Compare this to a *digital* system.

Attenuator — A device which reduces the amplitude of a signal

Average value — Obtained by recording or measuring *N* samples of a signal, adding up all of these values, and dividing this sum by *N*.

Bandwidth — A measure of how wide a signal is in frequency. If a signal covers 14,200 to 14,205 kHz its bandwidth is said to be 5 kHz.

BNC — A small connector used with coax cable.

Bridge circuit — Four passive elements, such as resistors, inductors, connected as a pair of voltage dividers with a meter or other measuring device across two opposite junctions. Used to indicate the relative values of the four passive elements. See the chapter discussion of Wheatstone bridges.

CMOS — A family of digital logic elements usually selected for their low power drain. See the **Digital Signal Theory** chapter of this *Handbook*.

Coaxial cable (coax) — A cable formed of two conductors that share the same axis. The center conductor may be a single wire or a stranded cable. The outer conductor is called the shield. The shield may be flexible braid, foil, semirigid or rigid metal. For more information, look in the **Transmission Lines** chapter.

Combiners — See ***Hybrid*****.**

D'Arsonval meter — A common mechanical meter consisting of a permanent magnet and a moving coil (with pointer attached).

DC (direct current) — The polarity is fixed for all time, as contrasted to ac (alternating current) where polarity constantly reverses.

Digital — A system that allows signals to assume a finite range of states. Binary logic is the most common example. Only two values are permitted in a binary system: one value is defined as a logical *1* and the other value as a logical *0*. See the *Handbook* chapter on **Digital Signal Theory**.

Divider — A network of components that produce an output signal that is a fraction of the input signal. The ratio of the output to the input is the division factor. An analog divider divides voltage (a string of series connected resistors) or current (parallel connected resistors). Digital dividers divide pulse trains or frequency.

DMM (Digital multimeter) — A test instrument that usually measures at least: voltage, current and resistance, and displays the result on a numeric digit display, rather than an analog meter.

Dummy antenna or dummy load — A resistor or set of resistors used in place of an antenna to test a transmitter without radiating any electromagnetic energy into the air.

DVM (digital volt meter) — See ***DMM***.

FET voltmeter — See also ***VTVM***. An updated version of a VTVM using field effect transistors (FETs) in place of vacuum tubes.

Flip-flop — A digital circuit that has two stable states. See the chapter on **Digital Signal Theory**.

Frequency marker — Test signals generated at selected intervals (such as 25 kHz, 50 kHz, 100 kHz) for calibrating the dials of receivers and transmitters.

Fundamental — The first signal or frequency in a series of harmonically related signals. This term is often used to describe an oscillator or transmitter's desired signal.

Harmonic — A signal occurring at some integral multiple (such as two, three, four) of a *fundamental* frequency.

Hybrid (hybrid combiners) — A device used to connect two signal generators to one receiver for test purposes, without the two generators affecting each other.

IC (integrated circuit) — A complete circuit built into a single electronic component.

LCD (liquid crystal display) — A low-power display device utilizing the physics of liquid crystals. They usually need either ambient light or backlighting to be seen.

LED (light emitting diode) — A diode that emits light when an appropriate voltage (usually 1.5 V at about 20 mA) is connected. They are used either as tiny pilot lights or in bar shapes to display letters and numbers.

Loran — A navigation system using very-low-frequency transmitters.

Marker — See ***Frequency marker*.**

Multiplier — A circuit that purposely creates some desired *harmonic* of its input signal. For example, a frequency multiplier that takes energy from a 3.5-MHz exciter and puts out RF at 7 MHz is a two times multiplier, usually called a frequency doubler.

N — A type of coaxial cable connector common at UHF and higher frequencies.

NAND — A digital element that performs the *not-and* function. See the **Digital Signal Theory** chapter.

Noise (noise figure) — Noise is generated in all electrical circuits. It is particularly critical in those stages of a receiver that are closest to the antenna (RF amplifier and mixer), because noise generated in these stages can mask a weak signal. The noise figure is a measure of this noise generation. Lower noise figures mean that less noise is generated and weaker signals can be heard.

NOR — A digital element that performs the *not-or* function. See the **Digital Signal Theory** chapter.

Null (nulling) — The process of adjusting a circuit for a minimum reading on a test meter or instrument. At a perfect null there is null, or no, energy to be seen.

Ohmmeter — A meter that measures the value of resistors. Usually part of a multimeter. See ***VOM*** and ***DMM***.

Peak value — The highest value of a signal during the measuring time. If a measured voltage varies in value from 1 to 10 V over a measuring period, the peak value would be the highest measured, 10 V.

PL-259 — A connector used for coaxial cable, usually at HF. It is also known as a male UHF connector. It is an inexpensive and common connector, but it is not weatherproof, nor is its impedance constant over frequency.

Prescaler — A circuit used ahead of a counter to extend the counter range to higher frequencies. A counter capable of operating up to 50 MHz can count up to 500 MHz when used with a ÷10 prescaler.

Q — The ratio of the reactance to the resistance of a component or circuit. It provides a measure of bandwidth. Lower resistive losses make for a higher Q, and a narrower bandwidth.

RMS (root mean square) — A measure of the value of a voltage or current obtained by taking values from successive small time slices over a complete cycle of the waveform, squaring those values, taking the mean of the squares, and then the square root of the mean. Very significant when working with good ac sine waves, where the RMS of the sine wave is 0.707 of the peak value.

Scope — Slang for oscilloscope. See the Oscilloscopes section of this chapter.

Shunt — Elements connected in parallel.

Sinusoidal (sine wave) — The nominal waveform for unmodulated RF energy and many other ac voltages.

Spectrum — Used to describe a range of frequencies or wavelengths. The RF spectrum starts at perhaps 10 kHz and extends up to several hundred gigahertz. The light spectrum goes from infrared to ultraviolet.

Spurious emissions, or spurs — Unwanted energy generated by a transmitter or other circuit. These emissions include, but are not limited to, *harmonics*.

Thermocouple — A device made up of two different metals joined at two places. If one joint is hot and the other cold a voltage may be developed, which is a measure of the temperature difference.

Time domain — A measurement technique where the results are plotted or shown against a scale of time. In contrast to the frequency domain, where the results are plotted against a scale of frequency.

TTL (Transistor-transistor-logic) — A logic IC family commonly used with 5 V supplies. See the chapter on **Digital Signal Theory**.

Vernier dial or vernier drive — A mechanical system of tuning dials, frequently used in older equipment, where the knob might turn 10 times for each single rotation of the control shaft.

VOM (volt-ohm-meter) — A multimeter whose design predates multiple scale meters (see **DMM**).

VTVM (vacuum tube voltmeter) — A meter that was developed to provide a high input resistance and therefore low current drain (loading) from the circuit being tested. Now replaced by the FET meter.

Wheatstone bridge — See ***Bridge circuit***.