Contents

Chapter 25 — CD-ROM Content

Supplemental Files
- Cathode ray tube theory
- Test and Measurement Bibliography

Project Files
- ARRL Lab Test Procedures Manual – 2010
- Gate Dip Oscillator articles and PCB artwork — Alan Bloom, N1AL
- Build a Return Loss Bridge — James Ford, N6JF
- Logic Probe — supporting photos and graphics — Alan Bloom, N1AL
- RF Power Meter — supporting files — William Kaune, W7IEQ
- Compensated RF Voltmeter articles — Sidney Cooper, K2QHE
- "Noise Instrumentation and Measurement" by Paul Wade W1GHZ
- RF Sampler Construction details — Thomas Thompson, WØIVJ
- RF Step Attenuator — Denton Bramwell, K7OWJ
- Tandem Match articles — John Grebenkemper, KI6WX
- Transistor Tester — PCB artwork and layout graphics — Alan Bloom, N1AL
- Two-Tone Oscillator — PCB artwork and layout graphics — ARRL Lab
- "A Low Frequency Adapter for your Vector Network Analyzer (VNA)" by Jacques Audet, VE2AZX

Chapter 25

Test Equipment and Measurements

This chapter, written by Alan Bloom, N1AL, covers test equipment and measurement techniques common to Amateur Radio. The initial sections discuss ac, dc and RF measurements and describe what to look for in commercially available test equipment. The concluding sections discuss transmitter and receiver performance testing, including the standard tests performed by the ARRL Lab. A selection of simple test equipment construction projects is presented as well.

25.1 Introduction

According to the great physicist Sir William Thomson, Lord Kelvin, (1824-1907) "When you can measure what you are speaking about, and express it in numbers, you know something about it; but when you cannot measure it, when you cannot express it in numbers, your knowledge is of a meager and unsatisfactory kind." This is the purpose of electronic test and measurement, to express the characteristics and performance of electronic devices in numbers that can be observed, recorded and compared.

The amateur should undertake to master basic electronic and test instruments: the multimeter, oscilloscope, signal generator, RF power meter and SWR and impedance analyzers. This chapter, written by Alan Bloom, N1AL is the reader's guide to these instruments and the physical parameters they measure. The chapter goes beyond basic instrumentation to advanced instruments and measurements that are often encountered by amateurs. The material does not attempt to be complete in covering all available instruments and measurements. The goal of the chapter is to instruct and educate, present useful projects, and encourage the reader to understand more about these important facets of the radio art.

25.1.1 Measurement Standards

The measurement process involves evaluating the characteristic being tested using a *standard*, which is a rule for determining the proper numbers to assign to the measurement. In the past, instruments were calibrated to standards represented by physical objects. For example, the reference standard for length was a platinum bar, exactly 1.0 meters long, stored in an environmentally-controlled vault in Paris, France. In the latter part of the 20th century, most measurements were redefined based on fundamental physical constants. In 1983 the meter was defined as the distance traveled by light in free space in 1/299,792,458 second.

While most common everyday measurements in the United States still use the old Imperial system (feet, inches, pounds, gallons, and so on), electronic measurements are based on the international system of units, called *SI* after the French name *Système International d'Unités*, which is the modern, revised version of the metric system. The SI defines seven base units, which are length (meter, m), time (second, s), mass (kilogram, kg), temperature (kelvin, K), amount of substance (mole, mol), current (ampere, A), and light intensity (candela, cd). All other units are derived from those seven. For example, the unit of electric charge, the coulomb, is proportional to s × A and the volt has the dimensions $m^2 \times kg \times s^{-3} \times A^{-1}$.

In the United States, measurement standards are managed by the *National Institute of Standards and Technology* (NIST), a non-regulatory agency of the federal government which until 1988 was called the National Bureau of Standards (NBS).[1] (Notes may be found in the References section.) NIST's services include calibration of *transfer standards*. The transfer standards in turn are used to calibrate *working standards* which are used by companies to calibrate and measure their products. Such products are said to be *NIST traceable* if the rules and procedures specified by NIST have been followed. Most low-cost instruments used by hobbyists are another level down in accuracy, being calibrated by instruments that themselves may or may not be NIST traceable.

25.1.2 Measurement Accuracy

Accuracy and resolution are different specifications that are often confused. *Resolution* is the smallest distinguishable difference in a measured value. *Accuracy* is the maximum expected error in the measurement. For example an 8-digit frequency counter can measure the frequency of a 100 MHz signal to a resolution of 1 Hz, which is 0.01 ppm (parts per million). However, accuracy is determined by the time-base oscillator used as a reference, which typically would be significantly less accurate than 0.01 ppm in a low-cost instrument. Similarly, many digital voltmeters can display measurements to more digits of resolution than their accuracy permits. The extra resolution can be useful when comparing two or more values that differ only slightly because closely-spaced measurements tend to have nearly the same error. The difference can then be measured more accurately than the individual values.

An instrument's accuracy can be specified in absolute or relative terms or sometimes in both. An example of an absolute specification is an RF power meter with an accuracy of "5% of full scale." If the full-scale reading is 100 W, then the accuracy is plus or minus 5 W at all power levels. Theoretically a reading of 10 W could represent an actual power of 5 W to 15 W.

An example of a relative accuracy specification is an analog voltmeter with an accuracy of 3%. That means a voltage of 1.000 V can be measured to an accuracy of plus or minus 0.030 V. A voltage of 10.000 V can be measured to an accuracy of plus or minus 0.3 V. An example of a combined absolute and relative accuracy specification is a frequency counter with an accuracy of "1 ppm plus one count." When measuring a 100 kHz signal with 1 Hz resolution, one count is 10 ppm of the 100 kHz signal. The total accuracy is 1 + 10 = 11 ppm, or ±1.1 Hz.

An important point is that measurement error is not a mistake but rather a natural result of the imperfections inherent in any measurement. The sources of error can be sorted into several general classes. *Systematic error* is repeatable; it is always the same when the measurement is taken in the same way. An example is the inaccuracy of the voltage reference in a digital voltmeter that causes all measurements to be off by the same percentage in the same direction. *Random error* is caused by noise and results in a different measured value each time it is measured. Receiver sensitivity measurements involve measuring the signal-to-noise ratio of the audio output which varies due to the random fluctuations of the noise level.

Dynamic error results when the value being measured changes with time. The peak-envelope power (PEP) of a single-sideband transmitter varies with the modulation that is present at the particular time the measurement is taken. *Instrument insertion error* (also called *loading error*) is an often-overlooked factor. For example, a voltmeter must draw at least a little current from the circuit under test in order to perform the measurement, which can affect circuit operation. When performing high-frequency measurements, the capacitance of the measurement probe often can be significant.

25.2 DC Measurements

In discussions of instruments and measurements, the abbreviation *dc* generally refers to currents and voltages that remain stable during the measurement. If the dc current or voltage changes rapidly, it is said to have an *ac component* that may be measured separately.

25.2.1 Basic Vocabulary and Units

The **Electrical Fundamentals** chapter introduces basic electrical units of measurement and how they relate to circuit operation. The following material discusses additional points of particular relevance to test and measurement.

VOLTAGE

Another name for voltage is *potential difference*, because it is a measure of the difference in electrical potential between two points in a circuit. That is a very important point that is widely misunderstood among radio amateurs, technicians and even many engineers. It makes no sense to speak of the voltage "at" a particular point in a circuit unless you specify the reference point to which you are measuring the difference in potential. Normally the reference point is understood to be the chassis or circuit common. However, it is worth remembering that if the other lead of your voltmeter is connected to some other point in the circuit the readings will likely be different. Voltage is not a property of only a single circuit node, but rather is a measure of the *difference* in potential between two points. This is why voltmeters have two leads.

Unfortunately, the circuit common connection is conventionally called the "ground" which leads people to believe that it must be connected to the Earth for proper operation. In fact, the voltage with respect to Earth has no effect on the circuit as long as all the differences of potential within the circuit are correct. For example, a low-pass filter may need to be "grounded" to the chassis or circuit common connection for proper operation but it does not need to be "grounded" to the Earth.

The simplest way to calibrate an inexpensive meter is simply to use a more accurate meter. For example, when constructing a home-built power supply, the analog panel meter may be calibrated with a digital voltmeter (DVM) because most DVMs have better accuracy than an analog meter. Perhaps the most practical voltage reference for the home workshop is an integrated circuit voltage reference. Special circuit techniques in the IC are used to generate a very stable, low-temperature-coefficient reference based on the band-gap voltage of silicon, approximately 1.25 V, on the chip. Inexpensive devices with specified accuracy of 0.1% and better are available from companies such as Analog Devices, Linear Technology, Maxim and National Semiconductor.

CURRENT

Conventional current is the flow of positive charge and is considered to flow in the direction of positive to negative voltage, or from the higher voltage point in a circuit toward the lower voltage point. The flow of negatively-charged electrons in the opposite direction from negative to positive voltage is sometimes called *electronic current*. Electronic and conventional current are equivalent, but flow in opposite directions. Conventional current is used in most circuit design and is reflected in the direction of the arrow in the symbol for diodes and bipolar transistors. An *ammeter* is an instrument for measuring current. An ammeter may be calibrated by measuring the attractive force between two electromagnets carrying the current to be tested, but in practice it is easier to place a known resistor in series, measure the voltage drop across the resistor, and calculate the current using Ohm's law. Resistance and voltage can both be calibrated accurately, so that method gives good accuracy.

The most common analog ammeter type is the *D'Arsonval galvanometer* in which the pointer is attached to a rotating electromagnet mounted between the poles of a fixed permanent magnet. The modern form of this meter was invented by Edward Weston and uses two spiral springs to provide the restoring force for the pointer, providing good scale linearity and accuracy. Most commonly-available meter movements of this type have a full-scale deflection between about 50 μA and several mA.

Digital ammeters use an analog-to-digital converter to measure the voltage drop across a low-value resistor and scale the result to display as a value of current. They are generally more accurate than analog meters and are more rugged due to the lack of delicate moving parts.

RESISTANCE

Measuring resistance requires one or more precision resistors to use as a reference, either in a bridge circuit or as part of an ohmmeter as described below. Resistors with a 1% rating cost only a few pennies and are readily available in a wide variety of values from many suppliers. *Precision* resistors are generally considered to be those with a tolerance of 0.1% or less and are typically available for less than a dollar or so in small quantities. Expect to pay a few dollars each for parts with a 0.01% tolerance.

An *ohmmeter* is a meter designed for measuring resistance. Various circuits can be used, but most are variations of the simplified schematics in **Fig 25.1**. In the circuit at A, the battery is in series with the resistor under test. If the resistance is zero (the test leads are shorted) then the battery is in parallel with the voltmeter and it reads a maximum value. If the resistance is infinite (test leads not connected) the meter reads zero. If the resistor equals R, the internal reference resistance, the meter reads half-scale.

In the circuit at B, the resistor under test is in parallel with the voltmeter. The meter indication is reversed from the series-connected circuit, that is, the meter reads zero for zero resistance, full-scale for infinite resistance, and mid-scale when the resistor equals R. In both circuits, an adjustment is normally provided to set the meter to full scale with the test leads shorted or open, as appropriate. In addition, a switch selects different values of the internal resistance R for measuring high or low-valued resistances.

A *Wheatstone bridge* is a method of measuring resistance that does not depend on the accuracy of the meter. Each arm of the bridge ($R_1 - R_S$ and $R_2 - R_X$) forms a voltage divider. In **Fig 25.2**, the meter is a zero-center type so that it can read both positive and negative voltages between the center connections of the two voltage dividers. When variable resistor R_S in Fig 25.2B is adjusted for a zero reading on the meter, then the two arms of the bridge have the same ratio,

$$\frac{R_1}{R_S} = \frac{R_2}{R_X}$$

The resistor under test, R_X, can be calculated from

$$R_X = R_S \frac{R_2}{R_1}$$

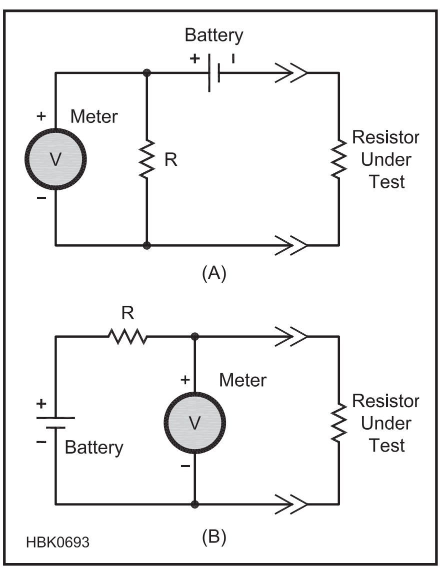


Fig 25.1 — Two ohmmeter circuits. At A, the meter reads full scale with a zero-ohm resistor and reads zero with no resistor connected. The circuit at B is the opposite; the meter reads full scale with no resistor connected (infinite resistance) and reads zero with zero resistance.

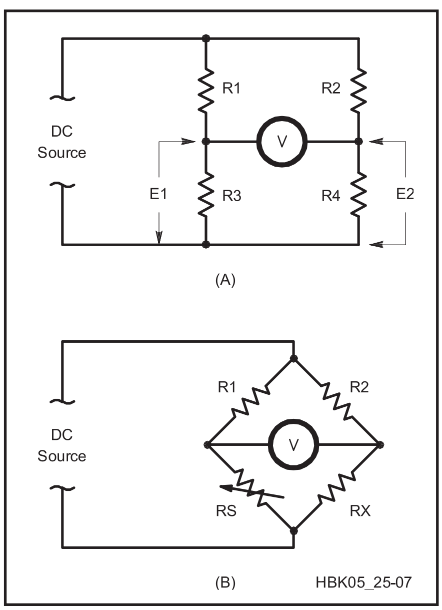


Fig 25.2 — A Wheatstone bridge circuit. A bridge circuit is actually a pair of voltage dividers (A). B shows how bridges are normally drawn.

Nowadays, the Wheatstone bridge is rarely used for measuring resistance since it is not as convenient as an ohmmeter, but the concept is important for several other types of measurement circuits that will be covered later.

25.2.2 Multimeters

A *multimeter* is probably the single most useful test equipment for an electronics experimenter. Besides measuring dc voltage and current, most also measure resistance and low-frequency ac voltage as well. Another common feature is a tone that sounds whenever the test leads are shorted, which is useful as a quick continuity tester. With most modern multimeters, the voltage they use for the resistance measurement is too low to forward-bias the junction of a silicon diode or transistor, which is a desirable characteristic for measuring in-circuit resistance. Many multimeters also have a special "diode" mode with a higher test voltage so that diodes and transistors can be tested as well. It is increasingly common for even low-cost digital multimeters to include functions such as frequency and capacitance measurement.

ANALOG MULTIMETERS

Nearly all analog multimeters use a D'Arsonval meter for the display. Compared to a digital display, analog meters have the advantage that it is easier to see whether the reading is increasing or decreasing as adjustments are made to the circuit under test. The experimenter may want to consider owning an inexpensive analog multimeter for that kind of testing as well as a higher-quality digital instrument for more precise measurements.

A *volt-ohm meter* (VOM) is a basic analog instrument that includes no electronic circuitry other than a switch and resistors to set the scale, a battery for the resistance-measuring circuit, and perhaps a diode to convert ac voltage to dc. Despite the name, most can also measure current as well as voltage and resistance.

A disadvantage of the VOM is that, when measuring voltage, the current to operate the meter must be drawn from the circuit under test. A figure of merit for a VOM it its *ohms-per-volt* (Ω/V) rating, which is just the reciprocal of the full-scale current of the meter movement. For example, if the VOM uses a meter that reads 50 μA full-scale then it has $1 / 50 \times 10^{-6} = 20{,}000$ ohms per volt. On the 1-V scale the meter has a resistance of 20 kΩ and on the 10-V scale, it is 200 kΩ.

Depending on the circuit being tested, drawing 50 μA may be enough to disrupt the measurement or the operation of the circuit. To solve that problem, some analog meters include a built-in amplifier with high input impedance. In the days of vacuum tubes, such meters were called *vacuum-tube volt-*

Fig 25.3 — This classic Hewlett-Packard HP412A vacuum-tube voltmeter (VTVM) has specifications that put many modern solid-state multimeters to shame.

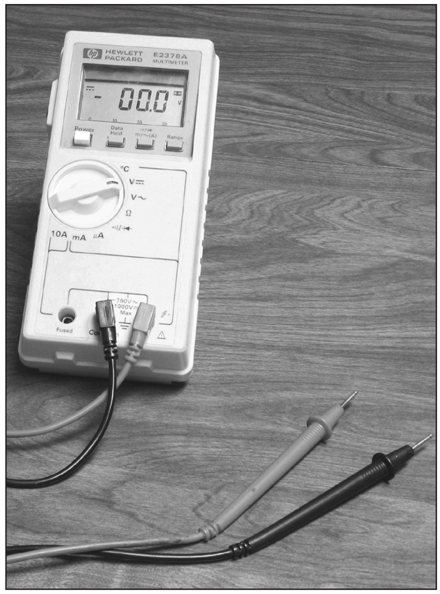


Fig 25.4 — A modern digital multimeter typically has a liquid crystal display readout.

meters (VTVM). An example is shown in **Fig 25.3**. The modern equivalent is called an *electronic voltmeter* and generally uses an amplifier with field-effect transistors at the input. Again, despite the name, VTVMs and electronic voltmeters usually can measure current and resistance as well.

DIGITAL MULTIMETERS

Instead of using an analog meter to display its measurements, a *digital multimeter* (DMM) has a digital readout, usually an LCD display. A microprocessor controls the measurement process and the display. (See **Fig 25.4**.) To convert the analog voltage or current being measured to a digital number requires an analog-to-digital converter (ADC) controlled by a microprocessor. Most use a dual-slope type of ADC, which trades off a relatively slow measuring speed for excellent accuracy and low cost. (Analog-to-digital conversion is discussed in the **Analog Basics** chapter.)

A digital voltmeter is constructed as shown in **Fig 25.5**. The input section is the same as in an analog electronic voltmeter. A range

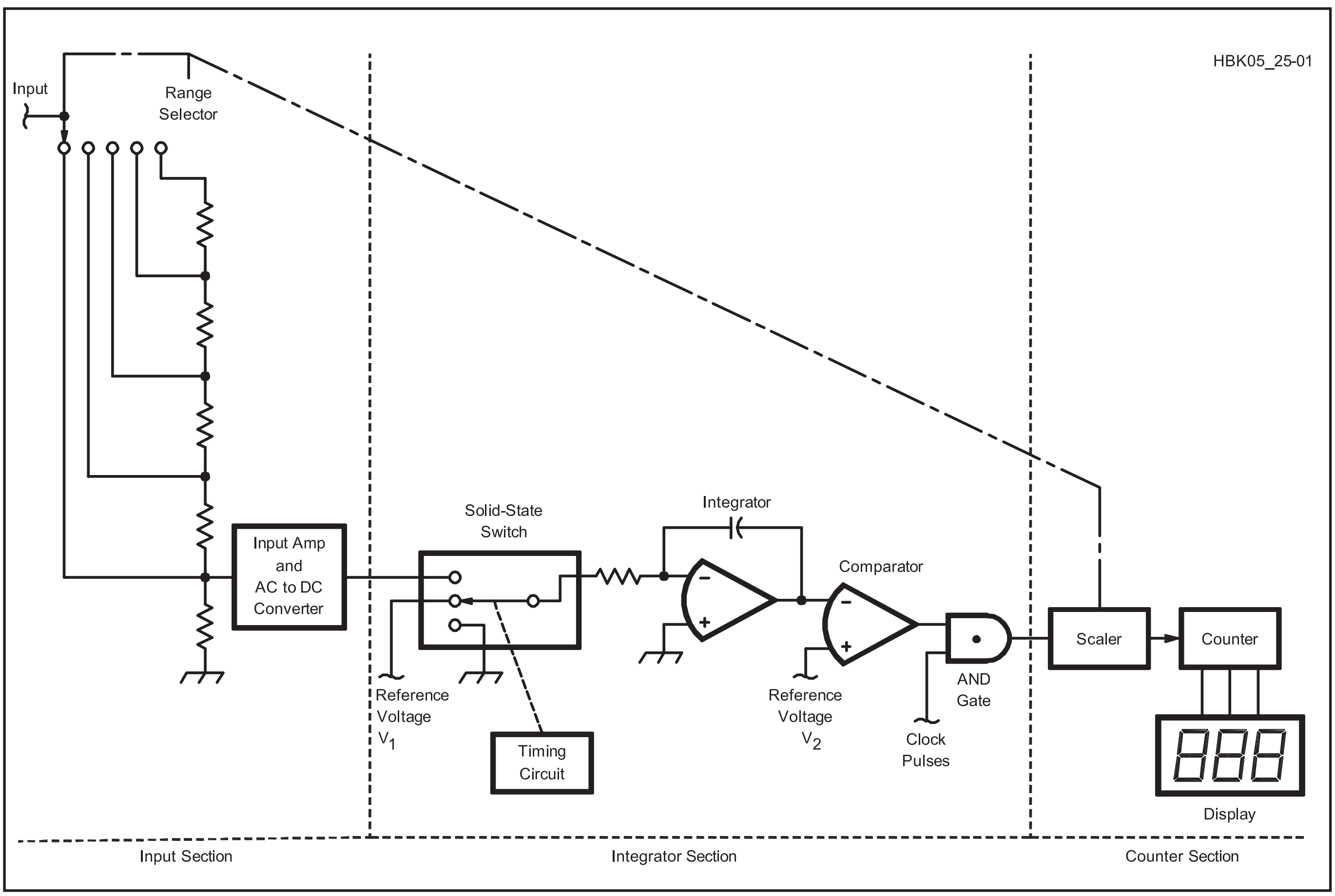


Fig 25.5 — A typical digital voltmeter consists of three parts: an input section for scaling, an integrator to convert voltage to a pulse whose width varies with voltage, and a counter to measure the width of the pulse and display the measured voltage.

selector switch scales the input signal appropriately for amplification by an input preamplifier, which also includes a rectifier circuit that is used when in ac mode to convert the ac signal into dc, suitable for conversion by the ADC.

An additional feature that becomes possible with digital multimeters is *autoranging*. The selector switch is used only to choose between voltage, current, resistance and any other available functions. The scale or range is selected automatically based on the amplitude of the signal being measured, which is a nice convenience. If the signal is fluctuating such that it causes frequent range changes, there is usually a way to turn off auto-ranging and select the range manually.

Digital multimeters that feature a serial data interface can also act as a *data logger*, taking and storing measurements for use by a PC or under the control of a PC.[2] This is a very useful feature for experimenters.

HOW TO USE A MULTIMETER

When preparing to make a measurement, the first thing to do is to select the proper function on the multimeter — voltage, current, resistance or some other function. If the meter is not an auto-ranging type, set the range to the lowest value that does not over-range the meter. If you're not sure of the voltage or current being measured, start with a high range, connect the test probes then switch the range down until a good reading can be obtained. Choosing the correct function and range may involve connecting the test leads to specific connectors on the meter.

When measuring high voltage, special precautions must be taken. As little as 35 V should be considered dangerous because it can produce lethal current in the human body under some conditions. Grasp the test probes by the insulated handles, being careful to keep fingers away from the metal probe tips. Pay attention to the multimeter manufacturer's maximum voltage ratings. Special test probes are generally available for measuring high voltage. Do not exceed the meter's rated maximum voltage.

High current can be dangerous as well. If a probe accidentally shorts a power supply to ground, sparks can fly, damaging the equipment and endangering the operator. Be careful of metal jewelry such as rings and bracelets. If connected across a high-current circuit you could get a nasty burn. Most meters have a fuse to protect the instrument from an over-current condition in current-measuring mode. If the multimeter turns on but always reads zero, consult the manual on how to replace the fuse.

To measure a current, the meter must be inserted in series with the circuit. In a series-connected circuit, all components carry the same current, so it doesn't matter which component is disconnected to allow inserting the ammeter. Select the one that is most convenient, or the one that is at a low-voltage point if you're measuring a high-voltage circuit. For measuring ac power circuits without disconnecting them, clamp-on current probes are used.

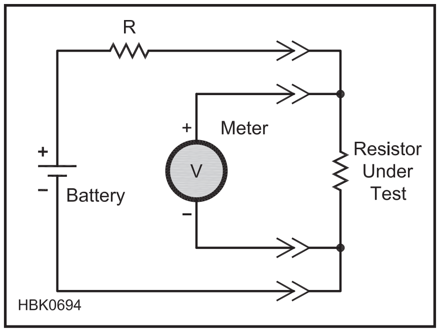


Fig 25.6 — The four-wire technique for measuring low-value resistors. By connecting the current source and the meter separately to the leads of the device under test, the error due to lead resistance is reduced.

It can be difficult to measure low-value resistors accurately because of the resistance of the meter leads, typically one or two tenths of an ohm. With the series-connected ohmmeter circuit described in the previous section, the calibration procedure compensates for that and some digital meters may have a means to compensate as well. However, accurate low-resistance measurement requires the *four-wire* technique, as illustrated in **Fig 25.6**. Two wires are connected to each end of the resistor, one to carry the test current and one to read the voltage with a high-input impedance voltmeter. In that way the resistance of the wires does not affect the measurement.

A similar technique can be used to measure high currents. Most inexpensive meters do not have a high-enough current range to measure the 20 A or so that is drawn from a 12-V power supply by a typical 100 W transceiver. The solution is to use an external *meter shunt*, which is a low value resistor placed in series with the current. See **Fig 25.7**. The multimeter reads the voltage, E, across the shunt, and then the current is calculated from Ohm's Law, $I = E/R_S$, where R_S is the resistance of the shunt. Resistors designed for this service may have four leads rather than two to allow a true four-wire measurement. The clamp-on current probes mentioned earlier may also be used.

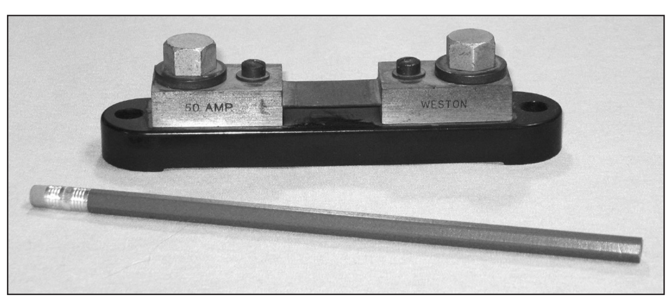


Fig 25.7 — This 50-A, 50-mV current shunt has a resistance of 0.05 / 50 = 0.001 Ω. The two large terminals are for connecting to the circuit under test and the two small terminals are for connecting to a voltmeter.

There are several issues that apply specifically to analog multimeters. An analog meter movement is a rather delicate device. It can be damaged if the case is dropped or mechanically shocked. When transporting the instrument, it is a good idea to place a low resistance or short in parallel with the terminals of the meter movement, which increases damping and reduces the amount of pointer movement as the case is bounced around. Some multimeters have a special OFF position on the selector switch for that purpose. On others, switching to the highest current range accomplishes the same thing.

Most D'Arsonval meters have an adjustment screw located near the pointer's pivot point that may be accessed from the front of the meter. It should be adjusted so that the meter reads zero with no signal applied.

Some older VOMs include a high-voltage battery that is used on the highest resistance ranges. When testing solid-state devices, that voltage can be high enough to cause damage. If there is any doubt, use another high-impedance multimeter to test the voltage on the test leads of the VOM when it is set to the highest resistance ranges.

MULTIMETER CRITICAL SPECIFICATIONS

The first decision when selecting a multimeter is whether you want an analog or digital type. That is largely a matter of personal taste. Digital meters are generally more accurate but analog meters may make it easier to tune a circuit for a specific voltage or current. The next decision is between a hand-held or bench-type instrument. The latter tend to have more features and better specifications, but obviously are less portable and usually are more expensive.

The most obvious selection criteria are the features provided. Nearly all multimeters measure dc voltage, current and resistance and most also measure low-frequency ac voltage. Other common features on digital meters include autoranging and automatic turn-off to

Probe Adapters for Multimeters

Multimeters come with test probes intended for precise contact with terminals, components, wires, and so forth. They work well if the item to be probed is easily exposed or otherwise available to the probe. Measuring signals on connector pins, however, is often a challenge. Inserting a probe into the miniature sockets on many connectors is often not possible and if the connector has exposed pins, trying to insure the probe does not slip to or between adjacent pins is nearly impossible. The solution is to build an adapter as shown in Figure 25.A's three examples.

Fig 25.A1 provides a convenient way to hold probes steady in a spring-loaded, push-button, two-wire speaker connector connected to a pair of Powerpole connectors. Keep the colors of the wires, buttons, and connectors consistent to prevent confusion. An enclosure such as an inexpensive plastic box protects the exposed terminals.

Fig 25.A2 shows how to adapt an automotive-style fuse to make a current-measuring adapter for your multimeter. First, remove the back of the fuseholder and the fuse element and pry the fuseholder open. Solder leads to the exposed terminals then glue or snap the fuseholder back together. Be sure to include the external in-line fuseholder — available in auto parts stores — so that the circuit is protected and you don't blow an expensive multimeter fuse. (Note — multimeter fuses are rated at the full voltage limits of the meter for your safety. Do not replace them with low-voltage fuses.)

Fig 25.A3 is a typical adapter for a multi-pin connector using a terminal strip. Take care to arrange the terminals in order of pin number and label them so you don't have to guess when using the adapter. Make an adapter for the common connectors in your station and you'll never regret it!

These are just three types of adapters — you will no doubt think of many more that will help you with your particular needs. Remember to protect yourself against exposed voltages and short-circuits when constructing and using the adapters. (Thanks to W4QO and KG4VHV and the *QRP Quarterly* for the suggestions.)

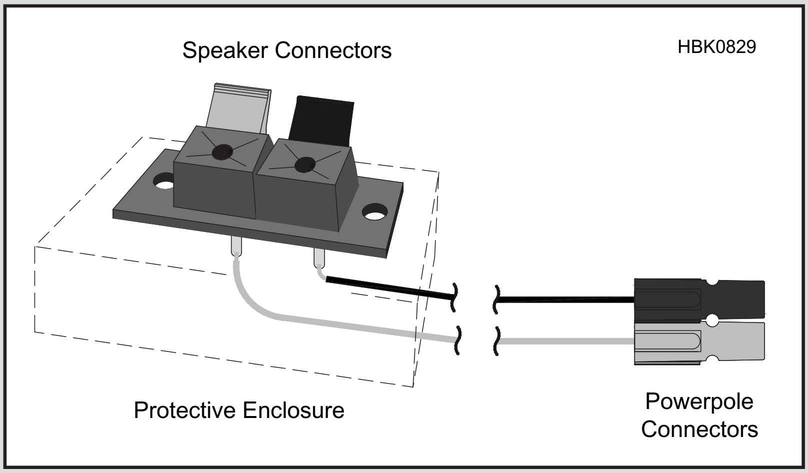


Fig 25.A1 — A convenient way to hold probes steady

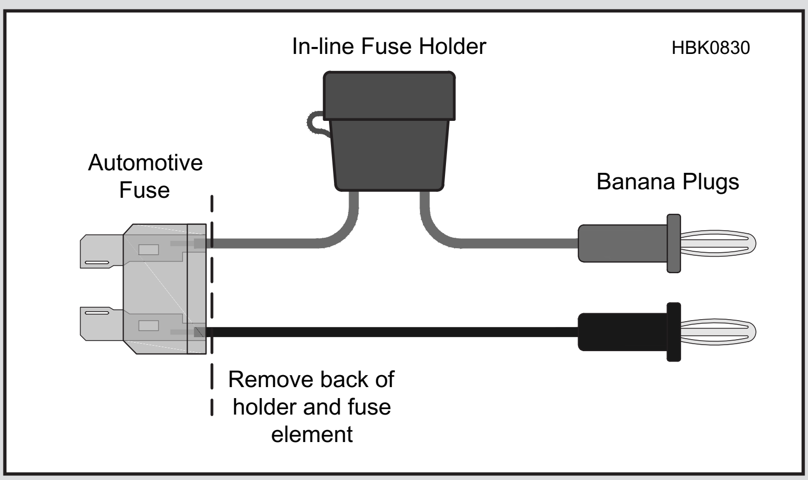


Fig 25.A2 — Adapting an automotive-style fuse to make a current-measuring adapter.

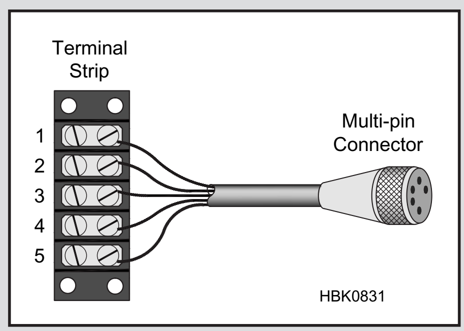


Fig 25.A3 — A typical adapter for a multi-pin connector.

save the battery. More capable meters include features such as data hold, peak voltage, true RMS voltage, 4-wire resistance, capacitor tester, inductor tester, diode and transistor tester, logic tester, frequency counter, computer data logging and a graphical display.

One important feature that is often not specified is over-voltage and over-current protection. On analog meters, there may be internal back-to-back diodes across the meter movement to protect it from over-voltage. Most digital meters include autoranging which should prevent damage from any voltage below the specified maximum. The current-measuring input on both types is normally protected with a fuse.

The next most critical specification is the available measurement ranges of the voltage, current, resistance and any other functions provided. Analog meters tend to space the voltage and current ranges by a factor of three. For example, the full-scale voltage readings might be 3 V, 10 V, 30 V, 100 V, 300 V or perhaps 5 V, 15 V, 50 V, 150 V, 500 V. The latter is a particularly good choice since two voltages that are commonly measured, 12 V dc and 120 V ac, are near the top of a range where accuracy is best. The more range selections provided, the greater the span of values that can be measured. With auto-ranging instruments the measurement minimum and maximum range may not be obvious without consulting the manual.

The input impedance is important in minimizing the effect of the multimeter on the circuit under test. Impedance should be high for voltage measurements and low for current measurements. For VOMs the figure of merit for voltage measurements is the ohms-per-volt rating. Multiply the full-scale voltage range by ohms-per-volt to get the input resistance. Both digital and analog electronic voltmeters usually have the same input impedance on all voltage ranges, typically between 1 and 11 MΩ. The input resistance for current measurements is often specified by the *burden voltage*, which is the voltage drop across the test leads with a full-scale signal. Typical values vary widely, from a few millivolts to more than a volt. The voltage drop can often be reduced by switching to a higher current range, at the expense of measurement resolution.

Measurement accuracy can be an important specification for some applications. An inexpensive analog meter may have voltage and current accuracy of 3% or so. The best analog meters have accuracy specifications in the range of 1%. At that level, accuracy may be limited by *parallax*, which causes the meter reading to appear to change as you change the angle of view. To mitigate that, some high-end analog meters have a *mirror scale*. The reflection of the pointer in the mirror has a parallax error equal and opposite to the unreflected pointer so that the correct reading is half way between the two.

You can't necessarily tell the accuracy of a digital meter from the number of digits in the display. Usually the accuracy is limited by the analog circuitry. However, the number of digits defines the resolution, which can be important when comparing two nearly-equal readings or when tuning a circuit for a peak or minimum value. A typical DMM might have a specified accuracy of 0.1% to 1% for dc measurements and perhaps 1% to 3% for ac voltage. (See the section on ac measurements.) Many inexpensive digital multimeters do not have published specifications and may not be very accurate.

Many bench-type multimeters and some hand-held units can be connected to a computer. That allows the computer to control the instrument to take automated readings and store the results in a computer file. Some older test equipment may have a GPIB (general purpose interface bus) interface, also known as IEEE-488 (HPIB on Hewlett-Packard equipment). GPIB-to-USB converters are available to allow connection to a PC.[3] Modern instruments typically have a USB or RS-232 interface.

USED AND SURPLUS MULTIMETERS

Fig 25.3 shows a typical multimeter that might show up at a ham flea market or on the Internet, a Hewlett-Packard model 412A VTVM. While obsolete, it is still a very useful instrument with capabilities that are rare in modern units. For example, it can measure dc voltage down to 1 mV (0.001 V) full scale and the burden voltage for current measurements is only 1 mV for currents up to 10 mA, rising to 100 mV at 1 A. The highest resistance range is 100 MΩ center scale, which allows reasonably-accurate resistance measurements up to about 1 GΩ. Don't overlook surplus equipment just because it is old. Some of it is a real bargain.

When buying a used analog multimeter, the most important thing to check is the meter movement itself. D'Arsonval meters are rather delicate and easy to damage. For a VOM or battery-operated multimeter, put the instrument in resistance-measuring mode then short and un-short the test leads. The needle should move smoothly between zero and full scale. If you can't do that test, at least make sure that the needle rests close to zero. Rotate the instrument back and forth to make the needle move and observe whether it appears to bind. It is difficult or impossible to repair a damaged meter movement.

Digital multimeters are less delicate than analog meters and many of their failures are electronic in nature so they sometimes can be repaired using normal troubleshooting techniques. Of course it is always a good idea to test used equipment before buying, if possible. If at all possible, measure voltage and resistance with the meter to be sure it is fundamentally sound.

25.2.3 Panel Meters

Analog panel meters are quite expensive to buy new so many experimenters keep an eye open for flea-market bargains. You often can find old "boat anchor" equipment with good salvageable panel meters selling for less than the value of the meters.

The scale markings on surplus meters often represent what the meter was measuring in the equipment rather than the actual current flowing through the meter itself. Sometimes the full-scale current of the meter movement will be shown in small text at the bottom of the scale. However, that may not be the same as the current measured at the meter terminals because some meters include an internal shunt.

The only sure way to know the full-scale current and resistance of a surplus meter is to measure it. The resistance can be measured with the ohmmeter function of a multimeter, but be careful. On the lowest resistance ranges, the multimeter may output enough voltage and current to damage the meter under test. If the multimeter is an auto-ranging type you have no way to control test current unless you can turn auto-ranging off. With a non-autoranging ohmmeter, start the measurement at the highest resistance scale and then reduce the scale one step at a time until a valid reading is obtained, while keeping an eye on the meter under test to be sure it is not over-ranged.

A safer way to measure both the full-scale current and resistance of a panel meter is to place a high-value resistor in series with it and connect the combination to a dc power supply, perhaps a battery. A 1.5-V battery in series with a 100 kΩ resistor (15 µA of current) is a good starting point. Keep trying smaller and smaller resistance values until a good reading is obtained on the meter. Assuming the scale that is marked on the meter's scale face is linear, the full-scale current is

$$I_{FS} = I_{TEST} \frac{D_{FS}}{D_{TEST}}$$

where I_{FS} is the full-scale meter current, D_{FS} is the scale's full-scale marking, D_{TEST} is the needle indication measured with the test current, and I_{TEST} is the test current, which is equal to the voltage across the resistor divided by the resistance. The meter's resistance is the voltage across the meter divided by I_{TEST}.

USING PANEL METERS

Whether surplus or new, it is rare that a panel meter measures exactly what you need for a particular application. Usually you must change the current or voltage sensitivity.

To increase the full-scale current, place a *current shunt* in parallel with the meter. This is simply a resistor whose value is

$$R_{SHUNT} = R_M \frac{I_M}{I_{FS} - I_M}$$

where R_M is the meter resistance, I_M is the meter full-scale current and I_{FS} is the desired full-scale current reading. The shunt resistance is very small for high-current shunts, such that the resistance of the wires or circuit traces can cause a significant error. To reduce that error, connect the meter directly to the leads of the shunt, with no wires or circuit traces in common with the high-current path.

You can make a low-value shunt by wrapping a length of copper wire around a resistor or other component used as a form. Unfortunately, however, the resistance of copper has a poor temperature coefficient, typically around 0.4 percent per degree C. That means the meter reading can change more than 10% between a warm and a cold day. As the wire self-heats from the high current flowing through, the meter reading can easily be in error by 20% or more. Commercial shunts are made from a metal with a low temperature coefficient such as nichrome. Copper-wire shunts should only be used where accuracy is not important. (A table of wire resistance in ohms per foot (Ω/ft) is available in the **Component Data and References** chapter.)

If the panel meter is to be used to measure voltage, a *voltage multiplier* resistor is inserted in series with the meter. The value is

$$R_{MULT} = \frac{V_{FS}}{I_M} R_M$$

where V_{FS} is the desired full-scale voltage, I_M is the meter full-scale current and R_M is the meter resistance. If the meter has an internal current shunt, it should normally be removed to maximize the value of the multiplier resistor. For high-voltage applications, be aware that in addition to a power rating a resistor also has a working voltage specification, perhaps 200 to 250 V or so for a typical ¼ W, through-hole resistor. Applying voltages higher than the rating — even if the rated power dissipation is not exceeded — can result in arcing across the body of the resistor. If you need to measure a voltage higher than the voltage rating, use several resistors in series. For example, to measure a 2000-V power supply, ten ¼ W resistors in series, each with a value of one-tenth the desired resistance, would be suitable.

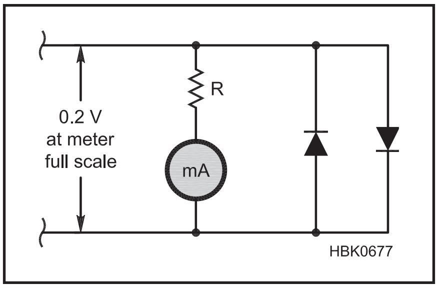


Fig 25.8 — Back-to-back silicon diodes protect the meter by limiting the maximum voltage. See the text for a discussion of how to select the value of R.

If you intend to use the same panel meter for several different purposes in your project, be sure to use a *break-before-make* switch to make the selection. That protects the meter by making sure it is never connected to two circuits at the same time, even for an instant.

Analog D'Arsonval meters are easily damaged if subjected to excessive current. A standard technique to protect them is to wire back-to-back diodes in parallel with the meter. Silicon junction diodes have the property that they act like an open circuit for low voltages and start to conduct when the forward voltage reaches about 0.4 V. Most meters can withstand up to two times the full-scale current without damage. So choose resistor R in **Fig 25.8** such that the voltage across the diodes is about 0.2 V when the meter current is at full scale, that is

$$R = \frac{0.2\ V}{I_M} - R_M$$

where I_M is the full-scale meter current and R_M is the meter resistance. If that equation results in a negative value for R, replace each diode with two diodes in series and recalculate using 0.4 V instead of 0.2 V in the equation.

MAKING NEW PANEL SCALES

It seems like the scale that is printed on the meter scale is never what you need for your project. You can usually disassemble the meter to remove the scale plate to modify the scale. Be sure you are working in a clean environment and temporarily reassemble the meter while you are working on the scale to prevent any dust or other contaminant from entering the delicate mechanism. Take particular care that magnetic metal debris cannot get into the meter movement where it is attracted by the permanent magnets.

There are several methods for modifying the scale markings for your own purposes. If only the labels are wrong and the tick marks are spaced appropriately, you can add new labels with a permanent marker or dry-transfer labels. If the old labels are not useful, sometimes they can be removed with a pencil eraser. If you need a completely new meter scale, one old trick is to turn the scale plate over and draw the new scale on the back. However, software is available to design and print custom scales. An Internet search will quickly find dozens of low-cost and free programs, such as *Meter Basic* by Jim Tonne, W4ENE, which is included on the CD-ROM that comes with this book.

DIGITAL PANEL METERS

Digital panel meters (DPMs) are available as preassembled modules that are almost as easy to use as analog meters. The displays are generally of the liquid crystal type, with or without a backlight, and typically have 3 to 4½ digits. A "½" digit is one that can display only a 1 or a blank. Displays with a half-digit usually have a full-scale input voltage of either 2 V or 200 mV minus one count, so that the full-scale voltage is 199.9 mV, for example. Most have programmable decimal points after each digit and some have indicators to indicate the units, such as µ, m, V, A and so forth. DPMs have a high input impedance so there is minimal loading on the circuit under test.

The required power supply voltage varies by model. Some require a floating supply, so if the power supply and the voltage being measured need a common ground connection, be sure the meter is capable of that.

Accuracy is typically 0.1% or better. The total accuracy is usually limited by the external circuitry that drives the meter, such as the amplifier, current shunt or attenuator that is required to get the signal within the input voltage range of the DPM.

25.3 AC Measurements

This section covers issues that affect all ac signals, while the following section on RF measurements concentrates on aspects of ac measurements that are particular to the higher frequencies.

25.3.1 Basic Vocabulary and Units

The **Electrical Fundamentals** chapter includes an introduction to basic electrical units of ac waveforms. The following material emphasizes some additional points that have particular relevance to test and measurement.

AC WAVEFORM VALUES

With ac signals, the voltage and current change periodically with time so, as you might expect, there are several ways to express their value. See **Fig 25.9**. The *average* voltage or current is the value averaged over one period and is equal to the *dc component* of the signal. For a symmetric, periodic ac signal with no dc component, the average is always zero. Non-periodic signals must be averaged over a long period of time to obtain a valid average.

The *peak* value of an ac signal is just as its name implies, the maximum value that the signal ever achieves. The *peak-to-peak* value is the difference between the positive and negative peaks. For a symmetrical-ac waveform such as a sine or square wave, the peak-to-peak is twice the peak.

The *RMS* (*root-mean-square*) value of voltage or current is that which would produce the same heating in a resistor as a dc voltage or current of the same value. For a sine wave, it is

$$V_{RMS} = \frac{1}{\sqrt{2}} V_{PK} \approx 0.707 \times V_{PK}$$

where V_{RMS} is the RMS voltage and V_{PK} is the peak voltage. A similar equation applies for RMS current. Don't forget that the equation only applies for sine waves. For example, the RMS voltage of a square wave is 1.0 times the peak. See **Fig 25.10**. Also be aware that the RMS value includes the effect of any dc component. If you wish to refer to the RMS value of the ac component only, be sure to state that explicitly.

The formulas for power apply for RMS voltage and current as well as for dc, $P = E_{RMS} I_{RMS}$, $P = E_{RMS}^2 / R$ and $P = I_{RMS}^2 R$, where P is power in watts, E_{RMS} is RMS potential difference in volts and I_{RMS} is RMS current in amperes. However, that assumes that the voltage and current are in phase, as in a resistor or an antenna at resonance. If they are not in phase, then the power in the above equations must be corrected by multiplying it by the *power factor*,

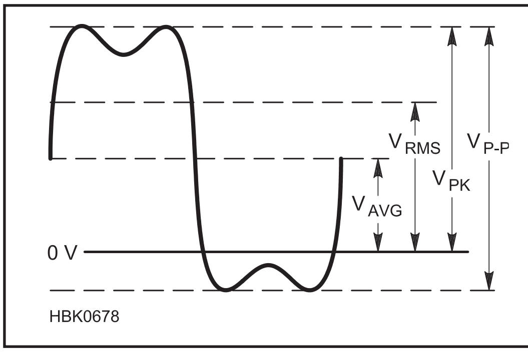


Fig 25.9 — Various methods to express the voltage or current of an ac signal. The average value, V_{AVG}, is the dc component. V_{RMS} is the root-mean-square, V_{PK} is the peak and V_{P-P} is the peak-to-peak value.

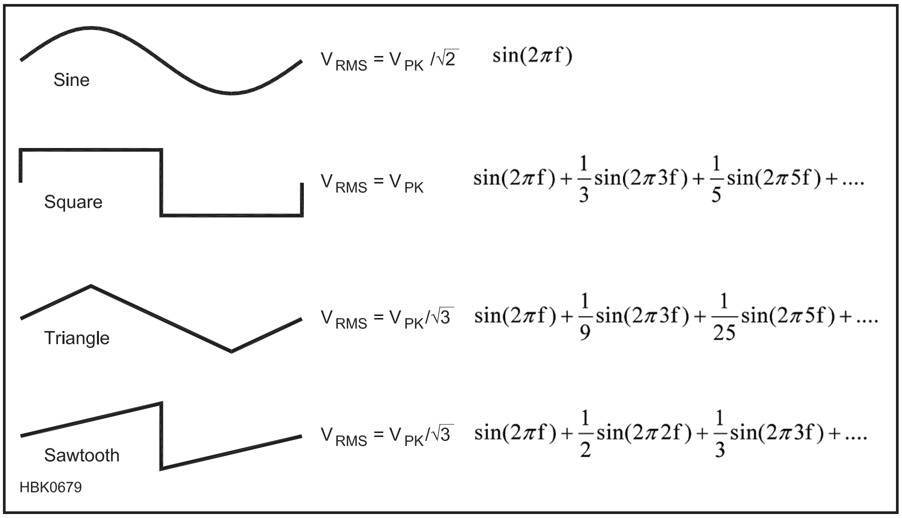


Fig 25.10 — Typical periodic ac waveforms seen in electronic circuitry. A single cycle of each example is shown. At the center are listed the relationships between V_{RMS}, the root-mean-square, and V_{PK}, the peak value, assuming the waveform has no dc component. Periodic waveforms are composed of sine waves at the fundamental and harmonic frequencies. At the right are the relative amplitudes of the various frequency components.

$$PF = \cos\theta$$

where θ = arctan (X / R), the phase angle caused by the reactance, X, and the resistance, R, of the circuit.

25.3.2 Using Multimeters for AC Measurements

Most multimeters can indicate the RMS value of ac voltage and current, but many do not measure the RMS value directly. Instead they measure the rectified average or peak voltage and then apply a correction factor so that the display reads RMS, assuming a sine wave. Unfortunately, that means that the RMS values are not accurate if the ac signal being measured does not have a sinusoidal waveform.

Some meters full-wave rectify the ac signal and then measure the average of that, internally correcting for the difference between the average of a rectified sine wave

$$V_{AVG} = \frac{2}{\pi} V_{PK} = 0.637 \times V_{PK}$$

and the RMS value

$$V_{RMS} = \frac{1}{\sqrt{2}} V_{PK} \approx 0.707 \times V_{PK}$$

so that the reading is in RMS. Very inexpensive analog meters may only use a half-wave rectifier which causes RMS readings for asymmetric waveforms to vary with the orientation of the test connections.

Additional considerations may apply to RMS readings. For example, the accuracy of the RMS reading for most meters varies with frequency of the applied signal. Check the specifications of the multimeter for the frequency range over which it may be used to measure RMS values.

The only way to accurately measure RMS values of non-sinusoidal signals is with a meter that has *true RMS* capability. Such a meter uses circuitry or software to compute the RMS value of the signal. Note that the measurement bandwidth of the meter must include the significant harmonics of the signal as well as the fundamental in order to give accurate RMS readings.

An example of a measurement that requires a true RMS voltmeter is receiver sensitivity. For that, you need to measure signal and noise levels at the receiver audio output. Standard multimeters using a rectifier and averaging circuit give inaccurate results when measuring noise because noise and sine waves have different peak-to-RMS ratios. Another advantage of true RMS meters is that they tend to have better scale linearity, even for sinusoidal signals. A diode detector is nonlinear, especially at the low end of the scale.

Frequency response is another limitation when making ac measurements with a multimeter. Most are specified from below 50 or 60 Hz, to cover power-line frequencies, up to a few hundred Hz. Many receiver measurements use a 1 kHz test frequency, so a meter specified up to at least that frequency is especially useful.

For all of these and other reasons, the ac accuracy is usually significantly worse than the dc accuracy. Generally, an oscilloscope makes more accurate ac measurements than a multimeter. Modern digital oscilloscopes often have built-in capability to indicate peak, average and true RMS voltage.

One final issue with multimeters is the ac impedance of the probes. While the dc input resistance of a modern electronic multimeter is typically over 1 MΩ, the capacitive reactance can be a significant factor at radio frequencies. Even if all you care about is the dc voltage, if ac signals are present, reactance of the probe can affect the circuit's operation.

25.3.3 Measuring Frequency and Time

FREQUENCY COUNTERS

The basic instrument for measuring frequency is the *frequency counter*. A block diagram of a very basic design is shown in **Fig 25.11.** Three digits are shown, but typically there are more. The signal to be measured is routed through three cascaded decade counters. For 1 Hz frequency resolution, the counters count for 1 second. For 10 Hz resolution, they count for 1⁄10 second, and so on. The count time is determined by a high-stability crystal oscillator, whose frequency is divided down to 1 Hz, 1⁄10 Hz, or whatever resolution is desired. At the end of each count time, control circuitry stores the final count in latches that drive the digit displays and at the same time resets the counters for the next count period.

With this scheme, the displays are updated once per second when 1 second resolution is chosen, 10 times per second with 10 Hz resolution, and so on. One issue is that if the frequency is part-way between two adjacent displayed values, the least-significant digit will flicker back and forth between the two values on successive counts. Sometimes the designer chooses not to show the least-significant digit for that reason, even though that slows down the display update rate by a factor of 10 for any given display resolution.

Some frequency counters include the ability to measure time as well. In the block diagram, the connections to the first divide-by-10 stage input and the control circuit input are swapped. In that way the signal being measured controls the count time, and the reference oscillator provides the signal being counted. If the divided-down reference oscillator has a frequency of 1 kHz, for example, then the period can be measured to a resolution of 1 ms. This same technique can be used to measure low-frequency signals as well. For example, when measuring the frequency of a subaudible tone encoder, you need at least 0.1 Hz measurement resolution. Normally, that would require a 10-second count time which can be inconvenient. Some counters are able to measure the period, calculate the reciprocal, and display the resulting frequency. Since the count time is only one cycle of the measured signal, the display updates in real time.

Many frequency counters include a *prescaler*, a frequency divider between the input and the main part of the circuitry, to allow operation at higher frequencies. Usually the prescaler has a 50-Ω input. For low frequencies, there is a separate high-impedance input, typically 1 MΩ, that bypasses the prescaler.

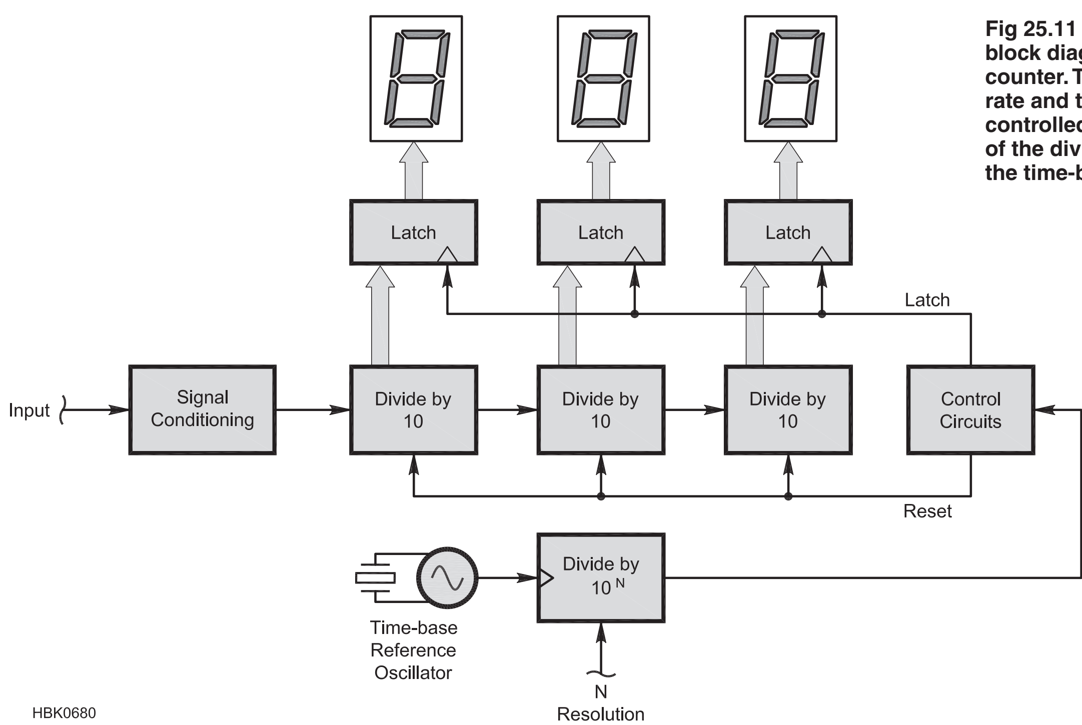


Fig 25.11 — A simplified block diagram of a frequency counter. The display update rate and the resolution are controlled by the divide ratio of the divider at the output of the time-base oscillator.

A switch selects between the two inputs.

It is important to realize that the so-called high-impedance input only has a high impedance at low frequencies. For example, if the stray input capacitance is 30 pF, then the impedance is only 177 Ω at 30 MHz. If you try to measure the frequency of an oscillator by connecting the frequency counter's input directly to the circuitry, it likely will alter the oscillator tuning enough to invalidate the measurement. If possible, connect the counter to the output of a buffer amplifier or at some other point in the circuitry that won't be adversely affected. If that isn't possible, another trick is to use a pickup coil placed near the oscillator. The coil could be a few turns of insulated wire soldered between the center conductor and shield of a coaxial cable that connects to the frequency counter input. Hold the coil just close enough to get a stable reading.

When connecting a frequency counter to a circuit, observe the maximum voltage and power ratings, both for dc and ac. An oscilloscope probe with a 10:1 attenuation ratio connected to the high-impedance input is a good method to reduce both the signal level as well as the capacitive loading.

Frequency counters tend to have sensitive inputs. Only a small fraction of a volt is typically enough for valid readings. It is quite practical to measure the frequency of a nearby transmitter off the air with a small whip antenna. The transmitter should not be modulated while measuring its frequency. SSB transmitters should be measured in CW mode.

The principle figure of merit for a frequency counter is its frequency accuracy, which is primarily determined by the reference oscillator, also known as the *time base*. The time base accuracy is affected by the temperature, power supply voltage, crystal aging and quality of calibration. For a temperature-compensated crystal oscillator (TCXO), the total accuracy is typically a few parts per million (ppm). If it is 10 ppm, for example, then the error at 144 MHz is 1.44 kHz.

Normally temperature is the factor with the greatest effect on the short-term stability. For best accuracy, calibrate the reference oscillator at the same temperature at which measurements will be taken.

Other important specifications are the number of digits in the display, frequency resolution, display update time, frequency range, input sensitivity, input impedance and, for portable units, power supply voltage and current. In choosing a frequency counter you'll need to decide if you want a desktop model or a portable handheld unit. Additional features to consider include the size and visibility of the display, the ability to measure period, a data hold feature, an external time base input, adjustable trigger level and polarity, input attenuator, switchable low-pass filter, and frequency ratio measurement.

Commercial frequency counters have become so common that it hardly pays to construct your own from scratch. Units with a wide range of prices, feature sets and performance levels are available both on-line and from local electronics distributors. Older used and surplus frequency counters tend to be less of a bargain than other types of test equipment because advances in solid-state electronics have made modern instruments inexpensive, lightweight and packed with features and performance.

FREQUENCY MARKER GENERATORS

Before the advent of digital frequency synthesizers, most shortwave amateur receivers and transceivers included a crystal calibrator, which is a low-frequency (typically 25 or 100 kHz) oscillator with strong harmonics throughout the HF spectrum, used to calibrate the analog dial on the radio. Early units used an actual 100 kHz crystal in the oscillator, but modern units use a higher-frequency crystal and a frequency divider to obtain the low-frequency signal.[4]

Most modern Amateur Radio equipment uses a crystal-controlled synthesizer or digital frequency display, so no crystal calibrator is needed. However, the idea can still be useful for testing homebrew gear. If the low-frequency signal consists of a series of narrow pulses rather than a square wave, then all the harmonics are of the same amplitude up to the point where the pulse width is a significant portion of a cycle. This can be approximated by placing a small-value capacitor in series with the output. The harmonics should have constant amplitude for frequencies below

$$f << \frac{1}{2\pi RC}$$

where f is the frequency in MHz, C is the capacitance in μF and R is the load resistance, usually 50 Ω.

WAVEMETERS

An *absorption wavemeter* is basically a tunable filter with some means of detecting the signal at the filter output. It allows crude spectrum analysis of a signal by manually tuning through the frequencies of interest. Commercial units designed for microwave frequencies often include a carefully-calibrated dial for reading the frequency with some precision. Typically a diode detector is used to indicate the output signal level. A wavemeter suitable for HF or VHF frequencies can be constructed with one or more coils and a variable capacitor.

DIP METERS

Old timers know this instrument as a grid-dip oscillator (GDO), so-called because the indicating "dip" was in the grid current of the vacuum-tube oscillator. Most dip meters these days are solid-state but the principle is the same. The oscillator coil is external, extending from the end of the instrument. A set of plug-in coils is provided to cover the unit's frequency range.

Along with resonant frequency measurements, the dip meter can also serve as a crude signal generator, capacitor and inductor meter, and antenna and transmission line tester, among other uses.[5] If you are purchasing a dip meter, look for one that is mechanically and electrically stable. On used units, the socket where the coils plug in is a common cause of intermittent operation. The coils should be in good condition. A headphone connection is helpful. Battery-operated models are convenient for antenna measurements.

If you hold the coil near a tuned circuit and adjust the dip meter tuning dial to the frequency of the tuned circuit, there is a dip in the meter reading as the resonant circuits interact with each other. To avoid detuning the circuit being tested, always use the minimum coupling that yields a noticeable indication.

Most dip meters can also serve as absorption wavemeters by turning off the oscillator and looking for a peak instead of a dip in the meter reading. Sometimes frequencies can be detected in this way that would be difficult to read on a frequency counter because of the presence of harmonics. Further, some dip meters have a connection for headphones. The operator can usually hear signals that do not register on the meter.

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 coil or a conductor with relatively high current is convenient. Maximum inductive coupling results when the axis of the pick-up coil is placed perpendicular to the current path and the coil is adjacent to the wire.

High-impedance circuits have high voltage and low current. 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. (For safety's sake, *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 strongest when the end of the pick-up coil is near a point of high impedance. In either case, the circuit under test is affected by the presence of the dip meter.

To measure resonance, use the following procedure. 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,

Table 25.1
Standard Frequency Stations
(Note: In recent years, frequent changes in these schedules have been common.)

Call Sign	*Location*	*Frequency (MHz)*
BPM	China	2.5, 5, 10, 15
BSF	Taiwan	5, 15
CHU	Ottawa, Canada	3.330, 7.850, 14.670
dcF	Germany	0.0775
HLA	South Korea	5.000
JJY	Japan	0.04, 0.06
MSF	Great Britain	0.06
RID	Irkutsk	5.004, 10.004, 15.004
RWM	Moscow	4.996, 9.996, 14.996
TDF	France	0.162
WWV	USA	2.5, 5, 10, 15, 20
WWVB	USA	0.06
WWVH	USA (Hawaii)	2.5, 5, 10, 15
ZSC	South Africa	4.291, 8.461, 12.724 (part time)

move the meter away from the circuit and confirm that the dip comes from the circuit under test (the depth of the dip should decrease with distance from the circuit until the dip is no longer noticeable). Finally, move the meter back toward the circuit until the dip just reappears. Retune the meter for minimum current and read the dip-meter frequency from the dial or with a calibrated receiver or frequency counter.

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, degrading the accuracy of the measurement. Increase the distance between the dip meter and the circuit until a shallow symmetrical dip is obtained.

FREQUENCY CALIBRATION

The best test equipment is of limited use if it is not well-calibrated. The traditional frequency calibration method is to zero-beat a crystal oscillator (or its harmonic) with a radio station of known frequency, preferably a standard frequency station such as WWV or WWVH. **Table 25.1** contains the locations and frequencies of some of those stations. A receiver is tuned to one of those frequencies and the oscillator is loosely coupled to the antenna. It may be necessary to use frequency multiplication or division to obtain a common frequency. The frequency difference between the two causes a *beat note*, a rapid variation in the strength of the tone received in the speaker that slows down as the frequencies are brought close together. Maximum beat-note modulation occurs when the off-the-air and oscillator signals are approximately equal in amplitude

While the transmitted frequencies from WWV and WWVH are highly accurate, better than 1 part in 10^{11}, after propagation via the ionosphere the received accuracy is significantly degraded by Doppler shift, typically to a few parts in 10^7. Also, due to fading of the received signal, it can be difficult to zero-beat the oscillator to better than about 1 Hz accuracy. Best results generally occur on the highest frequency that provides good reception.

VLF time standards and surplus rubidium standards can be used for frequency references.[6,7] The Global Positioning System (GPS) satellites offer further possibilities for very precise frequency calibration. Various companies sell *disciplined oscillator* units that correct the frequency using the cesium-clock-based signals from the GPS satellites. These can sometimes be found on the surplus market.[8] Amateur-level kits are also available or you can build one from scratch.[9]

25.3.4 Oscilloscopes

An *oscilloscope* ("scope" for short) is an instrument that displays voltage versus time on a screen, similar to 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 instrument to process signals for display.

ANALOG OSCILLOSCOPES

Fig 25.12 shows a simplified diagram of a triggered-sweep oscilloscope. At the heart of all analog scopes is a cathode-ray tube (CRT) display. An electron beam inside the CRT strikes the phosphorescent screen causing a glowing spot. Unlike a television CRT, an oscilloscope uses electrostatic deflection rather than magnetic deflection. The exact location of the spot is a result of the voltage applied to the vertical and horizontal deflection plates. 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 including the dc level, you would use the *dc-coupling* mode. In the *ac-coupled* mode, 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-V dc power supply. Note that you should not use ac coupling at frequencies below the low-frequency cutoff of the instrument in that mode, typically around 30 Hz, because the value of the blocking capacitor represents a high 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 within the range of the scope's vertical amplifier. The vertical amplifier boosts the signal to a level that can drive the CRT and also adds

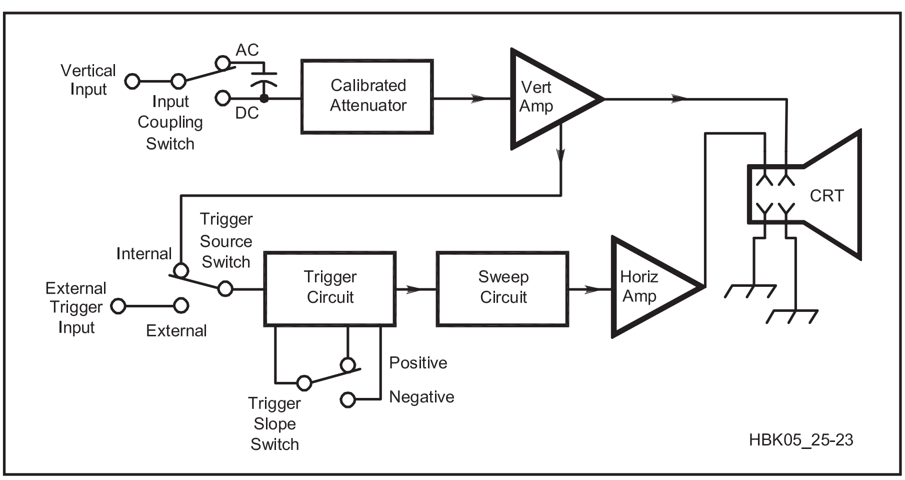


Fig 25.12 — Typical block diagram of a simple triggered-sweep oscilloscope.

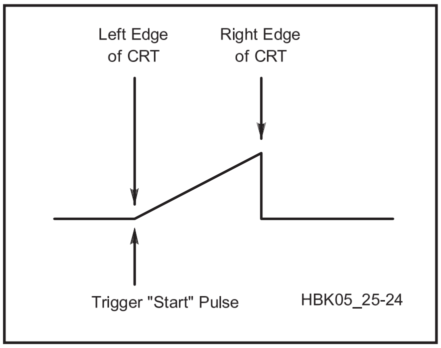


Fig 25.13 — The sweep trigger starts the ramp waveform that sweeps the CRT electron beam from side to side.

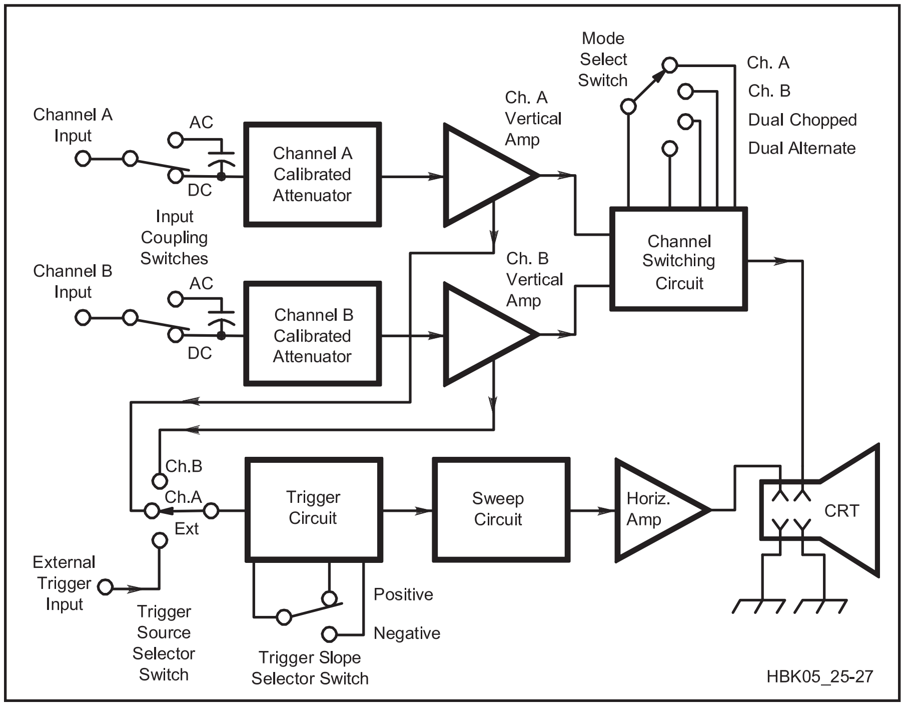


Fig 25.15 — Simplified dual-trace oscilloscope block diagram. Note the two identical input channels and amplifiers.

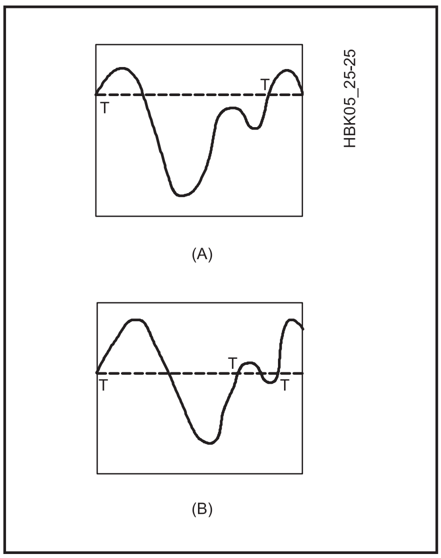


Fig 25.14 — In order to produce a stable display the selection of the trigger point is very important. Selecting the trigger point in A produces a stable display, but the trigger shown at B will produce a display that "jitters" from side to side.

a bias component to position the waveform on the screen. The result is that the vertical position of the beam on the CRT represents input voltage.

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 (*level triggering*) or exhibits a positive- or negative-going edge (*edge triggering*). The sweep generator gives a precisely timed voltage ramp (see **Fig 25.13**). The rising edge of the ramp signal feeds the horizontal amplifier that, in turn, drives the CRT. This 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 result is that the horizontal position of the beam on the CRT represents time. At the end of the ramp, the sharp edge of the ramp quickly moves the beam back to the left side of the screen.

The trigger circuit 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 25.14A** 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. There normally is a TRIGGER LEVEL control to move the trigger level up and down 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 25.14B 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 LEVEL control will fix that problem.

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 the external input jack with the trigger select switch in the EXTERNAL position.

DUAL-TRACE OSCILLOSCOPES

Dual-trace oscilloscopes have two vertical input channels that can be displayed together on the screen. Although the best dual-trace scopes use a CRT with two electron beams, it is possible to trick the eye into seeing two traces simultaneously using a single-beam CRT. **Fig 25.15** shows a simplified block diagram of a dual-trace oscilloscope using this method. 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-beam display but the appearance of dual traces is created by the scope using one of two methods, referred to as *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 switching from one channel to the other is so fast that each trace looks as though it were continuous. The chopped mode is essential for *single-shot* signals (signals that do not repeat periodically). It is most useful at slow sweep speeds. At fast sweep speeds, the switching from channel to channel becomes visible, making each trace into a dotted line.

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 very useful at very slow sweep speeds since the two traces no longer appear simultaneous. It also does not work for single-shot events.

Most dual-trace oscilloscopes also have a feature called "X-Y" mode. 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 single-trace oscilloscopes support this mode as well. 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

In recent years, semiconductor and display technology have advanced to the point that much of the analog signal processing in an oscilloscope can be replaced with low-cost microprocessor-based digital circuitry. This results in dramatically improved accuracy for both amplitude and time measurements as well as enabling sophisticated features that would be difficult or impossible in an analog scope. For example, a trace can be displayed with infinite persistence and stored as a computer file if desired.

In a digital oscilloscope the vertical amplifiers are replaced with an analog-to-digital converter (ADC), which samples the signal at regular time intervals and stores the samples in digital memory. The samples are stored with an assigned time, determined by the trigger circuits and the microprocessor clock. The samples are then retrieved and displayed on the screen with the correct vertical and horizontal position.

Early digital oscilloscopes used a CRT with electrostatic deflection, similar to an analog scope, with the horizontal and vertical deflection signals generated by digital-to-analog converters (DACs). Modern instruments usually use either a raster-scan CRT similar to a television picture tube or a solid state (LCD) display. The microprocessor determines which pixels to light up to draw the traces on the screen. Even though a digital scope does not have the same internal circuitry as an analog scope, many of the same terms are used to control operation. For example, "Sweep speed" still describes the amount of time per horizontal division, even though there is no electron beam to be swept across the display in the original sense.

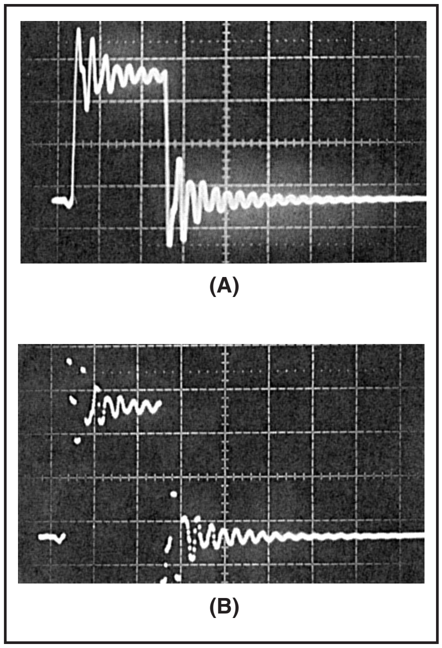


Fig 25.16 — Comparison of an analog scope waveform (A) and that produced by a digital oscilloscope (B). Notice that the digital samples in B are not continuous, which may leave the actual shape of the waveform in doubt for the fastest rise time displays the scope is capable of producing.

For the vertical signals you will see manufacturers refer to "8-bit digitizing," or perhaps "10-bit resolution." This is a measure of the number of 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-bit 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. (See the **DSP and Software Radio Design** chapter for more information on sampled signals.)

If the sample rate is not high enough, very fast signal changes between sampling points will not appear on the display. For example, **Fig 25.16** 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 25.16 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*. 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. Some modern oscilloscopes use a special technique to increase the effective sample rate for repetitive signals. The phase of the sample clock is adjusted slightly on each successive sweep, so that the new samples occur in between the previous ones. After several sweeps the missing data in the spaces between the original samples are filled in, producing a continuous trace. This only works with periodic signals that trigger at exactly the same point on each sweep.

OSCILLOSCOPE LIMITATIONS

Oscilloscopes have fundamental limits, primarily in frequency of operation and range of input voltages. For most purposes the voltage range 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. For example, a 100-MHz 1-V sine wave fed into an oscilloscope with a 100-MHz 3-dB bandwidth will read approximately 0.7 V on the display. The same instrument at frequencies below 30 MHz should be accurate to about 5%.

A parameter called *rise time* is 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 25.17**. Assuming the frequency response is primarily limited by a single-pole roll off in the amplifier circuitry, the mathematical definition of rise time is given by:

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

where t_r = rise time in μs and BW = bandwidth in MHz of the amplifier.

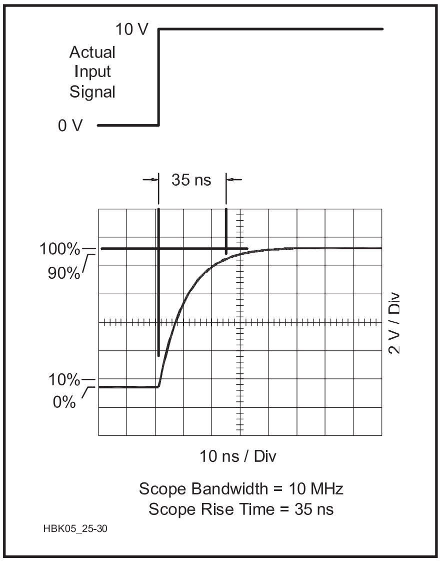


Fig 25.17 — The bandwidth of the oscilloscope vertical channel limits the rise time of the signals displayed on the scope.

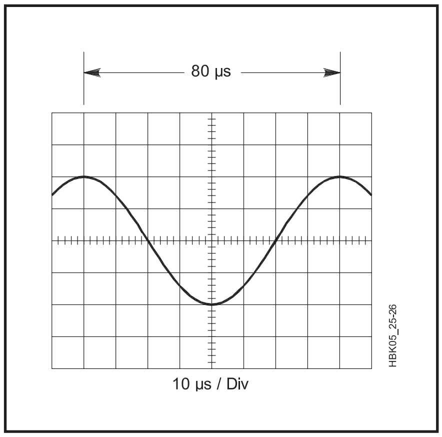


Fig 25.18 — An oscilloscope can measure frequency as well as amplitude. Here the waveform shown has a frequency of 80 microseconds (8 divisions × 10 µs per division) and therefore a frequency of 1/80 µs or 12.5 kHz.

It is also important to note that inexpensive analog oscilloscopes may not have better than 5% accuracy in these applications. Even moderately-priced oscilloscopes are still useful, however. The most important value of an oscilloscope is that it presents an image of what is going on in a circuit, which is very useful for troubleshooting waveforms and other time-varying phenomena. It can show modulation levels, relative gain between stages, clipping distortion, intermittent oscillations and other useful information.

USING AN OSCILLOSCOPE

An oscilloscope can measure a signal's shape, amplitude, frequency and whether it is dc, ac or a mixture of both. For example, in **Fig 25.18** it is clear from the shape that the signal is a sine wave. Assuming that the center horizontal line or axis represents zero volts, the signal has no dc component since there is as much above the axis as below it. If the vertical gain has been set to 1 V per division, then the peak value is 2 V and the peak-to-peak value is 4 V.

The horizontal travel of the trace is calibrated in units of time. If the sweep speed is known 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. The frequency is the reciprocal of the period. In Fig 25.18, for example, the distance between the peaks is 8 divisions. If the sweep speed is 10 µs/division then the period is 80 µs. That means that the frequency of the waveform is 1/80 µs, or 12,500 Hz. The accuracy of the measured frequency depends on the accuracy of the scope's ramp generator, typically a few percent for an analog instrument. 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.

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. However, at higher frequencies the capacitive reactance of the cable would be 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). The total capacitance causes problems when probing an RF circuit with relatively high impedance.

Most new oscilloscopes come with specially-designed *scope probes*, one for each vertical channel. They can also be purchased separately. The most common type is a *×10 probe* (called a "times ten" probe), which 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 were in use, the signals would be displayed on the scope face at 10 V/division. Some scopes can sense whether a ×10 probe is in use, and automatically change the scale of the scope's display.

Unfortunately a resistor alone in series with the scope input seriously degrades the scope's rise-time performance and bandwidth because of the low-pass filter formed by the series resistance along with the parallel capacitance of the cable and scope input. This may be corrected by using a compensating capacitor in parallel with the series resistor. If the capacitor value is chosen so that the R-C time constant is the same as the R-C network formed by the input resistance and capacitance of the scope, as

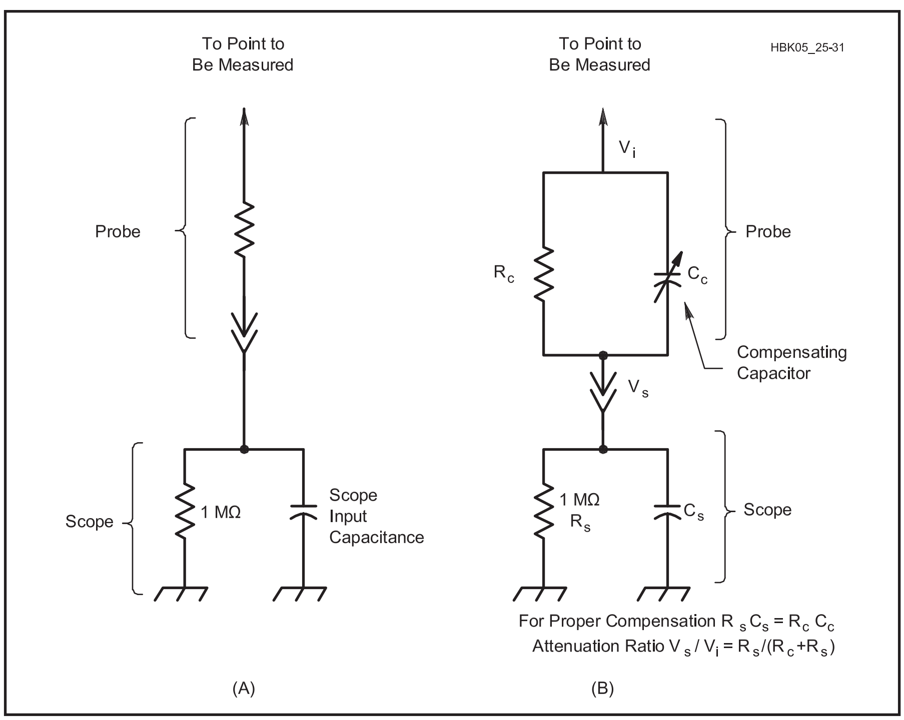


Fig 25.19 — Uncompensated probes such as the one at A are sufficient for low-frequency and slow-rise-time measurements. However, for accurate display of fast rise times with high-frequency components the compensated probe at B must be used. The variable capacitor is adjusted for proper compensation (see text for details).

shown in **Fig 25.19**, then the probe and scope should have a flat response curve throughout the whole bandwidth of the scope. A ×10 probe not only reduces the voltage by a factor of 10 but it increases the input resistance and reduces the capacitance as well, which reduces loading on the circuit under test.

To account for manufacturing tolerances in the scope and probe the compensating capacitor is made variable. Most scopes include a "calibrator" output that produces a fast-rise-time square wave for the purpose of adjusting the compensating capacitor in a probe. **Fig 25.20** shows possible responses when the probe is connected to the oscilloscope's calibrator jack. A misadjusted compensating capacitor can greatly affect the frequency response of the scope and create artifacts in signals that are not actually present.

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 compensating capacitor in the probe is chosen to compensate for the provided length of cable. It usually does 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 act as inductors at high frequencies where they create ringing and other undesirable artifacts in the displayed signal.

For the best high-frequency performance, the scope probe can be eliminated entirely and the oscilloscope input converted to a 50-Ω impedance. Some scopes have a switch to choose between a high-impedance or 50-Ω input. For others, you can purchase a 50-Ω *through-line* termination, which is just a connector with a male BNC on one end, a female BNC on the other, and an internal 50-Ω termination resistor in parallel. Plug the male connector to the scope's high-impedance vertical input and connect the 50-Ω cable from the device under test to the female connector.

Some situations may require the use of a scope to measure a *current* waveform, rather than a voltage. Specialized current probes are available that make this task possible, with some capable of measuring both dc and ac.

Be wary of using a scope to judge harmonic content or distortion of a sine wave. A waveform that "looks good" to the eye may have significant distortion or high-frequency components unsuitable for on-the-air signals. A spectrum analyzer (described below) should be used for determining the spectral content of signals.

CHOOSING AN OSCILLOSCOPE

For many years a scope (even a so-called portable) was big and heavy. In recent years, microprocessors and other ICs have reduced the size and weight. Modern scopes can take other forms than the traditional large cabinet with built-in CRT. Nearly all digital oscilloscopes use an LCD display for true portability. (See **Fig 25.21**) You can use your personal computer as an oscilloscope with an external signal digitizer that connects to the PC via a high-speed USB or Firewire interface. This saves money by using the cabinet, power supply, processor and display of the PC. Even stand-alone scopes can attach to a PC and download their data for storage and analysis or transfer it to memory storage devices. Many high-end scopes now incorporate non-traditional functions, such as the Fast Fourier Transforms (FFT). This allows spectrum analysis or other advanced mathematical techniques to be applied to the displayed waveform.

Features

When choosing an oscilloscope, the first decision is analog versus digital. Used and surplus instruments are usually analog but more modern units are shifting to digital as the prices of solid-state displays and other components come down. Digital models are generally more accurate and tend to have lots of features not found on analog oscilloscopes.

The next big decision is how many input channels you need. There are many situations in which having a second channel is extremely helpful and even more than two channels is often useful, especially when troubleshooting digital circuits. With two or more channels you typically also get X-Y mode.

Trace storage is essential for viewing very slow signals or single-shot transients. In the old days, high-end analog scopes used special CRTs that could store a trace in an analog fashion. They tended to be rather difficult to adjust properly and the CRTs are expensive and/or impossible to replace if they should ever fail. Trace storage is very easy to implement in a digital scope and nearly all of them have the feature. Ideally, one or more stored traces can be displayed simultaneously with the current trace for easy comparison.

Most scopes come with the ability to select internal, external or power-line synchronous triggering as well as an adjustable trigger level. Digital scopes may draw a horizontal line on the screen so you can see exactly where the trigger level is with respect to the signal, which is very handy. Automatic adjustment of trigger level is a common feature as well. Some scopes offer a noise-rejection feature that adds hysteresis to reduce false triggering. *Single sweep* is useful for looking at a non-repetitive signal. The single sweep can be *armed*, that is, reset and ready to start the next sweep, by pushing a button or sometimes with an external signal. *Trigger delay* allows horizontal centering of the display at a different place from the trigger point. *Trigger hold-off* inhibits the trigger for a selectable time after the sweep to prevent unwanted multiple triggers.

Some older tube-type scopes offered only

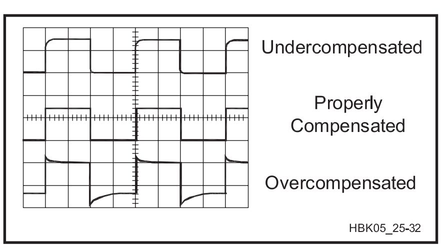


Fig 25.20 — Displays of a square-wave input illustrating undercompensated, properly compensated and overcompensated probes.

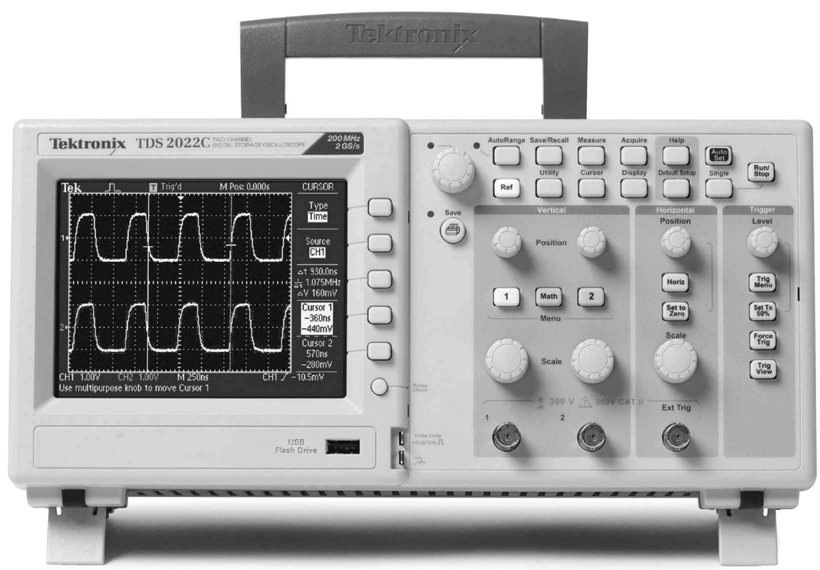


Fig 25.21 — The Tektronix TDS 2000C series is a typical example of digital oscilloscopes using an LCD display.

ac input coupling, but most nowadays have a switch to select between ac, dc and ground. As mentioned before, some include a switch to select 50-Ω or high input impedance. Some high-speed scopes include a switchable input low-pass filter to reduce wideband noise when measuring lower-frequency signals. A feature sometimes found on multiple-channel scopes is the ability to add or subtract two or more channels.

Digital oscilloscopes often include various data-analysis features. Averaging reduces noise and peak mode makes transient peaks visible. A common feature is the ability to calculate and display the peak, peak-to-peak and RMS values of a signal. For repetitive signals, the period and frequency can be calculated as well. Some scopes include amplitude and/or time markers that allow accurate measurements of specific points on a waveform. Sometimes the instrument state can be stored to one of several internal memories, which can be very handy on complicated instruments that take a long time to set up.

Computer connectivity used to be available only on high-end instruments but is now becoming more and more common. It allows storing traces for later reference and you can produce nice graphical screen shots for that magazine article or notebook entry about your latest creation. With the proper software it allows automated testing as well. Older instruments usually have a GPIB (general purpose interface bus) interface, also known as IEEE-488 or as HPIB on Hewlett-Packard equipment. GPIB-to-USB adapters are available from Prologix and a number of other companies. More modern scopes usually come with an RS-232 or USB interface.

Specifications

Bandwidth is the main "money spec" for an oscilloscope. The higher the frequency range, the more expensive it tends to be. Rise time is also sometimes specified. For digital scopes, the sample rate is just as important. Theoretically it must be at least twice the bandwidth but practically should be much higher than that to avoid aliasing. For repetitive waveforms, some scopes can do tricks with the sample phase to eliminate aliasing with a lower sample rate, as explained previously

The range of input signals is normally specified as the maximum and minimum volts per division. Amplitude accuracy is usually specified as a percent of the reading, typically 5-10% for inexpensive analog scopes and perhaps 1-2% for high-quality digital ones. Another important specification is the maximum input signal that can be accepted without damage. It is typically presented as a maximum dc plus peak ac voltage, that is, the maximum peak voltage of the complete signal. This is increased when using a ×10 probe, subject to the probe specifications.

For digital scopes, the resolution of the input analog-to-digital converter (ADC) is typically specified as a number of bits. Sometimes the lowest-voltage input ranges are obtained simply by using only the low end of the ADC range, which results in the displayed signal having a stepped response rather than a smooth curve.

The sweep speed is generally specified in seconds (milliseconds, microseconds) per division. The accuracy is typically a few percent for analog scopes and much better than that for digital scopes, often limited just by the screen resolution. The triggering system is a major factor that determines how useful a scope is in actual operation, but it can be hard to tell how well it works by studying the specifications. The trigger sensitivity is the main parameter to look for. It is specified in fractions of a division for internal trigger and in mV for external trigger.

BUYING A USED SCOPE

Many hams 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. The CRTs in older units may no longer be available. Many scopes made since about 1985 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 operator's manual also.

Older tube-type models are generally quite serviceable, often needing nothing more than a new tube or two. The massive lab-grade instruments from days of yore made by Tektronix and Hewlett-Packard can still give good service with a little care. They are so large and heavy that a special scope cart was often used to house them and allow easy movement from lab bench to lab bench.

Early generation digital scopes are now showing up on the surplus market at reasonable prices. Unfortunately, the user interface leaves much to be desired on some models, sometimes requiring the operator to use several layers of menus to access common functions. Look for one with an analog-like feel with separate buttons or knobs for most of the important functions.

25.3.5 Audio-Frequency Oscillators and Function Generators

There are a number of ways to generate an audio-frequency tone. If a square-wave output is acceptable, then a simple oscillator can be built with a 555 timer IC, two resistors and two capacitors. It is an inexpensive, time-tested design with good frequency stability.

If a sine-wave signal is needed, a *twin-T* oscillator made with a single bipolar transistor is about the simplest solution. The oscillator in **Fig 25.22** 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 that frequency. C1 should be about twice the capacitance of C2 or C3. 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.

While the twin-T oscillator does give a

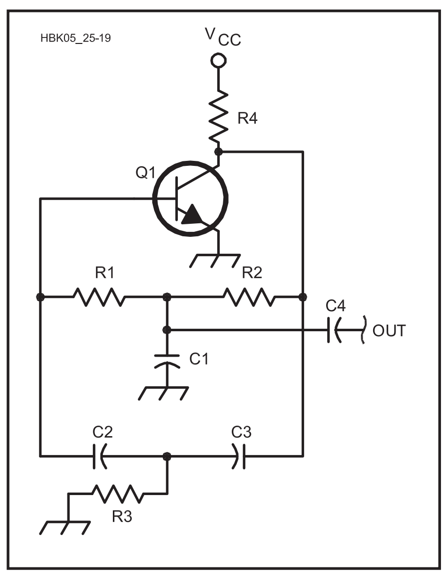


Fig 25.22 — Values for the twin-T audio oscillator circuit range from 18 kΩ for R1-R2 and 0.05 µF for C1 (750 Hz) to 15 kΩ and 0.02 µF for 1800 Hz. For the same frequency range, R3 and C2-C3 vary from 1800 Ω and 0.02 µF to 1500 Ω and 0.01 µF. R4 is 3300 Ω and C4, the output coupling capacitor, can be 0.05 µF for high-impedance loads.

roughly sinusoidal output, the distortion is rather high and the frequency is not easily tunable. Both of those problems are addressed in the *Wien bridge* oscillator illustrated in **Fig 25.23** The key to low distortion is to prevent the amplifier from going into saturation. If equal values of R and C are used in both sections of the positive feedback network, then the gain through that network is ⅓. If the negative feedback network also has a 3:1 ratio, then the total loop gain is exactly 1, the condition for oscillation. The light bulb in the bottom leg has a positive temperature coefficient. (Note that a true incandescent bulb must be used — an LED may not be substituted.) As the signal level increases, its resistance goes up, lowering the gain. In this way the amplitude is held steady so that the amplifier does not saturate. The frequency may be tuned with a single control by using a two-section potentiometer for the two resistors labeled R. Typically small fixed-value resistors are placed in series with each potentiometer section to give about a 10:1 tuning range. Additional ranges can be had by switching the capacitors, selected using the equation

$$C = \frac{1}{2\pi fR}$$

A suitable bulb is a number 327 or 1819. Under normal operation, it should be lit to a bit less than full brightness, which is determined by the value of the negative feedback resistor R_F, typically around 400 Ω.

The idea of using the positive temperature coefficient of resistance of a light bulb to stabilize the output of a tunable oscillator was used in the very first instrument produced by the *Hewlett-Packard* company (now Agilent Technologies). (The Model 200A audio oscillator circuit was first described in company founder Hewlett's 1939 college thesis.)

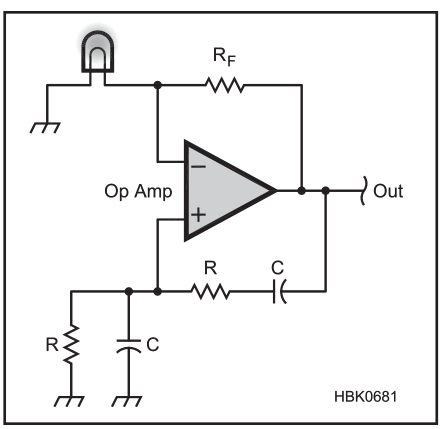


Fig 25.23 — A Wien bridge oscillator has lower distortion than a twin-T oscillator. The light bulb acts to stabilize the feedback to prevent distortion due to amplifier clipping.

FUNCTION GENERATORS

A *function generator*, also known as a *waveform generator*, is a type of audio oscillator that can generate several waveforms. In addition to sine waves, most can also generate square waves, triangle waves and sawtooth ramp waveforms. A pulse output with variable duty factor is another common feature. Many models can also linearly sweep the frequency. Frequency coverage is typically from below audio frequencies up to a few MHz.

The heart of most analog function generators is a triangle wave oscillator. To create the sine wave, a diode shaping circuit rounds off the top and bottom of the triangles, resulting in a reasonably-accurate sine wave with distortion on the order of a percent or two. Circuitry that determines when the triangle wave is rising or falling is used to generate the square wave as well. The rise/fall duty factor of the triangle wave can be varied, which also varies the duty factor of the square wave. If the triangle wave duty factor is set near 100% or 0%, it becomes either a rising or falling sawtooth wave.

The easiest way to build your own analog function generator is to use a waveform generator IC that includes most of what you need in one package. The two most common such ICs available today are the MAX038 and XR2206. The ICL8038 is seen in many older circuits but is obsolete and no longer in production. The AD9833 direct digital synthesis (DDS) IC (see **Fig 25.24**) is also designed to be used as a low frequency function generator.[10] (DDS is discussed in the **Oscillators and Synthesizers** and **DSP and Software Radio Design** chapters.)

An *arbitrary waveform generator* (AWG or ARB) has capabilities that are a superset of a waveform generator. The sinusoidal waveform ROM look-up table in Fig 25.24 is replaced by RAM so that data which creates waveforms other than a sine wave can be used. Usually a large amount of memory is included so that long non-repetitive waveforms can be generated in addition to periodic signals. ARBs generally include a microprocessor and provide means to generate complicated sequences that may combine repeating and nonrepeating segments.

CHOOSING A FUNCTION GENERATOR

For many applications, the basic analog sine/square wave generator based on a Wien bridge oscillator works fine. The sine wave is adequate for testing the gain, frequency response, and maximum signal level of an audio-frequency circuit. The square wave is useful for checking the transient response of circuits.

The additional wave shapes offered by a function generator are useful for more esoteric applications. Sawtooth and triangle waves can be used to sweep a voltage through a range to test the response of a circuit to various voltages. The pulse output is useful for testing digital circuitry, especially if the instrument has provision for adjusting the on and off voltage levels. Frequency sweep capability is very handy for testing the frequency response of an audio circuit, especially if a sweep ramp or sweep trigger output is provided for synchronizing an oscilloscope. Some instruments provide an input to control the frequency by means of an external voltage.

When selecting a function generator, the first specification to look at is the frequency range. Many units cover low radio frequencies as well as audio. The frequency accuracy may be important for many applications. Digitally-synthesized models use a quartz crystal clock oscillator so are much more accurate than analog models. Some analog models do have a built-in frequency counter, but the accuracy may still be limited by frequency drift.

The amplitude range and accuracy should also be considered. An amplitude of several volts is useful for testing power devices and the ability to accurately set the amplitude to a few millivolts may be needed for driving the microphone input of a transmitter. The output impedance is usually either 600 Ω or 50 Ω.

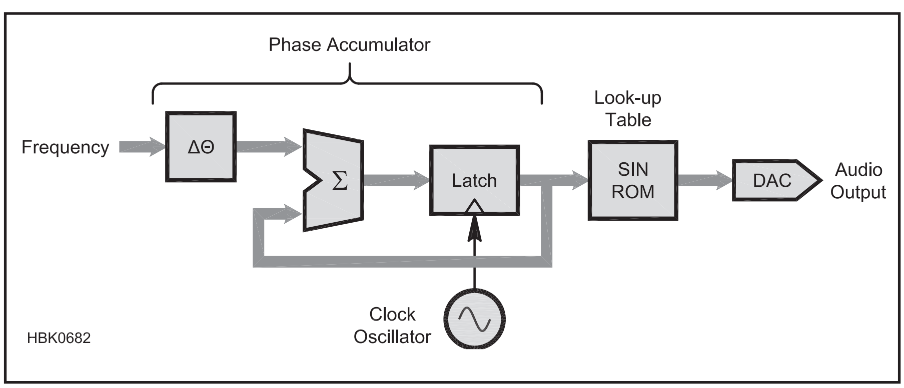


Fig 25.24 — A direct digital synthesizer (DDS) sine-wave generator. Not shown is a low-pass filter at the output that attenuates unwanted spurious frequencies above one-half the clock frequency.

Function generators are available for a wide range of prices, from inexpensive hand-held units to sophisticated bench instruments costing thousands of dollars. Surplus tube-type models can be good values, starting with the venerable Hewlett-Packard 200A, manufactured until the early 1970s. More-recent models can also be found on the used-equipment market made by HP, its successor company Agilent Technologies, Tektronix, B&K, Keithley, Wavetek, Leader, Fluke and others.

An even less-expensive solution is to use a computer sound card as a function generator. A variety of free function generator software can be found on the Internet. Since the sound card output is ac-coupled it is not possible to adjust the dc offset voltage as it is with most function generators. Also, the frequency range, the level range and accuracy, and the output drive capability are not as good as you would expect from a special-purpose instrument. However, quite sophisticated waveforms may be generated with the right software and the price is right.

25.3.6 Measuring Inductance and Capacitance

The traditional way to measure inductance (L), capacitance (C), or resistance (R) is with an *LCR bridge* consisting of a Wheatstone bridge driven by a sine-wave voltage and with an ac voltmeter for the null detector. In **Fig 25.25,** the box labeled Z is the reference component. It must be the same type as the device under test, an inductor, capacitor or resistor. The bridge is nulled when the ratio of the variable resistor R_V to R is the same as the ratio of the impedance of the device under test to that of Z. The value of R_V is proportional to the resistance or inductance and inversely proportional to the capacitance. To make it proportional to C, simply swap the positions of R_V and R in the circuit. In all cases, when $R_V = R$, the null is achieved when the value of the device under test equals the value of Z.

Most commercial LCR bridges include a switch to select various values of Z. Typically they are in decade steps so that the dial calibration works on all scales. The frequency of the sine-wave source should be such that the reactance of the capacitor or inductor under test is not so low as to present too small a load impedance to the sine-wave source and not so high that the load presented by the ac voltmeter reduces the depth of the null. Often several frequencies may be selected, the lower frequencies being used for large-value L and C and higher frequencies used for the smaller values.

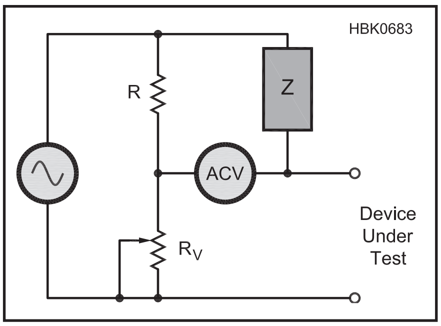


Fig 25.25 — An LCR bridge, which is a type of Wheatstone bridge used to measure inductors (L), capacitors (C) and resistors (R). The box labeled "Z" is the reference component to which the device under test is compared.

A dip meter may be used for measuring either L or C as long as a component of the opposite type is available whose value is accurately known. The technique is to make a tuned circuit by connecting the inductor and capacitor in parallel and then measure the resonant frequency with a grid dip meter. The inductance or capacitance can be determined from the dip frequency, f, using the formula

$$L = \frac{1}{(2\pi f)^2 C} \quad \text{or}$$

$$C = \frac{1}{(2\pi f)^2 L}$$

Some multimeters have built-in capability to measure inductors and capacitors. It is also possible to build an adapter to allow measuring L and C with any multimeter. Circuit examples are given in the Construction Projects section at the end of this chapter.

The use of capacitors in high-frequency switchmode supplies (see the **Power Sources** chapter) makes it important to measure their *equivalent series resistance (ESR)* and *equivalent series inductance (ESL)*. ESR and ESL cause loss and affect the switching circuit's ability to regulate properly. An ESR meter measures a capacitor's ESR by using short pulses or ac signals. Some ESR meters can be used with the capacitor in-circuit although they should not be used with the capacitor charged or energized. ESL is typically measured with an inductance meter as descibed above.

25.4 RF Measurements

RF measurements are a special case of ac measurements. While everything in the previous section about measuring low-frequency ac applies, there are additional factors to take into consideration for high frequency measurements, such as parasitic values and operational bandwidth. This section concentrates on equipment and techniques especially suited to measuring at radio frequencies. The **RF Techniques** chapter has additional information on circuits at high frequencies.

25.4.1 Measuring RF Voltage and Current

An *RF probe* rectifies a radio-frequency signal so that its amplitude can be measured with a dc instrument such as a multimeter. The circuit is typically quite simple, consisting of a diode, a resistor and a couple capacitors, as in the example of **Fig 25.26**. The resistor value is chosen so that the rectified output voltage is approximately equal to the RMS value of a sine wave input signal. The 390 kΩ value shown assumes the meter has a 1 MΩ input resistance. The diode can be a high-speed switching diode such as a 1N914 or 1N4148, or it can be a Schottky diode for greater sensitivity.

The detector is located as close to the measuring point as possible to minimize stray inductance and capacitance. The leads to the dc voltmeter can be longer. The RF probe can be housed in any convenient enclosure that fits easily in the hand and the circuitry can be constructed on a scrap of perforated phenolic board, as long as the leads are kept short. A much more elaborate version with a coaxial input and an integrated compensated voltmeter is described on the CD-ROM that accompanies this book.

Measuring RF current is a little more dif-

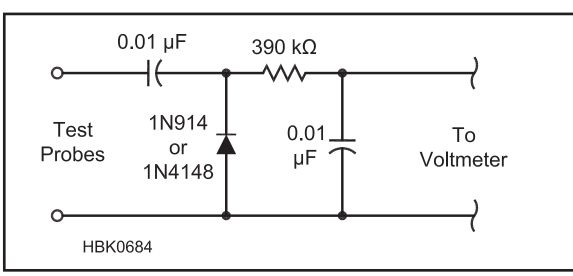


Fig 25.26 — A basic RF probe, used to convert an RF signal into a dc voltage that can be measured by a voltmeter.

ficult. When measuring the current on an antenna or feed line, it may not be practical to break the connection to insert an ammeter. Even when that can be done, the meter itself often upsets the measurement, either because it must be plugged into ac power or simply because of the instrument's size.

A time-honored technique is to wire a small incandescent lamp in series with the antenna or feed line to be measured. Although you can get a rough idea of the amount of RF current by comparing the brightness to the brightness with a known dc current, this method is only useful for relative measurements where the absolute value does not need to be known.

Another alternative is to construct an RF ammeter using a current transformer that clamps over the wire.[11] That method has the advantage that the wire does not need to be disconnected. It is important that the leads to the meter not couple to the wire or cable being measured.

Highly-accurate commercial RF ammeters made to measure the base current of AM broadcast station antennas can sometimes be found on the surplus market. If you can find one it would give much better accuracy than a homebrew device.

25.4.2 Measuring RF Power

Transmitter power is normally given in units of watts. Lower-power RF signals found in receiver and transmitter circuits may be specified in units of milliwatts (mW) or microwatts (µW) but it is perhaps more common to see units of *dBm*, decibels with respect to one milliwatt. For example, it is much easier to express the power of an S1 signal as –121 dBm than 0.00000000076 µW. The formula is dBm = 10 log (1000 P), where P is the power in watts. One milliwatt (P = 0.001) is 0 dBm, 10 mW is +10 dBm, 0.01 mW is –20 dBm and so on. **Table 25.2** lists common dBm and power equivalent.

Measuring RF power can be a little confusing because there are several ways to do it. We have already covered the difference between peak and RMS voltage and current. RF power is always based on RMS values. For example, if the RF voltage into a 50-Ω dummy load is 70.7 V RMS (100 V at the peak of the RF sine waves), then the power is $P = E^2 / R = 70.7^2 / 50 = 100$ W. (See **www.eznec.com/Amateur/RMS_Power.pdf** for a more extensive explanation of power and RMS.)

Peak envelope power (PEP) has nothing to do with the difference between the peak and average voltage of a sine wave. It is a measure of the power of an RF signal at the modulation peak, averaged over one RF cycle. For a CW signal, the PEP is simply the power when the key is closed, as read on any wattmeter. However, for an SSB signal, the power is constantly changing as you speak. An average-reading wattmeter will read a value much lower than the PEP.

Measuring the peak envelope power is more difficult than measuring the average power. An oscilloscope can be a highly-accurate method within its bandwidth limitations if the load impedance is known accurately. Don't forget that the oscilloscope shows peak-to-peak rather than RMS voltage, so you must divide the maximum reading by $2\sqrt{2}$ or 2.828. Some wattmeters do have PEP-reading capability. Their circuitry must have very fast response to the detected RF signal to give an accurate reading of the peaks.

A *directional wattmeter* is a device that measures power flowing in each direction on a transmission line. The manner in which RF signals propagate on transmission lines is covered in the **Transmission Lines** chapter. Many amateurs keep a wattmeter permanently connected at their station to monitor the condition of their transmitter and antennas. See the **Station Accessories** chapter for a further discussion of directional wattmeters.

A *bolometer* is a device for measuring transmitter power by measuring the heat dissipated in a resistive load. A thermistor or other device measures the temperature. The device is calibrated with a dc voltage, since dc voltage and current can be measured very precisely. With careful construction and calibration, very high accuracy can be obtained. However, the response time of the measurement is very slow, so a bolometer is normally used to calibrate another wattmeter rather than being used directly for measurements. Bolometers are made commercially, however it is possible to homebrew one using a plastic picnic cooler.[12]

Commercial laboratory power meters are generally intended for measuring power levels in the milliwatt or microwatt range. There are two basic types, based on either diode or thermocouple detectors. Below a certain power level, the dc output from a diode detector is directly proportional to power, that is, the square of the RF voltage. With a suitable dc amplifier, the detector output can drive a meter with a linear scale to read power directly.

Thermocouple-type power meters feed the RF signal into a resistor that heats up in proportion to the power level. The temperature is measured with a *thermocouple*, which consists of a pair of junctions of dissimilar metals. A voltage is generated based on the temperature difference of the two junctions, which is proportional to the RF power. The thermocouple method gives high accuracy and wide bandwidth, but the measurement time can be up to several seconds at low power levels, rather than being nearly instantaneous as with a diode detector. Older-model surplus diode and thermocouple-type meters can often be found for reasonable prices, but the proper sensors for the desired frequency range and power level often cost more than the power meter itself.

Analog Devices makes a series of integrated circuits that can detect RF signals and output a dc voltage proportional to the logarithm of the power level. That makes it easy to construct an RF power meter that reads directly in dBm. For example, the AD8307 covers dc to 500 MHz with 1-dB accuracy over an 88-dB (nearly 1 billion-to-one) power range.

25.4.3 Spectrum Analyzers

A spectrum analyzer is similar to an oscilloscope. Both present a graphical view of an electrical signal. The oscilloscope is used to observe electrical signals in the *time domain* (amplitude versus time). The time domain,

Table 25.2
Power and dBm Equivalents

dBm	*Power*	*dBm*	*Power*
–60 dBm	1 nW	3 dBm	2 mW
–30 dBm	1 µW	6 dBm	4 mW
–20 dBm	10 µW	10 dBm	10 mW
–10 dBm	100 µW	20 dBm	100 mW
–6 dBm	¼ mW	30 dBm	1 W
–3 dBm	½ mW	60 dBm	1 kW
0 dBm	1 mW	61.7 dBm	1.5 kW

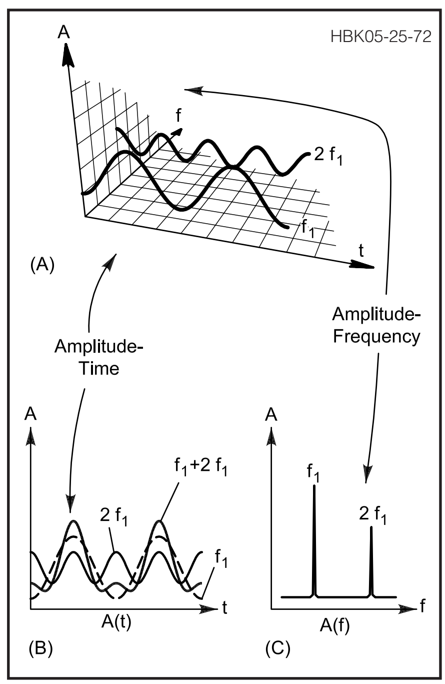


Fig 25.27 — A complex signal in the time and frequency domains. A is a three-dimensional display of amplitude, time and frequency. B is an oscilloscope display of time vs amplitude. C is spectrum analyzer display of the frequency domain and shows frequency vs amplitude.

however, gives little information about the frequencies that make up complex signals, which are best characterized in terms of their frequency response. This information is obtained by viewing electrical signals in the *frequency domain* (amplitude versus 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 25.27**. The three-dimensional coordinates in Fig 25.27A show time as the 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 in Fig 25.27B, 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 the sum of f1 and 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, such as harmonic content or distortion as mentioned previously. For measurements that are best made in the time domain, the oscilloscope is the tool of choice.

SPECTRUM ANALYZER BASICS

There are several different types of spectrum analyzer, but 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 25.28**. 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 in which a wide-band input is converted to an IF higher than the highest input frequency. Up-conversion is used so that a fixed-tuned input filter can remove any image signals and only the first local oscillator needs to be tuned to tune the receiver. 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.

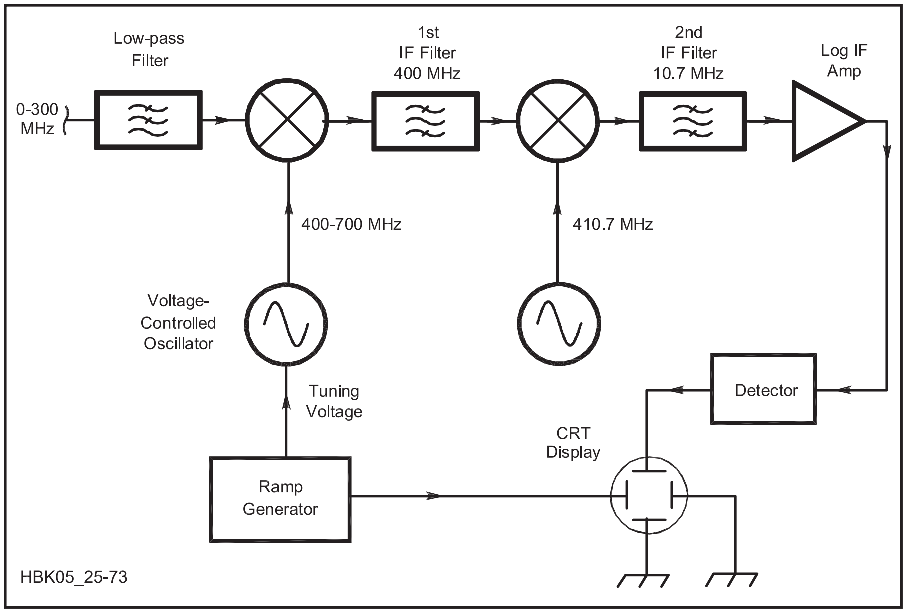


Fig 25.28 — A block diagram of a typical superheterodyne spectrum analyzer. Input frequencies of up to 300 MHz are up-converted by the local oscillator and mixer to a fixed frequency of 400 MHz.

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 range of image frequency extends from 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 large, on the order of 60 to 80 dB. This requires a first IF filter with a high Q, which 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, 10.7 MHz in this example. 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, consider two signals that differ by 60 dB, which is a thousand to one difference in voltage or a million to one in power. That means 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

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* (MDS), 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, one factor that affects dynamic range is *second- and third-order intermodulation distortion* (IMD). 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. (See the **Receivers** chapter for more information on IMD.)

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 25.28, 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. For some applications, it may be useful to add an external wide-band, low-noise preamplifier to improve the noise figure.

The selectivity of the analyzer is called the *resolution bandwidth* (RBW). This term refers to the minimum frequency separation of two signals of equal level so that the signals are separated by a drop in amplitude of 3 dB between them. The IF filters used in a spectrum analyzer differ from a communications receiver in that they have very gentle skirts and rounded passbands, rather than the flat passband and very steep skirts of an IF filter in a high-quality communications receiver. The rounded passband is necessary because the signals pass through the filter passband as the spectrum analyzer sweeps the desired frequency range. If the signals suddenly pop into the passband (as they do if the filter has steep skirts), the filter tends to ring. A filter with gentle skirts has less ringing. Another effect, which occurs even with rounded-passband filters, is that the signal amplitudes are reduced at fast sweep rates, which distorts the display and requires that the analyzer not sweep frequency too quickly. When adjusting the resolution bandwidth or the width of the *frequency span* (the range of frequencies being measured), the scan rate may need to be reduced so that the signal amplitude is not affected by fast sweeping.

The signal produced by the detector is known as the *video* signal. Most spectrum analyzers include a low-pass video filter to reduce the displayed noise level. The *video bandwidth* (VBW) is the bandwidth of this filter. Like the RBW, the VBW must also be taken into consideration when setting the sweep speed.

USING A SPECTRUM ANALYZER

Spectrum analyzers are used in situations where the signals to be analyzed are complex, for very low-level signals, or 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.

Unlike the oscilloscope which is a wide-bandwidth instrument, the spectrum analyzer measures the waveform using a narrow bandwidth; thus it is capable of reducing the noise power displayed.

Probably the most common Amateur Radio application of a spectrum analyzer is the measurement of the harmonic content and other spurious signals in the output of a radio transmitter. **Fig 25.29** 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 input mixers can be damaged by only a few milliwatts, so most analyzers have an adjustable input attenuator to provide a reasonable amount of attenuation for protection. The power limitation of the attenuator itself is usually on the order of a watt or so, however. This means that the power attenuator must have 20 dB of attenuation 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 proper 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.

Spurious Responses in Spectrum Analyzers

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

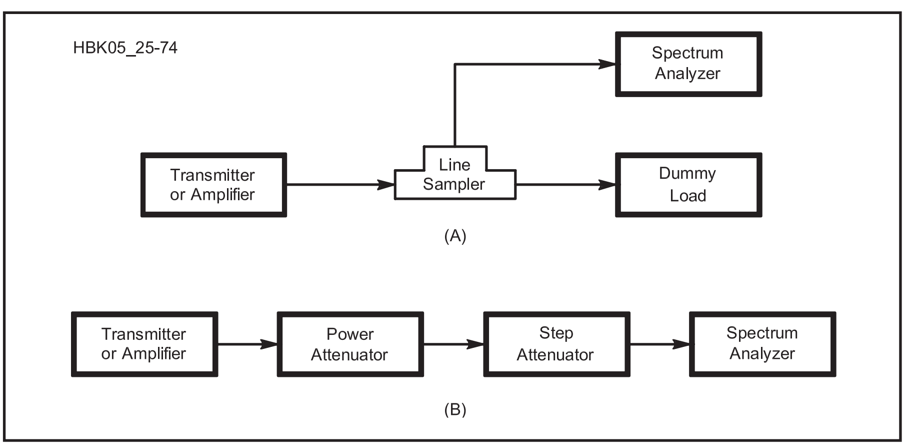


Fig 25.29 — Alternate bench setups for viewing the output of a high power transmitter or oscillator on a spectrum analyzer. A uses a line sampler to pick off a small amount of the transmitter or amplifier power. In B, most of the transmitter power is dissipated in the power attenuator.

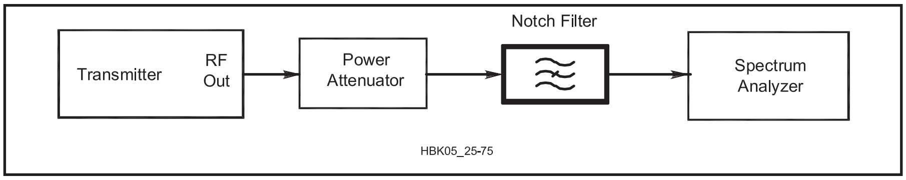


Fig 25.30 — A notch filter is another way to reduce the level of a transmitter's fundamental signal so that the fundamental does not generate harmonics within the analyzer. However, in order to know the amplitude relationship between the fundamental and the transmitter's actual harmonics and spurs, the attenuation of the fundamental in the notch filter must be known.

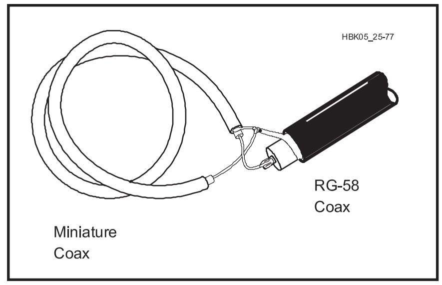


Fig 25.32 — A "sniffer" probe consisting of an inductive pick-up. It has the advantage of not directly contacting the circuit under test.

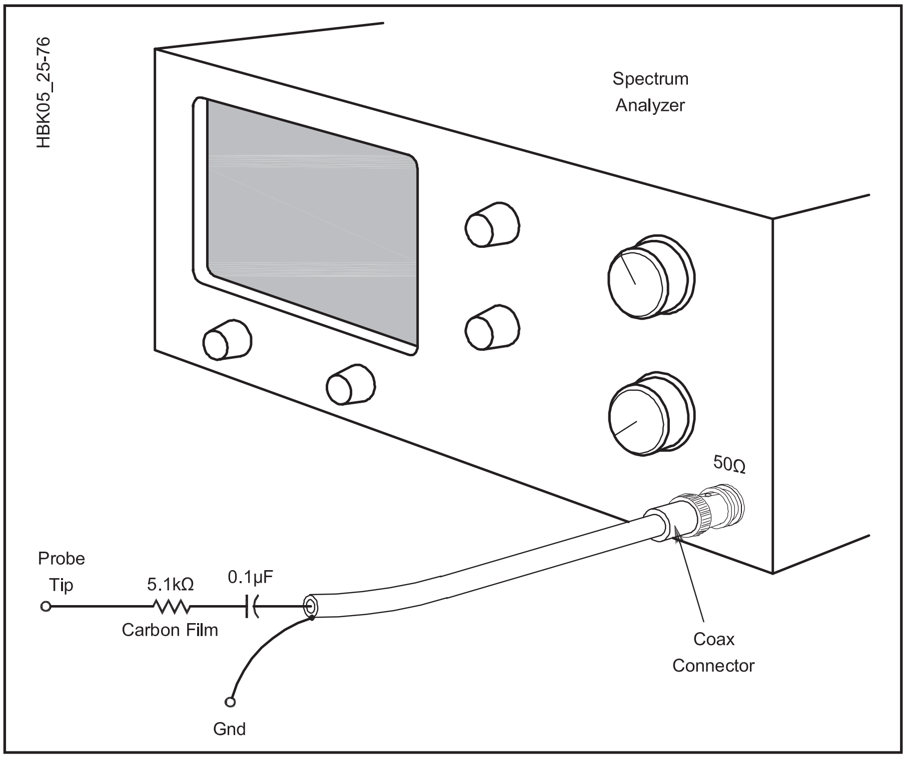


Fig 25.31 — A schematic representation of a voltage probe designed for use with a spectrum analyzer. Keep the probe tip (resistor and capacitor) and ground leads as short as possible.

a spectrum analyzer are harmonics and IMD. It is good practice to check for the generation of spurious signals within the spectrum analyzer. When an input signal causes the spectrum analyzer to generate a spurious signal, 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.

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 25.30**. 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.

The input impedance for most RF spectrum analyzers is 50 Ω, however not all circuits have convenient 50-Ω connections that can be accessed for testing purposes. Using a probe such as the one shown in **Fig 25.31** allows the analyzer to be used as a troubleshooting tool. The probe can be used to track 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 about 5000 Ω. A different type of probe is shown in **Fig 25.32**. 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 pickup loop and soldered to a larger piece of coax that connects to the spectrum analyzer. The dimensions of the loop are not critical, but smaller loop dimensions enable the loop to more precisely locate the source of radiated RF. Coax is used for the loop to provide shielding from the electric field component (capacitive coupling). Connecting the coax shield on only one end provides a complete electrostatic shield without introducing a shorted turn.

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 startup 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 pickup 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.

Measuring Very Low-Level Signals

One very powerful characteristic of the spectrum analyzer is the instrument's capa-

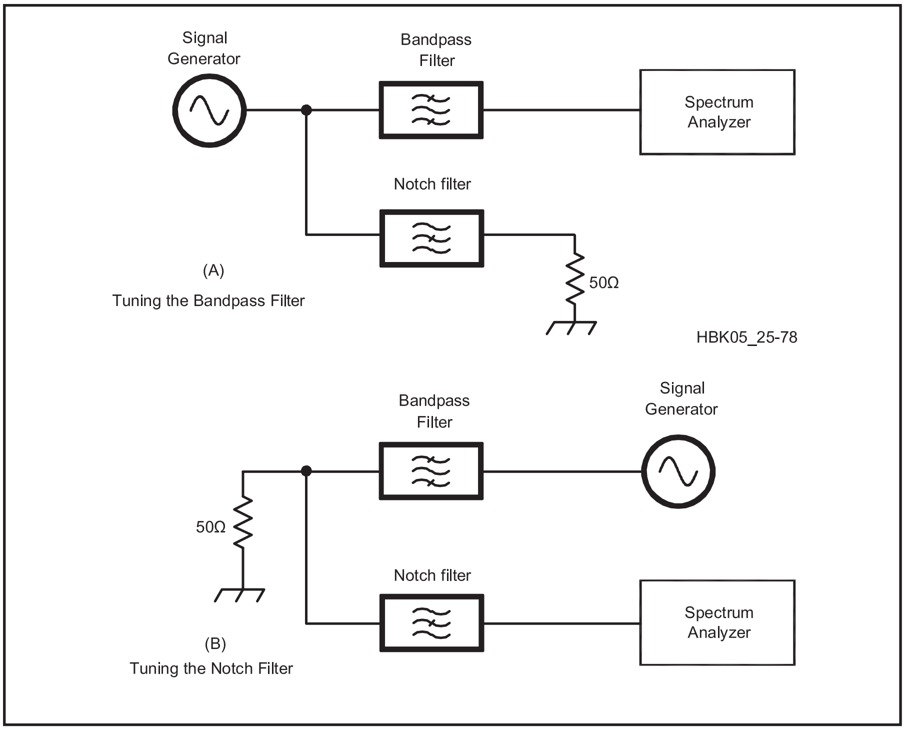


Fig 25.33 — Block diagram of a spectrum analyzer and signal generator being used to tune the band-pass and notch filters of a duplexer. All ports of the duplexer must be properly terminated and good quality coax with intact shielding used to reduce leakage.

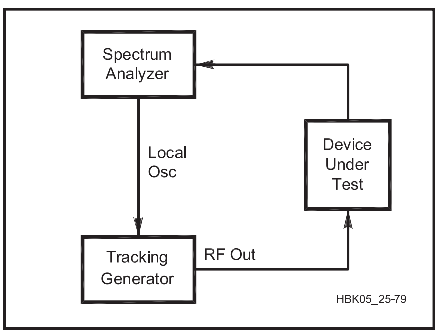


Fig 25.34 — A signal generator (shown in the figure as the "Tracking Generator") locked to the local oscillator of a spectrum analyzer can be used to determine filter response over a range of frequencies.

bility to measure very low-level signals. This characteristic is very advantageous when very high levels of attenuation are measured. **Fig 25.33** 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. 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

Many measurements involve a signal generator tuned to the same frequency as the spectrum analyzer. It would 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 25.34 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. That is the function of a *vector network analyzer*, covered in a following section.

CHOOSING A SPECTRUM ANALYZER

The frequency range is perhaps the most important specification when choosing a spectrum analyzer. Of course it must cover the amateur bands you wish to test, but it must also cover harmonics of those frequencies if you wish to test transmitter harmonics. Many higher-frequency spectrum analyzers only cover down to 10 kHz, so they are not useful for audio spectrum analysis (spectrum analyzers for audio use are also available). Some instruments use a harmonic sampler on the higher microwave bands. You may need to use an external filter to remove any unwanted signals at harmonics or sub-harmonics of the signal you wish to examine.

Frequency stability is vital for narrow-band measurements such as looking at modulation spectra. Some older instruments include a frequency lock scheme that stabilizes the display after the analyzer is tuned to the desired signal.

The range of resolution bandwidths (RBW) available determines what kinds of measurements are possible. To measure two-tone intermodulation distortion of an SSB signal, the minimum RBW must be narrow enough to resolve the two tones, that is, no more than about 100 Hz. If you wish to view the demodulated time-domain video from a fast-scan television signal, the RBW must be wide enough to include the entire signal, perhaps 5 or 6 MHz.

Input sensitivity is normally not very important. An external preamplifier can always be added if needed for a particular measurement. However, dynamic range is a key specification, as previously discussed. Also, the screen display range is important. For example, if there are 8 divisions on the display and the maximum dB-per-division setting is 10 dB/div, then the maximum display range is 80 dB. Many spectrum analyzers can display signals in linear as well as logarithmic (dB) mode, which can be useful to look at modulation.

For accurate power measurements, the amplitude accuracy is important. For doing relative measurements, which are important for measuring spurious signals relative to the carrier, logarithmic linearity is the key specification.

There are several features that expand the spectrum analyzer's utility. *Markers* are pips that can be moved with a knob back and forth across the signals on the display. By placing a marker at the top of a signal you can read the frequency and amplitude directly. *Peak search* is a feature that automatically places a marker at the peak of the strongest signal. *Marker delta* measures the difference in frequency and amplitude of two markers.

Zero span means setting the sweep width to zero hertz to view the change in signal versus time. This is useful for looking at modulation. Many spectrum analyzers include a sweep triggering feature for this purpose, similar to the triggering circuit in an oscilloscope. Some include a speaker so you can hear what the modulating signal sounds like.

Modern spectrum analyzers have many of the features found in digital oscilloscopes, such as trace storage and retrieval. Most include computer connectivity, both to store measured data and to control the instrument with the proper software.

In addition to traditional swept-frequency spectrum analyzers, some modern units use the *Fast Fourier Transform* (FFT) either in place of, or in addition to, the swept-frequency architecture. The FFT allows for much faster screen updates when using narrow spans. Some RF and microwave instruments have a *digital IF* based on the FFT. For narrow spans, the local oscillator is held at a constant frequency and the "sweep" is performed mathematically by the FFT.

There are many used spectrum analyzers on the surplus market, at all different vintages and price points. As usual, try before you buy if at all possible and obtain the operator's and service manuals, if available. Another alternative is a spectrum analyzer built on a card that plugs into a personal computer or that consists of an external digitizing pod connected to the PC via a serial data link. As with PC-based oscilloscopes, they can represent excellent value because they save the cost of the display, cabinet, power supply, and microprocessor provided by the computer.

The least-expensive alternative is an audio spectrum analyzer based on a PC's sound card. Free spectrum analysis software is available on the Internet. The main limitation is the frequency coverage, which is typically about 40-45% of the sample rate of the sound card. The sound card that comes standard in a PC typically has a 48 kHz sample rate, so that the frequency response is limited to about 20 kHz. That is plenty for checking out the audio circuits in a communications receiver or transmitter.

25.4.4 Measuring RF Impedance

Most RF impedance-measuring devices are based on the Wheatstone bridge, as previously described in the section on LCR bridges. The difference is that an impedance bridge does not require that the device under test be a pure inductance, capacitance or resistance but may be a complex impedance that includes both resistance and reactance. That means that two adjustments are required to balance the bridge, rather than just one as with an LCR bridge. Both the resistance and reactance must match.

In addition, since an impedance bridge is typically designed to operate at higher frequencies, circuit layout and the connection to the device under test are more important. Usually a coaxial connector is used to connect to the device and the impedance measurements are referenced to 50 Ω.

A *return-loss bridge* (RLB) is an RF bridge with fixed, usually 50-Ω, resistors in each leg except the one connected to the device under test. *Return loss* is the ratio of the reflected signal to the signal incident on a component, usually expressed in dB, which is always a positive number for a passive device.[13] (See the **RF Techniques** chapter for a discussion of return loss.)

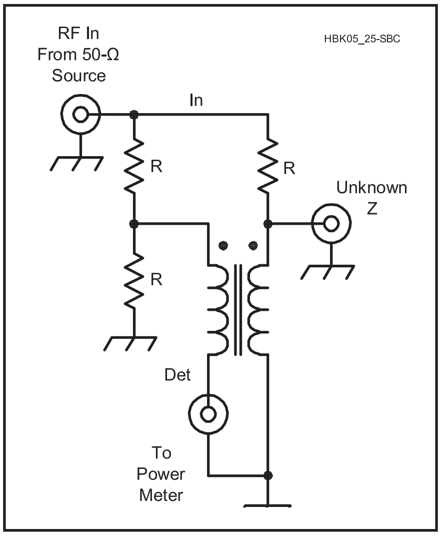


Fig 25.35 — A return-loss bridge for RF. See text for details.

The schematic of a simple return-loss bridge is shown in **Fig 25.35**. The circuit can also be used as a *hybrid combiner* (see the RF and Microwave Test Accessories section of this chapter), where the port labeled UNKNOWN is the common port and the other two are isolated from each other. For good results at high frequencies, it should be built in a small box with short leads to the coax connectors.

Apply the output of a signal generator or other signal source to the RF IN port of the RLB. The power level should be appropriate for driving the device connected to the UNKNOWN port, after accounting for the 6-dB loss of the bridge. Connect the bridge POWER METER port to a power meter or other power-measuring device 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 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.

ANTENNA ANALYZERS

A return-loss bridge measures only the magnitude of the reflection coefficient. Among radio amateurs, an RF impedance analyzer capable of measuring both the magnitude and the phase is often called an *antenna analyzer*, since it is mostly used for measuring antennas.

To keep size and cost to a minimum, portable antenna analyzers usually use a narrow-band source (an internal oscillator) and wide-band detector (a diode). Some manufactured units include a microprocessor and can read out SWR, return loss, resistance, reactance and the magnitude and phase of the impedance on an LCD display. Antenna analyzers suitable for amateur use are available from a number of manufacturers — search for "antenna analyzers" on the Internet to find them. When shopping for an antenna analyzer, pay careful attention to the capabilities and limitations. Be aware that some units measure only the SWR or impedance magnitude while others do measure both the resistive and reactive parts of the impedance. Some units give the magnitude of the reactance but not the sign.

Some users have reported difficulties in obtaining accurate impedance measurements on low-band antennas when there is a nearby AM broadcast station. This is due to the wide-band detector responding to the incoming signal from the AM station.

Noise Bridges

Rather than use a narrow-band source and wide-band detector, another technique is to use a wide-band noise generator for the source and a receiver for the detector. Since most hams already have a receiver, this eliminates the need for an accurate, stable signal generator. You tune the receiver to the desired frequency and adjust the resistance and reactance controls for minimum noise in the receiver. If the receiver is fitted with a *panadapter* (a spectrum display for frequencies near the receive frequency using the receiver's IF circuits), you can locate the null as you adjust the controls, speeding the adjustment.

25.4.5 Network Analyzers

An electronic *port* is a pair of terminals with equal and opposite current flows. An antenna analyzer can measure the impedance of a single-port network such as an antenna, the input port of an antenna tuner, or a two-terminal component like a resistor or capacitor. An example of a two-port network is an amplifier. The input is one port, the output the other. A low-pass filter is another example. An example of a three-port network is a diplexer, used to connect one transceiver to two antennas. The theory behind two-port networks is covered in the **RF Techniques** chapter.

A *scalar network analyzer* (SNA) is an instrument that can measure, as a function of frequency, the magnitude of the gain or loss between the two ports of a two-port device as well as the magnitude of the return loss (equivalent to the SWR) of each port. The term "scalar" means that only magnitudes of those quantities are measured and not phase angles. In **Fig 25.36**, the signal labeled "REF" is proportional to the sweep oscillator output and the one labeled "A" is proportional to the signal that is reflected back from the device under test due to an imperfect 50-Ω match. The ratio of A to REF can be used to calculate the return loss and SWR of the device. Signal "B" is proportional to the signal that travels between the two ports of the device. The ratio of B to REF gives the device gain. To find the gain in the opposite direction and the return loss of the other port, simply turn the device around and swap the two ports. Some network analyzers come with a *test set* that includes RF switches to do that automatically.[14]

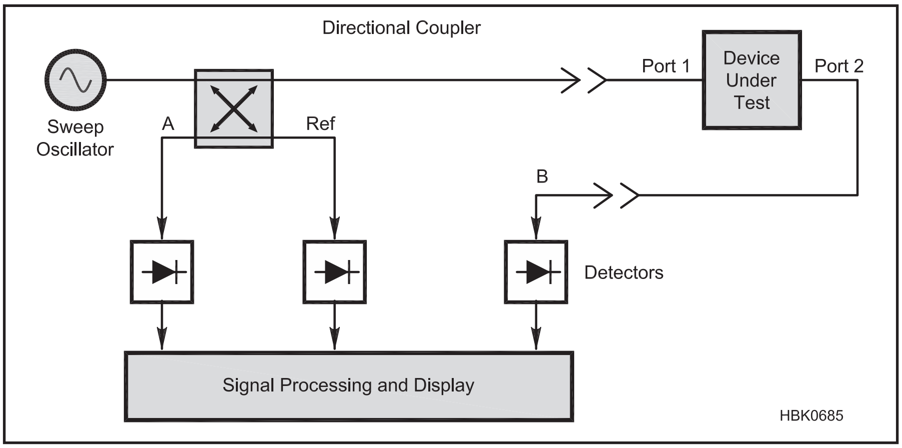


Fig 25.36 — Block diagram of a scalar network analyzer (SNA). A vector network analyzer has the same block diagram except that the detectors can measure phase as well as amplitude.

The box labeled "Signal processing and display" includes circuitry to take the logarithm of the signals so they can be displayed in dB. In modern instruments it normally includes a microprocessor, which does much of the signal processing as well as controlling the sweep oscillator and other circuitry.

A *vector network analyzer* (VNA) is an instrument that can measure both the magnitude and phase of the gain and return loss of a two-port network. The block diagram is identical to Fig 25.36 except that the detectors can measure not only the magnitude, but also the phase of each signal with respect to the reference, REF. The measured data are usually presented in terms of *scattering parameters*, or *S parameters*. S_{11} is the ratio of the incident to the reflected signal at port 1, S_{22} is the equivalent on port 2, S_{21} is the gain from port 1 to 2, and S_{12} is the gain from port 2 to 1. Each S parameter is a complex number, representing both magnitude and phase. Further discussions of S parameters can be found in the **Computer-Aided Circuit Design** and the **RF Techniques** chapters. Usually the VNA provides the capability to plot the S parameters on a Smith chart overlaid on the display for easy evaluation. The Smith chart is discussed in the **Transmission Lines** chapter.

The accuracy of a VNA depends on the performance of the directional couplers, detectors, and other circuitry. The coupler directivity is a key contributor to measurement errors. For example, if the directivity is 20 dB, then even a perfect 50-Ω load will show 20 dB of return loss, equivalent to a 1.22:1 SWR. Fortunately it is possible to eliminate most errors by measuring several terminations of known impedance and calculating a series of correction factors. The mathematics involved is quite complex, but it is all handled by the instrument software and is invisible to the user. The most popular technique is called a Short-Open-Load-Through (SOLT) calibration because it uses four calibration standards, a short-circuit, an open-circuit, a 50-Ω load and a through connection between the two ports. One of the sources of error removed by the calibration is any error due to the length of the coaxial cables that connect to the device under test. For that reason, the calibration standards should always be connected at the end of the cables that are to be used for the test. A thorough discussion of VNA error-correction techniques can be found in an Agilent Technologies application note.[15]

CHOOSING A NETWORK ANALYZER

A spectrum analyzer with tracking generator can be the basis of a high-performance scalar network analyzer. Referring to Fig 25.36, the only pieces that need to be added are the directional coupler and a pair of 50-Ω loads. The tracking generator takes the place of the sweep oscillator. Connect the spectrum analyzer input to A, B or REF and terminate the other two with 50-Ω loads. Unlike a conventional scalar network analyzer, a spectrum analyzer has a tuned receiver so it does not respond to spurious frequencies generated by the sweep oscillator or the device under test.

At the other end of the performance "spectrum" a simple SNA for measuring gain or attenuation can be made with any sweep oscillator, perhaps a function generator with sweep capability, plus a detector of some kind and an oscilloscope to display the detected voltage. If the oscilloscope bandwidth is great enough to cover the desired frequency range, no detector is needed and the RF signal can be displayed directly. Dynamic range is limited with this technique since the displayed signal is linear and not logarithmic. However it is sufficient for many purposes, such as measuring the passband ripple of a filter or the gain versus frequency of an amplifier.

Used scalar network analyzers are readily available on the surplus market. The sweep oscillator is normally a separate unit. Vector network analyzers also may be purchased surplus. Some require an external sweep oscillator while others include an internal signal source. The tests sets may be internal or external as well.

In recent years, several amateur VNA designs have appeared. These tend to be much smaller, lighter and lower-cost than commercial units. One by DG8SAQ down-converts the signals to be detected to frequencies in the audio range so they can be fed to the sound card on a personal computer.[16] It uses a pair of Analog Devices DDS chips for the sweep oscillator and the down-converter's local oscillator. All signal processing and control is done in the microprocessor in the PC. Paul Kiciak, N2PK, also developed a VNA using DDS chips for the source and a direct-conversion architecture which performs the detection function at dc. Like the DG8SAQ design, it relies on a PC for the processing, control, and display. His design has become available in kit form from several suppliers, and numerous third-parties have written software to support it. A number of VNA kits and products are available, most using a PC as a basis.[16,17,18]

Many VNAs have a low-frequency limit in the range of 100 kHz to 1 MHz. This limit is imposed by the variable frequency oscillator of the analyzer. The range of the VNA can be extended lower to the audio frequency range by using an external adapter that shifts the output RF signal from the VNA, applies it to the circuit under test, and then re-shifts the circuit's output signal into the VNA's RF range. A low frequency adapter project is described later in the projects section at the end of this chapter.

25.4.6 RF Signal Generators

An *RF signal generator* is a test oscillator that generates a sine-wave signal that has an accurately-calibrated frequency and amplitude over a wide radio-frequency range. Usually AM and FM modulation capability is provided and sometimes other modulation types as well.

Some instruments also have built-in sweep capability. If not, narrow-band sweep can be obtained by feeding a sawtooth waveform into the FM input. Another feature that is very useful for transceiver testing is *reverse power protection*. It protects the sensitive output circuits in the event that the transceiver accidentally goes into transmit mode while connected to the signal generator.

Before the days of modern digital electronics, all signal generators used free-running oscillators. There is a wide variation in frequency stability and accuracy between the best and the worst. Nearly any modern synthesized signal generator has good enough accuracy for most amateur purposes. However that accuracy sometimes comes at the price of phase noise, a type of wideband noise caused by short-term fluctuations in phase that usually drops off gradually from the carrier frequency. The old tube-type Hewlett-Packard 608-series signal generators have better performance in this respect than some modern synthesized instruments. See **Fig 25.37**. Phase noise is discussed in the **Oscillators and Synthesizers** chapter.

Some signal generators, even professional laboratory-grade types, may not have enough output attenuation to accurately measure the minimum discernible signal (MDS) on a sensitive communications receiver. A 10-dB noise figure in a 500 Hz bandwidth implies an MDS of –137 dBm, which is well below the minimum amplitude level on many signal generators. The problem is easily resolved by adding an external fixed 20-dB attenuator. Just remember to subtract 20 dB from all amplitude readings.

Many signal generators include some way to connect them to a computer so they can be controlled for automated testing. That is useful for testing a receiver at many frequencies across a band to be able to plot sensitivity or dynamic range as a function of frequency, for example. Older instruments usually have a GPIB interface while modern ones are more likely to have RS-232, USB or Ethernet.

Probably the most common use for signal generators is for receiver testing. That is covered in detail in a later section. They are also useful for general-purpose test signals. For example, when developing a new receiver design, you could test the RF, IF and audio stages before the VFO is completed by using a signal generator as the local oscillator.

Critical specifications include the frequency range, frequency accuracy and stability, amplitude range, amplitude accuracy, and the spectral purity, including phase noise, harmonics and non-harmonic spurious emissions. Modulation accuracy and the capabilities of the internal modulation generator, if present, may be important as well, depending on the application.

Many inexpensive signal generators made for the hobbyist and consumer electronics service industry are not suitable for measuring receiver sensitivity because when the output attenuator is at maximum, there is more signal leaking out the cabinet and radiating from the power cord than is coming out the coax connector. Such instruments also tend to have poor frequency stability and the tuning dial tunes so fast that they are difficult to set to a precise frequency. They are useful for simple troubleshooting but are not capable of making accurate measurements.

Better-quality signal generators intended for servicing land-mobile and other communications equipment sometimes become available on the surplus market. They don't have the precision specifications of lab instruments but are generally rugged, reliable and easy to use.

Laboratory-grade signal generators are generally too expensive for the home hobbyist if purchased new but older models such as the HP8640B or its military version (AN/USM323) are widely available and have excellent specifications. Of course they may be less reliable than new instruments but in many cases the older technology is repairable without special equipment. Service manuals are sometimes available from the manufacturer on their website.

Fig 25.37 — A classic Hewlett-Packard 608F tube-type signal generator. While not as stable as modern synthesized signal generators, in other respects the performance is quite up to date.

25.4.7 Using Noise Sources

Most of the time, noise is something to be avoided in electronic circuits. It can be quite useful for testing, however, precisely because the calibrated noise that you create on purpose has the same properties as the unwanted circuit noise that you are trying to minimize. Thus, calibrated sources of noise are used as test instruments for various measurements. (See the **RF Techniques** chapter for a discussion of noise and its associated terminology.)

MEASURING RECEIVER NOISE FIGURE

A noise source can be used to measure the noise figure of an SSB or CW receiver. The *excess noise ratio (ENR)* is the ratio of the noise added by the noise source to the

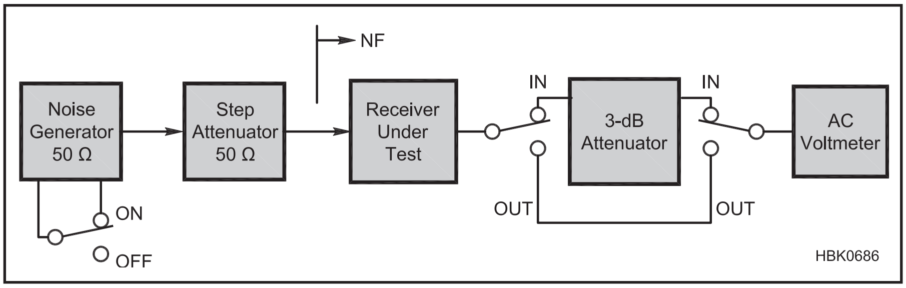


Fig 25.38 — Setup for measuring receiver noise figure.

thermal noise level, normally expressed in dB. An explanation of noise figure is given in the **RF Techniques** chapter. The basic idea is to connect the noise source to the receiver antenna input and then measure the difference in the noise level at the receiver speaker terminals when the noise source is turned on or off. When it is off, the measured noise contains only thermal noise and the noise contributed by the receiver. When it is on, the noise includes the excess noise from the noise source. The noise figure can then be calculated from the ratio of those two values,

$$NF_M = ENR - 10\log\left(\frac{E_{ON}{}^2}{E_{OFF}{}^2} - 1\right)$$

where NF_M is the measured noise figure in dB, ENR is the excess noise ratio in dB, E_{ON} is the RMS output noise voltage with the noise source on, and E_{OFF} is the RMS output noise voltage with the noise source off.

Rather than using the equation, another technique is to include a step attenuator with 1-dB steps at the noise source output. See **Fig 25.38**. Adjust the attenuator so that the noise increases by 3 dB (1.414 times the voltage) when the noise source is turned on. The noise figure is then simply the ENR of the noise source minus the attenuation.

The nice thing about measuring noise with noise is that the accuracy of the voltage reading is not important so long as the ratio of the two readings is accurate. A non-RMS ac voltmeter is not accurate when measuring noise, but so long as the inaccuracy is the same for both voltage levels the ratio is correct. When using a non-RMS instrument, follow the step attenuator technique described in the previous paragraph along with a 3-dB attenuator (1/1.414 voltage divider) in front of the ac voltmeter. First take a voltmeter reading with the noise source off and the 3-dB attenuator bypassed. Then turn on the 3-dB attenuator and the noise source and adjust the step attenuator for the same reading. The noise figure is the ENR of the noise source minus the attenuation.

There are some potential errors to watch out for. The receiver AGC must be turned off for this test so that the gain does not change when the noise source is turned on or off. In addition, the audio gain and perhaps the IF gain must be adjusted so that there is no clipping of the noise peaks under any condition. With the noise source turned on, the RMS value of the noise at the speaker output should be adjusted to no more than ⅓ the maximum output level to prevent the peaks from clipping.

OTHER USES FOR A NOISE SOURCE

You can take advantage of the flat spectrum of a noise source to measure the frequency response of RF devices. It requires either a spectrum analyzer or a receiver that can be tuned manually across the frequency range of interest. With sufficient averaging, the output noise level at each frequency accurately reflects the response of the device at that frequency. For example it is quite easy to measure the total response of a receiver from antenna to speaker output using an RF noise source and free software running on a PC that turns the computer sound card into an audio spectrum analyzer. With the noise source connected to the receiver antenna and the audio output connected to the sound card input, the total response, including the IF crystal filter and all the audio stages, is shown by the average noise level in the spectrum analyzer window.

In addition to receivers, noise sources can also be used to measure the noise figure of other devices such as amplifiers. The technique is basically the same, except that the output signal to be measured is at a radio frequency instead of audio. Another difference is that the noise figure of the measuring instrument can affect the results. That is almost never a concern when measuring receivers since the gain from the antenna to the speaker output is normally so high as to swamp out the effect. The corrected noise figure is

$$NF = 10\log\left[10^{NF_M/10} - 10^{(NF_{NMI} - G)/10}\right]$$

where NF_M is the measured noise figure in dB, NF_{NMI} is the noise figure of the noise-measuring instrument in dB, and G is the gain of the device under test in dB. For that equation to be valid, the bandwidth of the measuring instrument must be no greater than the bandwidth of the device under test.

Commercial noise figure meters combine everything you need to perform noise figure measurements of amplifiers and other devices into one box. While the noise source itself is usually a separate module, the switched power is provided by the noise figure meter. Calculations are performed internally and the noise figure reads out directly on a meter or display. The noise source is repeatedly switched on and off to produce a continuously-updated noise figure reading, which is handy for adjusting or tuning the device under test. Older units typically operate at a small number of fixed intermediate frequencies and it is up to the user to heterodyne the signal to the IF. Many modern units operate over a wide range of frequencies and some can sweep to provide a plot of noise figure versus frequency. Agilent Technologies has a good application note on noise figure measurement that is available for free download.[19]

26.4.8 RF and Microwave Test Accessories

The following section assumes that all coaxial devices are designed for 50-Ω characteristic impedance. While that is by far the most common value for Amateur Radio systems, most of the information applies as well for other impedances, such as 75 Ω which is

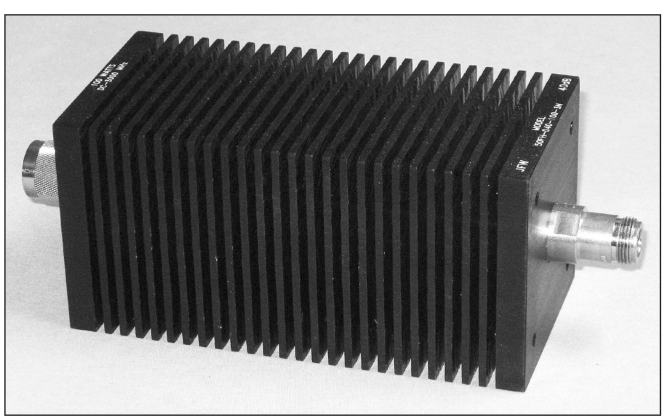


Fig 25.39 — A 100-watt power attenuator rated for dc-3000 MHz. The transmitter is connected to the male type-N connector. The female connector is for the low-power connection.

common in the video and television industry.

ATTENUATORS AND TERMINATIONS

Most amateurs are familiar with the concept of a *dummy load*, which is nothing more than a high-power 50-Ω resistor with a coaxial connector. The term comes from the fact that it is often used as a dummy antenna to provide a low-SWR load for transmitter testing. The resistor must have low stray reactance throughout the desired frequency range. Common wire-wound power resistors are not suitable for a dummy load because they have too much inductance. Commercial loads, such as the power attenuator in **Fig 25.39**, use special resistors encased in a finned enclosure for dissipating the heat generated by the power absorbed from the transmitter. One technique used in some amateur dummy loads is to enhance the power dissipation capability of less-expensive, lower-power resistors by submerging them in a bath of oil.

Low-power dummy loads, normally called *terminations*, are used whenever a device being tested needs to have one of its coaxial connectors terminated in a load that has good return loss over a wide band of frequencies. A *feed-through termination*, such as the one in **Fig 25.40**, is one that has two coaxial connectors with a straight-through connection between them as well as a 50-Ω load resistor connected between the center conductors and the shield. It is used to provide a 50-Ω load when using a measuring device with a high-impedance input, such as an oscilloscope. To avoid standing waves on the coaxial transmission line connected to the device under test, it is important to place the feed-through termination directly at the oscilloscope's input connector so that the transmission line is properly terminated.

Another technique to maintain a low SWR is to use a *T connector* at the oscilloscope input. The device under test is connected to one side of the T and some other device that provides a good 50-Ω load is connected to the other side. The high-impedance oscilloscope input does not excessively disturb the 50-Ω system. This is a good way to "tap into" a signal traveling between two devices on a coaxial line while maintaining the connection between the devices.

A *fixed attenuator* is useful both for reducing a signal level as well as for improving the 50-Ω match. For example, a 10-dB attenuator guarantees a minimum of 20 dB of return loss (1.22:1 SWR) even if the load SWR is infinite. A 20-dB attenuator makes a high-quality 50-Ω termination, even with nothing connected to its output.

A *step attenuator* is used when you need to adjust the signal level in fixed steps. The term "10-dB step attenuator" means that the step size is 10 dB. The maximum attenuation would typically be perhaps 70 to 120 dB. An attenuator with 1 dB steps and 10 dB of maximum attenuation would be called a "1 dB step attenuator" or a "0 to 10 dB step attenuator." Two attenuators, one with 10 dB steps and one with 1 dB steps, can be connected in series to obtain 1 dB resolution over a very wide range of attenuation. At microwave frequencies, continuously-variable attenuators are available for waveguide transmission lines. They can sometimes be obtained used at reasonable prices.

OTHER TEST ACCESSORIES

Other accessories available for waveguide include coax-to-waveguide adapters, detectors, directional couplers, isolators, absorption wavemeters, mixers and terminations of various kinds. Each size of waveguide covers about a 1.5:1 frequency range. Be aware that over the years there have been several systems for assigning letters to the various microwave frequency bands. For example, "X" band in the Agilent Technologies (formerly part of Hewlett-Packard) catalog is 8.2 to 12.4 GHz, but the old US Navy definition was 6.2 to 10.9 GHz, and the ITU assignment for X-band radar is 8.5 to 10.68 GHz. The IEEE standard definition is 8.0 to 12.0 GHz. When buying surplus waveguide accessories be sure they cover the frequency range you need.

A *coaxial detector* is just what the name implies, a diode detector in a coaxial package, usually with a 50-Ω RF load impedance. Silicon Schottky diodes are usually used to obtain good sensitivity although gallium arsenide devices are sometimes employed for the high microwave and millimeter-wave frequencies. Below a certain signal level, typically about –15 dBm, the detected output voltage is proportional to the square of the input voltage, that is, it is proportional to the input power. Sensitivity is typically specified in mV/mW, assuming a high-impedance load for the detected signal.

Wideband amplifiers can often be useful in test systems. They are available with various connector types in packages sized appropriately for the power level. A low-noise amplifier is useful as a preamplifier for a spectrum analyzer, for example. A higher-

Fig 25.40 — Various RF test accessories. At the rear is a 10 dB step attenuator with a 0-120 dB range. In the foreground from left to right are a BNC "T" connector, a 50-Ω feed-through termination, a 0.1-500 MHz low-noise amplifier, and a 10-dB fixed attenuator.

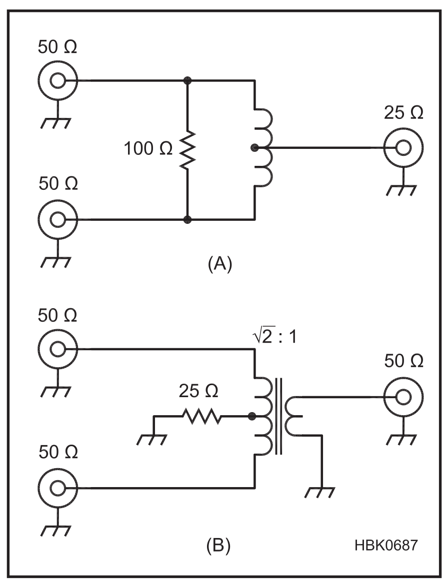


Fig 25.41 — Two forms of a lumped-element Wilkinson hybrid combiner. The one at B is 50 Ω on all three ports. In both circuits, the two input ports on the left are isolated from each other so long as the output port sees a low-SWR termination.

power amplifier can be used at the output of a signal generator either for receiver dynamic range testing or as an input signal for testing an RF power amplifier.

Directional couplers and bridges are useful network analyzer accessories as previously described but can also be used to acquire a small amount of signal for other test purposes. For example, a 30-dB coupler at the output of a 100-W transmitter produces a 100-mW signal that can be fed to a coaxial detector or a spectrum analyzer, while passing the rest of the 100-W signal through to the antenna or dummy load.

A *power divider*, also known as a *splitter*, is a device that divides an input signal equally among its output ports. A *combiner* is basically a power divider hooked up backward; the signals from the input ports are combined at the output port. Some power dividers consist only of a network of resistors. The insertion loss is typically at least 6 dB for 2-port splitters, meaning that a quarter of the power comes out each port and half the power is absorbed in the resistors.

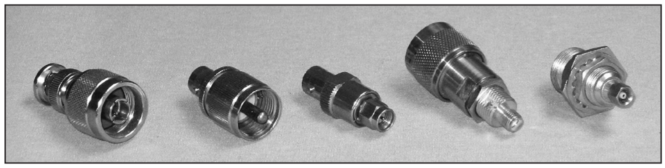


Fig 25.42 — Coaxial connector adapters. From the left are: BNC male to type-N male, BNC female to UHF male, BNC female to SMA male, type-N male to SMA female, and type-N female to SMC male.

A *hybrid combiner*, such as the *Wilkinson combiner* shown in **Fig 25.41**, includes a transformer as part of the network. It has the advantage that, so long as there is a good, low-SWR match on the output port, there is excellent isolation between the two input ports. That is very useful when combining the signals from two signal generators for high-level dynamic range testing of a receiver. The port-to-port isolation prevents the two signals from combining in the output amplifiers of the signal generators, where they could generate distortion products greater than the ones generated by the receiver that you are trying to measure. The insertion loss is nominally 3 dB.

The traditional Wilkinson combiner has a 2:1 ratio of input to output impedance. The variation in Fig 25.41B uses a transformer with a $\sqrt{2}$: 1 turns ratio (2:1 impedance ratio) to obtain 50 Ω on all ports.

Adapters such as those shown in **Fig 25.42** are useful for making connections among various pieces of test equipment.

25.5 Receiver Measurements

Receivers present a special measurement challenge because of the wide range of signal levels they must be able to handle. In a state-of-the-art receiver, the difference between the minimum discernible signal and the blocking level can be as high as 130 or 140 dB. It requires special care and high-quality test equipment to measure that level of performance.

25.5.1 Standard ARRL Lab Tests

The ARRL laboratory staff has standardized on a suite of tests for receivers, transmitters and several other types of RF equipment. The exact procedures for each test are meticulously specified in a 163-page document to ensure that tests performed at different times by different people are done in the same way. We won't attempt to repeat that level of detail here but rather describe the tests in general terms suitable for use by the home experimenter. (The full set of procedures is available online in the Product Review Testing area of the ARRL website at **www.arrl.org/product-review**.) Additional information on receiver design and performance is available in the **Receivers** chapter.

RECEIVER SENSITIVITY

Several methods are used to determine receiver sensitivity. The modulation mode often determines the best choice. One of the most common sensitivity measurements is *minimum discernible signal* (MDS) or *noise floor*, which is suitable for CW and SSB receivers. The minimum discernible signal 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. Be certain that the receiver is peaked on the generator signal. In **Fig 25.43** a step attenuator is included at the receiver input since most signal generators cannot accurately generate the low signal levels required for MDS measurements. An audio-frequency ac voltmeter is connected to the receiver's speaker terminals. If no speaker is connected, then a resistor of the same resistance as the speaker impedance should be substituted. Set the receiver to CW mode with a bandwidth of 500 Hz, or the nearest to that bandwidth that is available. Since the noise power is directly proportional to the bandwidth, always use identical filter bandwidths when comparing readings. Turn off the AGC, if possible. With the generator output turned off, you should hear nothing but noise in the speaker. Note the voltmeter reading at the receiver audio output. Next, turn on the generator and increase the output level until the voltmeter shows a 3-dB increase (1.414 times the voltage). The signal input at this point is the minimum discernible signal, which can be expressed in μV or dBm.

In the hypothetical example of Fig 25.43, the signal generator was adjusted to –133 dBm to cause the 3-dB increase in audio output power and the step attenuator is set to 4 dB. MDS is calculated with this equation:

MDS = –133 dBm – 4 dB = –137 dBm

where the MDS is the minimum discernible signal and 4 dB is the loss through the attenuator.

Note that the voltmeter really should be a true-RMS type to accurately measure the RMS voltage of the noise. A typical average-reading multimeter is calibrated to indicate the RMS voltage of a sine wave but reads low on Gaussian noise. The error is small for the MDS test, however. To correct for the error, adjust the signal generator for a 3.2-dB increase (1.445 times the voltage) instead of 3.0 dB.

Another issue to watch out for is that the

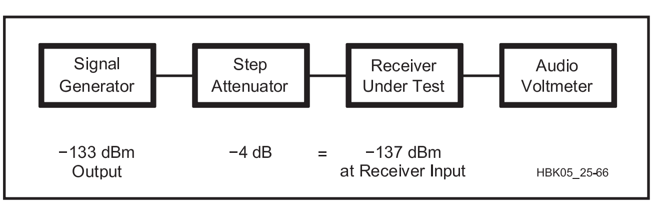


Fig 25.43 — A general test setup for measuring receiver MDS, or noise floor. Signal levels shown are for an example discussed in the text.

peak noise voltage is much greater than the RMS, typically by a factor of 4 or 5. The receiver volume should be adjusted so that the RMS output voltage is no more than about one-fifth the clipping level of the audio amplifier.

For AM modulation, receiver sensitivity is expressed as the RF signal level that results in a 10 dB signal plus noise to noise ratio at the audio output. The AM signal is 30% modulated with a 1000 Hz tone. Otherwise, the test setup is the same as for the MDS test. The signal generator output level is adjusted until there is a 10 dB (3.16 voltage ratio) increase in audio output voltage when the modulation is switched from off to on. Again, there is an error if a typical average-reading multimeter is used instead of a true-RMS voltmeter. For the AM sensitivity measurement, the error is 0.8 dB. Adjust the signal generator output to give a 10.8 dB (3.47 voltage ratio) increase in signal when the modulation is turned on. If an audio distortion meter is used in place of the RMS voltmeter, you can leave the modulation turned on and adjust the signal generator output level for 31.6% distortion.

For FM modulation, receiver sensitivity is expressed as the RF signal level that results in 12 dB SINAD. SINAD stands for "signal plus noise and distortion" and is calculated from

$$\text{SINAD} = 10 \log \left[\frac{\text{signal} + \text{noise} + \text{distortion}}{\text{noise} + \text{distortion}} \right] \text{dB}$$

where signal, noise and distortion are all entered in units of power (watts or milliwatts). It is very similar to the signal plus noise to noise ratio used for AM testing except that any distortion in the audio signal is added to the noise measurement. It means, however, that the signal, the noise and the distortion must all be measured at the same time, without turning off the modulation. For that, a distortion analyzer is needed as shown in **Fig 25.44**. A distortion analyzer includes a switchable band-reject filter to null out the 1 kHz tone for the noise-and-distortion measurement. The filter is bypassed for the signal-plus-noise-and-distortion measurement. For the test, the signal generator is set for FM modulation with a 1 kHz tone and the desired FM deviation, normally 3 kHz for VHF repeater operation. Adjust the signal generator output level until the distortion analyzer indicates 25% distortion, which is equivalent to 12 dB SINAD. Don't forget to subtract the attenuation of the step attenuator from the signal generator output level, as was done for the MDS test.

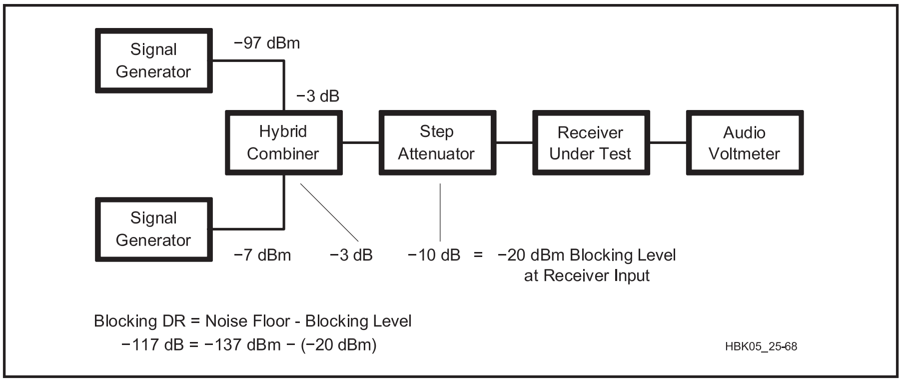


Fig 25.45 — Receiver blocking dynamic range is measured with this equipment and arrangement. Signal levels shown are for the example discussed in the text.

Noise figure is a measure of receiver sensitivity that, unlike the other methods presented so far, is independent of the receiver bandwidth and operating mode. It can be calculated from the MDS so long as the noise bandwidth of the receiver filtering is known. *Noise bandwidth* is the bandwidth of a hypothetical perfect filter with a rectangular spectrum shape that would produce the same total noise power as the receiver filter. Accurate measurement of a filter's noise bandwidth requires integrating the spectral response using a swept signal source, but the filter's 3 dB bandwidth can be used as a reasonable approximation. The formula is

$$\text{NF} = \text{MDS} - \left(10 \log (\text{BW}) - 174\right) \text{ dB}$$

where MDS is the minimum discernible signal in dBm and BW is the noise bandwidth in Hz. Assuming the noise bandwidth is 500 Hz, we get NF = MDS + 147 dB. For example, if the MDS is –137 dBm then NF = –137 + 147 = 10 dB.

RECEIVER DYNAMIC RANGE

Dynamic range is a measure of a receiver's ability to receive weak signals without being overloaded by strong signals. It is easy to design a receiver with good sensitivity to weak signals. It is also easy to design a receiver that is not overloaded by strong signals. It is much more difficult to design a receiver that can do both at the same time.

Blocking dynamic range (BDR) is the difference between the noise floor and the signal level at which *blocking* occurs, that is, the level that causes a 1 dB reduction in gain for nearby weaker signals. The noise floor is just the MDS and can be measured using the technique described in the previous section.

The blocking level is measured using a test setup similar to the one used for measuring MDS except that two signal generators are connected to the input through a hybrid combiner, as shown in **Fig 25.45**. The receiver AGC should be turned off for this test. The mode is set to CW and the bandwidth to 500 Hz or the closest available. Two signal generators are used. One generates the weak signal that the receiver is tuned to. The ARRL standard specifies –110 dBm at the receiver input, which requires –97 dBm at the input to the hybrid combiner, assuming it has 3 dB loss. The other signal generator generates the strong interfering signal on a nearby frequency. Standard frequency separations are plus and minus 20, 5 and 2 kHz. The level of the strong signal is increased until the level of the weaker signal measured at the receiver audio output decreases by 1 dB.

Referring to Fig 25.45, the blocking level is the level from the signal generator minus the loss of the hybrid combiner and attenuator,

$$\text{BL} = -7 - 3 - 10 = -20 \text{ dBm}$$

The blocking dynamic range is given by

$$\text{BDR} = \text{BL} - \text{MDS} = -20 - (-137) = 117 \text{ dB}$$

assuming an MDS of –137 dBm as in the previous examples.

One complication is that it may be difficult to measure the amplitude of the audio tone because of the presence of noise caused by the phase noise of the signal generator and the receiver's local oscillator, especially at the 2 and 5 kHz frequency spacings. The solution is to use an audio-frequency spectrum analyzer to measure the change in tone amplitude. The

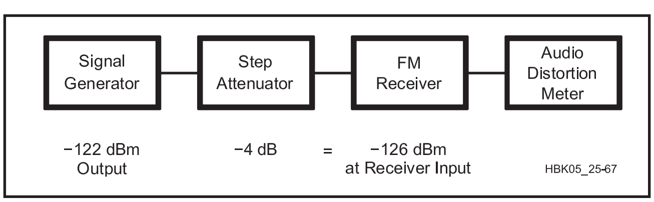


Fig 25.44 — FM SINAD test setup.

absolute accuracy of the spectrum analyzer is not important so long as it can accurately show a 1-dB change in signal level. An instrument based on a computer sound card and free spectrum analysis software should be adequate.

Reciprocal mixing is the name for the mixing of a nearby interfering signal with the phase noise of the receiver's local oscillator, which causes noise in the audio output. Although the ARRL BDR test eliminates this effect from the measurement, in actual on-the-air operation the phase noise is often the factor that limits the effective dynamic range. To address this issue, the ARRL test suite includes a separate measurement for reciprocal mixing. The test setup is the same as for MDS except that the signal generator in Fig 25.43 is replaced with a low-phase-noise crystal oscillator with an output power level of +15 dBm. The step attenuator block should include both a unit with 10-dB steps to be able to adjust the signal level over a wide range as well as a 1-dB step attenuator for fine adjustment. The receiver is tuned to 20, 5 or 2 kHz above or below the oscillator frequency. The output noise level is first measured with the oscillator turned off, then the oscillator is turned on and the step attenuator gradually reduced until the noise increases by 3 dB. Reciprocal mixing is expressed as a negative number:

$$\text{reciprocal mixing} = \text{MDS} - (+15 - \text{A})\ \text{dB}$$

where MDS is the noise floor, +15 is the signal level of the crystal oscillator in dBm, and A is the total attenuation in dB.

Intermodulation distortion (IMD) means the creation of unwanted signals at new frequencies because of two or more strong interfering signals modulating each other. If there are two interfering signals at frequencies f1 and f2, then *second-order IMD* products occur at f1 + f2. If f1 and f2 are close together, then the second-order products occur near the second harmonics. *Third-order IMD* products occur at 2f1 – f2 and 2f2 – f1. If f1 and f2 are close together, then the third-order products occur close by. For example, if f1 and f2 differ by 10 kHz, then the third-order IMD products are 10 kHz above the higher frequency and 10 kHz below the lower.

The two-tone IMD test setup shown in **Fig 25.46** is the same for second and third-order IMD. To obtain sufficient output power and isolation of the two signal generators, it may be necessary to follow each one with a wide-band power amplifier, not shown. The receiver under test is set to receive CW with the same bandwidth as for the MDS test and is tuned to the frequency of the distortion product to be measured. The two signal generators are always set to the same output amplitude level, which is increased until IMD products equal in amplitude to the noise floor appear, resulting in a 3-dB increase in the audio voltmeter reading. The *IMD dynamic range* is then the difference in dB between the level of the interfering signals and the MDS.

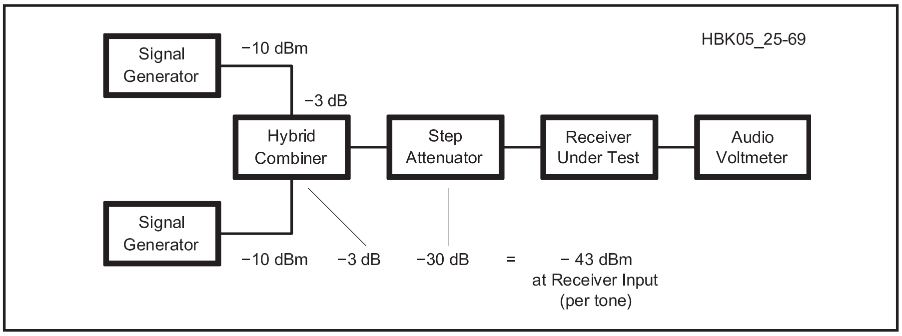


Fig 25.46 — The test setup for receiver intermodulation distortion dynamic range. Signal levels shown are for the example discussed in the text.

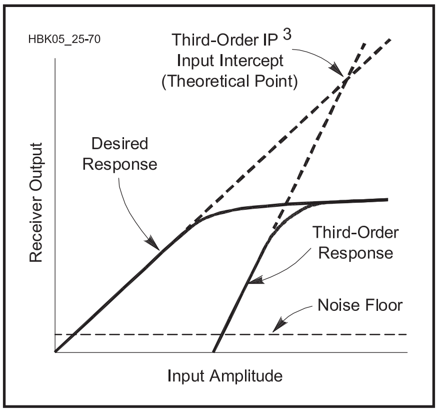


Fig 25.47 — The third-order intercept point can be determined by extending the lines representing the interfering signal level and the third-order intermodulation products on a plot of the signal levels in dB.

For the second-order IMD measurement, the standard ARRL test sets the two signal generators to 6.000 MHz and 8.020 MHz and the receiver is tuned to 14.020 MHz. For the third-order IMD measurements, the two signal generators are set to frequencies 20, 5 and 2 kHz apart, separated from the receiver frequency by plus and minus 20, 5 and 2 kHz so that the lower or upper IMD product falls within the receiver passband.

For example, tune the receiver to 14.020 MHz and the two signal generators to 13.980 MHz and 14.000 MHz. With the signal generators turned off, measure the noise level with the audio voltmeter. Turn on the signal generators and increase their amplitudes until the voltmeter shows a 3-dB increase. If the signal generator amplitudes are –10 dBm, the loss in the hybrid combiner is 3 dB, and the step attenuator is set to 30 dB attenuation, then the third-order IMD level is

$$\text{IMD} = -10 - 3 - 30 = -43\ \text{dBm}$$

If the MDS is –137 dBm, then the third-order IMD dynamic range is

$$\text{IMD\_DR} = \text{IMD} - \text{MDS} =$$
$$-43 - (-137) = 94\ \text{dB}$$

As with the blocking dynamic range test, the phase noise of the signal generators and the receiver LO may obscure the IMD product being measured. Again, the solution is to use an audio spectrum analyzer to measure the tone amplitude. Calibrate the amplitude by temporarily tuning one of the signal generators to the receiver frequency and setting the amplitude level so that the signal level at the receiver is the MDS. Note the level on the spectrum analyzer. Then return the signal generator to the interfering frequency and adjust the signal generators' amplitudes until the IMD product is at the same level as the MDS signal. The ARRL test bench actually uses a third signal generator and a second hybrid combiner at the receiver input to generate the calibration signal so that it and the IMD product can be seen on the spectrum analyzer at the same time for a more-accurate comparison.

In most analog components such as mixers and amplifiers, the second-order products increase in amplitude by 2 dB for each 1 dB increase in the interfering signals and third-order products increase 3 dB per dB. If the output signal levels are plotted versus the input levels on a log-log chart (that is, in units of dB), the desired signal and the undesired IMD products theoretically trace out straight lines as shown for third-order products in **Fig 25.47**. Although the IMD products increase more rapidly than the desired signal, the lines never actually cross because blocking occurs before that level, however, the point where the extensions of those two lines cross is called the *third-order intercept point* (IP3).

Although the third-order intercept is an artificial point, it is a useful measure of the strong-signal-handling capability of a receiver. It can be calculated from the equation

$$IP3 = MDS + 1.5 \times IMD\_DR \text{ dBm}$$

where IP3 is the third-order intercept point, MDS is the minimum discernible signal in dBm and IMD_DR is the third-order IMD dynamic range in dB. Using the numbers from the previous example, IP3 = –137 + 1.5 × 94 = +4 dBm.

The second-order intercept may be calculated in an analogous way.

$$IP2 = MDS + 2 \times IMD\_DR$$

where IP2 is the second-order intercept and IMD_DR is the second-order IMD dynamic range in this case.

An alternate method of measuring third-order IMD is to use S5 (–97 dBm) instead of the MDS as the reference level to which the IMD products are adjusted. That results in a higher IMD level but a smaller value of IMD dynamic range (the difference between the IMD level and the reference). It may be a more accurate method of determining the third-order intercept because the signal levels do not have to be measured in the presence of noise that is at the same level as the signal.

The third-order intercept is generally not a valid concept for software-defined receivers (SDRs) that do not use an analog front end. Some SDRs do not use a mixer but feed the signal from the antenna directly to an analog-to-digital converter (ADC). ADCs usually do not exhibit the 3 dB per dB relationship between signal level and third-order products, at least over major portions of their operating range. Comparing third-order dynamic range measurements of an SDR and a conventional analog radio may give misleading results.

Fig 25.48 shows the relationship between the various dynamic range values. The base line represents different power levels, with very small signals at the left and large signals at the right. The numbers listed are from the previous examples for a typical receiver. The thermal noise and the noise floor are referenced to a 500 Hz bandwidth. The third-order IMD dynamic range is less than the blocking dynamic range, which means that signals as low as –43 dBm may cause IMD interference while signals must exceed –20 dBm to cause blocking. However, the intermodulation distortion may actually cause fewer problems since the IMD products only appear at certain discrete frequencies. A signal that exceeds the blocking level can cause interference across the entire band.

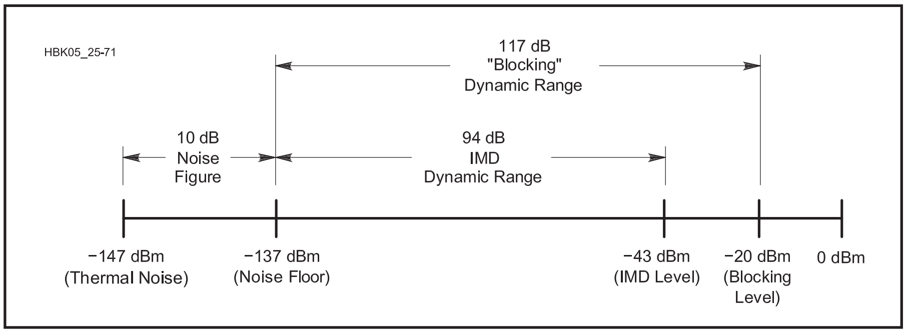


Fig 25.48 — Performance plot of the example receiver discussed in the text.

Third-order IMD dynamic range may also be measured on an FM receiver. The test setup is the same as Fig 25.46 except that the audio voltmeter is replaced with an audio distortion meter, as in Fig 25.44. The frequencies involved are the same as in that example. For this test, one of the signal generators (the one tuned to the frequency farthest from the receiver under test) is FM-modulated with a 1000 Hz tone at 3 kHz deviation. That causes the IMD products also to be FM-modulated, which can be measured with the distortion meter. The signal generator amplitudes are increased until the distortion product produces 12 dB SINAD (25% distortion) on the meter. The FM third-order IMD dynamic range is calculated using the same equation as for SSB and CW except that the SINAD sensitivity is substituted for the MDS. For example, if the 12 dB SINAD sensitivity is –120 dBm, the signal generator outputs are –10 dBm, the combiner loss is 3 dB, and the step attenuator is set to 30 dB, then

$$FM\_IMD\_DR = -10 - 3 - 30 - (-120) = 77 \text{ dB}$$

where FM_IMD_DR is the third-order FM IMD dynamic range.

OTHER ARRL RECEIVER TESTS

Most modern communications receivers are superheterodyne types. The first *IF rejection* and *image rejection* test measures the signal levels at the intermediate frequency and the image frequency that produces an audio output signal equivalent to the MDS, or noise floor. The test setup is the same as for the MDS test, as shown in Fig 25.43. The receiver is set to CW mode and 500 Hz bandwidth. The signal generator is tuned to the receiver first intermediate frequency or to the image frequency, which is the receiver frequency plus or minus two times the IF. The signal generator output level is gradually increased until there is a 3-dB increase in the audio voltmeter reading. As with the MDS test, if a multimeter is used instead of a true-RMS voltmeter, the signal generator should be adjusted for a 3.2-dB increase. The IF or image rejection is just the signal generator level at the receiver input less the MDS. For example, if the MDS is –137 dBm, the signal generator level is –40 dBm, and the attenuator is set for 10 dB, then the IF or image rejection is IR = –40 – 10 – (–137) = 87 dB.

FM adjacent-channel rejection is a measure of an FM receiver's ability to detect a weak signal in the presence of a strong interfering FM-modulated signal on an adjacent frequency channel. The test setup is the same as for the two-tone IMD dynamic range test shown in Fig 25.46 except that a distortion meter is substituted for the audio voltmeter, as was done for the FM SINAD test illustrated in Fig 25.44. The standard channel spacing is 20 kHz. The weak signal is modulated with a 1000 Hz tone and the interfering signal with a 400 Hz tone, both with 3 kHz deviation. The receiver under test is set for FM modulation and is tuned to the frequency of the weak signal. The weak signal is adjusted for 12 dB SINAD (25% distortion) on the distortion meter and then the interfering signal level is increased until the SINAD drops to 6 dB (50% distortion). The adjacent channel rejection is the difference between the power of the interfering signal and the 12 dB SINAD sensitivity, which is just the difference in level between the two signal generators.

Another test that applies especially to FM receivers is the squelch sensitivity test. As usual, the signal generator is set for FM modulation with a 1 kHz tone and 3 kHz deviation. If squelch is available for SSB, it can be tested in that mode as well. With the signal generator off, the squelch control on the receiver is adjusted just to the point where the noise is squelched. Then the signal generator is turned on and the level increased until the signal is heard.

Audio power output is tested with the same setup as the FM SINAD test illustrated in Fig 25.44. The receiver under test is set for SSB mode with the widest bandwidth available. A load resistor of the specified resistance, usually 8 Ω, is connected to the speaker output in place of the speaker. The signal generator level is set for an S9 level and the receiver is tuned for a 1 kHz output tone frequency. The receiver volume control is increased until the specified distortion level, usually 10%, is indicated on the distortion meter. The output power is given by the equation

$$P = V_{RMS}^2 / R$$

where P is the power in watts, V_{RMS} is the RMS output voltage, and R is the load resistance.

The audio and IF frequency response measures the audio frequencies at which receiver audio output drops by 6 dB from the peak. It includes the total response of the entire receiver, from the antenna connector to the speaker output. The test setup is the same as for the audio power output test except that some method must be included to measure the output audio frequency, such as a digital oscilloscope or frequency counter. The receiver is set for SSB mode at the bandwidth to be tested and the AGC is turned off. First tune the signal generator for a peak audio output signal and record the level of the audio signal. Then tune the signal generator downward until the signal drops 6 dB (½ the voltage) and record the audio frequency. Then tune the signal generator upward and record the high-end frequency at which the signal drops by 6 dB. The 6-dB bandwidth is the difference between the two frequencies.

There are several other miscellaneous ARRL receiver tests that are commonly reported in product reviews. The S meter test measures the signal level required to produce an S9 indication on the meter. The notch filter test uses a setup similar to the IMD dynamic range test in Fig 25.46 with an audio spectrum analyzer at the output. One signal is notched and the other is used as a level reference; the notch depth is the amplitude difference between the two tones. The DSP noise reduction test uses a similar setup except that one signal generator is replaced with a wideband noise generator. The signal generator is adjusted for S9 and the noise source for a 3-dB increase in the audio voltmeter so that the noise and signal are at the same level. The DSP noise reduction is then turned on and the reduction in noise level recorded.

25.5.2 Other Common Receiver Measurements

There are a few other useful receiver tests that are not covered in the standard ARRL test suite. For example, in addition to the IF and image response, it is possible for a receiver to have spurious responses at other frequencies as well. Testing for that can be a time-consuming process since it involves tuning the signal generator through a wide frequency range with the receiver tuned to each of a number of representative frequencies, typically at least one on each band. Start with the signal generator at maximum output power. When you find a response, reduce the level until the received signal level is at the noise floor (MDS). The spurious response amplitude is the difference between the signal generator level and the MDS.

Another time-consuming test is for internally-generated spurious signals, sometimes called *birdies* because they sometimes sound like a bird chirping as you tune through the signal. Birdies are typically caused by harmonics of the local oscillator(s) and BFO and their IMD products. For this test, connect a 50-Ω termination to the antenna connector and tune the receiver through its entire frequency range, writing down the frequency and S meter reading for each spurious signal found. You must tune very slowly because many birdies pass through the IF passband much faster than regular signals, sometimes in the opposite direction.

The ARRL S meter test only measures the response at the S9 level. If the S meter is accurate throughout its range it can be used to measure signal levels off the air. The standard definition is that, on the HF bands, S9 should correspond to –73 dBm (50 μV into 50 Ω) and each S unit corresponds to 6 dB, or a doubling of RF voltage. So S8 is –79 dBm or 25 μV, S7 is –85 dBm or 12.5 μV, and so on. The S unit calibration of most commercial equipment varies considerably from 6 dB per S unit, but the S meters accuracy can easily be measured with a calibrated signal generator.

The automatic gain control circuitry has a major effect on the operation of a receiver. The static AGC performance can be measured on a CW or SSB receiver with a signal generator and an audio voltmeter connected to the speaker output. You can plot a graph of audio output in dBV (decibels with respect to one volt) versus RF input level in dBm to see how good a job the receiver does in keeping the output level constant. Some receivers have menu settings to set the slope of the curve and the threshold, which is the small signal level at which the AGC circuitry starts to reduce the gain. The dynamic response is also important, although it is difficult to measure. A short attack time is important to reduce transients on sudden strong signals, but if it is too short then in-channel intermodulation distortion between strong signals can make signals sound "mushy."

The ARRL test suite includes a measurement of the 6 dB bandwidth at the audio output, but the sound is also affected by slope and ripple in the passband response. Using the same test setup as for the 6 dB measurement, you can measure the output voltage at a number of equally-spaced frequencies and plot the results on a graph. Output power is measured at the 10% distortion level, but it also is useful to measure the distortion at a volume level closer to what is used in actual practice. An output level of 1 V_{RMS} is commonly used for that test.

Frequency accuracy and stability are important performance criteria for a receiver. If a signal generator of sufficient frequency accuracy is not available, a standard time signal such as from radio station WWV can be used. The traditional method is to put the receiver in SSB mode and *zero-beat* the carrier, which means to tune the receiver until the audio tone (the beat note) is at zero Hz. Since most receivers' audio frequency response only extends down to a couple hundred Hertz, it is difficult to get good accuracy using that method. If a frequency counter or other means of measuring the audio output frequency is available, you can tune the receiver to obtain a 1000 Hz tone and add (for LSB) or subtract (for USB) 1000 Hz from the receiver's indicated frequency. The frequency drift from cold turn-on can be measured by plotting the audio output frequency versus time.

25.6 Transmitter Measurements

The signal levels found in a transmitter do not vary nearly as much as in a receiver. All the signals are generated internally, rather than being received off the air, so are much better controlled. Partly for that reason transmitter measurements tend to be simpler than receiver measurements and there are fewer of them.

25.6.1 Standard ARRL Lab Tests

For a transmitter, the RF power output is probably the first measurement that comes to mind. The test setup is straightforward: connect the transmitter output to the input of an RF wattmeter and connect the wattmeter output to a suitable dummy load. For CW, AM and FM modes, simply key the transmitter and measure the power on the wattmeter. For AM and FM the modulation should be turned off. For SSB, a two-tone audio oscillator should be connected to the microphone input to take the place of the voice signal. For the SSB test the wattmeter must be a peak-reading type to measure the PEP power

TRANSMITTER SPECTRAL PURITY

Ideally a transmitter confines its emissions to a narrow frequency range around the desired signal. Unwanted emissions can be divided into two categories, those that fall close to the desired signal and those that extend far away.

In the latter category are included harmonics and other discrete spurious frequencies. The measurement is done with the transmit-

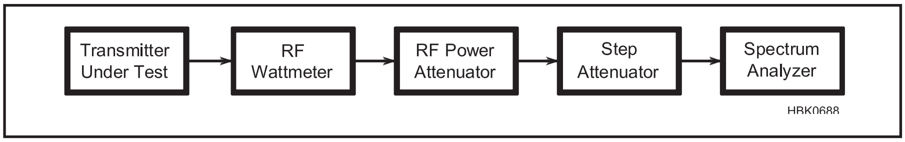


Fig 25.49 — The transmitter spectral purity test uses a spectrum analyzer to display the spurious frequencies.

ter in CW mode using the test setup shown in **Fig 25.49**. Tune up the transmitter as specified in the manual, set it for the desired power level, and adjust the frequency from one end of the band to the other while observing the spectrum analyzer. This should be done on each band. It may be necessary to retune the transmitter occasionally as the frequency is adjusted. The spurious-signal and harmonic suppression is the difference in dB between the carrier and the maximum spurious signal. It is important that the power level into the spectrum analyzer be maintained at a low enough level that spurious signals are not generated in the spectrum analyzer itself. To test for that, try changing the step attenuator setting; the desired carrier and all spurious signals should change by the same amount. If not, increase the attenuation until they do.

Representative spectrum analyzer plots are shown in **Fig 25.50**. The horizontal (frequency) scale is 5 MHz per division. The desired carrier frequency at 7 MHz appears 1.4 divisions from the left of the plot. Although not shown, a large apparent signal is often seen at the extreme left edge. This signal at zero Hz is caused by leakage of the spectrum analyzer local oscillator frequency and should be ignored.

In addition to the discrete spurious signals measured by the previous test, a transmitter may also generate broadband noise. It can be due both to the phase noise of the local oscillator as well as AM noise from all the devices in the amplifier chain. Generally the phase noise predominates, at least for frequencies close to the carrier. The composite noise test measures the total noise from both sources as well as any close-in spurious frequencies that don't show up in the wideband spurious signal and harmonic suppression test. Measuring phase noise requires high-performance test equipment and special measuring techniques.[20] The ARRL lab uses a special low-noise oscillator and a Hewlett-Packard (now Agilent Technologies) phase noise test set under computer control to perform this sophisticated measurement, which takes about 15 minutes to perform once it is set up. The result is a plot of the noise spectrum such as the one in **Fig 25.51**.

The most important unwanted emissions close to the carrier frequency are caused by distortion in the transmitter amplifier stages. In an SSB transmitter, this distortion creates

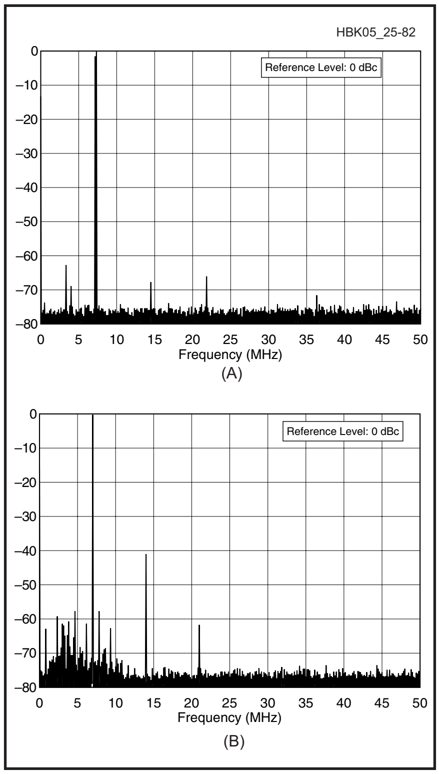


Fig 25.50 — Comparison of the spurious signal levels of two 100-watt transmitters, as shown on the spectrum analyzer display. The display on the top shows about 63 dB worst-case spurious signal suppression while the one on the bottom has a second harmonic suppressed approximately 42 dB. For transmitters below 30 MHz installed after January 1, 2003, the worst-case spurious emission must be at least 43 dB below the carrier power.

a signal that is wider than the bandwidth of the original modulation and causes interference to other stations. For this test, the same test setup is used as for the harmonics and spurious frequencies test shown in Fig 25.49 except that a two-tone audio generator is connected to the transmitter microphone input to simulate a voice signal, which contains many frequency components. The two tone frequencies must be non-harmonically related to prevent tone harmonics from being confused with IMD products. The ARRL Lab uses 700 Hz and 1900 Hz for these tests. Because many transmitters' modulation frequency response is not flat, the relative am-

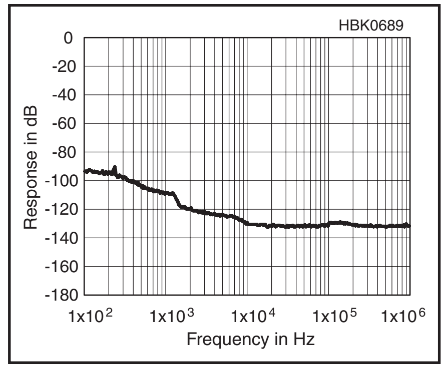


Fig 25.51 — Spectral display of an amateur transmitter output during composite-noise testing in the ARRL Lab. Power output was 200 W on the 14 MHz band. The carrier, off the left edge of the plot, is not shown. This plot shows composite transmitted noise 100 Hz to 1 MHz from the carrier on a logarithmic scale. The vertical scale is in dB with respect to the carrier.

plitudes of the two audio-frequency tones must be adjusted to obtain equal-amplitude RF tones on the spectrum analyzer. The test should be done for both lower and upper-sideband modes.

Two-tone IMD products are measured with respect to the transmitter peak-envelope power (PEP) which is 6 dB greater than the amplitude of either of the two tones. By adjusting the spectrum analyzer reference level to place the two tones 6 dB below the zero-dB reference line, as shown in **Fig 25.52**, the IMD distortion may be read out directly. For the signal shown, the third-order products are at –30 dB from PEP.

Carrier and unwanted sideband suppression may be measured with the same setup. In this case the 700 Hz tone of the two-tone generator is turned off and only the 1900 Hz tone is used. The single tone is set to the 0-dB reference line of the spectrum analyzer. For USB, the suppressed carrier shows as a small pip 1900 Hz below the desired signal and the unwanted sideband is 3800 Hz below. For LSB the unwanted signal frequencies are above the desired signal.

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 waveform and SSB/FM transmit-to-receive turnaround tests (important for many digital modes).

A typical setup for measuring CW keying waveform and time delay is shown in

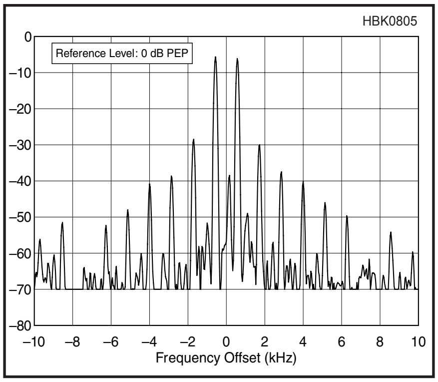


Fig 25.52 — An SSB transmitter two-tone test as seen on a spectrum analyzer. Each horizontal division represents 2 kHz and each vertical division is 10 dB. The third-order products are 30 dB below the PEP (top line), the fifth-order products are down 38 dB and seventh-order products are down 40 dB.

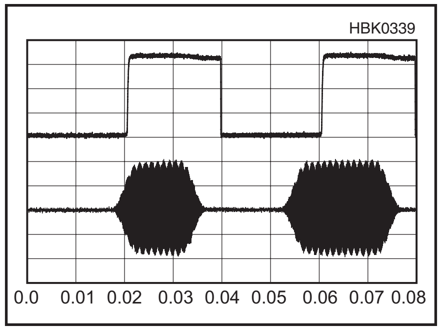


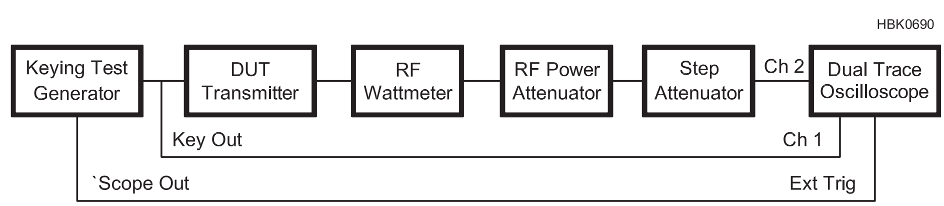
Fig 25.54 — Typical CW keying waveform for a modern Amateur Radio transceiver during testing in the ARRL Lab. This plot shows the first two dits in full break-in (QSK) mode using external keying. Equivalent keying speed is 60 WPM. The upper trace is the actual key closure; the lower trace is the RF envelope. (Note that the first key closure starts at the left edge of the figure.) Horizontal divisions are 10 ms. The transceiver was being operated at 100 W output on the 14 MHz band.

Fig 25.53. A keying test generator is used to key the transmitter at a controlled rate. The generator 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), which corresponds to a series of dits at 60 WPM. **Fig 25.54** shows a typical display. The first two dits at the beginning of the transmission are displayed in order to show any transients or truncations that occur when the transceiver transitions from receive to the transmit state. 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 **Receivers** and **Transmitters and Transceivers** chapters for further discussion of CW keying issues.

For voice modes, 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 25.55** for the test setup. For SSB the transmitter is modulated with a single 700-Hz tone. For FM the transmitter is unmodulated. The keying generator is set to a speed that allows plenty of time for the transceiver to recover between dits. The ON or OFF delay times are measured from the 50% point of the falling or rising edge of the key out line to the 50% point of the RF waveform.

The transmit-receive turnaround time is the time it takes for a transceiver to switch from the 50% rise time of the key line to 50% rise of audio output. Turnaround time is an important consideration in some digital modes with required turnaround times of less than 50 ms in some cases. The test setup is shown in **Fig 25.56**. This test requires extreme care to prevent excessive transmitter power from reaching the signal generator and exceeding its specifications. The step attenuator is preset to maximum and gradually decreased until the receiver's S meter reads S9. Receiver AGC is usually on and set for the fastest response for this test but

Fig 25.53 — CW keying waveform test setup.



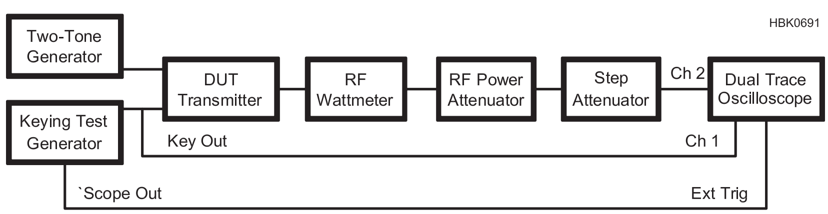


Fig 25.55 — PTT-to-RF-output test setup for voice-mode transmitters.

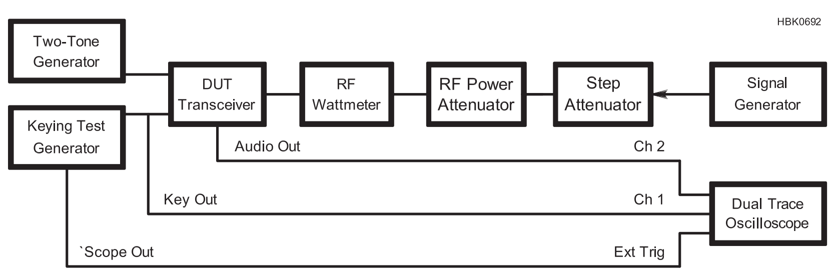


Fig 25.56 — Transmit-receive turnaround time test setup.

experimentation with AGC and signal input level can reveal surprising variations. As for the PTT-to-RF output test, the transmitter is tuned to full power output with a single 700 Hz tone. The keying rate must be considerably slower than the turnaround time; rates of 200 ms on/200 ms off or faster, have been used with success in Product Review tests at the ARRL Lab.

25.6.2 Other Common Transmitter Measurements

The peak envelope power (PEP) of a 100%-modulated AM signal is four times the average power. For that reason, the specified AM power level of most SSB transmitters is generally about 25% of the PEP SSB power rating.

The modulation percentage can most easily be measured with a wide-band oscilloscope connected to the 50-Ω dummy load. It is OK to exceed the specified bandwidth somewhat as long as a clean signal is displayed since this is a relative measurement only. With 100% modulation, the negative modulation peaks just reach zero signal level and the positive peaks are twice the amplitude of the unmodulated carrier. The exact value can be calculated with the equation,

$$M = 100\frac{\max - \min}{\max + \min}$$

where M is the modulation percentage, max is the signal amplitude at the peaks and min is the amplitude at the troughs. An alternative is to use an RF spectrum analyzer. With 100% sine-wave modulation, the two sidebands are each –6 dB with respect to the carrier. In this case, $M = 100 \times 10^{(S+6)/20}$, where S is the sideband level with respect to the carrier (a negative number) and M is the modulation percentage.

The equivalent measurement for an FM transmitter is the *deviation*, which is the amount the RF frequency deviates from the center carrier frequency. It is possible to purchase an instrument that measures FM deviation directly; the function is generally included in two-way radio test sets. Another way is to use slope detection with a standard analog spectrum analyzer. Start by choosing a resolution bandwidth (RBW) on the spectrum analyzer such that as you tune away from the unmodulated carrier the signal level changes approximately linearly (the same number of kHz per dB) over at least a 10 kHz range. An RBW of 10 kHz is usually about right. Record the kHz per dB sensitivity value. Then adjust the frequency for the middle of the linear range and set the spectrum analyzer for zero span. With modulation applied to the transmitter, the deviation is equal to one-half the peak-to-peak dB variation of the signal on the screen, times the kHz per dB value determined previously.

Yet another way to measure FM deviation with a spectrum analyzer is to use the fact that the carrier disappears for a modulation index of 2.405, as explained in the **Modulation** chapter. To set the deviation to 3 kHz, for example, apply sine wave modulation at a frequency of 3.0/2.405 = 1.25 kHz and adjust the modulation level to null the carrier.

25.7 Miscellaneous Measurements

While receiver and transmitter measurements are perhaps the type of testing most associated with Amateur Radio, the home experimenter will have occasion to do other measurements as well.

25.7.1 Testing Digital Circuitry

Virtually every electronic device these days includes digital circuitry. Even a simple QRP transceiver usually either includes or is used with a digital keyer and it may include other digital circuitry, such as a display, as well. A multimeter is generally of little use for testing digital circuits because it cannot respond fast enough to indicate the high and low transitions.

An oscilloscope is a very useful tool because it gives a visual display of the digital signal versus time. It can show impairments such as transient glitches, overshoot, slow rise times, and high or low logic levels that are out of spec. A scope with at least two channels is greatly desired because you often to need to see the time relationships between different signals. A separate external trigger input is also very useful for the same reason. The oscilloscope's bandwidth must be high enough to accurately display the signals to be tested. Bandwidths in the GHz range are needed for state-of-the-art digital circuitry, but 100 MHz should be adequate for most needs.

A hand-held *logic probe* is a device that indicates whether a circuit node is high, low or toggling in between the two states (even with very narrow pulses). While it does not give as much information as an oscilloscope, it is much smaller, cheaper and easier to use. In many cases it is all that is needed to troubleshoot a circuit of moderate complexity. For example, many circuit faults result in a "stuck bit" that remains high or low all the time, which is easy to detect with a logic probe. Typically the logic probe has two wires with clip leads on the ends that connect to the ground and power supply of the circuit under test. The operator then touches the probe tip to the point to be tested and LEDs light up to indicate the logic state.

At the other end of the complexity spectrum is the *logic analyzer*. This is typically the size of an oscilloscope with one or more external pods that can connect to dozens or hundreds of circuit nodes at the same time. Models that use an external digitizing pod and connect to a PC via USB are also available.

A graphics screen shows individual signals with an oscilloscope-like display or multiple signals can be treated as a bus, with the display reading out the values as a series of hexadecimal numbers versus time. Unlike an oscilloscope, a logic analyzer does not indicate actual voltage levels, but only whether a signal is high or low and the timing of the transitions. Sophisticated triggering modes allow synchronization to various clocks or data patterns. Usually a large on-board memory allows capturing long data traces for display or later analysis. Some models include a pattern generator to generate a long series of multi-bit test vectors. Logic analyzers are quite expensive to buy new but are commonly available on the surplus market.

25.7.2 Service Monitors

A *service monitor* is a "one-box tester" for transceivers. It includes a signal generator for testing the receiver and a spectrum analyzer for testing the transmitter, using the same RF connector so that only one connection to the transceiver's antenna jack is required. Other common features are an RF wattmeter and dummy load, a frequency counter, an FM deviation meter, audio tone generators to connect to the microphone, and an audio voltmeter and distortion analyzer/SINAD meter to connect to the speaker output. Some units contain additional features such as DTMF (touch-tone) and CTCSS (sub-audible tone) generators, an audio frequency counter and adjacent-channel power (ACP) measurement capability.

Older service monitors found on the surplus market were designed for testing analog two-way radios and repeaters. Many are portable for easy transportation to a mountaintop repeater site. Later units may be more oriented to testing cellular telephone base stations. Modern instruments cover the latest digital modes, with bit error rate (BER) testers for the receiver and various modulation quality tests for the transmitter.

While all the functions of a service monitor are available in separate instruments, having everything integrated in one box is more convenient and allows faster testing, which is something a commercial enterprise is willing to pay extra for. A brand-new service monitor is not inexpensive, but older used units made by such companies as Singer-Gertsch, Cushman and IFR that are suitable for testing analog radios can sometimes be found for reasonable prices.

25.7.3 Antenna Measurements

Traditionally, amateurs have measured coax-fed antennas using a standing wave ratio (SWR) meter. It is a good method for determining an antenna's resonant frequency and how well-matched it is to the characteristic impedance of the coaxial feed line. To obtain more detailed information about the complex impedance versus frequency, you can use an antenna analyzer to read out the magnitude and phase (or the resistance and reactance) of the antenna impedance. Standing wave ratio and SWR meters are covered in the **Transmission Lines** and **Station Accessories** chapters. Antenna analyzers are discussed in the Measuring RF Impedance section of this chapter.

Those instruments can tell you about the antenna's impedance but they say nothing about how well it actually radiates. For that, you need a way of detecting the signal level at some distance from the antenna. Probably the most common way to do that is to ask for signal reports on the air. Unfortunately, propagation is so variable on most amateur bands that a signal report gives only a very rough idea of how well your antenna is working, unless it can be compared with another antenna (perhaps at another local amateur's station) at the same time.

A *field-strength meter* is a device, generally with a built-in antenna, that picks up the radiated signal off the air and indicates the level on a meter or display. Professional field-strength meters use a carefully-calibrated antenna and circuitry that can read out the actual radiated signal strength in volts per meter or watts per square meter. Most amateur field-strength meters are not calibrated but give a relative indication only. They are useful for tuning an antenna, antenna tuner or transmitter for maximum signal as well as for comparing different antennas. A field strength meter is a simple one-evening construction project.[21] If you have a reference antenna (such as a half-wave dipole) that has a known gain, then the gain of a second antenna can be calculated by alternately transmitting with each antenna and measuring the difference in signal level at a receiver whose antenna is located far enough away to be outside the *near field*, typically up to several wavelengths from the antenna under test. At first glance, it seems like this should be an easy measurement to make but in practice there are a number of devilish details that can ruin the measurement accuracy.

The biggest issue is reflections. If there is any significant reflector of RF signals between the transmitting and receiving antennas, the signal level will not be accurate. Even if there are no wires, fences or other conducting objects in the vicinity, the ground reflection can result in an apparent additional gain of 6 dB or conversely a loss of many dB if the receive antenna happens to be in a null. At microwave frequencies it is sometimes possible to use a high-gain, narrow-beamwidth receive antenna and mount the antenna under test high enough so that the ground reflection is outside the receive antenna's beamwidth.[22] At HF frequencies that is rarely possible. Commercial antenna companies use elaborate *antenna test ranges* that employ various techniques to assure measurement accuracy. Absent a proper test range, the best solution is probably to measure the reference antenna and antenna under test at various heights above ground to get an idea of how much ground reflections are affecting the measurement.

The subject of antenna measurements is addressed by Paul Wade, W1GHZ in his "Microwavelengths" *QST* columns for Oct 2012 (covering the antenna range) and Jan 2013 (discussing measurements and equipment).

25.7.4 Testing Digital Modulation

As digital modulation modes become more and more important in Amateur Radio, it is increasingly important to have ways of testing the performance. There are dozens of different digital formats in use, from traditional radio teleprinting (RTTY) using frequency-shift keying (FSK) to the latest systems that employ sophisticated error detection and correction along with various modulation types that pack multiple bits into each symbol. Despite the wide differences in modulation and coding, nearly all have in common a relatively-narrow bandwidth suitable for use with a voice transceiver using SSB or FM modulation.

If you're having trouble with reception or transmission of digital signals using a PC sound card, one straightforward troubleshooting technique is to install the software on two computers and see if you can transmit data from one computer to the other by connecting the sound card output of one to the input of the other and vice versa. If you don't get perfect reception, that indicates a problem with the computer software or hardware.

The next step is to transmit into a dummy load and receive the signal with a separate receiver located close by so it picks up the stray radiation from the dummy load. A piece of wire plugged into the receiver antenna connector can be moved around to adjust the signal level. Many software programs for receiving digital signals include a spectrum display, which can indicate faults in the transmitted signal such as distortion and *skew*, the amplitude imbalance among the tones of a multi-tone modulation signal. To see what the signal is supposed to look like, connect the two computers directly, as previously described. Then when you examine the RF signal transmitted into the dummy load, any additional bandwidth due to distortion or skew in the spectrum shape should be apparent.

If the demodulation software does not include a spectrum display, there are separate programs available that can display the spectrum of the signal at the sound-card input, as discussed in the Spectrum Analyzer section of this chapter. An RF spectrum analyzer measuring the RF output signal directly would give an even better idea of modulation quality because it is not affected by the filters and other circuitry of the receiver. A receiver panadapter as described earlier is a less-expensive substitute.

Comprehensive testing of a digital communications system is quite complicated because of all the variables involved. The *bit error ratio* (BER) is the number of single-bit errors divided by the number of bits sent in a certain time interval. It requires special test equipment to measure because the individual bits are typically decoded deep inside the demodulation software where they are difficult to access. The *packet error ratio* (PER) is easier to measure. In a packetized data system it is the number of incorrect packets divided by the number of packets sent. It can be measured either before or after error correction. In a non-packetized system like PSK31 the character error ratio is a useful figure of merit. BER is affected by the signal-to-noise ratio (SNR), interference, distortion, synchronization errors and multipath fading. PER is further affected by the effectiveness of the coding and error correction of the particular digital mode used.

It is interesting to measure BER or PER as a function of the SNR. For some digital systems with lots of error correction the errors are nearly zero down to a certain signal level and then degrade very sharply below that. However, in real-world operation the SNR is almost never constant. The signal is constantly changing, both in amplitude and phase, as propagation changes due to movement of the ionosphere (on HF) or of the vehicle (VHF and above), as explained in the **Propagation of Radio Signals** chapter. Measuring actual on-the-air performance is not a good way to compare systems because propagation varies so much at different times. For a repeatable test, you need a *channel simulator,* which is a device that intentionally degrades a test signal in a precise way as to simulate an over-the-air radio channel. Moe Wheatley, AE4JY offers a free software HF channel simulator which can be downloaded from the Internet.[23] A hardware HF channel simulator has been described by Johann B Forrer, KC7WW.[24]

25.7.5 Software-based Test Equipment

Most amateurs these days own a personal computer with a powerful microprocessor, tons of memory and mass data storage, a large color display and a sound card that provides stereo high-fidelity audio input/output. It doesn't take a great deal of imagination to realize that these resources can be harnessed to make low-cost measuring instruments of various types.

Audio-frequency instruments can be implemented directly using the sound card, which typically has a frequency response from perhaps 50 Hz up to about 20 kHz. (Professional and audiophile sound cards with higher performance are also available.) Free software is available on the Internet for instruments such as audio function generators, DTMF and CTCSS tone generators, DTMF and CTCSS decoders, two-tone generators for SSB transmitter testing, distortion/SINAD analyzers, oscilloscopes and audio spectrum analyzers. In addition to the frequency-response limitations of a typical sound card, another issue is that the device can be damaged by applying excessive voltage to the inputs or outputs. It is wise to add external buffer amplifiers that include overvoltage protection.

Radio-frequency test equipment can also use the sound card inputs by means of some type of frequency converter, consisting of a local oscillator and mixer. If the mixer is a quadrature type, the two outputs may be fed to the stereo sound card inputs so that software can treat the left and right channels as the in-phase and quadrature signals. A common application is a narrow-band RF spectrum analyzer. The RF bandwidth is typically limited to twice the sound card's audio bandwidth. Low-cost hardware is available in kit form that can be used with free software downloaded from the Internet.[25]

One problem with using a sound card is that the signals may be susceptible to ground loops and radio-frequency interference (RFI). Since the computer and the device under test are grounded separately to the ac power system, hum and noise can be generated from currents flowing in the ground connection between the two. It is helpful to use a short, low-resistance ground connection between the sound card and the device under test. It is also possible to use isolation transformers or differential amplifiers to isolate the grounds and thus break the ground loop. Good quality cables and attention to proper shielding and grounding help prevent hum and noise pickup.

RFI may be an issue if you wish to make measurements while transmitting. Again, it may be helpful to use the shortest possible connections between the device under test and the sound card. For longer connections use shielded cable or twisted-pair wires such as found in Ethernet LAN cable. To prevent common-mode currents from entering the sound card, wrap several turns of the coax or twisted-pair wire through a ferrite toroid or clamp-on core as described in the **RF Techniques** and **RF Interference** chapters.

25.8 Construction Projects

25.8.1 Bipolar Transistor Tester

Here is a basic "good/bad" tester for bipolar transistors, designed by Alan Bloom, N1AL. This tester is small enough to carry in your pocket to a flea market (**Fig 25.57**). A printed-circuit board is available from FAR Circuits (**www.farcircuits.net**) but the simple circuit can easily be hand-wired on perfboard. (Printed-circuit board layout graphics are also available on this book's CD-ROM.)

To test an unknown NPN or PNP transistor, just remove the tester's working NPN or PNP transistor from its socket and replace it with the device to be tested. If you hear a tone in the headphones the transistor is good, otherwise it is bad. More elaborate instruments can measure various transistor parameters such as current gain, breakdown voltage and high-frequency performance, however this simple tester suffices in most situations. It

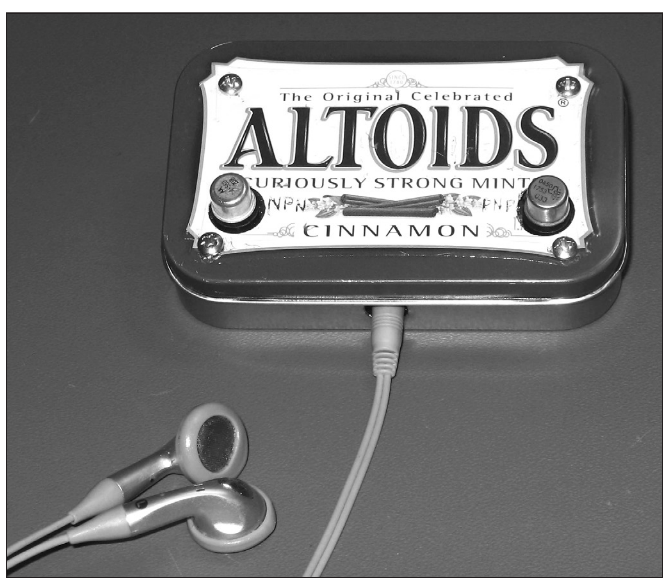


Fig 25.57 — A transistor tester built into an Altoids tin.

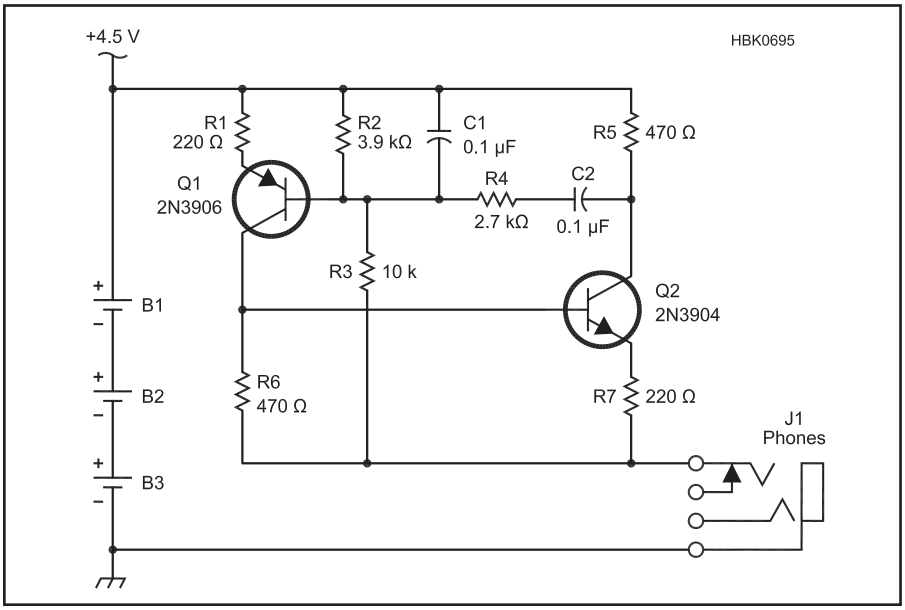


Fig 25.58 — Schematic diagram and parts list of the transistor tester. All parts can be obtained from Digi-Key at www.digikey.com except for the printed circuit board, available from FAR Circuits at www.farcircuits.net.

C1, C2 — 0.1 µF ceramic capacitor (Digi-Key 490-3873-ND)
J1 — Stereo 3.5 mm phone jack (Digi-Key CP1-3554NG-ND)
Q1 — 2N3906 PNP transistor (Digi-Key 2N3906FS-ND)
Q2 — 2N3904 NPN transistor (Digi-Key 2N3904FS-ND)
R1, R7 — 220 Ω, ¼ W, 5% resistor (Digi-Key 220QBK-ND)
R2 — 3.9 kΩ, ¼ W, 5% resistor (Digi-Key 3.9KQBK-ND)
R3 — 10 kΩ, ¼ W, 5% resistor (Digi-Key 10KQBK-ND)
R4 — 2.7 kΩ, ¼ W, 5% resistor (Digi-Key 2.7KQBK-ND)
R5, R6 — 470 Ω, ¼ W, 5% resistor (Digi-Key 470QBK-ND)
Quantity 2 — 3-pin, TO-5 transistor socket (Digi-Key ED2150-ND)
Quantity 6 — AAA-size battery clips (Digi-Key 82K-ND)
PCB — Printed circuit board (FAR Circuits)

Fig 25.59 — Mounting of the circuit board inside the case.

is rare for a transistor to be damaged in such a way that it still works but no longer meets its specifications.

When testing a batch of transistors of unknown condition you can use this tester to quickly sort them into a "good" and a "bad" pile and be fairly confident that the ones in the "good" pile are working correctly.

Metal-can TO-5 transistors are shown here, but small plastic transistors work just as well if you bend the leads a little. The TO-5 parts can be pressed down flat against the socket and so are less likely to fall out in a pocket.

The circuit in **Fig 25.58** is simply two transistors connected with positive feedback through a frequency-selective network, forming an oscillator at a frequency of approximately 500 Hz. Each transistor is configured for a voltage gain of about 2.0 and the feedback network has a gain of about ⅓, so that the total loop gain is a little greater than unity, the condition required for oscillation.

It should be nearly impossible to damage an unknown transistor by plugging it in wrong or into the wrong socket because the supply voltage is less than the base-emitter breakdown voltage of a bipolar transistor and the current is limited to a few milliamps. No on/off switch is included; simply unplug the headphones when you're done testing.

The prototype was built in an Altoids tin, as shown in **Fig 25.59**, but any handy enclosure would do. You'll need four mounting holes for the circuit board and two clearance holes for the transistor sockets. The headphone jack is best mounted on the side of the enclosure. If such a shallow enclosure is used, line the inside bottom with some insulating material such as electrical tape.

All components are mounted on the top side of the printed circuit board except the battery clips and the headphone jack, which go on the bottom. Leave a little extra lead length on the two 0.1 µF capacitors if they need to be bent over to clear the enclosure cover. The PC board from FAR Circuits does not have plated-through holes, so the leads of R5 and R6 must be soldered on both sides.

The transistor sockets are designed for TO-5 metal-can transistors but the smaller TO-18 or TO-92 plastic-cased devices can also be tested by bending the leads to fit. Nearly all TO-92 bipolar transistors have the base lead in the middle. Bend the center (base) lead toward the flat side of the transistor body, spread the three leads a little, and it should plug right in. Additional solder pads are provided for the base, emitter and collector of each transistor in case you wish to wire up additional sockets for other case types such as TO-220 or TO-3 power transistors. A transistor cross-reference guide is also handy to have to determine lead assignments.

25.8.2 Logic Probe

This simple logic probe (**Fig 25.60**) was designed by Alan Bloom, N1AL. It works with several different logic types, including TTL, 5 V CMOS and 3.3 V CMOS. A printed circuit board is available from FAR Circuits at **www.farcircuits.net** or the simple circuit may be hand-wired on perfboard. (Printed-circuit board layout graphics are also available on this book's CD-ROM.)

The purpose of a logic probe is to indicate whether signals are present at various circuit nodes. That's most of what troubleshooting a simple digital circuit requires. The probe has indicators to show whether the signal is high, low or toggling between the two states. It doesn't give as much information as an oscilloscope or logic analyzer but it is much smaller, cheaper and easier to use.

CIRCUIT OPERATION

This logic probe features a seven-segment LED display that forms letters to indicate the state of the signal at the probe's tip. A capital "L" is displayed if the signal is low and "H" if the signal is high. If it is toggling between low and high with roughly a 50% duty factor, the letter "B" is displayed to indicate that "Both" high and low are present. If the signal is mostly low with short-duration positive pulses, then a "C" is displayed. If the signal is mostly high with low-going pulses, the LED indicates an "A". "C" and "A" can be remembered as Cathode (mostly low) and Anode (mostly high), respectively.

The circuit is shown in **Fig 25.61**. Each of the common-anode seven-segment LED display segments is lit when its pin is low. The 74ACT04 inverters are arranged so that a continuous low input signal lights up the proper segments to form an "L" and a continuous high forms an "H". The 74LS122 retriggerable multivibrator outputs a 33-ms pulse whenever there is a positive-going tran-

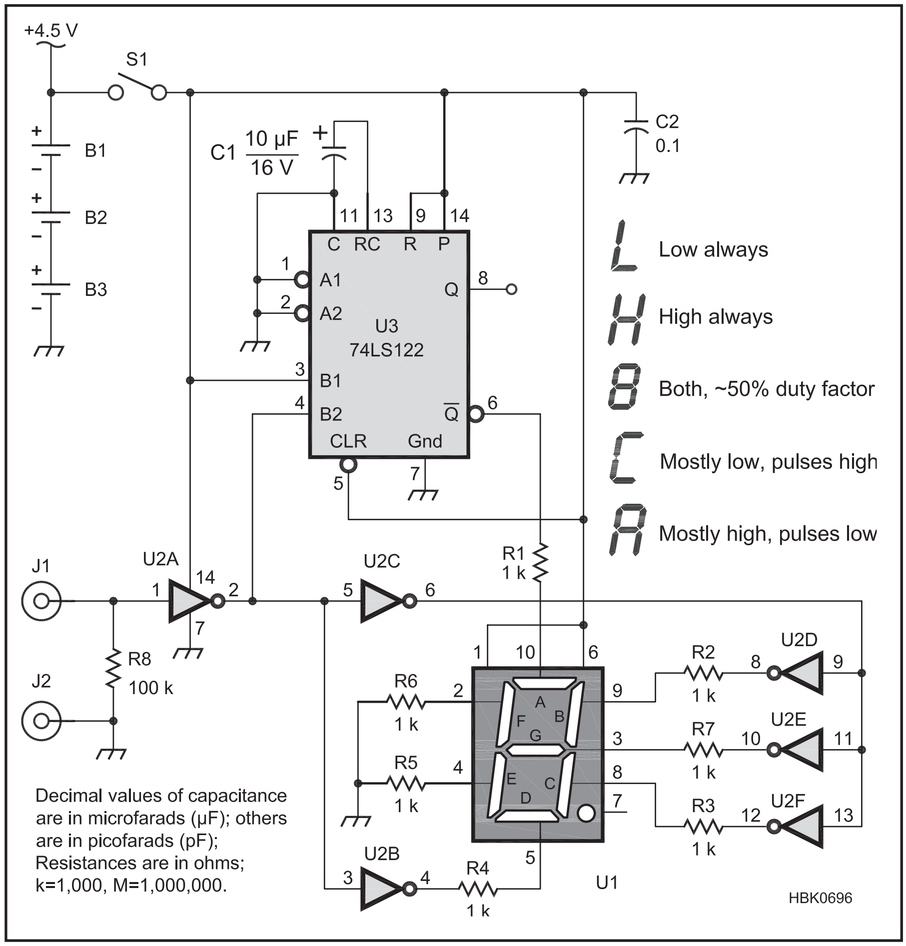


Fig 25.61 — Schematic diagram and parts list of the logic probe. All parts can be obtained from Digi-Key at www.digikey.com except for the printed circuit board, available from FAR Circuits at www.farcircuits.net.

C1 — 10 µF electrolytic capacitor (Digi-Key P807-ND)
C2 — 0.1 µF ceramic capacitor (Digi-Key 490-3873-ND)
J1 — Horizontal tip jack (Digi-Key J110-ND)
R1-R7 — 1 kΩ, ¼ W, 5% resistor (Digi-Key 1.0KQBK-ND)
R8 — 100 kΩ, ¼ W, 5% resistor (Digi-Key CF1/4100KJRCT-ND)
S1 — SPDT, right-angle slide switch (Digi-Key CKN9559-ND)
U1 — 7-segment common-anode LED (Digi-Key 160-1525-5-ND)
U2 — 74ACT04 hex inverter (Digi-Key 296-4351-5-ND)
U3 — 74LS122 monostable multivibrator (Digi-Key 296-3639-5-ND)
Qty 6 — Battery clips (Digi-Key 82K-ND)
Qty 2 — L-bracket (Digi-Key 621K-ND)
Printed circuit board (FAR Circuits)

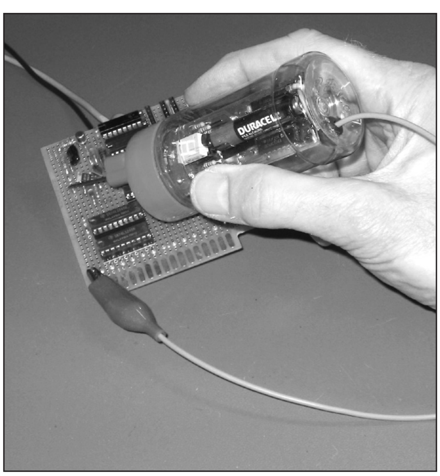


Fig 25.60 — A logic probe is small and easy to use.

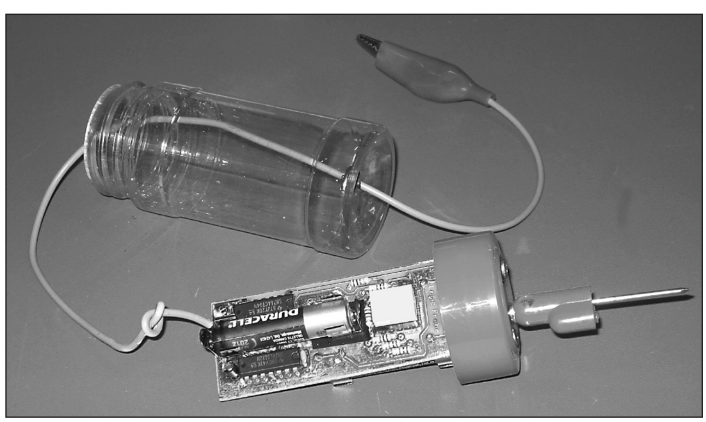


Fig 25.62 — The logic probe was built into a spice bottle.

sition on its B2 input. Repetitive transitions with a period less than about 33 ms (30 Hz or greater) assert the 74LS122 output continuously, which causes the top segment to light. If the signal is mostly low, the "L" turns into a "C" and if the signal is mostly high, the "H" turns into an "A".

CONSTRUCTION

The logic probe can be built in a plastic spice jar as in **Fig 25.62** but any clear plastic container big enough to accommodate the 1.2 × 3.5-inch printed-circuit board would also work. Batteries B1 and B2 and the test lead jack, J1, are mounted on the bottom side; all other components are on the top. Place a piece of tape under the top terminal of B3 to keep it from shorting to the ground plane. The PC board from FAR circuits does not have plated-through holes so at least the following must be soldered on both sides: R5, R6, R8, B1 and B3 (inner contacts), U2 pins 1 and 14, and U3 pins 1, 2, 3, 7 and 11. Also solder a wire to both sides of the via hole to the left of pin 1 of the display (U1). If you wish, you can use a socket for the LED display to raise it up higher, but don't use sockets for U2 and U3 since some of their leads must be soldered on both sides. Note that the L-brackets that attach the board to the bottle cap have a long and a short side. The long side is placed against the board.

The socket for the probe is a standard 0.08 inch diameter tip jack. The probe is just a nail soldered into the end of a mating tip plug. You can also use a standard test lead plugged into the jack. The component labeled "J2" on the schematic is just a solder pad for the ground lead. Tie a knot in the ground lead where it exits the bottle for strain relief.

USING THE LOGIC PROBE

The probe includes a self-contained, battery-operated power supply so only the ground lead needs to be connected to circuit ground.

It's amazing how much useful information you can get from such a simple device. One of the most common faults in a digital circuit is a node that is "stuck" low or high due to a short circuit or a faulty component. The logic probe is perfect for detecting that condition.

You can also get a rough idea of what percentage of time the signal is low or high by looking at the relative brightness of the segments. If the signal is low 100% of the time, an "L" is displayed. If there are narrow positive pulses, a "C" is shown. As the pulse width increases, the center and the two right segments start to glow dimly, finally forming a "B" when the duty factor is near 50%. As the duty factor increases further, the bottom segment dims, eventually turning the "B" into an "A". And finally, if the signal is high 100% of the time, the top segment goes out, leaving an "H".

25.8.3 Inductance Tester

Many inexpensive DVMs offer a capacitance measurement feature but measuring inductance is much less common. This project describes a simple test fixture for using a signal generator to measure inductance. It was originally published as "Mystery Inductor Box," by Robert J. Rogers, WA1PIO, in the January 2011 issue of *QST*.

MEASURING INDUCTANCE

One way to measure the value of an inductor is to connect it in parallel with a known capacitance and measure the resulting resonant frequency. In the past, grid-dip oscillators (GDO) have been used to determine the resonant frequency of such parallel tuned circuits. (See the project "Gate-Dip Oscillator" elsewhere in this section.) Using a dip oscillator, a dip, or drop in meter current was noted at the point of resonance.

This test fixture provides a convenient substitute for the dip meter method. A separate signal generator is used to provide the needed signal source and switched internal capacitors are used to resonate the inductors. An internal meter is provided as is a port for an external detector to indicate resonance. The schematic with parts list is shown in **Fig 25.63**.

BUILDING THE BOX

Fig 25.64 shows the layout of the internal components. Construction is straightforward with no critical dimensions or layout requirements. The 470-Ω resistor is used to provide a better peak in the voltage seen at resonance by the oscilloscope or meter movement. Most of the generator voltage will appear across this resistor until the point of resonance when the highest fraction of the applied signal voltage will then appear across the parallel circuit.

An oscilloscope is connected to the right BNC connector and its high input impedance, typically 1 MΩ, will not load the parallel tuned circuit at the relatively low frequencies used with this test fixture. Instead of the oscilloscope, the internal meter movement may be used to give a peak indication at resonance. Large capacitance values were used to minimize the effects of lead length, a concern while operating test equipment at higher frequencies.

USING THE TEST FIXTURE

The test fixture can be used to accurately determine the value of inductors in the mH and µH range. Selecting the 0.01 µF capacitor allows measurement of inductors in the µH range using a signal generator with a 159 kHz to 5 MHz output. Selecting the 0.22 µF capacitor measures larger mH-range inductors using signal generator frequencies in the 1 to 34 kHz region.

Fig 25.65 shows a typical test setup using the built-in meter as the peak indicator. A signal generator's sine wave output is connected to J1, the input BNC connector, and the unknown inductor to the terminal posts.

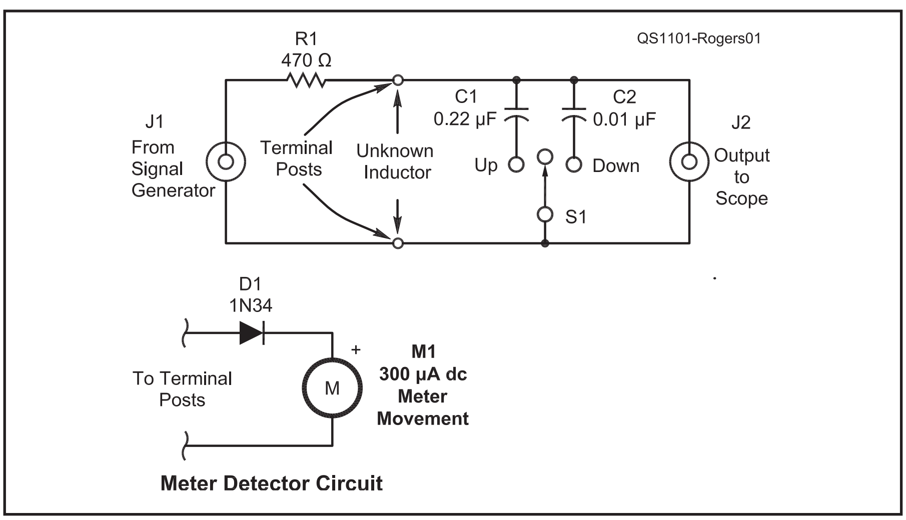


Fig 25.63 — Schematic diagram and parts list of the inductor test fixture. The switch's center-off position is used to disconnect the internal capacitors so that a desired value of capacitance can be placed in parallel with the inductor under test at the terminal posts. Measurement accuracy will depend on the actual value of C1 and C2 — determining the capacitor values with a capacitance meter will result in more accurate inductance values.

C1 — 0.22 µF ceramic capacitor.
C2 — 0.01 µF ceramic capacitor.
D1 — Germanium diode, 1N34 or 1N34A.
J1, J2 — BNC chassis-mount connectors.
M1 — 300 µA meter movement in a plastic case
R1 — 470 Ω, ¼ W resistor.
S1 — SPDT, center-off, toggle switch.

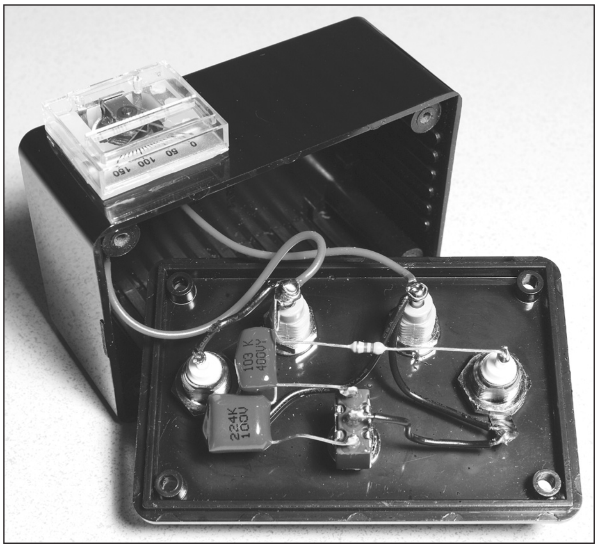


Fig 25.64 — Layout of the internal components.

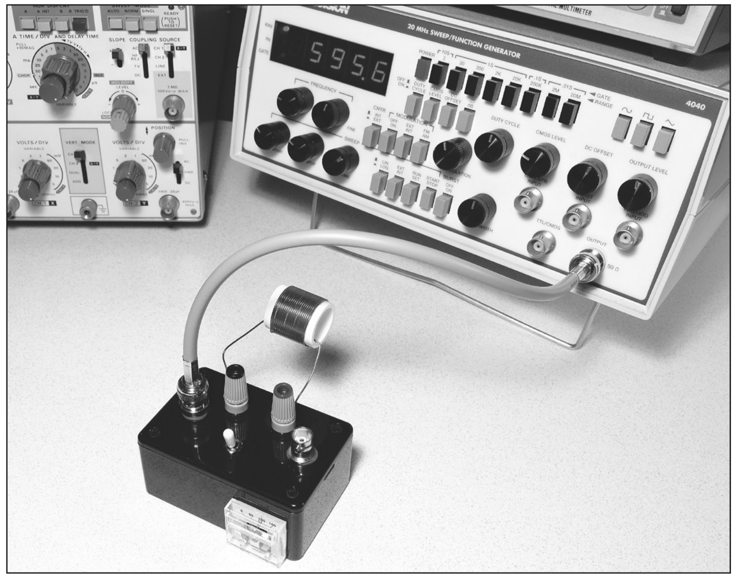


Fig 25.65 — The signal generator's output is 596 kHz (0.596 MHz) with the switch in the position selecting the 0.01 µF capacitor. Using the formula given results in an calculated inductance of 7.1 µH.

Select one of the capacitors with the switch. Connect an oscilloscope to J2, the output BNC connector, or use the built-in meter. Adjust the signal generator output to give some indication on the meter. Adjust the frequency of the signal generator until a peak is obtained on the oscilloscope or meter.

If the switch is in the wrong position, there will not be a sharp peak in the voltage seen by the oscilloscope or built-in meter. Avoid the band edges of the generator frequency range where output voltages tend to fall off from the mid-range values. Use the capacitor value selected, C, and the frequency of the signal generator at resonance, f, in the following formula:

$$L = \frac{1}{4\pi^2 f^2 C} = \frac{1}{39.5 f^2 C}$$

where f is in hertz, C is in farads and the resulting L is in henrys. A chart of frequency versus inductance for each of the two internal capacitors used to measure small and large inductors is shown in **Table 25.3**.

The SPDT switch also has a center-off position which disconnects the internal capacitors. This allows testing a particular capacitor-inductor combination connected to the terminal posts. The internal meter movement is always connected across the banana terminal posts.

Table 25.3
Inductance Value at Resonant Frequency

0.22 µF Selected Inductance (mH)	Frequency (kHz)	*0.01 µF Selected* Inductance (µH)	Frequency (MHz)
0.10	33.931	0.118	4.633
0.25	21.460	0.25	3.183
0.50	15.174	0.50	2.251
0.75	12.390	0.75	1.838
1.00	10.730	1.00	1.592
5.00	4.796	2.00	1.125
10.00	3.393	5.00	0.712
20.00	2.399	10.00	0.503
30.00	1.959	20.00	0.356
40.00	1.696	30.00	0.291
50.00	1.517	40.00	0.252
60.00	1.385	50.00	0.225
70.00	1.282	60.00	0.205
80.00	1.199	70.00	0.190
90.00	1.131	80.00	0.178
100.00	1.073	90.00	0.168
		100.00	0.159

25.8.4 Fixed-Frequency Audio Oscillator

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

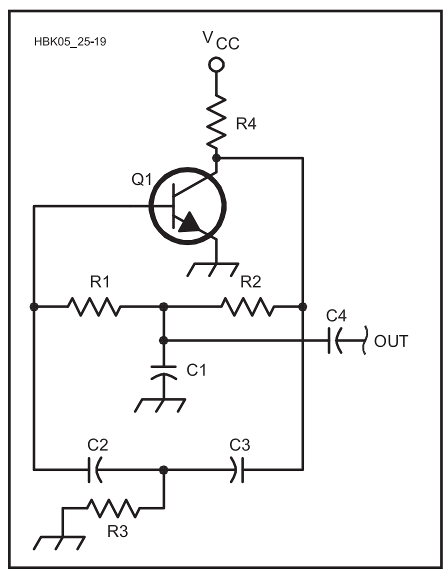
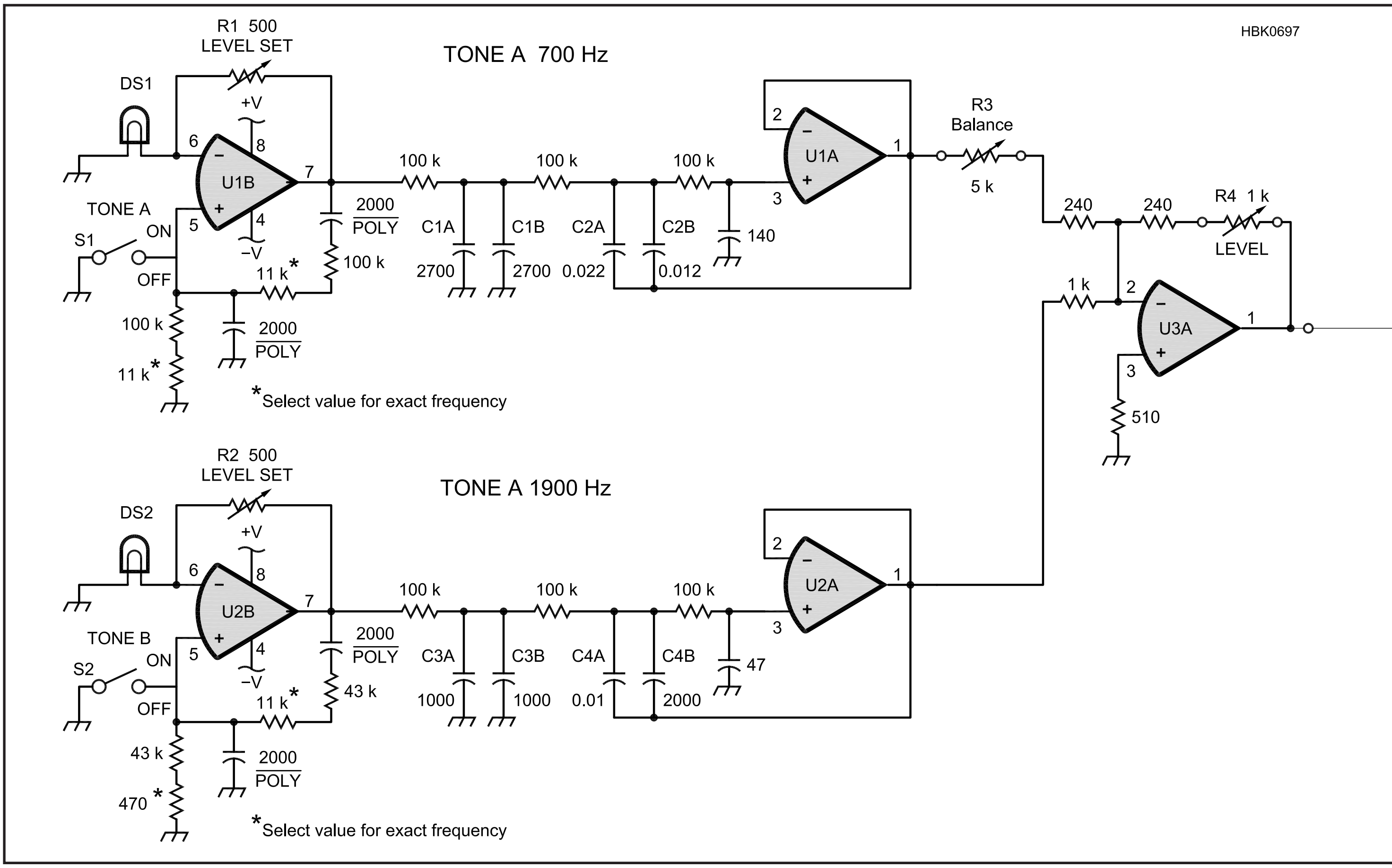


Fig 25.66 — Values for the twin-T audio oscillator circuit range from 18 kΩ for R1-R2 and 0.05 µF for C1 (750 Hz) to 15 kΩ and 0.02 µF for 1800 Hz. For the same frequency range, R3 and C2-C3 vary from 1800 Ω and 0.02 µF to 1500 Ω and 0.01 µF. R4 is 3300 Ω and C4, the output coupling capacitor, can be 0.05 µF for high-impedance loads.



A circuit of a simple RC oscillator that is useful for general testing is given in **Fig 25.66**. 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.

25.8.5 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 (see

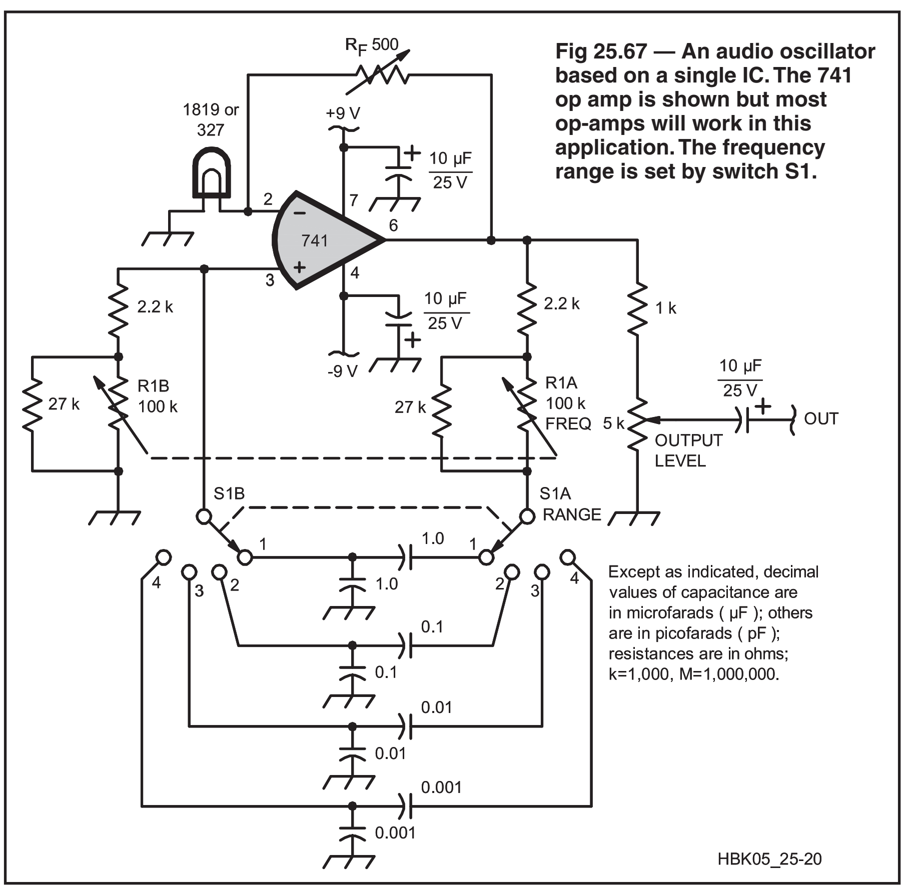


Fig 25.67 — An audio oscillator based on a single IC. The 741 op amp is shown but most op-amps will work in this application. The frequency range is set by switch S1.

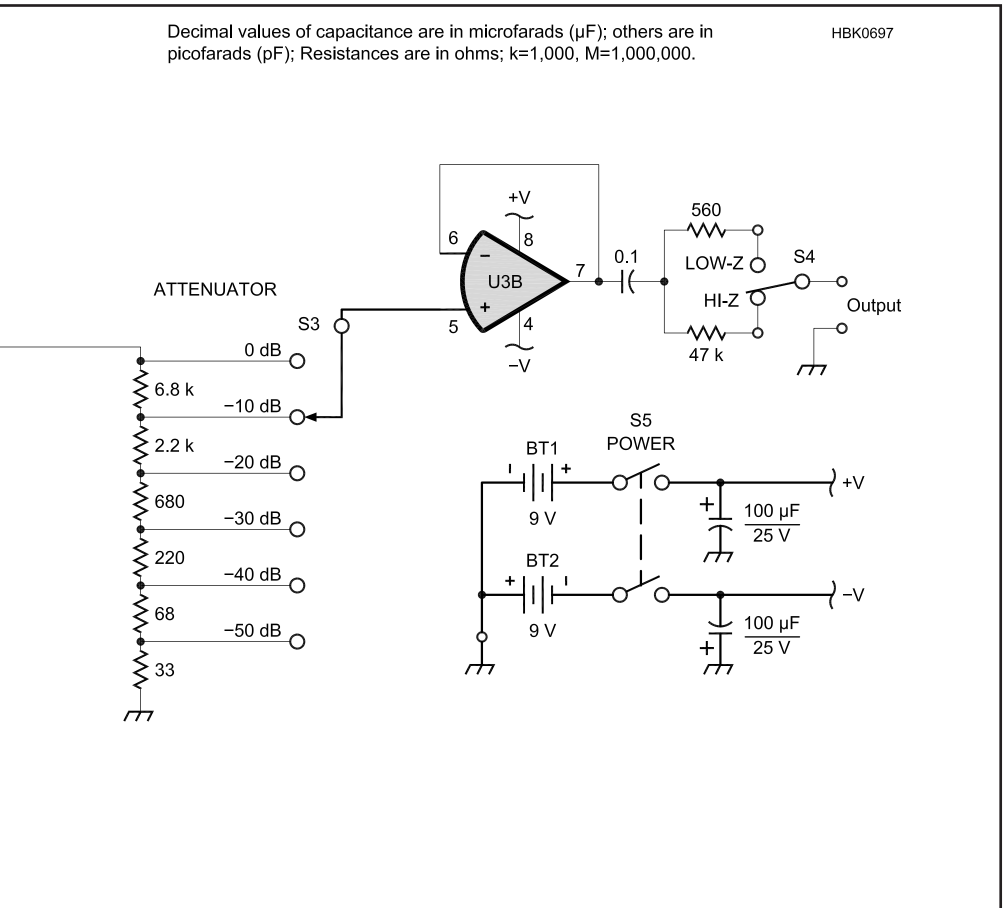


Fig 25.68 — Two-tone audio generator schematic.

BT1, BT2 — 9 V alkaline.
C1A,B — Total capacitance of 0.0054 μF, ±5%.
C2A,B — Total capacitance of 0.034 μF, ±5%.
C3A,B — Total capacitance of 0.002 μF, ±5%.
C4A,B — Total capacitance of 0.012 μF, ±5%.
DS1, DS2 — 12 V, 25 mA lamp.
R1, R2 — 500 Ω, 10-turn trim potentiometer.
R3 — 500 Ω, panel mount potentiometer.
R4 — 1 kΩ, panel mount potentiometer.
S1, S2 — SPST toggle switch.
S3 — Single pole, 6-position rotary switch.
S4 — SPDT toggle switch.
S5 — DPDT toggle switch.
U1, U2, U3 — Dual JFET op amp, type LF353N or TL082.

Fig 25.67). Power is supplied by two 9-V batteries from which the circuit draws 4 mA. The frequency range is selectable from about 7 Hz to around 70 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 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*.

25.8.6 Two-Tone Audio Generator

This generator is used in the ARRL Laboratory to test SSB transmitters for ARRL Product Reviews and makes a very convenient signal source for testing the linearity of a single-sideband transmitter. To be suitable for transmitter evaluation, a generator of this type must produce two non-harmonically related tones of equal amplitude. The level of harmonic and intermodulation distortion must be sufficiently low so as not to confuse the measurement. The frequencies used in this generator are 700 and 1900 Hz, both well inside the normal audio passband of an SSB transmitter. Spectral analysis and practical application with many different transmitters has shown this generator to meet all of the requirements mentioned above. While designed specifically for transmitter testing it is also useful any time a fixed-frequency, low-level audio tone is needed.

CIRCUIT DETAILS

Each of the two tones is generated by a separate Wien bridge oscillator, U1B and U2B. (see **Fig 25.68**) The oscillators are followed by RC active low-pass filters, U1A and U2A. Because the filters require nonstandard capacitor values, provisions have been made on the circuit board for placing two capacitors in parallel in those cases where standard values cannot be used. (The circuit board artwork for layout and part placement is available as graphics on the CD-ROM accompanying this book.) The oscillator and filter capacitors should be polystyrene or Mylar film types if available.

The two tones are combined by op amp U3A, a summing amplifier. This amplifier has a variable resistor, R4, in its feedback loop which serves as the output LEVEL control. While R4 varies the amplitude of both tones together, R3, the BALANCE control, allows the level of tone A to be changed without affecting the level of tone B. This is necessary because some transmitters do not have equal audio response at both frequencies. Multi-turn pots are recommended for both R3 and R4 so that fine adjustments can be made. Following the summing amplifier is a step attenuator; S3 controls the output level in 10-dB steps. The use of two output level controls, R4 and S3, allows the output to cover a wide range and still be easy to set to a specific level.

The remaining op amp, U3B is connected as a voltage follower and serves to buffer the output while providing a high-impedance load for the step attenuator. Either high or low output impedance can be selected by S4. The values shown are suitable for most transmitters using either high- or low- impedance microphones.

CONSTRUCTION AND ADJUSTMENT

Component layout and wiring are not critical, and any type of construction can be used with good results. Because the generator will normally be used near a transmitter, it should be enclosed in some type of metal case for shielding. Battery power was chosen to reduce the possibility of RF entering the unit through the ac line. With careful shielding and filtering, the builder should be able to use an ac power supply in place of the batteries.

The only adjustment required before use is the setting of the oscillator feedback trimmers, R1 and R2. These should be set so

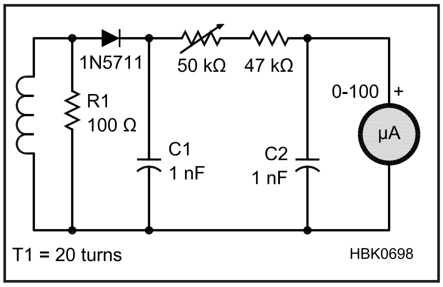


Fig 25.69 — The schematic of the RF current probe. See text for component information.

Fig 25.70 — Assembly of the RF current probe. Use an all-plastic meter and mount the circuits and toroid directly on the back of the meter case.

that the output of each oscillator, measured at pin 7 of U1 and U2, is about 0.5 volts RMS. A DVM or oscilloscope can be used for this measurement. If neither of these is available, the feedback should be adjusted to the minimum level that allows the oscillators to start reliably and stabilize quickly. When the oscillators are first turned on, they take a few seconds before they will have stable output amplitude. This is caused by the lamps, DS1 and DS2, used in the oscillator feedback circuit. This is normal and should cause no difficulty. The connection to the transmitter should be through a shielded cable.

25.8.7 RF Current Meter

The following project was designed by Tom Rauch, W8JI (**http://w8ji.com/building_a_current_meter.htm**). The circuit of **Fig 25.69** is based on a current transformer (T1) consisting of a T157-2 powdered-iron toroid core with a 20-turn winding. The meter is used with the current-carrying wire or antenna inserted through the middle of the core as a one-turn primary

When 1 A is flowing in the single-turn primary, the secondary current will be 50 mA = primary current divided by the turns ratio of 20:1. R1 across the transformer flattens the frequency response and limits the output voltage. The RF voltage is then detected and filtered by the D1 (a low-threshold Schottky diode for minimum voltage drop) and C1. The adjustable sum of R2 and R3 allow for full-scale (FS) calibration of the 100 µA meter. C2 provides additional filtering. The toroid core and all circuitry are glued to the back of the meter case with only R2 exposed — a screwdriver-adjustable calibration pot.

It is important to minimize stray capacitance by using a meter with all-plastic construction except for the electrical parts. The meter in **Fig 25.70** has an all-plastic case including the meter scale. The meter movement and all metallic areas are small. The lack of large metallic components minimizes stray capacitance from the proximity of the meter. Low stray capacitance ensures the instrument has the least possible effect on the circuit being tested.

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

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

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

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

25.8.8 RF Ammeters

When it comes to getting your own RF ammeter, there's good news and bad news as related by John Stanley, K4ERO. First, the bad news. New RF ammeters are expensive, and even surplus pricing can vary widely between $10 and $100 in today's market. AM radio stations are the main users of new units. The FCC defines the output power of AM stations based on the RF current in the antenna, so new RF ammeters are made mainly for that market. They are quite accurate, and their prices reflect that!

The good news is that used RF ammeters are often available. For example, Fair Radio Sales in Lima, Ohio has been a consistent source of RF ammeters. Ham flea markets are also worth trying. Some grubbing around in your nearest surplus store or some older ham's junk box may provide just the RF ammeter you need. Be sure you are really buying an RF ammeter as meters labeled "RF Amps" may just be regular current meters intended for use with an external RF current sensing unit.

RF AMMETER SUBSTITUTES

Don't despair if you can't find a used RF ammeter. It's possible to construct your own. Both hot-wire and thermocouple units can be homemade. Pilot lamps in series with antenna wires, or coupled to them in various ways, can indicate antenna current (F. Sutter, "What, No Meters?," *QST*, Oct 1938, p 49) or even forward and reflected power (C. Wright, "The Twin-Lamp," *QST*, Oct 1947, pp 22-23, 110 and 112).

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 (Z. Lau, "A Relative RF Ammeter for Open-Wire Lines," *QST*, Oct 1988, pp 15-17).

25.8.9 RF Step Attenuator

A good RF step attenuator is one of the key pieces of equipment that belongs on your workbench. The attenuator in this project offers good performance yet can be built with a few basic tools. The attenuator is designed for use in 50-Ω systems, provides a total attenuation of 71 dB in 1-dB steps, offers respectable accuracy and insertion loss through 225 MHz and can be used at

450 MHz as shown in **Table 25.4**. This material was originally published as "An RF Step Attenuator" by Denton Bramwell, K7OWJ, in the June 1995 *QST*.

The attenuator consists of 10 resistive π-attenuator sections such as the one in **Fig 25.71**. Each section consists of a DPDT slide switch and three ¼-W, 1 %-tolerance metal-film resistors. The complete unit contains single 1, 2, 3 and 5-dB sections, and six 10-dB sections. **Table 25.5** lists the resistor values required for each section.

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

Table 25.4
Step Attenuator Performance at 148, 225 and 450 MHz
Measurements made in the ARRL Laboratory

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

Note: Laboratory-specified measurement tolerance of ±1 dB

Table 25.5
Closest 1%-Tolerance Resistor Values

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

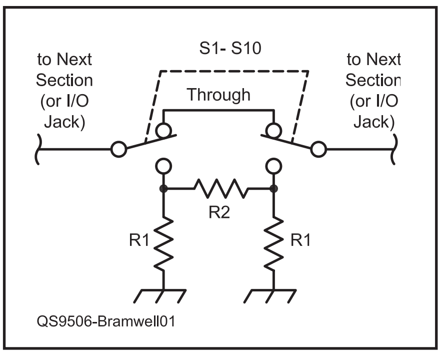


Fig 25.71 — Schematic of one section of the attenuator. All resistors are ¼-W, 1%-tolerance metal-film units. See Table 25.5 for the resistor values required for each attenuator section. There are six 10 dB sections and one each of 1, 2, 3 and 5 dB.

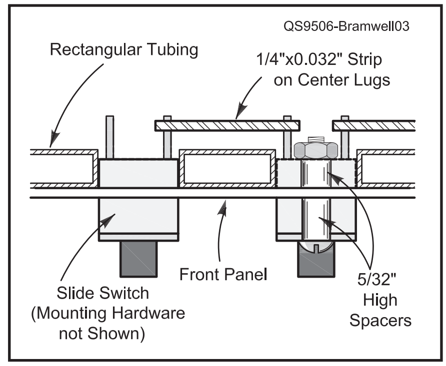


Fig 25.72 — Key to obtaining acceptable insertion loss in the "through" position is to make the whole device look as much as possible like 50-Ω coax. The rectangular tubing and the ¼ × 0.032-inch brass strip between the switch sections form a 50-Ω stripline.

CONSTRUCTION

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

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

For the front panel, nibble or shear a piece of 2-inch-wide brass to a length of about 9½ inches. Space the switches from each other so that a piece of the rectangular brass tubing lies flat and snugly between them (see **Fig 25.72**). Drill holes for the #4-40 mounting screws and nibble or punch rectangular holes for the bodies of the slide switches.

Before mounting any parts, solder in place one of the 1-inch-wide chassis side pieces to make the assembly more rigid. Solder the side piece to the edge of the top plate that faces the "through" side of the switches; this makes later assembly easier (see **Fig 25.73**). Although the BNC input and output connectors are shown mounted on the top (front) panel, better lead dress and high-frequency performance may result from mounting the connectors at the ends of the enclosure.

DPDT slide switches designed for sub-panel mounting often have mounting holes tapped for #4-40 screws. Enlarge the holes to allow a #4-40 screw to slide through. Before mounting the switches, make the "through" switch connection (see Fig 25.71) by bending the two lugs at one end of each switch toward each other and soldering the lugs together or solder a small strip of brass between the lugs and clip off the lug ends. Mount the switches above the front panel, using 5⁄32-inch-high by 7⁄32-inch-OD spacers. Use the same size spacer on the inside. On the inside, the spacer creates a small post that helps reduce capacitive coupling from one side of the attenuator to the other. The spacers position the switch so that the 50-Ω stripline can be formed later.

The trick to getting acceptable insertion loss in the "through" position is to make the attenuator look as much as possible like 50-Ω coax. That's where the rectangular tubing and the ¼ × 0.032-inch brass strip come into the picture (see Fig 25.72); they form a 50-Ω stripline. (See the **Transmission Lines** chapter for information on stripline.)

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

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

Next, solder in place the three 1%-tolerance resistors of each section, keeping the leads as short as possible. Use a generous blob of solder on ground leads to make the lead less inductive. Install a ½-inch-square

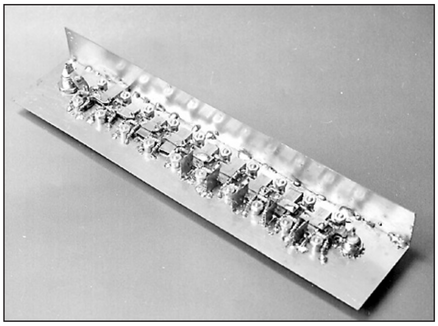


Fig 25.73 — Solder one of the 1-inch-wide chassis side pieces in place to make the assembly more rigid during construction. Solder the side piece to the edge of the top plate that faces the "through" side of the switches; this makes the rest of the assembly easier.

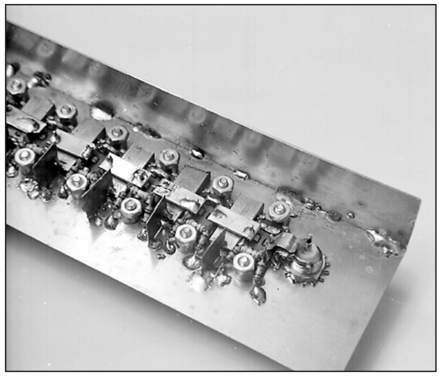


Fig 25.74 — The attenuator before final mechanical assembly. The ¼-inch strips are spaced 0.033 inch apart to form a 50-Ω connection from the BNC connector to the stripline. There are ½-inch square shields between 10-dB sections. The square shields have a notch in one corner to accommodate the end of the rectangular tubing.

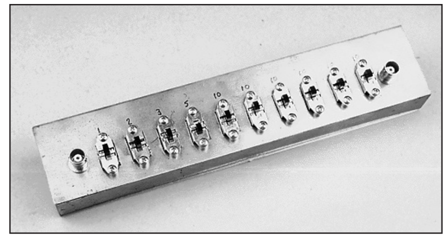


Fig 25.75 — The completed step attenuator in the enclosure of brass sheet. The BNC connectors may be mounted on the front panel at the end of the switches or on the end panels.

brass shield between each 10-dB section to ensure that signals don't couple around the sections at higher frequencies.

Use parallel ¼-inch strips of 0.005-inch-thick brass spaced 0.033 inch apart to form 50-Ω feed lines from the BNC connectors to the switch contacts at each end of the stripline as shown in Fig 25.73. (Use the undrilled piece of 0.032-inch-thick brass to insure the proper line spacing.) The attenuator with all switches and shields in place is shown ready for final mechanical assembly in **Fig 25.74**

Finally, solder in place the remaining enclosure side, cut and solder the end pieces, and solder brass #4-40 nuts to the inside walls of the case to hold the rear (or bottom) panel. Drill and attach the rear panel and round off the sharp corners to prevent scratching or cutting anyone or anything. Add stick-on feet and labels and your step attenuator of **Fig 25.75** is ready for use.

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

25.8.10 High-Power RF Samplers

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

The problem with high-powered attenuators is that they are expensive to buy or build since the front end of the attenuator must handle the output power of the transmitter or amplifier. If one already has a dummy load, an RF sampler may be used to produce a replica of the signal at a reduced power level. The sampler described here was originally presented in *QST* Technical Correspondence for May 2011 by Tom Thompson, WØIVJ.

A transformer sampler passes a single conductor (usually the insulated center conductor from a piece of coaxial cable) from the transmitter or amplifier to the dummy load through a toroidal inductor forming a transformer with a single turn primary. The secondary of the transformer is connected to a resistor network and then to the test equipment as shown in **Fig 25.76**. Assume that the source, whether a transmitter or amplifier, is a pure voltage source in series with a 50-Ω resistor. This most likely is not exactly the case but is sufficient for analysis.

If a current, I, flows into the dummy load, then a current, I / N flows in the secondary of the transformer, where N is the number of turns on the secondary. Fig 25.76 also shows the equivalent circuit, substituting a current source for the transformer. The attenuation is 40 dB and 15 turns for the secondary of the transformer. If R_{SHUNT} = 15 Ω, and R_{SERIES} = 35 Ω, then the voltage across a 50-Ω load resistor, R_{SAMPLE}, is 1/100 of the voltage across the dummy load, which is 40 dB of attenuation.

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

The samplers described here use an FT37-61 ferrite core followed by two resistors as described above. The through-line SWR is good up to 200 MHz, the SWR is fair looking into the sampled port, and the useful bandwidth extends from 0.5 MHz to about 100 MHz. If you are interested in an accurate representation of the third harmonic of your HF transmitter or amplifier, it is important for the sampler to give accurate attenuation into the VHF range.

Fig 25.77 shows a photo of a sampler built into a 1.3 × 1.3 × 1 inch (inside dimensions) box constructed from single-sided circuit board material. The through-line connection is made with a short piece of UT-141 semi-rigid coax with the shield grounded only on one side to provide electrostatic shielding between the toroid and the center conductor of the coax. (Do not ground both ends of the shield or a shorted turn is created.) R_{SHUNT} is hidden under the toroid, and R_{SERIES} is shown connected to the sample port. This construction technique looks like a short piece of 200-Ω transmission line in the through-line which affects the SWR at higher frequencies. This can be corrected by compensating with two 3 pF capacitors connected to the through-line input and output connectors as shown in the photo. The through-line SWR was reduced from 1.43:1 to 1.09:1 at 180 MHz by adding the capacitors. This compensation, however, causes the attenuation to differ at high frequencies depending on the direction of the through-line connection. A sampler constructed using the box technique is useable from below 1 MHz through 30 MHz.

Fig 25.78 shows a different approach using 9⁄16 inch diameter, 0.014 inch wall thickness, hobby brass tubing. This lowers the imped-

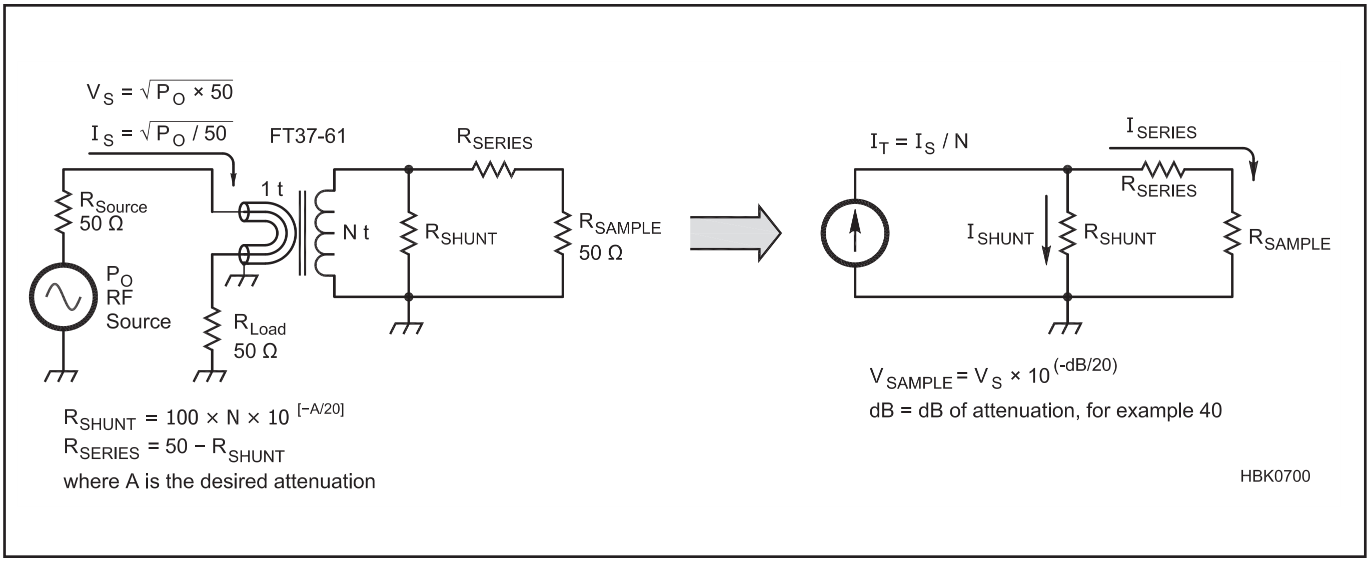


Fig 25.76 — RF sampler circuit diagram and equivalent circuit showing calculations.

ance of the through-line so that no compensation is needed. The through-line SWR for the tube sampler is 1.08:1 at 180 MHz which is as good as the box sampler and the sensitivity to through-line direction is reduced. Although the high frequency attenuation is not as good as the box sampler, the construction technique provides a more consistent result. A sampler constructed using the tube technique should be useable through 200 MHz.

CONSTRUCTION OF THE TUBE SAMPLER

Both samplers use 15 turns of #28 AWG wire on an FT37-61 core, which just fits over the UT-141 semi-rigid coax. R_{SHUNT} is a 15 Ω, 2 W, non-inductive metal oxide resistor and R_{SERIES} is a 34.8 Ω, ¼ W, 1% non-inductive metal film resistor. The power dissipation of the resistors and the flux handling capability of the ferrite core are adequate for sampling a 1500-W source. For those uncomfortable using BNC connectors at high power, an SO-239 version may be constructed using an FT50A-61 core and larger diameter tubing. Construction details are included on this book's CD-ROM.

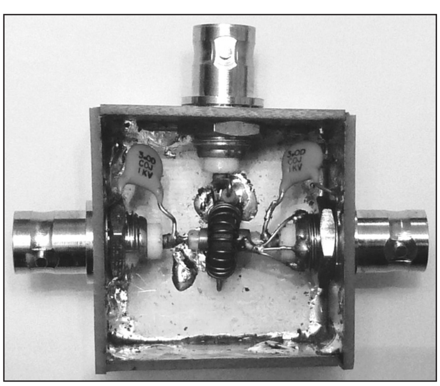
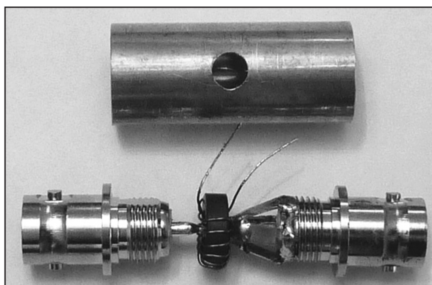


Fig 25.77 — RF sampler using box construction.



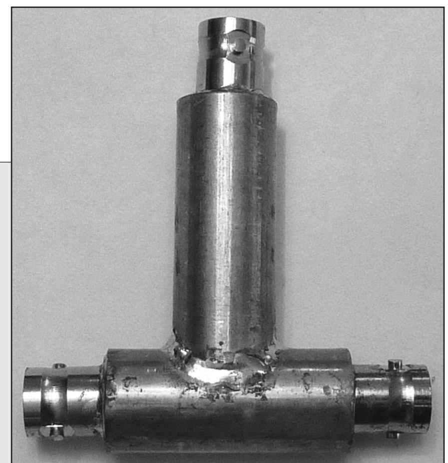


Fig 25.78 — RF sampler using tube construction.

25.8.11 RF Oscillators for Circuit Alignment

Receiver testing and alignment can make use of inexpensive RF signal generators which are available as complete units and in kit form. Any source of signal that is weak enough to avoid overloading the receiver usually will serve for alignment work and troubleshooting.

A crystal oscillator is often a satisfactory signal source for amplifier testing and receiver repair or alignment. Several example circuits can be found in the **Oscillators and Synthesizers** chapter. The output frequencies of crystal oscillators, while not adjustable, are quite precise and very stable. The Elecraft XG2 (**www.elecraft.com**) and NorCal S9 (**www.norcalqrp.org**) are good example of a simple fixed-frequency signal source kits. The harmonics of the output signals are on known frequencies and can also be used as low-level signal sources. The fundamental signals have known output amplitudes for calibrating S meters and other gain stages.

Variable frequency oscillators can be used as signal generators and there are several kits or assembled units available based on a direct digital synthesis (DDS) integrated circuit. (See the **DSP and Software Radio Design** chapter.) The Elecraft XG3 is a programmable signal source that operates from 1.5 to 200 MHz with four programmable output levels between –107 and 0 dBm. Another similar product is the DDSv4 by RMT Engineering (**www.rmt-tech.com**) with a 0-100 MHz range.

For receiver performance testing, precise frequency control, signal purity, noise, and low-level signal leakage become very important. A lab-quality instrument is required to make these measurements. Commercial and military-surplus units such as the HP608-series are big and stable, and they may be inexpensive. Recently, the HP8640-series of signal generators have become widely available at very attractive prices. 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.

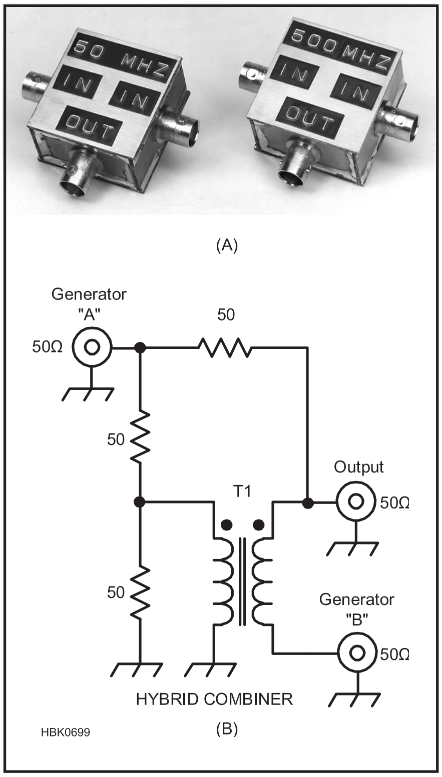


Fig 25.79 —The hybrid combiner on the left of A is designed to cover the 1 to 50-MHz range; the one on the right 50 to 500 MHz. B shows the circuit diagram of the hybrid combiner. Transformer T1 is wound with 10 bifilar turns of #30 AWG enameled wire. For the 1 to 50-MHz model, T1 is an FT-23-77 ferrite core. For the 50 to 500-MHz model, use an FT-23-67 ferrite core. Keep all leads as short as possible when constructing these units.

25.8.12 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 (**www.minicircuits.com**).

Alternatively, a hybrid combiner is not difficult to construct. The combiners described here (see **Fig 25.79**) provide 40 to 50 dB of isolation between ports, assuming the common port is terminated in a 50-ohm load. Attenuation in the desired signal paths (each input to output) is 6 dB. Loads with low return loss typical of receiver inputs will reduce isolation.

The combiners are constructed in small boxes made from double-sided circuit-board material as shown in Fig 25.79A. 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 non-inductive resistors (or matched units from the junk box) should be used. The circuit diagram for the combiners is shown in Fig 25.79B

The combiner may also be constructed and used as a *return loss bridge* as described in the *QST* article by Jim Ford, N6JF, "Build a Return Loss Bridge," from September 1997 and included on the CD-ROM included with this book. Return loss is discussed in the **RF Techniques** chapter.

25.8.13 Gate-Dip Oscillator (GDO)

The project is adapted from the May 2003 *QST* article, "A Modern GDO — the "Gate" Dip Oscillator" by Alan Bloom, N1AL. A GDO is a tunable oscillator with the coil mounted outside of the chassis. The external coil allows you to measure the resonant frequency of a tuned circuit without any electrical connection to it. Just place the GDO coil near the tuned circuit and tune the GDO while watching for a dip in its meter reading.

This "no connection" measurement capability is handy in other applications. For example, you can measure the resonant frequency of a Yagi's parasitic elements that don't have a feed line connection or measure the resonant frequency of antenna traps. The GDO can "sniff out" spurious resonances in a linear amplifier's tank circuit with the amplifier powered down. (Be sure no high voltage is present!)

The GDO has other uses on the workbench. To measure inductance, temporarily connect a known-value capacitor in parallel with the unknown inductor and find the circuit's resonant frequency. Use the formula $L = 1/(2\pi f)^2C$ or the reactance vs frequency chart in the **Electrical Fundamentals** chapter to find the value of the inductor. (If C is measured in µF and f in MHz, the resulting value of L is in µH.) The same process works to find the value of a capacitor, such as that of an unmarked air-variable capacitor at a hamfest, by using the formula $C = 1/(2\pi f)^2L$.

A GDO can be used as a simple signal generator to test amplifiers, mixers, and filters. To troubleshoot a receiver, tune the GDO to each of the IF frequencies, starting with the last IF stage and hold the coil close to that part of the circuitry. If you can hear a signal at the receiver's output, then that IF stage and all circuitry after it are working. This is especially handy on densely packed, surface-mount boards that are difficult to probe.

With the oscillator turned off, the GDO functions as a tuned RF detector known as an *absorption frequency meter* or *wavemeter*. An obvious use is to determine if RF energy in a tuned circuit has the right frequency. Using the capacity probe as an antenna, it can be used as a frequency-selective field-strength meter. By holding the coil near a cable, you can detect RF current flowing on the outside of a shield and the GDO makes an excellent "sniffer" to detect RF leakage from a shielded transmitter at the fundamental and harmonic frequencies.

A headphone output is provided to listen for key clicks, hum or buzz, and low-frequency parasitic oscillations in a transmitted signal. By coupling the coil to an antenna, you can even use the GDO as a tunable "crystal radio"!

CIRCUIT DESIGN

Fig 25.80 shows the completed GDO. **Fig 25.81** is a view inside the case and **Fig 25.82** pictures the entire set of coils. The schematic and parts list are shown in **Fig 25.83**. A pair of source-coupled N-channel JFETs (Q1 and Q2) form the oscillator portion of the circuit. No RF chokes are required. This eliminates false resonances resulting from self-resonance of the chokes.

Q4, a bipolar 2N3904 transistor, serves a dual purpose. Its base-emitter junction acts as the RF detector. Further, it amplifies the rectified current flowing in the base and sends it to the emitter-follower Q5, a 2N2907. Transistor Q3 is a JFET source-follower amplifier for the output RF connector. The RF output may be used as a signal source or to drive a frequency counter for more accurate frequency display.

Determine the value of R5 using the formula $R5 = (1.5/I_m) - R_m$, where I_m is the full-scale meter current and R_m is the meter resistance. To measure your meter's resistance, be sure to use an ohmmeter that does not use more than the GDO meter's full-scale current. Most modern DVMs use test currents in the range off 50-200 µA for measuring resistance. The value of R5 in the parts list is for a 200 µA full-scale meter used by the author. Meters from 50 to 500 µA full-scale are suitable.

Fig 25.80 — The "gate dipper" fits comfortably in the hand.

In detector mode, the power to the oscillator and RF buffer is turned off. Q4 detects and amplifies any signals picked up by the coil. The meter sensitivity control works in both oscillator and wavemeter mode and also controls the volume to the headphones. Battery current is 3-5 mA with the oscillator on and zero in wavemeter mode when no signal is being received.

CONSTRUCTION

Feel free to substitute parts on hand for those in the parts list. The exceptions are transistors Q1, Q2, and Q4, which should be the types specified or their equivalents.

The 2¼ × 2¼ × 5 inch chassis is a compromise; it's large enough to allow all parts to fit easily and small enough to fit comfortably in the hand. The coil is mounted off-center, both to allow the shortest connection to the tuning capacitor and to afford easier coil coupling to an external circuit.

The prototype used a perforated test board for the RF portion of the circuitry and a solder terminal strip for the meter circuitry. If you prefer, a printed circuit board pattern and parts layout are available on this book's CD-ROM and circuit boards are available from FAR Circuits (**www.farcircuits.net**). Note that the battery's *positive* terminal is connected to ground which is backwards from the normal arrangement.

The coil forms are "⅜ inch" flexible plastic water tubing. The inside diameter is actually slightly less than ⅜ inch, which makes a nice force-fit onto the ⅜-inch threads of a chassis-mount BNC plug (not the more common chassis-mount receptacle). The lowest-frequency coil is wound on a pill bottle with two 1-inch diameter aluminum washers at the connector mount for added strength.

To construct the coils, start by running the wire through the center of each form and soldering it to the BNC connector's center pin connection. Then press the form onto the connector threads and cut a small notch in the form's opposite end to hold the wire in place for the winding. After winding the coil, cut and tin the wire end, then solder it to the ground lug mounted on the connector. Cover each coil with a layer of heat-shrink tubing or electrical tape to hold the turns in place and protect the wire. Clear tubing allows you to see the coil and a small label with the coil's frequency range can be placed under the tubing before shrinking. Fig 25.82 shows the coils before the heat shrink tubing is added.

Winding data is listed in **Table 25.6**. Unless you happen to duplicate the prototype exactly, the frequency range of each coil is likely to be different. However, this data can be used as a starting point for your own coil designs.

The heavy wire for the two highest-frequency coils is bare copper scrap from house wiring cable. The other coils are wound with

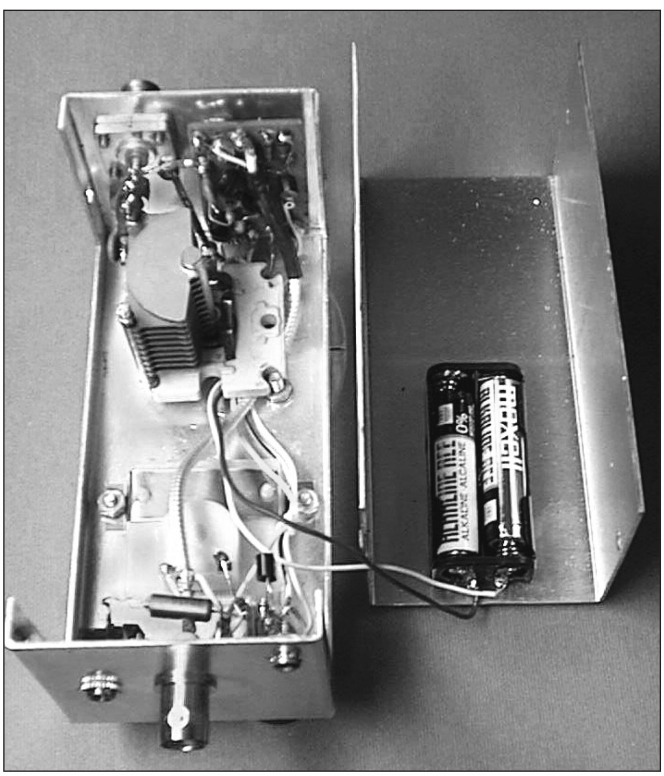


Fig 25.81 — Photograph of the GDO with the cover off. The tuning capacitor was oriented for the shortest possible connection to the coil connector. The prototype was constructed with two batteries but three result in better operation.

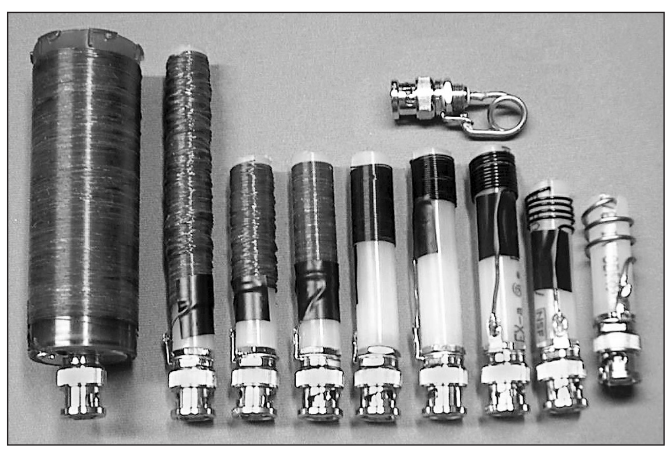


Fig 25.82 — This is how the coils look before covering them with heat-shrink tubing or tape. Small pieces of electrical tape are used to hold the turns in place during construction.

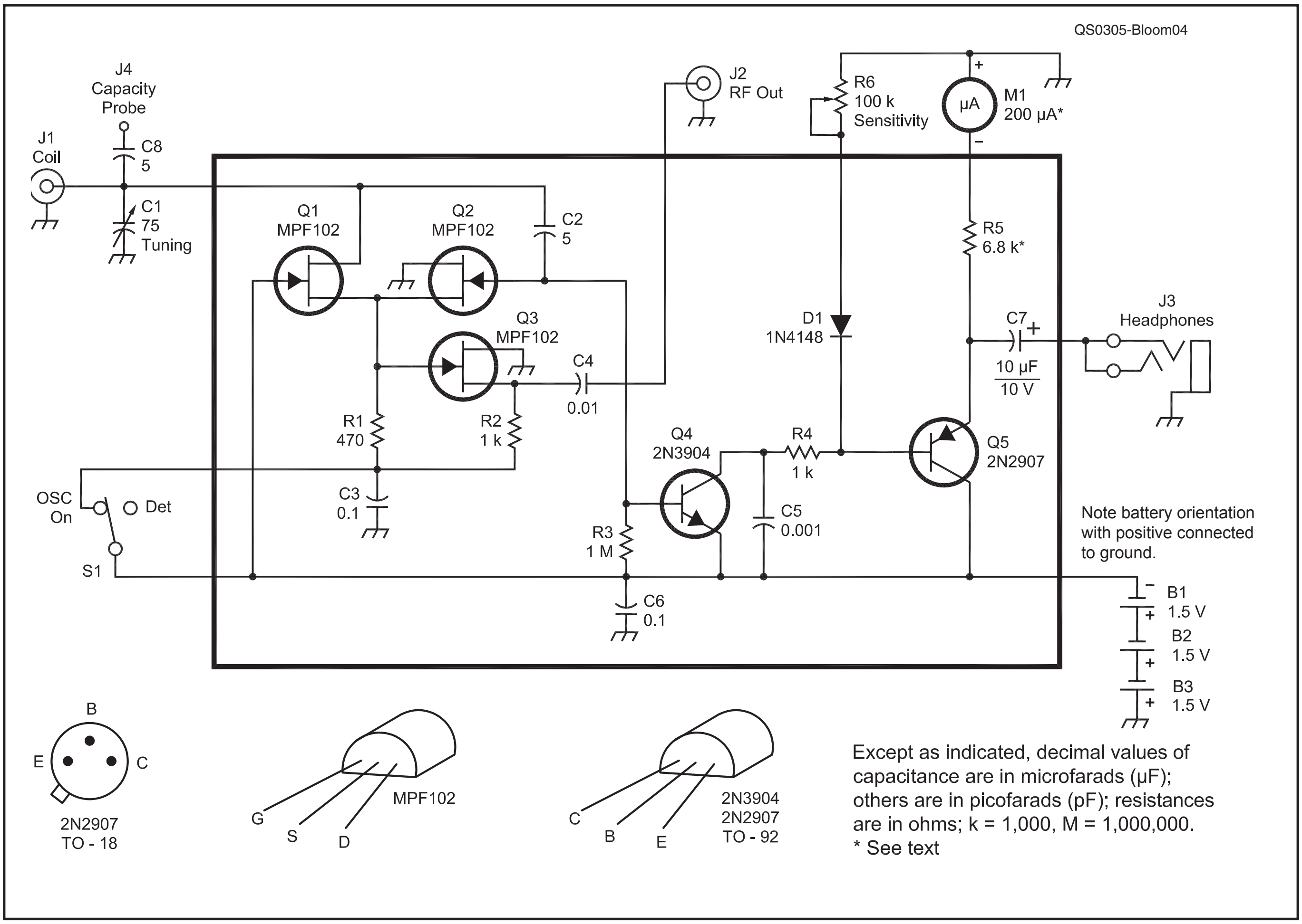


Fig 25.83 — Schematic diagram of the GDO.

B1, B2, B3 — 1.5 V AA or AAA cell.
C1 — 75 pF or 365 pF variable.
C2, C8 — 5 pF disc capacitor.
C3, C6 — 0.1 µF capacitor.
C4 — 0.01 µF capacitor.
C5 — 0.001 µF capacitor.
C7 — 10 µF, 10 V capacitor.
D1 — 1N4148 diode.
J1, J2 — BNC female chassis-mount connector.
J3 — 3.5 mm stereo phone jack.
J4 — Phone tip jack, phone tip plug.
M1 — 0-200µA, 1⅝ inch square panel meter, (see text).
Q1, Q2, Q3 — MPF102 transistor, JFET.
Q4 — 2N3904 transistor.
Q5 — 2N2907A transistor.
R1 — 470 Ω, ¼ W resistor.
R2, R4 — 1 kΩ, ¼ W resistor.
R3 — 1 MΩ, ¼ W resistor.
R5 — See text.
R6—100 kΩ potentiometer, linear taper.
S1 — SPST toggle switch.
Misc: #18, #22 and #26 gauge enameled wire; 10 BNC male chassis-mount connector (for coils); dual AA or AAA battery holder; 2¼ × 2¼ × 5 inch chassis; 3 ft of ⅜ inch PVC water tubing for coils.

Table 25.6
Coil Winding Data for the GDO

Wire Gauge	*Form Diameter (in.)*	*Form Length (in.)*	*Coil Length (in.)*	*Number of Turns*	*Frequency Range*
#12	0.375	—	0.125	2	130-150 MHz
#14	0.5	1.5	.5	3	62-108 MHz
#18	0.5	2.0	0.4	5	29.5-62 MHz
#18	0.5	2.5	0.5	10.5	16.5-35 MHz
#22	0.5	2.5	0.6	21	9.2-19 MHz
#26	0.5	2.5	1.0	46	5.1-10.5 MHz
#30	0.5	2.5	1.5	100	2.8-5.6 MHz
#30	0.5	2.5	1.6	180	1.5-2.8 MHz
#30	0.5	3.9	3.3	390	0.9-1.5 MHz
#30	1.25	3.9	3.4	230	620-980 kHz
				(+53 pF)	505-640 kHz
				(+106 pF)	440-525 kHz

enamel-insulated magnet wire. The exact wire gauge is not critical although it may affect the number of turns required.

The three smallest coils (29.5-150 MHz) were space wound to the lengths listed. The remaining coils were close-wound. On the 0.9-1.5 and 1.5-2.8 MHz coils, there are too many turns to fit in a single layer. Wind these coils by overlapping turns a few at a time. Winding two complete overlapping layers increases the inter-winding capacitance which can cause spurious resonances and reduced tuning range.

The range of the lowest-frequency coil can be extended by connecting it to the GDO through a BNC "T" connector to which one or two 53 pF capacitors are attached using clip leads (see Table 25.6). Adding extra capacitance is a useful way to slow down the tuning rate for measuring narrow-band devices such as crystals.

With the smallest coil, the GDO oscillates over only a small portion of the capacitor tuning range. Fortunately, this range covers the 2 meter amateur band. With the second-smallest coil, the oscillation amplitude drops off at lower frequencies but is useable down to about 62 MHz in the prototype. A lower-inductance tuning capacitor would improve performance at VHF.

The small 75-pF tuning capacitor results in a tuning range of only about 2:1. The advantage of the low ratio is that tuning is not so critical and the frequency dial is easier to read. The disadvantage is that it requires more coils to cover the overall frequency range. Substituting a 365-pF AM broadcast radio tuning capacitor will increase the tuning range to nearly 3:1 with a consequent sharpening of adjustment.

The tuning dial in **Fig 25.84** was made from a circular piece of ¼-inch clear plastic glued to a knob. Making the diameter slightly larger than the chassis width allows one handed operation with a thumb adjusting the tuning as in Fig 25.80. The scale, cemented to the chassis under the dial, was drawn using a computer graphics program although measuring the frequency and creating the scales was very time-consuming. A better solution would be to include a small frequency counter with the GDO using the RF output. 4- or 5-digit accuracy is sufficient.

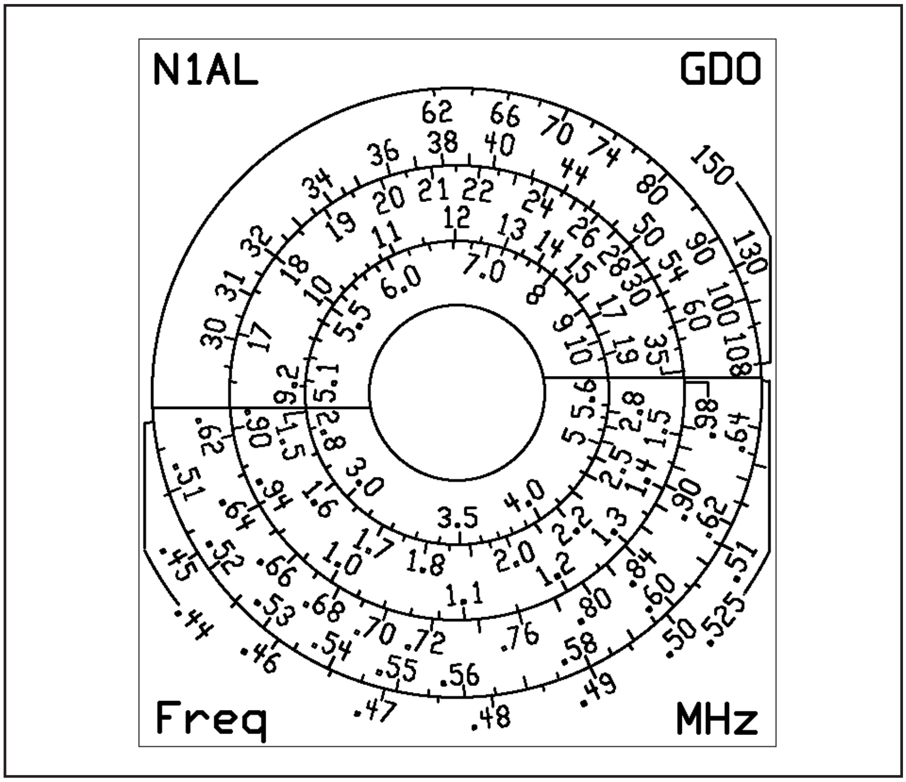


Fig 25.84 — The dial scale for the prototype GDO. Builders of this project should expect to create and calibrate a custom dial for the finished instrument or use a frequency counter as noted in the text.

OPERATION

To measure the resonant frequency of a tuned circuit, switch to oscillator mode and adjust the meter sensitivity to about ¾ scale. Then orient the GDO coil close to and approximately parallel to the coil under test and tune the dial until you get a strong dip on the meter. Tune slowly or you may miss the dip. Overly-close coupling to the circuit under test causes such a strong dip that the oscillator is pulled far off-frequency. Once a dip is found, move the GDO coil farther from the circuit under test until the dip is barely visible — this results in the most accurate measurement.

Coupling to toroids or shielded coils can be difficult. One solution is to connect a wire from the capacitance probe to the "hot" end of the circuit to be tested. For looser coupling, just place the wire close to the circuit under test instead of connecting it to the circuit. When measuring the inductance of a toroid with a test capacitor, the capacitor leads often form enough of a loop to allow coupling to the GDO coil. Some authors recommend coupling the GDO to a one-turn loop through the toroid, but that creates a shorted secondary winding, changing the inductance of the toroid winding.

Antenna measurements are best made at the high-current point of the antenna conductor. For a half-wave dipole, this is near the center. Orient the coil perpendicular to the conductor for maximum coupling. Be sure to short the feed point of the antenna before making the measurement. If you can't find the dip, make the shorting wire into a 1-turn loop for better coupling to the GDO coil You should also see resonance at the odd harmonic frequencies as well as the fundamental.

To measure the electrical length of a transmission line, couple the GDO to a small wire loop; connect it to one end of the line and leave the other end disconnected. (For best accuracy use the smallest loop that gives sufficient coupling.) The line is ¼ wavelength long at the lowest resonant frequency, so the electrical length in meters is 75/f, where f is the resonant frequency in MHz. Again, you will also see resonance at the odd harmonics.

The voltage level at the RF output connector varies from coil to coil but typically runs about 250 mV_{RMS} into an open circuit and 50 mV_{RMS} into a 50-Ω load. That is sufficient for a typical frequency counter or to serve as a test signal for troubleshooting.

Additional uses of the GDO are presented in the following articles on this book's CD-ROM: "The Art of Dipping" in Jan 1974 *QST*, "Add-Ons for Greater Dipper Versatility" in Feb 1981 *QST*, and "What Can You Do with a Dip Meter?" in May 2002 *QST.*

25.8.14 RF Power Meter

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

The primary use for this unit is to monitor

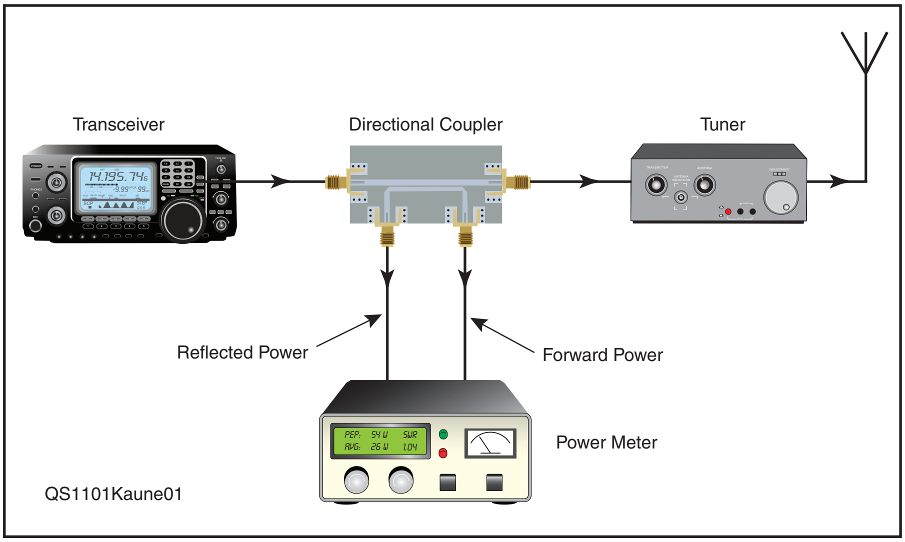


Fig 25.85 — W7IEQ station setup, including the power meter being described here.

the output power and tuning of a transceiver. The author's station configuration is shown in **Fig 25.85**. RF power generated by the transmitter is routed via RG-8 coaxial cable through a directional coupler to an antenna tuner, which is connected to the antenna with RG-8. The directional coupler contains circuits that sample the RF power flowing from the transmitter to the tuner (the forward power) and the RF power reflected back from the tuner to the transmitter (the reflected power). These samples are sent via RG-58 cable to the two input channels of the power meter. This project includes the directional coupler and the power meter. Enough detail is provided in the full article so that an amateur can duplicate the device or modify the design.

DIRECTIONAL COUPLER

The directional coupler is based on the unit described in "The Tandem Match" by John Grebenkemper, KI6WX in the Jan 1987 issue of *QST* and also included on this book's CD-ROM. A pair of FT-82-67 toroids with 31 turns of #26 AWG magnet wire over lengths of RG-8 form the basis of the directional coupler shown in **Fig 25.86**.

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

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

POWER/SWR METER — CIRCUIT DESCRIPTION

Fig 25.87 shows a front panel view of the power meter. An LCD displays the measured peak (PEP) and average (AEP) envelope powers as well as the standing wave ratio (SWR). The power meter calculates either the peak and average envelope power traveling from the transceiver to load (the forward power) or the peak and average envelope powers actually delivered to the load (the forward power minus reflected power). The average envelope power (AEP) represents an average of the forward or load powers over an averaging period of either 1.6 or 4.8 seconds.

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

The forward and reflected power samples from the directional coupler are applied to a pair of Analog Devices AD8307 logarithmic detectors. External 20 dB attenuators (Mini-Circuits HAT-20) reduce the signals from the directional coupler to levels compatible with the AD8307. As noted earlier, the directional coupler has an internal attenuation of about

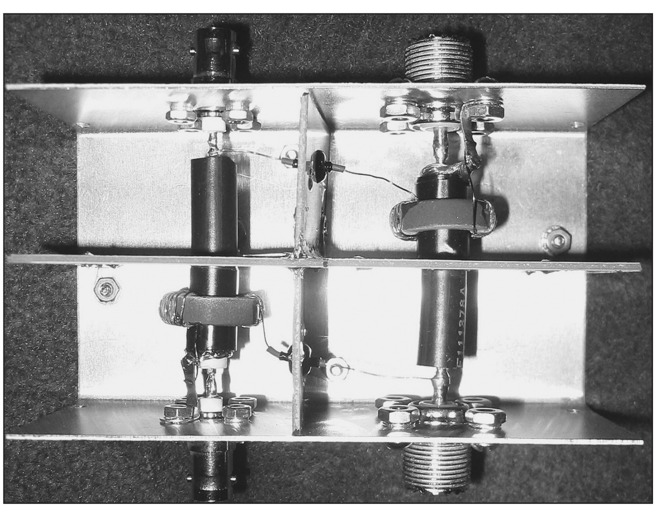


Fig 25.86 — Completed directional coupler.

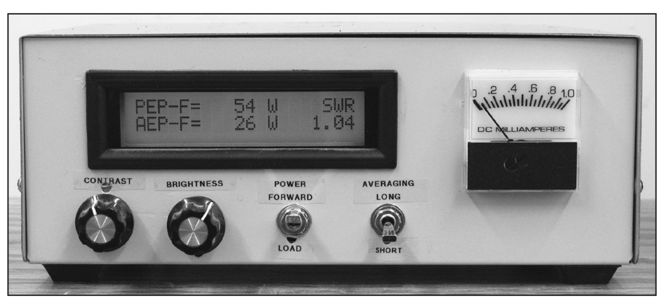


Fig 25.87 — Front panel of power meter. The LCD shows the peak envelope power (PEP), the average envelope power (AEP) and the SWR. The two knobs control the contrast and back lighting of the LCD. One toggle switch determines whether forward or load powers are displayed. A second switch sets the averaging time for the AEP calculation. The meter displays SWR and is used for tuning purposes.

30 dB, so the total attenuation in each channel is about 50 dB. Thus, a rig operating at a power level of 1 kW (60 dBm) will result in an input to the forward power channel of about 10 dBm. (The schematic diagram and parts list of the power meter are provided on the CD-ROM version of the article.) The detectors are configured such that the time constant of their output follows the modulation envelope of the RF signal.

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

25.8.15 RF Voltmeter

Anthony Langton, GM4THU, designed this simple RF voltmeter to measure low RF voltages from oscillators and other low-power RF circuits through several MHz. It has an approximately 10 kΩ input impedance to avoid loading the circuit under test but not so high that it is unduly affected by stray capacitance.

Fig 25.88 shows the meter's circuit. C1 blocks dc voltage and R1 sets the input impedance. The upper section of the NE5532 op-amp is a precision rectifier. D1-D4 should be 1N4148. The lower section of the NE5532 is a supply splitter and can be replaced by a low-bandwidth op-amp, such as a 741 or equivalent. If substituted, the upper op-amp should have a gain-bandwidth product of at least 10 MHz.

R3 adjusts the voltage-to-current conversion ratio. With a 100 µA full-scale meter, maximum voltage range is approximately 1 V. There is enough adjustment range to calibrate the meter for peak, RMS, or peak-to-peak voltage readings.

Fig 25.89 shows the meter circuit construction. Leads should be kept short and the entire circuit can be constructed as point-to-point wiring and mounted directly on the back of the analog meter. The original meter used a BNC connector for input connections.

25.8.16 A Low-Frequency VNA Adapter

(The following project is based on the *QEX* article "A Low Frequency Adapter for you Vector Network Analyzer (VNA)" by Jacques Audet, VE2AZX. An overview is given here while the full description, including links to construction files and other information, is provided in PDF on this book's CD-ROM.)

The Low Frequency Adapter adds low frequency capability to a vector network analyzer (VNA), as well as adding audio frequency generation, 1 MΩ probe amplified interface, and direct conversion receiver capability. In the VNA application, an IF bandwidth of 10 Hz must be used to extend the lowest frequency down to 20 Hz. The low frequency adapter has allowed the author to measure R, L and C components down to 20 Hz, using a Hewlett Packard 8753D VNA. He has also been able to accurately characterize and document the response of many audio type amplifiers that otherwise would have required tedious measurement methods. More information on this adapter is available at **ve2azx.net/technical/LFA/LowFreqAdapter.htm**.

All S_{21} measurements are performed within the 10 to 15 MHz frequency range of the VNA. The low frequency measurements include:

- S_{21} magnitude and phase
- Group delay
- Compression point at a single frequency
- TRU calibration, with the device under test bypassed with a short circuit, to set a reference amplitude and phase frequency response.

The adapter can also be used as probe buffer/amplifier, a low frequency signal generator, a direct conversion receiver, and a vector voltmeter at low frequencies. S_{11} (reflection coefficient) measurements cannot be made directly with this adapter but the author provides a method and supporting spreadsheet for converting transmission measurements to impedance values. Construction of the adapter is shown in **Fig 25.90** and **Fig 25.91**.

The VNA is configured to use frequencies from 10 MHz to 15 MHz on its output (port 1) when measuring S_{21} (attenuation or gain as well as phase shift between ports 1 and 2). **Fig 25.92** shows the adapter's block diagram. The output signal is mixed in a double balanced mixer (DBM) with a local oscillator (LO) of 10 MHz. The difference signals from 20 Hz to 5 MHz are passed through a low-pass filter (LPF) and are available at the transmit (TX) output port for frequency response testing.

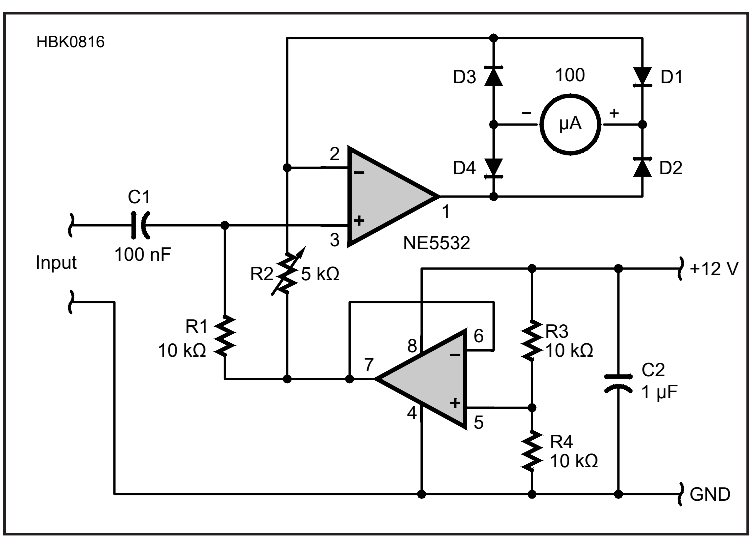


Fig 25.88 — RF voltmeter schematic. See text for op-amp substitution requirements. D1 – D4 are 1N4148.

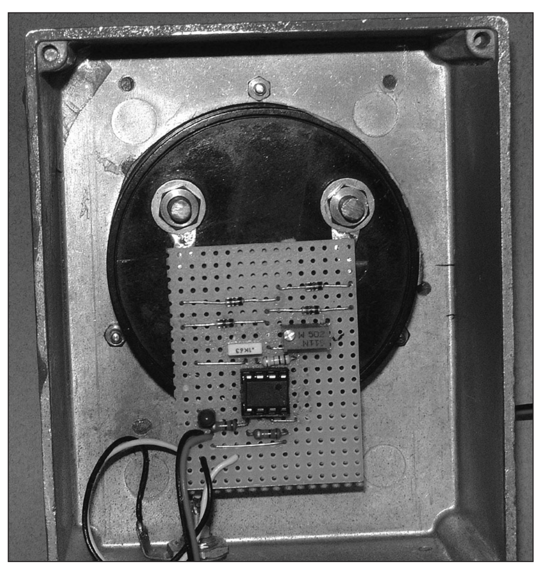


Fig 25.89 — RF voltmeter construction. Keep leads short. Point-to-point construction can be used and mounted directly on the meter for mechanical simplicity.

Fig 25.90 — The adapter is built in a standard Hammond extruded aluminum cabinet. Front and rear panels are pictured.

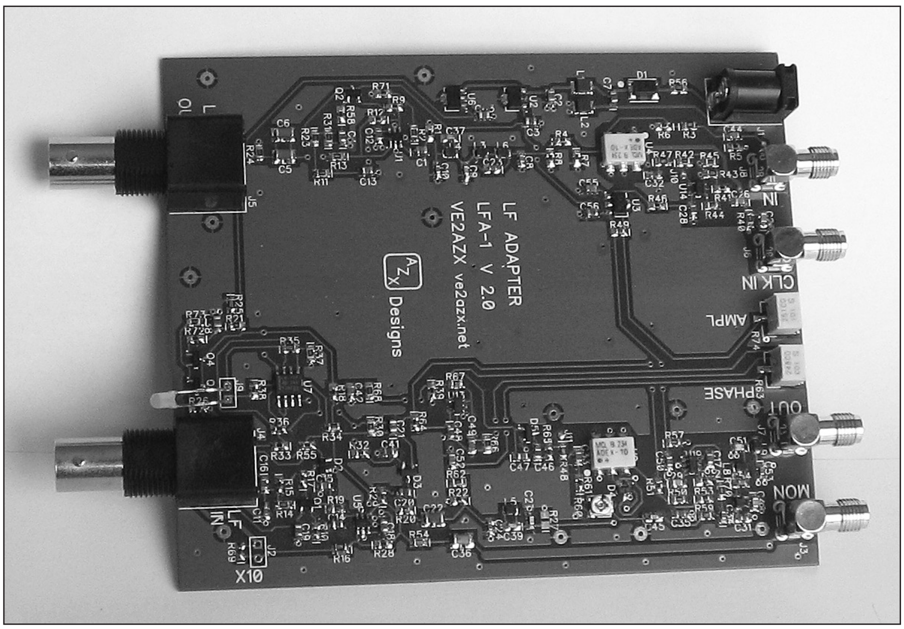


Fig 25.91 — The assembled circuit board for the adapter. R, L, and C components are 0805 SMT packages. Smaller 0603 components may also be used and will fit on the pads. See the CD-ROM version of this article and the author's website for additional information on the PCB.

The output of the device under test (DUT) is fed to a high impedance buffer and to a 5 MHz low pass filter before being re-multiplexed in the 10 to 15 MHz range by a second DBM.

The signal at the RF output of the first mixer generates sidebands (mixing products) above and below the 10 MHz LO frequency. The VNA synchronously demodulates the upper sideband and uses this signal to compute the attenuation or gain of the device under test. Both the first and second 5 MHz filters (LPF1 and LPF2) provide attenuation of the 10 MHz LO signal, so it does not pass thru the device under test. These filters also greatly attenuate the sum frequencies in the 20 to 25 MHz range, which could decrease the accuracy if these were present at the second DBM IF input.

Since the VNA does coherent detection of the signal (in order to measure the phase), it is necessary to synchronize its internal clock with the low frequency adapter LO signal. This is normally done by using a common external 10 MHz clock feeding the VNA and the low frequency adapter. This also enables the low frequency adapter to do S_{21} phase

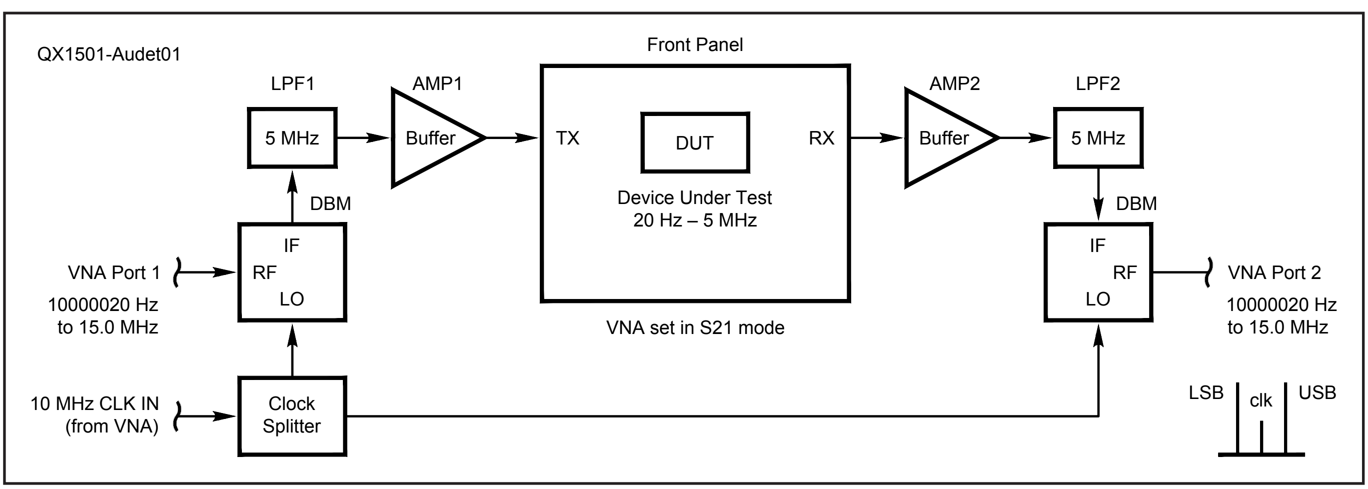


Fig 25.92 — Basic block diagram of the low frequency adapter for vector network analyzers.

measurements from 20 Hz to 5 MHz.

In order for the low frequency adapter to be as transparent as possible to the VNA, it has unity gain from its input and output. Also the TX output port has a 50 Ω impedance to drive low impedance loads. On the receive side, the input impedance consists of 1 MΩ in parallel with 8 pF so that it is compatible with oscilloscope probes. An additional capacitor may be added to match a specific scope's input capacitance, thus providing a flat frequency response with an external ×10 probe. The high impedance allows the user to terminate the device under test by shunting a parallel termination across the receive (RX) input.

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25.10 Test and Measurement Glossary

Accuracy — The maximum expected error in a measurement.

Ammeter — A device for measuring electrical current.

Antenna analyzer — A device to measure the RF impedance of a one-port network such as an antenna.

Antenna test range — An area designed to minimize the effect of RF reflections to permit accurate antenna gain and pattern testing.

Attenuator — A broadband device that reduces the amplitude of a signal by a specified, well-controlled amount.

Autoranging — The ability of a *multimeter* to set its range automatically based on the signal level.

AWG (arbitrary waveform generator, also ARB) — An instrument that can generate a complex signal based on waveforms stored in memory.

Birdies — Slang term for internally-generated spurious signals in a receiver that may be steady, warble, or pulsate.

Blocking dynamic range — The difference between the *blocking level* and the *MDS*.

Blocking level — The level of an interfering signal that causes weak signals to be reduced in amplitude by 1 dB.

Bolometer — A device for measuring RF power by measuring the heat dissipated in a dummy load.

Bridge — A circuit used to indicate the relative values of passive circuit elements by observing the null on a meter or other indicator. See this chapter's discussion of Wheatstone bridges.

Burden voltage — The full-scale voltage drop of an ammeter.

Coaxial detector — An RF detector with a coaxial connector.

Combiner — A device to combine signals from two sources. A *power splitter* in reverse. See ***Hybrid combiner.***

D'Arsonval meter — The most common type of mechanical meter, consisting of a permanent magnet and a moving coil with pointer attached.

Dip meter (dip oscillator) — An instrument with an oscillator whose coil is external to the enclosure so it can be coupled to the circuit under test. The meter value drops (dips) when the oscillator frequency is tuned to the frequency of resonance of the circuit under test.

Directional coupler, directional bridge — A device that senses the RF power flowing in one direction on a transmission line.

DMM (digital multimeter) — A test instrument that measures voltage, current, resistance and possibly other quantities and displays the result on a numeric digit display, rather than on an analog meter.

DVM (digital volt meter) — See *DMM*.

Dynamic error — An error whose value depends on the time of measurement.

Dynamic range — The difference in dB between the strongest and weakest signals that a receiver can handle.

Electronic voltmeter — An amplified analog *multimeter*.

ENR (excess noise ratio) — The ratio of the noise added by a noise source to the thermal noise, normally expressed in dB.

Feed-through termination — See *Through-line termination*.

Field-strength meter — A device to indicate radiated RF field strength.

Four wire — A technique for measuring small resistances or large currents that uses four wires rather than two to reduce errors due to test lead resistance.

Frequency counter — A device that measures the frequency of a periodic signal and displays the result on a digital readout.

Frequency domain — Description of signals as a function of frequency, as opposed to as a function of time.

Function generator — An audio tone generator that generates tones in the shape of various functions, such as sine, square, triangle and ramp waveforms.

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

Gaussian distribution — The distribution of the probability of the instantaneous voltage of a noise signal as a function of the voltage. It forms a bell-shaped curve with a maximum at the dc value of the signal (typically zero volts).

GDO (grid-dip oscillator) — A *dip meter* that uses a vacuum tube.

Harmonic signal — A periodic signal. Its frequency spectrum consists only of the fundamental and harmonics of the repetition rate.

Hybrid combiner — A passive device used to combine two signals in such a way as to reduce the interaction between the signal sources.

IF and image rejection — The difference in level between an interfering signal at a receiver's IF or image frequency and the level of the desired signal that produces the same response.

IMD (intermodulation distortion) — The creation of unwanted frequencies because of two or more strong signals modulating each other.

IMD dynamic range — The difference in signal level between two signals that cause IMD products at the level of the MDS and the MDS.

LCR Bridge — A device for measuring inductors, capacitors or resistors using a Wheatstone bridge.

Lissajous Pattern — As used by amateurs, the combined display of two sine wave signals on an oscilloscope to determine the relationship between their frequencies. (Also called Lissajous figure.)

Logic analyzer — A sophisticated instrument for analyzing digital circuitry.

Logic probe — A simple device for sensing and displaying a digital signal's state.
Marker — An indicator on the screen of a *spectrum analyzer* that allows reading out the frequency and amplitude of a specific signal. See also ***Frequency marker.***
Marker delta — A *spectrum analyzer* feature that reads out the difference in frequency and amplitude of two markers.
MDS (minimum discernible signal) — The level of the *noise floor* at the antenna connector of a receiver. Depends on the measurement bandwidth.
Multimeter — A device that measures several electrical quantities, such as voltage, current and resistance, and displays them on an analog meter or digital display.
Multiplier (voltage multiplier) — A resistor placed in series with a meter to increase the full-scale voltage reading.
Near field — The area close to an antenna where the electromagnetic wave is not completely formed. Antenna gain and pattern measurements are not valid in this region.
Network analyzer — An instrument to measure the return loss and transmission gain between the two ports of a two-port network.
NIST (National Institute of Standards and Technology) — A non-regulatory agency of the US federal government that manages measurement standards. It used to be called NBS, the National Bureau of Standards.
NIST traceable — Refers to a device whose accuracy is based on NIST standards using rules and procedures specified by NIST.
Noise bandwidth — The bandwidth of a rectangular-spectrum filter that would produce the same total noise power as the filter under consideration.
Noise density — Noise power per hertz.
Noise figure — A figure of merit for the sensitivity of receivers and other RF devices. It is the ratio of the effective noise level to the thermal noise level, usually expressed in dB.
Noise floor — The noise received by a receiver in a specified bandwidth, referenced to the antenna connector. It is the thermal noise in dBm plus the noise figure in dB.
Noise source — An instrument that generates well-calibrated white noise for test purposes.
OCXO (oven-controlled crystal oscillator) — A crystal oscillator mounted in a temperature-controlled oven to improve the frequency stability.
Ohmmeter — A meter that measures resistance. Usually part of a *multimeter*. See ***VOM*** and ***DMM***.
Ohms per volt — A measure of the sensitivity of a voltmeter that has multiple scales. It is the reciprocal of the current drawn by the meter.
Oscilloscope — An instrument that displays signals in the *time domain*. Has a graphical display that shows amplitude on the vertical axis and time on the horizontal axis.
Peak search — A feature of a *spectrum analyzer* in which a marker is automatically placed on the strongest signal in the display.
Peak to peak value — The difference between the most positive and most negative signal values. For a symmetrical signal it is twice the peak value.
Peak value — The highest value of a signal during the measuring time.
Periodic — Refers to a signal that repeats exactly at a regular time interval, the period.
Phase noise — Wideband noise on an RF signal caused by random fluctuations of the phase.
Port — A pair of connections to a network, typically via a coaxial connector
Power divider (power splitter) — A device to divide an RF signal between two loads. A *combiner* in reverse.
Prescaler — A circuit used ahead of a counter to extend the frequency range. A counter capable of operating up to 50 MHz can count up to 500 MHz when used with a divide-by-10 prescaler.
Random error — A non-repeatable error caused by noise in the measurement system.
Reciprocal mixing — The mixing of a nearby interfering signal with the phase noise of the receiver local oscillator, which causes noise in the audio output.
Resolution — The smallest distinguishable difference in a measured value.
Resolution bandwidth — The smallest frequency separation between two RF signals that a *spectrum analyzer* can resolve. It is determined by the IF filters in the *spectrum analyzer*.
Return loss — The ratio, usually expressed in dB, between the incident and reflected RF signals.
Return-loss bridge (RLB) — A bridge used for measuring the return loss of an RF circuit, transmission line or antenna.
Reverse power protection — A *signal generator* feature to protect the instrument from accidental transmission by a transceiver under test.
RF probe — A hand-held probe with a detector to allow measuring RF signals with a dc voltmeter.
Scalar network analyzer (SNA) — A *network analyzer* that measures only the magnitude of the gain and return loss.
Scattering parameters (S parameters) — A set of four parameters to characterize the complex return loss and transmission gain in both directions of a two-port network.
Scope — Slang for *oscilloscope*.
Second-order IMD — IMD caused by strong signals at frequencies f1 and f2 that occurs at a frequency of f1 + f2.
Service monitor — An integrated package of test equipment packaged as a single instrument for testing receivers and transmitters.
Shunt (meter shunt) — A resistor connected in parallel with a meter to increase the full-scale current reading.
SI (Système International d'Unités) — The modern, revised version of the metric system.
Signal generator — An instrument that generates a calibrated variable-frequency RF signal, usually with adjustable amplitude and modulation capability.
SINAD (signal plus noise and distortion) — A measure of the relative level of signal compared to the noise and distortion at the audio output of an FM receiver.
Single-shot — Refers to a non-periodic signal that can be measured in a single time interval.
Spectrum analyzer — An instrument that displays signals in the *frequency domain*. Has a graphical display that shows amplitude (normally in logarithmic, or dB, form) on the vertical axis and frequency on the horizontal axis.
Spurious emissions, or spurs — Unwanted energy generated by a transmitter or other circuit. These emissions include, but are not limited to, harmonics.
Standard — A rule for the proper method to measure some quantity. A standard may involve a standard artifact that defines the unit.
Step attenuator — An attenuator that can be switched between different attenuation values.
Systematic error — A repeatable error due to some characteristic of the measurement system.
TCXO (temperature-compensated crystal oscillator) — A crystal oscillator that includes circuitry to compensate for frequency drift with temperature.
Termination — A resistor with a coaxial connector used to terminate a transmission line in its characteristic impedance.
Test set — A *network analyzer* accessory that automatically configures the connections to the device under test for various measurements.

Third-order IMD — IMD caused by strong signals at frequencies f1 and f2 that appears at frequencies 2f1 – f2 or 2f2 – f1.

Third-order IMD intercept point (IP3) — The power level representing the intersection of plots on a dBm scale of the power level of the two strong signals and their IMD products.

Through-line termination — A termination with two coaxial connectors that connect straight through and a terminating resistor to ground.

Time base — A highly-accurate reference oscillator used in a frequency counter or other device that needs an accurate time or frequency reference.

Time domain — Refers to the variation in time of electronic signals, as opposed to their frequency spectrum.

Tracking generator — A *spectrum analyzer* accessory that consists of a signal generator whose frequency tracks the frequency of the analyzer.

True RMS — Refers to a meter that measures the actual RMS voltage or current instead of measuring the average or peak value and calculating the RMS value assuming a sinusoidal waveform.

VCXO (voltage-controlled crystal oscillator) — A crystal oscillator whose frequency can be adjusted slightly using an external applied voltage.

Vector network analyzer (VNA) — A *network analyzer* that measures both the magnitude and phase of the gain and return loss of a two-port network.

Video bandwidth — The bandwidth of the post-detection filter in a *spectrum analyzer*. It limits the maximum sweep speed.

VOM (volt-ohm meter) — A *multimeter* that does not include active circuitry such as an amplifier.

VTVM (vacuum-tube voltmeter) — A *multimeter* that uses one or more vacuum tubes.

Wheatstone bridge — See *Bridge circuit*.

Wilkinson combiner — A type of *hybrid combiner*.