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Chapter 25

Test Equipment and Measurement

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 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.

The chapter has two major sections: Instruments and Measurements. Reorganized from the original chapter written by Alan Bloom, N1AL, the first section covers the different types of instruments used in amateur stations. (Basic ac, dc, and RF measurement units of measurements are covered in the chapters on **Electronic Fundamentals**, **Radio Fundamentals**, and **Circuits and Components**.) The second section describes the types of measurements performed on RF electronics and in amateur radio stations. Transmitter and receiver performance tests are provided, including the standard tests performed by the ARRL Lab for *QST* Product Review and other evaluations. Transceiver tests have been updated based on material provided by Adam Farson, AB4OJ/VA7OJ. Material on using antenna and vector network analyzers (VNA) was updated by Jim Brown, K9YC. Finally, a selection of test equipment construction projects is presented as well.

25.1 Measurement Fundamentals

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 $s \times A$ 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: **www.nist.gov**). (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 Accuracy, Precision, and Uncertainty

Accuracy and precision are different measurement specifications that are often confused. *Accuracy* is the maximum expected error in the measurement. *Precision* is the ability of a device to make consecutive measurements that are close to one another.

Figure 25.1A illustrates the difference by placing hits on a target. Higher accuracy means being able to hit a specific location on any single attempt. Higher precision means being able to hit the same location on multiple attempts.

Resolution is the smallest distinguishable difference in a measured value. In Figure 25.1B, the target hits are all in the same location but the increased number of rings allows for higher resolution in stating their location. Another measurement characteristic is *sensitivity* — the smallest absolute value of change that can be detected by the instrument.

Expressing these characteristics in terms of measurements, assume an 8-digit frequency

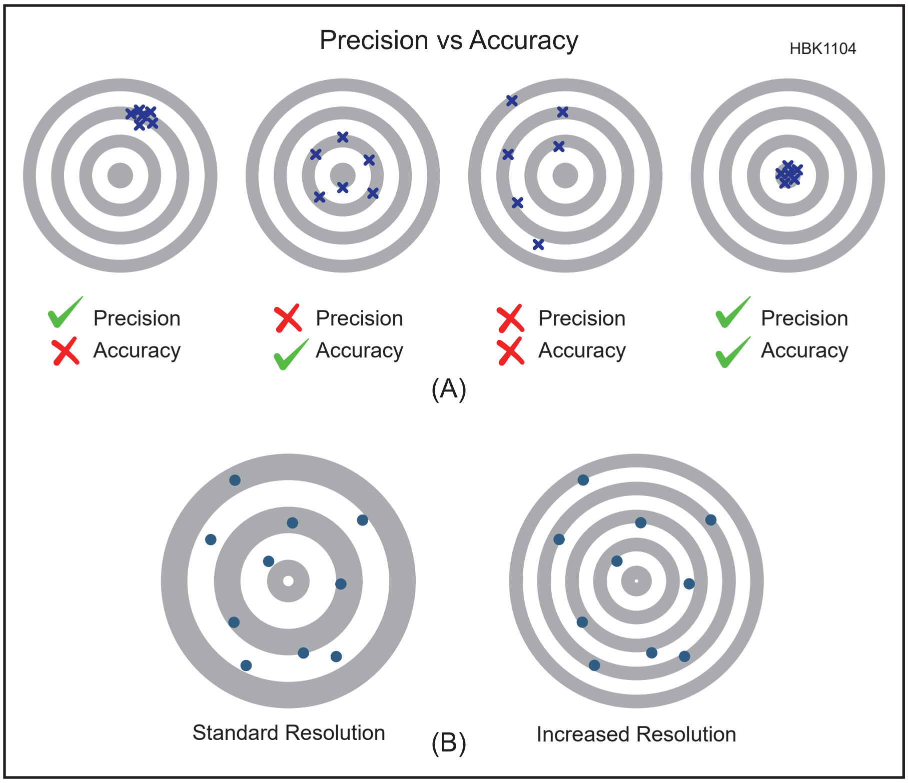


Figure 25.1 — An illustration of precision versus accuracy (A) and of resolution (B).

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. The instrument's precision would be determined by its ability to produce the same result if the same signal was measured multiple times. In the case of a frequency counter, the sensitivity would be determined by the smallest signal voltage level for which a measurement could be made.

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.

25.1.3 Measurement Error and Uncertainty

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.

Measurement uncertainty must be considered when the test system uses multiple instruments. Each instrument has its own accuracy limitations. Different instruments may have different precision, as well. As a result, even if the parameter being measured has exactly the same value, the measurements are likely to vary. Even successive measurements with the same instrument in the same circumstances will vary. Measurement uncertainty is a statistical expression of how the measured values will fall within a range of accuracy.

25.2 Basic Test Meters

Before beginning, be aware that both multi-function test instruments and standalone "panel" meters are both referred to as "meters." To make things even more confusing, multi-function instruments are often called "voltmeters," as well! Avoid confusion by considering the context of the material.

25.2.1 Voltmeters

Voltage is always measured "with respect to" some other point. Voltage is a measure of the *difference* in potential between two points, not a property of a single point. Most circuits have a *reference point* and voltages are specified or measured with respect to that point. Normally the reference point is understood to be the chassis or circuit common.

Unfortunately, the common connection for a circuit is often referred to as "ground" which suggests that it must be connected to a ground conductor or the Earth for proper operation. This use of the word "ground" only means "the local reference voltage." 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.

25.2.2 Ammeters

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.

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.

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.

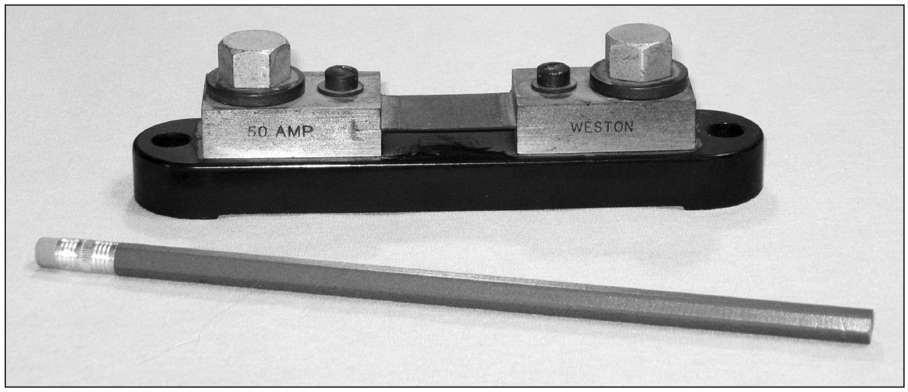


Figure 25.2 — This 50-A, 50-mV current shunt has a resistance of 0.05 / 50 = 0.001 Ω. The two large terminals are for connecting to the circuit under test and the two small terminals are for connecting to a voltmeter.

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 **Figure 25.2**. 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. (This technique is discussed later in the chapter.)

25.2.3 Ohmmeters

An *ohmmeter* is a meter designed for measuring resistance. Various circuits can be used, but most are variations of the simplified schematics in **Figure 25.3**. 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.

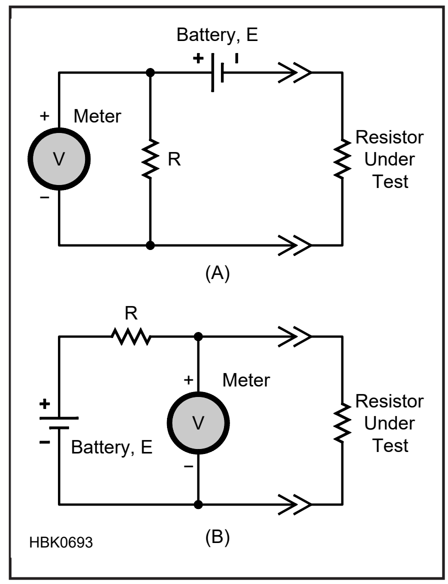


Figure 25.3 — Two ohmmeter circuits. At A, the meter reads full scale with a zero-ohm resistor and reads zero with no resistor connected. The circuit at B is the opposite; the meter reads full scale with no resistor connected (infinite resistance) and reads zero with zero resistance.

25.2.4 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 by measuring the forward voltage drop across them. It is increasingly common for even low-cost digital multimeters to include functions such as frequency and capacitance measurement.

ANALOG INSTRUMENTS

An analog volt/ohm/current meter (referred to as a "VOM" here) contains 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 analog instrument 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 voltmeters* (VTVM). An example is shown in **Figure 25.4**. The modern equivalent is called an *electronic voltmeter* and generally uses an amplifier with field-effect transistors at the input.

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 semiconductors to conduct, giving erroneous readings of resistance. In some cases, the applied measurement voltage can be high enough to damage the circuit being measured. If there is any doubt, use an instrument with a diode-test function because that meter will use only a low voltage to measure resistance.

Figure 25.4 — This classic Hewlett-Packard HP412A vacuum-tube voltmeter (VTVM) has specifications that put many modern solid-state multimeters to shame.

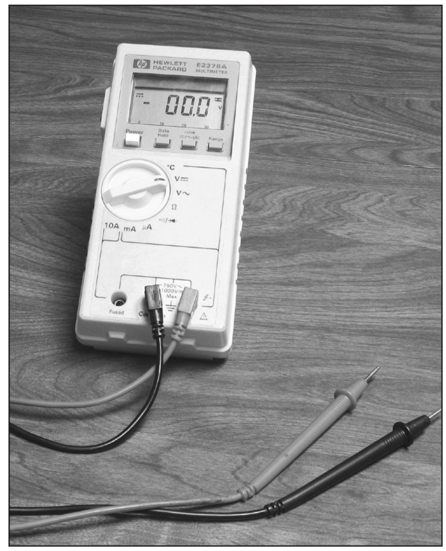


Figure 25.5 — A modern digital multimeter typically has a liquid crystal display readout.

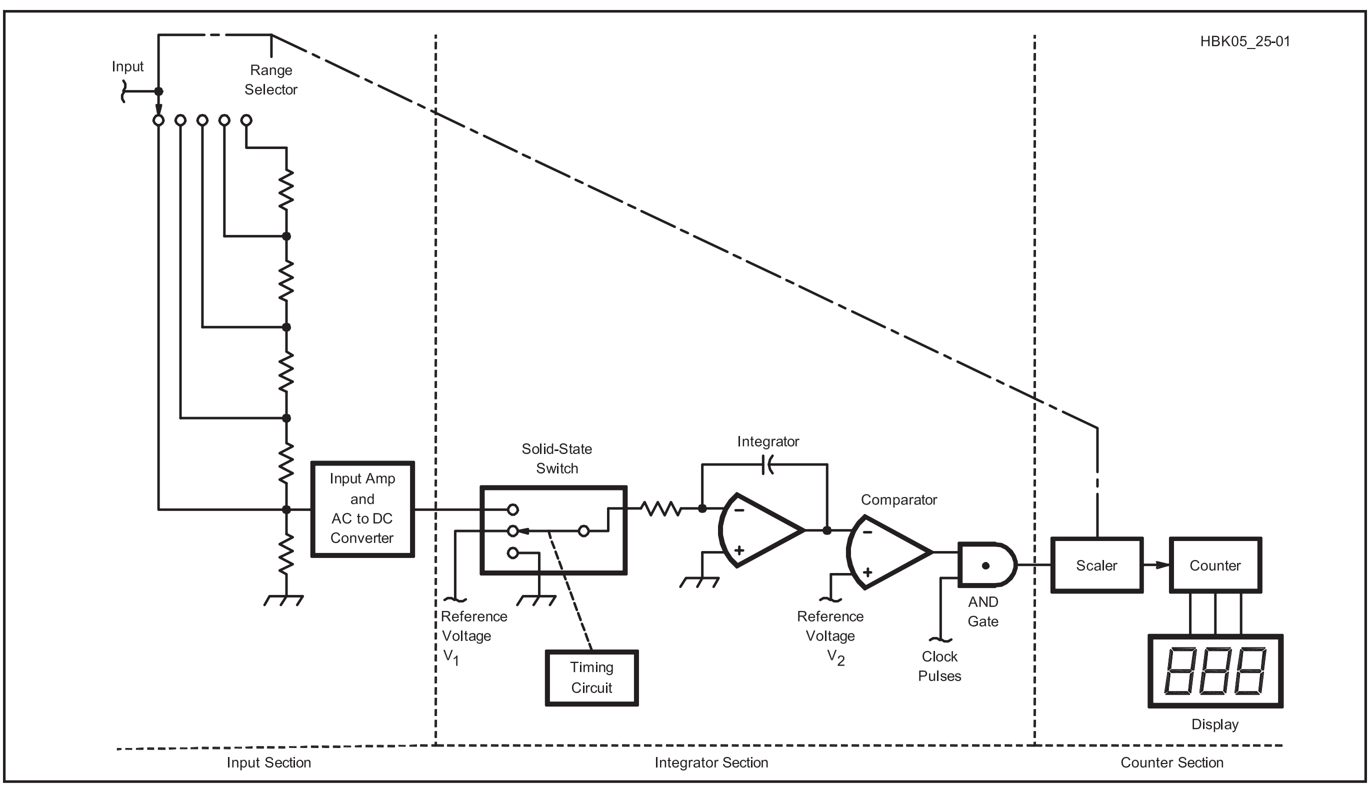


Figure 25.6 — A typical digital voltmeter consists of three parts: an input section for scaling, an integrator to convert voltage to a pulse whose width varies with voltage, and a counter to measure the width of the pulse and display the measured voltage.

DIGITAL MULTIMETER (DMM)

A *digital multimeter* (DMM) has a digital readout, usually an LCD display. A microprocessor controls the measurement process and the display. (See **Figure 25.5**.) 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 **DSP and SDR Fundamentals** chapter.)

A digital voltmeter is constructed as shown in **Figure 25.6**. The input section is the same as in an analog electronic voltmeter. A range 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 or value of resistance 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.

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. (See Reference for Hageman) This is a very useful feature for experimenters or for troubleshooting.

DMMs generally have greater precision than analog meters, often showing results to 4 or more significant figures. This is a detriment, though, for measuring signals that are varying over time or for trying to adjust for a peak or minimum value. The rapidly changing numerical display can be difficult to interpret as the signal changes. If autoranging is enabled, that can also cause the display to change format erratically. Many amateurs keep an analog meter available for such uses.

MULTIMETER SPECIFICATIONS AND FEATURES

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 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 highest. 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

Meter Safety

The following paragraphs are not a comprehensive treatment of safety, but provide some examples of important safety practices.

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.

On your meter is printed a maximum voltage rating (See **Figure 25.A1**). Heed it! That is the maximum voltage the probes, connectors, and body of the meter can withstand. A higher voltage may cause a *flashover* from the wiring to anything touching the case, such as your hand. Electrocution caused by using meters beyond their voltage rating happens regularly. In the photo, note that the probe connectors are recessed for extra insulation — use the right probes for the full rating.

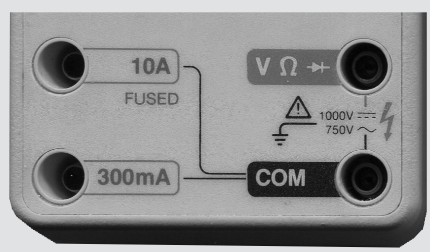


Figure 25.A1 — Voltage ratings are shown next to a voltmeter's probe jacks (1000 Vdc and 750 Vac). Exceeding these voltages can result in a flashover, presenting a severe electrocution hazard.

Special high-voltage probes, such as the Fluke 80K-40 or B&K HV44-A are available for measuring the multi-kV voltages sometimes found in tube RF amplifiers. When testing high-voltage equipment, don't touch the meter or oscilloscope when the equipment is energized and the probes are attached.

As probes, fixtures, and connectors age, their insulation can become brittle and crack. That compromises the insulation, and you can even come in direct contact with the internal wires. Before beginning to work on a piece of equipment using high voltage, carefully inspect your test probes and other equipment. Make sure the insulation is clean, flexible, and not scuffed or cracked so it can protect you as you expect. Replace anything that doesn't look safe.

High current can be dangerous as well. Make sure cables and connectors can carry the current without significant heating. If a probe accidentally shorts a power supply, 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.

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 evaluating 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. (See **www.prologix.biz**) Modern instruments typically have a USB or RS-232 interface.

25.2.5 Panel Meters

STANDALONE ANALOG METERS

The most common type of standalone analog meter 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. The D'Arsonval meter movement can be used to measure ac and dc voltage and current as well as resistance.

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 that the meter does not exceed the meter's current rating which can be very low for some meters. (You may want to measure the current from the meter before using it to test a panel meter.) Do not use a diode test function on anything except diodes and transistors since the test current of that function may be high enough to damage many sensitive meters.

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}.

DIGITAL PANEL METERS (DPM)

Digital panel meters (DPMs) are 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 4-½ 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.

25.2.6 Calibration of Meters

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.

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.

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 ensure the probe does not slip to or between adjacent pins is nearly impossible.

The solution is to build an adapter as shown in the following three examples.

Figure 25.A2 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.

Figure 25.A3 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 a 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.)

Figure 25.A4 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.)

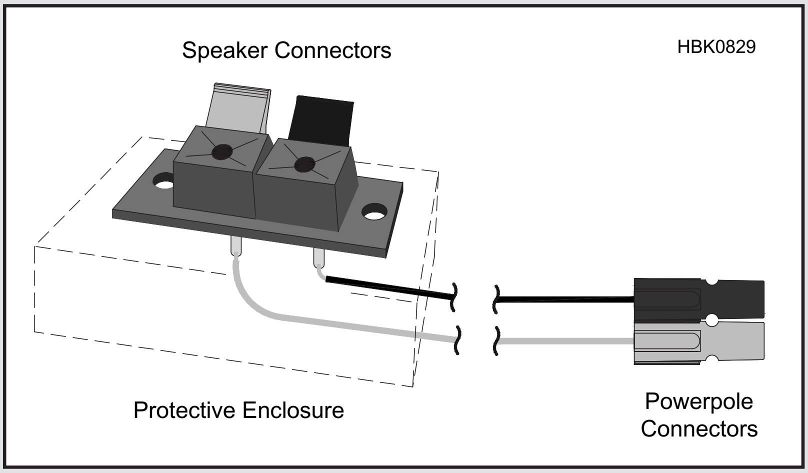


Figure 25.A2 — A convenient way to hold probes steady

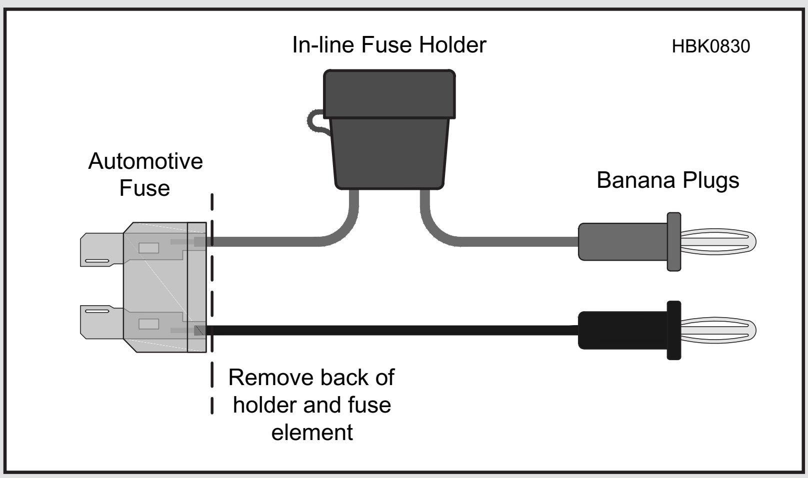


Figure 25.A3 — Adapting an automotive-style fuse to make a current-measuring adapter.

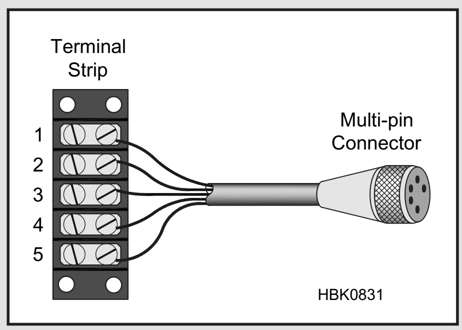


Figure 25.A4 — A typical adapter for a multi-pin connector.

25.3 Frequency Counters

The basic instrument for measuring frequency is the *frequency counter*. A block diagram of a very basic design is shown in **Figure 25.7.** 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.

MEASURING PERIOD

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.

PRESCALERS

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. A switch selects between the two inputs. (Note that for older equipment, many divide-by-10 prescalar ICs once widely used are mostly no longer available.)

INPUT IMPEDANCE AND SENSITIVITY

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 — even this may detune the oscillator slightly.

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

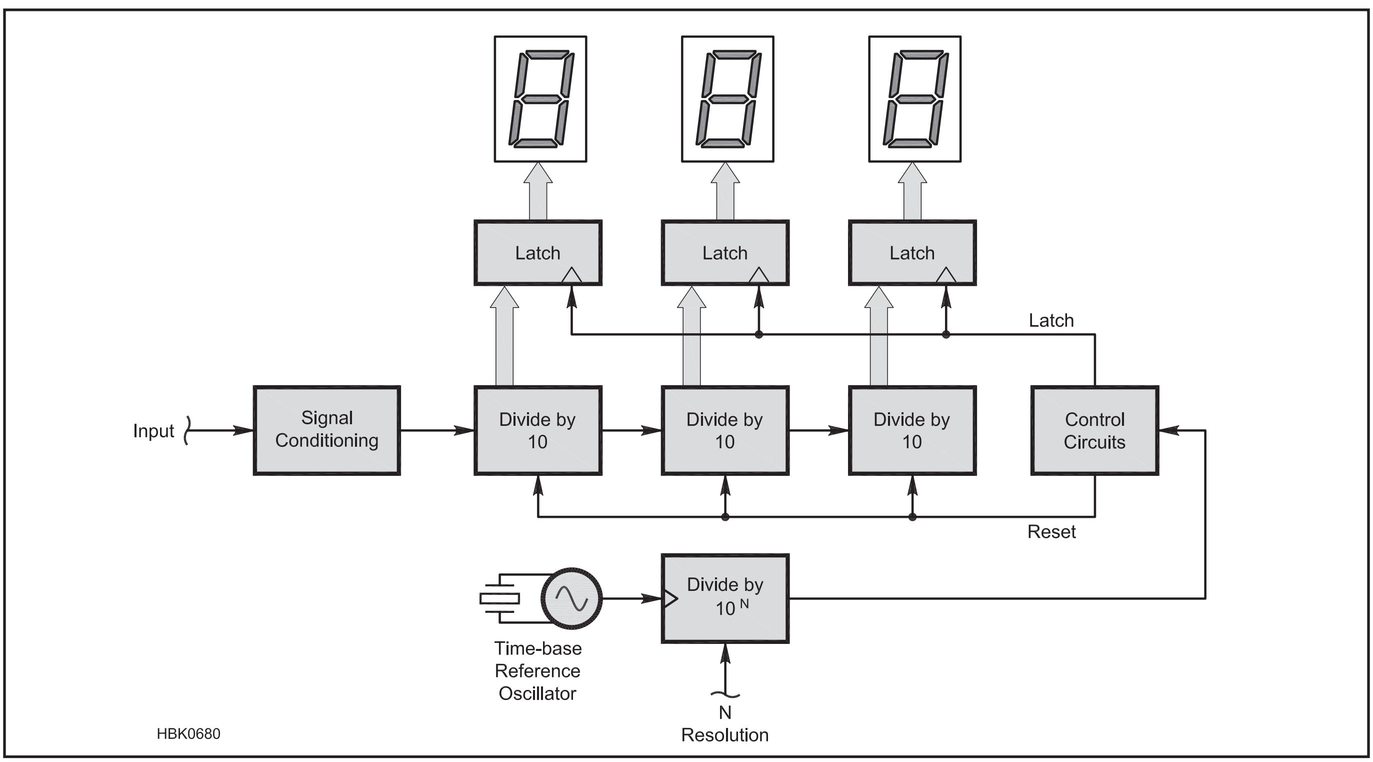


Figure 25.7 — A simplified block diagram of a frequency counter. The display update rate and the resolution are controlled by the divide ratio of the divider at the output of the time-base oscillator.

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. (Don't place the antenna close to the transmit antenna or it may pick up enough energy to damage the counter.) The transmitter should not be modulated while measuring its frequency. SSB transmitters should be measured in CW mode.

ACCURACY AND TIME BASE

The principal 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.

CHOOSING A COUNTER

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.

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.

25.4 Signal Generators

25.4.1 Function Generators

A *function generator*, also known as a *waveform generator*, is a type of 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. Function generators are generally less accurate than the signal generators discussed in the next section.

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 **Figure 25.8**) is also designed to be used as a low frequency function generator. (See the Reference for Richardson. DDS is discussed in the **Oscillators and Synthesizers** chapter.)

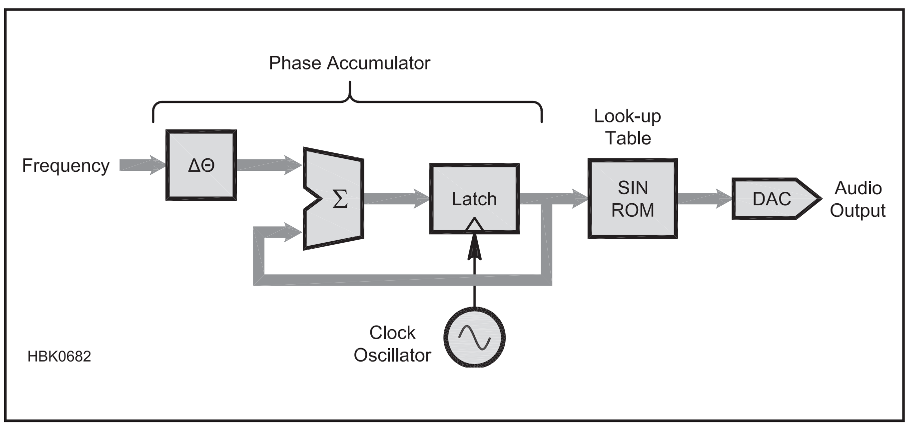


Figure 25.8 — A direct digital synthesizer (DDS) sine-wave generator. Not shown is a low-pass filter at the output that attenuates unwanted spurious frequencies above one-half the clock frequency.

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 Figure 25.8 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.

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 Keysight 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.

Figure 25.9 — 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.

25.4.2 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* (RPP). It protects the sensitive output circuits in the event that the transceiver accidentally goes into transmit mode while connected to the signal generator. RPP circuits typically have a protection range of 1 to as much as 50 watts. Few RPP circuits will protect a signal generator from a 100-watt signal. Be completely aware of all of the ways that a transceiver can be put into transmit and remove all microphones, keying, and PTT lines before connecting it to any test instrument other than a watt meter or SWR meter. If possible, turn the transmit power to as low as possible, as well.

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 or synthesized HP8640B signal generators have better performance in this respect than some modern synthesized instruments. See **Figure 25.9**. 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 20-dB fixed or step attenuator. Just remember to subtract any attenuation used 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.

25.5 Inductance and Capacitance Testers

25.5.1 LCR Measurement Using a DMM

Some multimeters have built-in capability to measure capacitance. They apply a dc pulse and measure the resulting RC time constant as the capacitor under test charges. This type of measurement has an accuracy of about 1%. Measurements are generally made without a calibration option to compensate for the effect of probes and stray capacitance. Inductance is more difficult for the low-power electronics of a DMM to measure and most models do not offer inductance measurement. A guide to measuring capacitance with a DMM is available at **www.fluke.com/en-us/learn/blog/digital-multimeters/how-to-measure-capacitance**.

25.5.2 LCR Meters and Bridges

There is a wide variety of instruments to measure component values. These range from low-cost handheld instruments for hobbyist use to benchtop units that support manufacturing. The accuracy of even simple handheld meters is suitable for amateur radio applications. **Figure 25.10** shows the popular Peak Atlas LCR 45 model that can measure 0–2 MΩ, 0–10,000 µF, and 0-10 H with accuracies of 1 to 1.5%.

Figure 25.10 — Peak Atlas LCR45 LCR and Impedance Meter.

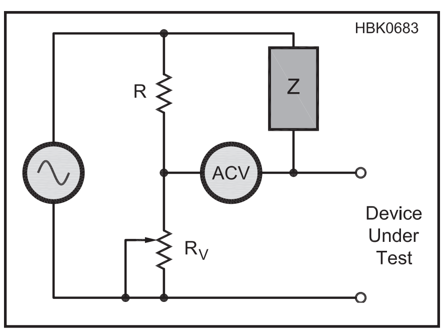


Figure 25.11 — An LCR bridge, which is a type of Wheatstone bridge used to measure inductors (L), capacitors (C) and resistors (R). The box labeled "Z" is the reference component to which the device under test is compared.

Both handheld and benchtop LCR meters measure inductance and capacitance by using a phase-shift method. An ac signal is applied to the component under test through a resistor internal to the meter. The resulting phase shift between voltage and current waveforms is measured as a time interval that is converted to a value of phase. Since the internal resistor value is known, the L or C value can be calculated. There are several values of resistance available so that a wide range of inductance can be measured. The manual will explain the measurement process, which includes "calibrating out" the effects of leads and test fixtures.

Benchtop LCR meters generally offer features such as programmable frequencies, measurement accuracy to 0.01%, computer control and data collect for automated applications. Advanced feature such as dc bias voltage, and dc bias current and sweep capability are common. LCR meters in this category are used for ac calibration of inductance, capacitance and resistance standards, dielectric constant measurements with a variety of dielectric cells, and production testing of components and sensors.

LCR BRIDGES

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 **Figure 25.11**, 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.

MEASURING LCR WITH A DIP METER

A dip meter (described in the RF and Microwave Test Accessories section) 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 dip meter. The inductance or capacitance can be determined from the dip frequency, f, using the formula

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

25.5.3 Component Test Frequency

Electrical components need to be tested over the frequency range at which the component will be used. An instrument with a wide frequency range and multiple programmable frequencies provides this capability. Common measurement frequencies are dc, 50/60 Hz, 120 Hz, 1 kHz, 100 kHz, and 1 MHz. LCR meters with programmable frequencies provide the most flexibility, in matching frequency of measurement to the frequency at which the component will be used. In R&D applications, variable measurement frequency is useful to determine useful frequency range or resonance. Most LCR meters today use an AC test signal over a frequency range of 10 Hz to 2 MHz. The measured inductance of coils using high permeability ferrite cores will often vary significantly with frequency.

25.5.4 Equivalent Series Inductance (ESL) and Resistance (ESR)

The use of capacitors in switchmode supplies 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 described above.

25.6 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.

While scopes being sold today are digital, much of the terminology used to describe their operation was defined for analog scopes. Describing how analog scopes work will help explain the function of controls with the same name on digital scopes, even though a digital scope does not have the same internal circuitry as an analog scope. 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. It is the scope function that is useful, regardless of how it is implemented.

It is also important to note that inexpensive analog oscilloscopes may not have better than 5% accuracy. Lab-quality analog scopes, while they might be fairly old, are still available and can be calibrated to close to their original specifications. Such scopes are economically at almost every hamfest flea market. Regardless of price and age, analog scopes are still useful, with many models suitable for HF and even VHF use.

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.

25.6.1 Analog Oscilloscopes

Figure 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. 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 component of an ac/dc signal or the complete 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. Solid state dip meters are widely available used. Be sure the plug-in coils are included. A project to build your own dip meter is included in the online material.

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 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

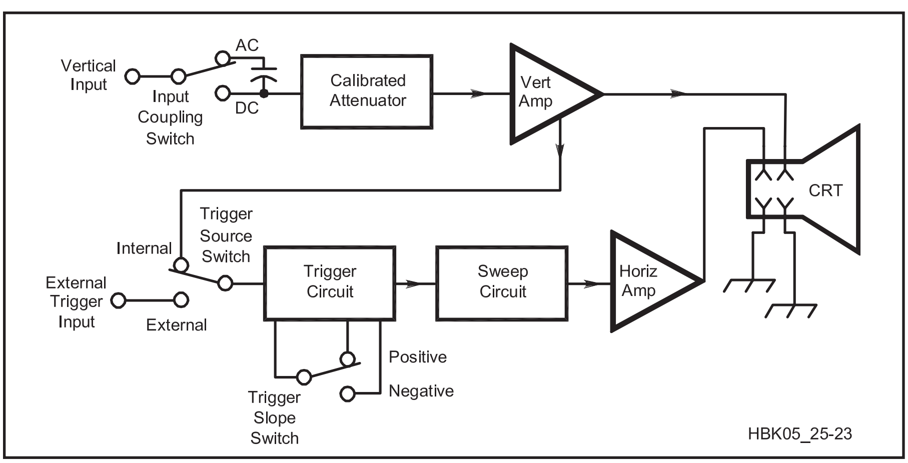


Figure 25.12 — Typical block diagram of a simple triggered-sweep oscilloscope.

timed voltage ramp (see **Figure 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. **Figure 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.

Figure 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.

Single sweep is useful for looking at a nonrepetitive 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.

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

25.6.2 Dual-Trace Functions

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. **Figure 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).

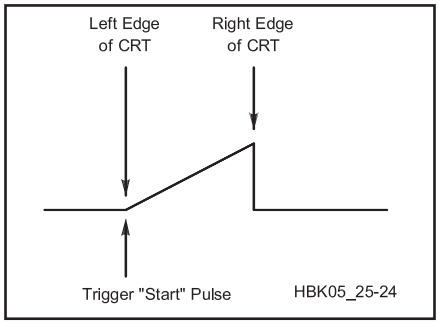


Figure 25.13 — The sweep trigger starts the ramp waveform that sweeps the CRT electron beam from side to side.

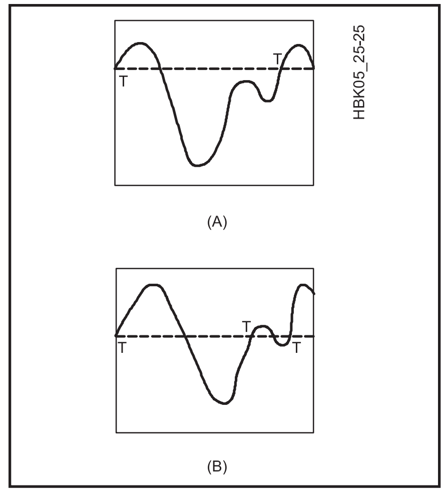


Figure 25.14 — In order to produce a stable display the selection of the trigger point is very important. Selecting the trigger point in A produces a stable display, but the trigger shown at B will produce a display that "jitters" from side to side.

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)

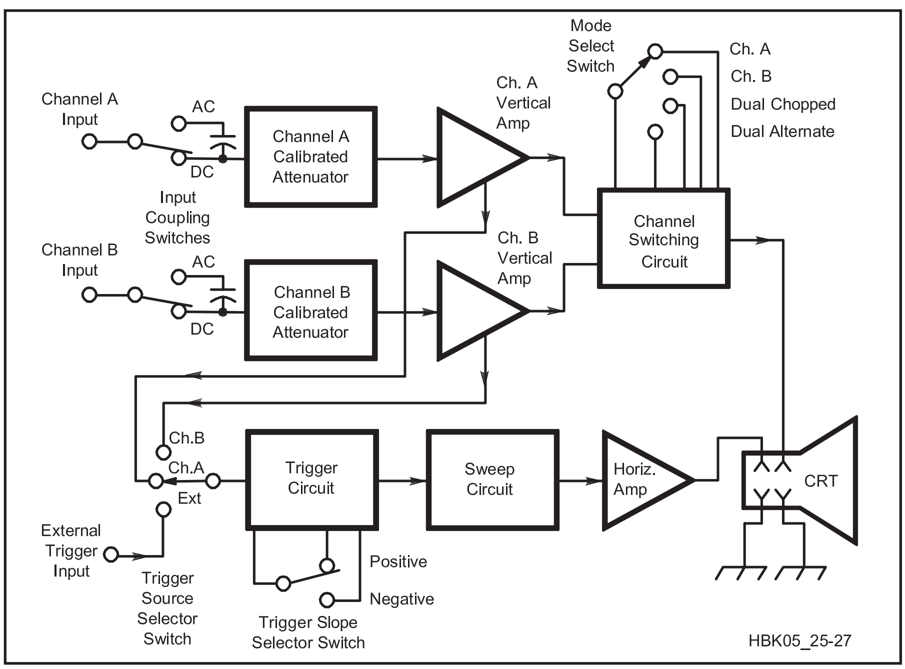


Figure 25.15 — Simplified dual-trace oscilloscope block diagram. Note the two identical input channels and amplifiers.

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.

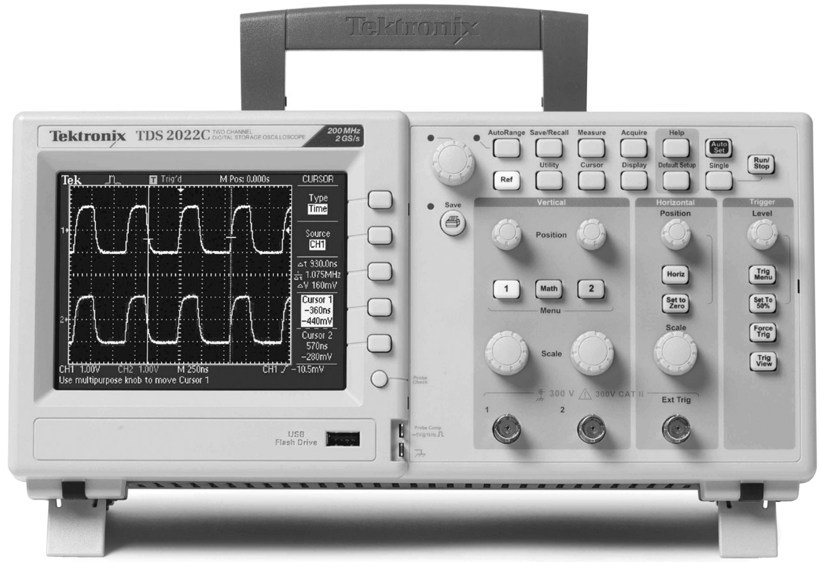
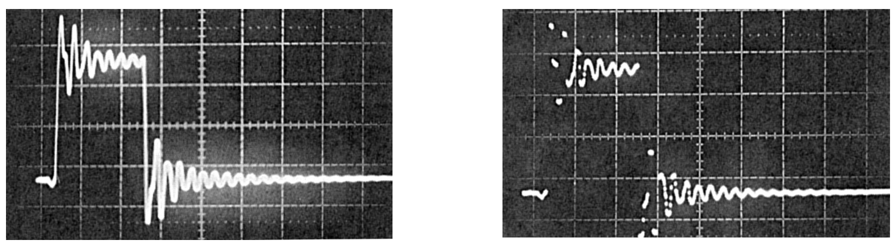


Figure 25.16 — The Tektronix TDS 2000C series is a typical example of digital oscilloscopes using an LCD display.



(A) (B)

Figure 25.17 — Comparison of an analog scope waveform (A) and that produced by a digital oscilloscope with a low sampling rate (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 signal changes the scope is capable of displaying.

25.6.3 Digital Oscilloscopes

Signal processing in a digital oscilloscope is performed with microprocessor-based digital circuitry. A lightweight LCD display is used for true portability. (See **Figure 25.16**) The microprocessor determines which pixels to light up to draw the traces on the screen. 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 data file if desired.

You can also use your personal computer as an oscilloscope with an external signal digitizer module (a high-speed, multi-channel analog-to-digital converter) 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.

Most stand-alone digital scopes can save configuration and trace data to a memory card or thumb drive. Many models can download their data to a PC for storage and analysis. Many high-end scopes now incorporate non-traditional functions, such as the Fast Fourier Transform (FFT). This allows spectrum analysis or other advanced mathematical techniques to be applied to the displayed waveform.

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.

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 displayed. 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 highest frequency component of the input signal. (See the **DSP and SDR Fundamentals** 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, **Figure 25.17** 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, called the *Nyquist frequency*, the reconstructed signal's apparent frequency will be wrong; this is referred to as *aliasing*. In Figure 25.17 you can see that there is about one sample taken per cycle of the signal's ringing frequency. This does not meet the 2:1 rule established above. The result is that the scope displays an incomplete waveform or one with a different apparent frequency.

A simple 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.

25.6.4 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 **Figure 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 Figure 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 or to insure that the counter is not displaying the frequency of a harmonic or other unintended signal.

LIMITATIONS OF OSCILLOSCOPES

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 a scope's 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 **Figure 25.19**.

For an analog scope, assuming the frequency response is primarily limited by a 1- or 2-pole filter 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.

For digital scopes, substitute 0.45 for 0.35 in the equation because digital scopes have a much steeper roll-off above the specified cut-off frequency. The steeper roll-off is used to reduce aliasing by eliminating any signal above the Nyquist frequency as discussed in the previous sections.

25.6.5 Oscilloscope Probes

Oscilloscopes are usually connected to a circuit under test with a short length of shielded cable and a probe. At low frequencies, a piece of small-diameter coax cable and some sort of insulated test probe might do. 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.

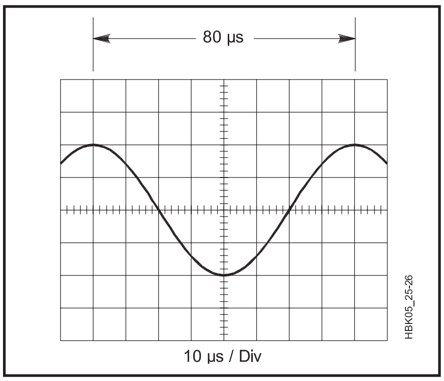


Figure 25.18 — An oscilloscope can measure frequency as well as amplitude. Here the waveform shown has a period of 80 microseconds (8 divisions × 10 µs per division) and therefore a frequency of 1/80 µs or 12.5 kHz.

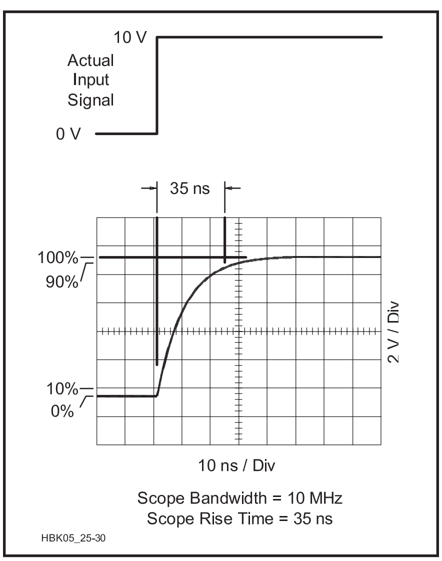


Figure 25.19 — The bandwidth of the oscillo-scope vertical channel limits the rise time of the signals displayed on the scope.

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 was 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. (A good overview of oscilloscope probes, how they are designed and intended to be used, along with links to descriptions of different types of probes is published by Tektronix at **www.tek.com/en/blog/what-is-an-oscilloscope-probe**.)

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 shown in **Figure 25.20**, 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. **Figure 25.21** 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.

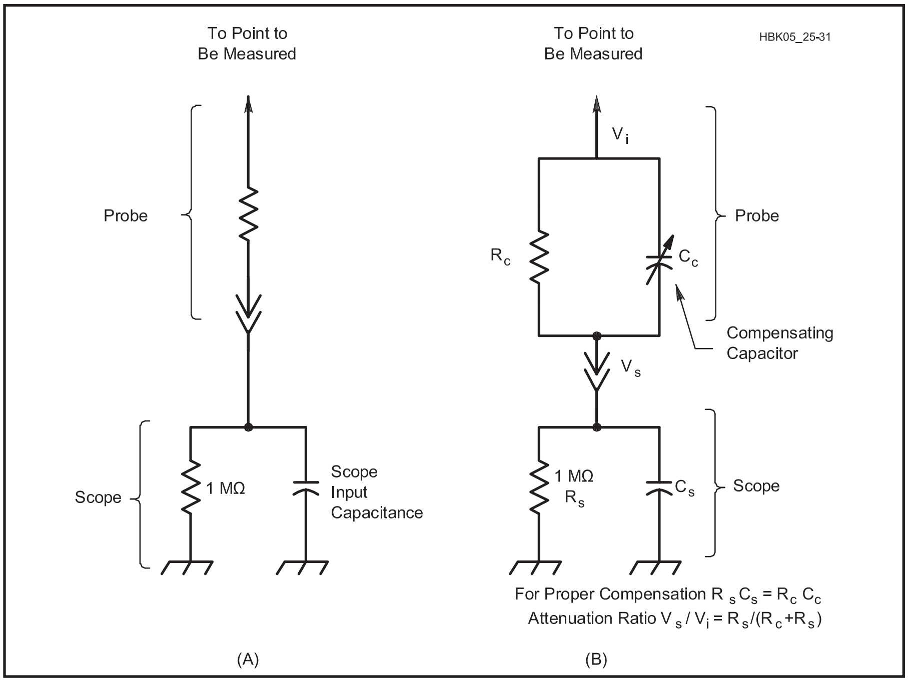


Figure 25.20 — Uncompensated probes such as the one at A are sufficient for low-frequency and slow-rise-time measurements. However, for accurate display of fast rise times with high-frequency components the compensated probe at B must be used. The variable capacitor is adjusted for proper compensation (see text for details).

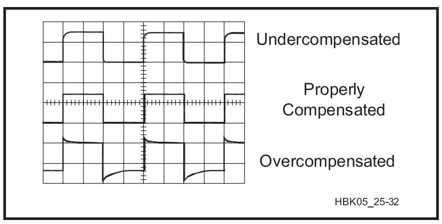


Figure 25.21 — Displays of a square-wave input illustrating undercompensated, properly compensated and

25.6.6. Choosing an Oscilloscope

FEATURES

When choosing an oscilloscope, the first decision is analog versus digital. Used and surplus instruments are usually analog but digital scopes are now widely available.

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. 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.

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. (See the Comparators section of the **Circuits and Components** chapter for more information about hysteresis.)

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 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 and high-end analog scopes are often quite affordable. Digital scopes are now showing up on the surplus market at reasonable prices. Early digital models had somewhat cumbersome user interfaces, so look for one with an analog-like feel with separate buttons or knobs for most of the important functions. One an older analog scope be sure the CRT trace is bright and there are no burned-in spots or lines on the CRT from extended use.

If you buy an analog 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.

25.7 Spectrum Analyzers

25.7.1 Time and Frequency Domain

A spectrum analyzer is similar in appearance 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, 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.

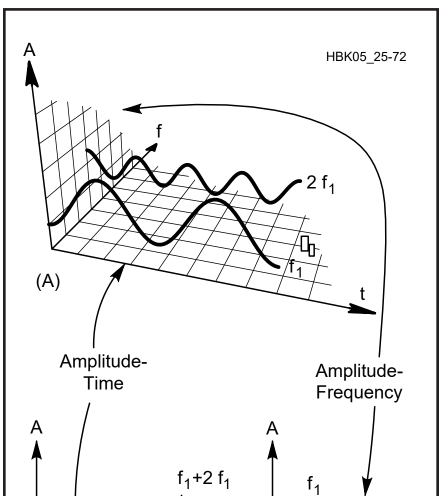


Figure 25.22 — A complex signal in the time and frequency domains. A is a three-dimensional display of amplitude, time and frequency. B is an oscilloscope display of time vs amplitude. C is spectrum analyzer display of the frequency domain and shows frequency vs amplitude.

To better understand the concepts of time and frequency domain, see **Figure 25.22**. The three-dimensional coordinates in Figure 25.22A 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 Figure 25.22B, 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.

25.7.2 Spectrum Analyzer Functions

As with oscilloscopes, analog spectrum analyzers have been replaced by digital models that use DSP/SDR techniques to perform similar signal processing and display functions. Just as for oscilloscopes, most of the operation and display terminology used today was defined for analog instruments. We will describe analog spectrum analyzer functions in order to explain the controls and functions used on digital analyzers today.

ANALOG SPECTRUM ANALYZERS

The most common analog spectrum analyzer is basically 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 **Figure 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 analog spectrum analyzers use an up-converting technique in which a wideband 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.

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 linear display, as for an oscilloscope. 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.

25.7.3 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

SDR Receivers as Spectrum Analyzers

SDR receivers can also be used as spectrum analyzers, although they do not have the same specifications as a spectrum analyzer designed for use as a test instrument. Even inexpensive SDRs can make absolute signal level measurements although the difference in levels between signals is probably more accurate. They can measure frequency and frequency response, as well. Portable SDRs can serve as portable spectrum analyzers when searching for interference sources— see the **RFI and EMC** chapter for more information about this application.

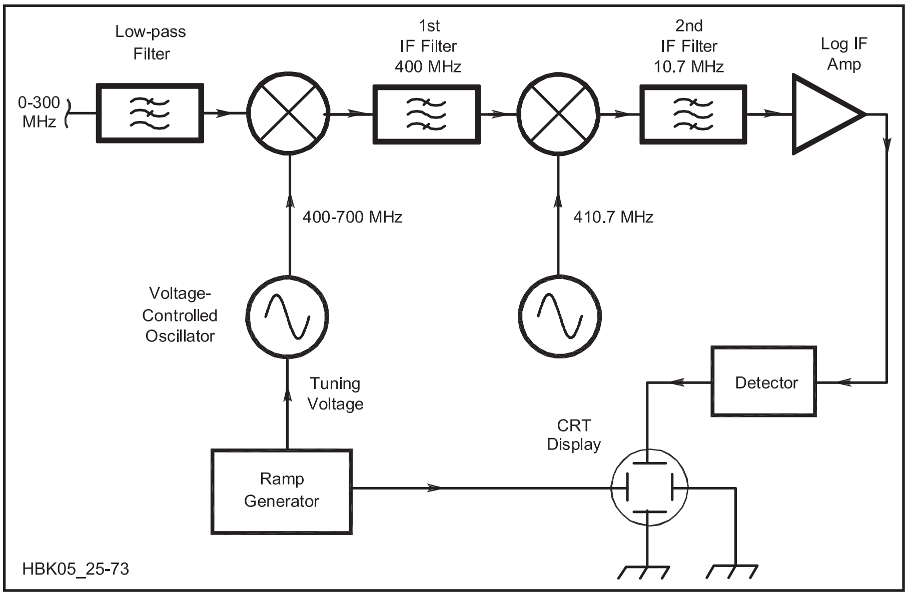


Figure 25.23 — A block diagram of a typical superheterodyne spectrum analyzer. Input frequencies of up to 300 MHz are up-converted by the local oscillator and mixer to a fixed frequency of 400 MHz.

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 **Receiving** 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 Figure 25.23, 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 necessary 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.

25.7.4 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 with the selected resolution bandwidth. 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 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 that uses 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.

25.8 Impedance, Antenna, and Network Analyzers

Another type of analyzer measures parameters of components, antennas, and circuits over a frequency range. These analyzers are often used to measure impedance and are sometimes referred to as *impedance analyzers*. Analyzers optimized for use with antenna systems are called *antenna analyzers*. More capable analyzers are used to measure gain, filter response, isolation, and other properties of circuits and systems that are referred to by the general name of *networks*. These analyzers are called *network analyzers*.

Over the last decade, a new generation of relatively low-cost analyzers has emerged, both as free-standing handheld instruments, and as instruments that interface to a computer via a USB port and require software running on the computer to complete the measurement system. Some of the free-standing instruments can operate under computer control and with the computer software greatly expanding their capabilities.

These instruments make measurements of the *scattering parameters (S-parameters)* of a single-port or two-port network. They allow any circuit or component to be treated as a "black box" with the S-parameters describing its behavior in terms of amplitude and phase.

The scattering parameters for a two-port network, are:

S_{11}, the reflection coefficient of the input port;

S_{21}, the voltage gain from input to output, also known as Forward Transmission or Forward Gain;

S_{12}, the voltage gain from output to input, also known as Reverse Transmission;

S_{22}, the reflection coefficient of the output port.

Scattering parameters are explained in the **RF Techniques** chapter's Two-Port Networks section.

25.8.1 Network Analyzer Basics

Connections to the analyzer are made at a *port* which is a pair of terminals that serve as an input or output of a network. The port terminals can be anything from alligator clips to a coaxial connector. A network can have more than one port. A dummy load is a *single-port network*, for example, and an amplifier or filter or transmission line are examples of a *two-port network* with an input and output. See the **RF Techniques** chapter section on Two-Port Networks for more information about networks and ports.

A *scalar analyzer* measures only the magnitude of a parameter (a scalar is a value expressed as a single number). Both oscilloscopes and spectrum analyzers are scalar instruments. Scalar parameters include SWR, attenuation, or power. A *vector analyzer* measures the magnitude of two parameters (typically voltages) and the phase difference between them. That combination creates a vector with a magnitude and angle. (See the **Radio Mathematics** online supplement for more information about vectors, scalars, and phasors.) The measurements are repeated over a range of frequencies.

25.8.2 Antenna Analyzers

Because measurements of antenna systems, including feed lines, are so common, a class of analyzers has been designed with features specifically for that task. The popular name for these instruments is *antenna analyzers*, whether they use analog or digital circuitry.

In many products available today, the measurements and graphs are displayed on a built-in screen about the size of those on mobile phones. The real-time graphing functions are especially helpful when making tuning adjustments in the field. Data is stored internally and can be downloaded to an external PC for processing and storage. (See the sidebar on Touchstone data files later in this section.)

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. Several basic designs are available with tradeoffs between performance and cost.

Be aware that some inexpensive or older units measure only the SWR or impedance magnitude while others measure both the resistive and reactive parts of the impedance. Some of these units may give the magnitude of the reactance but not the sign, requiring the operator to change frequency a small amount and watch the change in impedance magnitude to determine the sign and thus the type of the impedance, inductive or capacitive.

These analyzers have many more uses than impedance and SWR measurements. Paul Wade, W1GHZ, has contributed the paper "Antenna Analyzer Pet Tricks" in this book's online information. It includes how-to guides for a number of useful antenna system measurements.

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 wideband detector responding to the incoming signal from the AM station. Some manufacturers offer external high-pass broadcast-reject filters to allow the analyzers to be used in the presence of these strong signals. The filter can affect measurements near the broadcast band, particularly in the 3.5 MHz and lower-frequency bands. Check with the manufacturer about the limitations of using filters with the analyzer.

Determining Reactance Type

Some antenna analyzers can show the positive or negative sign of the reactance. Less expensive units may only show a reactance magnitude. To determine whether reactance is inductive or capacitive on a unit that doesn't indicate the sign, make a slight increase in frequency. If the reactance increases, it is inductive, and if reactance decreases, it is capacitive. This is a helpful trick when adjusting antennas with a portable analyzer that doesn't indicate reactance sign.

Protecting Antenna Analyzers

Portable antenna analyzers are sensitive instruments with low-power components at their input. They can be damaged by static electricity if care is not exercised. Antennas can collect a significant static charge from rain or wind. Be sure to momentarily ground the transmission line before connecting it to an analyzer to reduce the risk of damage to the sensitive components on the input.

Static buildup can be minimized by providing a dc path between center conductor and the grounded shield for the antenna being measured, offering some protection both for the analyzer and for station equipment. The dc path can be a shunt inductor (typically an RF choke) or a high-value resistor (10 kΩ or more). If the resistor is to be a permanent part of the antenna, it must have a power rating suitable for the transmitter power.

Antennas with non-insulated elements mounted directly on a grounded boom or support are already protected against static. Insulated elements need protection against static. Measure resistance between the feed line center conductor and shield to be sure of whether there is a dc path to drain static.

Using the analyzer when another transmitter is active can also cause damage if enough signal is picked up by the antenna under test. Some analyzers include an isolation relay that protects the input when a measurement is not in progress.

SCALAR ANTENNA ANALYZERS

The simplest way to measure the frequency response of a coax-fed antenna to show its resonant frequency or frequency of minimum SWR is to use a standing wave ratio (SWR) meter. By measuring SWR, the meter shows how well-matched the load or the antenna's feed point impedance is to the characteristic impedance of the feed line. The frequency of the measurement is controlled by changing the frequency of the transmitter. Because SWR is a single numeric value, the combination of a transmitter and SWR meter creates a simple scalar antenna analyzer.

SWR meters are usually calibrated for 50 Ω since most transmitters are designed to drive a 50-Ω load. (Some can be switched to operate in a 75-Ω system.) An SWR meter by itself, however, cannot provide any information about the values of resistance and reactance that make up the impedance.

The first self-contained antenna analyzers replaced the transmitter with a manually tuned low-power signal source and added a detector and analog meter in a single piece of portable equipment. These simple scalar analyzers display SWR and/or impedance magnitude, but not separate values of resistance and reactance. While they have been made obsolete by the more advanced equipment available today, they are still quite suitable for simple adjustments and troubleshooting antennas.

Today's scalar antenna analyzers add more signal processing capabilities to measure both resistance and reactance, along with SWR. Some units have a manually tuned signal source while others make swept frequency measurements under the control of a microprocessor. **Figure 25.24** shows the block diagram of a typical scalar antenna analyzer.

Figure 25.25 shows a relatively simple circuit that uses diodes for detecting voltages corresponding to voltage and current at the external load. This inexpensive circuit is useful but at low signal levels the diodes introduce some non-linearity and temperature drift which may be an issue. This type of diode detector responds to signals over a wide frequency range and there may be stray pickup from nearby broadcasting stations that makes the measurement results inaccurate. For a more detailed analysis of the antenna system, an instrument with a narrowband detector such as for a radio receiver gives much better performance than the broadband diode detector.

The rectified voltages can be digitized by a microprocessor and the results displayed numerically. The signal source is typically a varactor-tuned LC oscillator in the analog units or a direct digital synthesizer (DDS) in more sophisticated models. The DDS signal source is very stable since it is controlled by a crystal oscillator and it can be set to the desired frequency quickly with a keyboard entry. The DDS version costs somewhat more but compared to the varactor-tuned oscillator, it has significant performance and operating conveniences. (See the **Oscillators and Synthesizers** chapter for information on DDS signal sources.)

VECTOR ANTENNA ANALYZERS

Today, the most common antenna analyzer is a *vector impedance analyzer* (VIA) that measures both the magnitude and the phase angle of impedance at a single-port across a range of frequencies. The impedance is then displayed numerically (R + *j*X), as a phasor (|Z|∠θ), or some computed value, such as SWR. (Magnitude is denoted by vertical bars, | |, and the angle by ∠.) The measurements can also be displayed on a graph and many instruments can transfer the measurements to a computer for further processing and storage. A VIA 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.

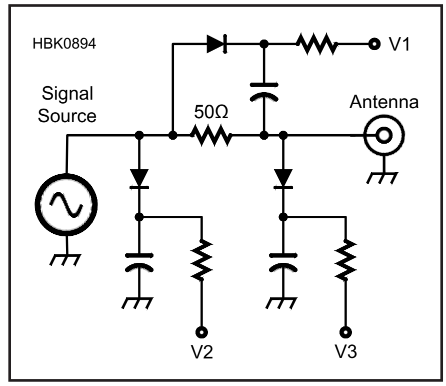


Figure 25.25 — A relatively simple diode detector circuit for measuring impedance that gives adequate results for many applications.

An older analog instrument called a *vector impedance meter* made a single-frequency measurement of complex impedance and displayed resistance and reactance on a pair of analog meters. The HP4800 series of instruments was a typical example.

The impedance data can be used to calculate several parameters for the antenna system. Using the specified value for the system reference impedance, which can be any value (it doesn't have to be 50 Ω), For example, the reflection coefficient (ρ or rho) can be calculated as:

$$\rho = (Z_L - Z_0) / (Z_L + Z_0)$$

where Z_L is the measured impedance of the load and Z_0 is the specified impedance of the transmission line, which can be any value. Z_L is a complex number; therefore, ρ is, in general, a complex number with a magnitude between zero and one. Rho is approximately equal to zero when the line is matched to the antenna because there is no reflection, all the transmitter power is absorbed by the antenna. When the antenna is poorly matched to the line, ρ is larger and it approaches 1 when the mismatch is large.

$$SWR = (1 + |\rho|) / (1 - |\rho|)$$

Note that SWR only depends on the magnitude of ρ denoted by the vertical bars as |ρ| so it is not a complex number.

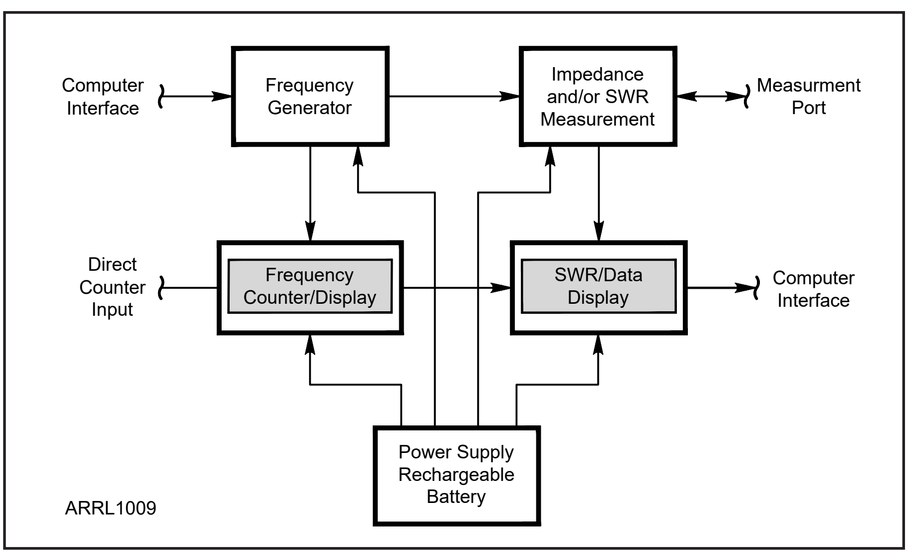


Figure 25.24 — Block diagram showing the elements of a scalar antenna analyzer.

25.8.3 Network Analyzers

A *scalar network analyzer* (SNA) is an instrument that can measure, as a function of frequency, the magnitude of the gain or loss

Touchstone Data Files

The most useful analyzers can store sets of measurements in a file for review later and transfer to a PC for further processing and display. The standard format for swept impedance data (frequency and impedance) is called *Touchstone* (**en.wikipedia.org/wiki/Touchstone_file**). Files in the Touchstone format can be read and processed by a wide variety of software, including as an input to design or simulation software.

Touchstone files are plain text with a line for each data point. A single-line header defines the sub-format of the data. The sub-format most widely supported by hardware and software for impedance data is s1p in which the parameter is complex S11, and each line consists of Frequency, Real S11, and Imaginary S11, with spaces as the separator.

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. (Because this is a scalar analyzer, only magnitude is measured.) In **Figure 25.26A**, 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* or include RF switches to do that automatically.

The box in the figure 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 in Figure 25.26B shows an example of a typical lab instrument VNA. The VNA's signal source has a 50-Ω impedance that drives the network port and detector circuits.

The generator can drive port 1 as the input and measure port 2 as the output or the generator output switch can reverse the signal flow. The first configuration (port 1 as input) allows the VNA to measure S_{11} and S_{21}. The signal flow is reversed to measure S_{12} and S_{22}. The splitters also supply the signal generator output to the detector circuits.

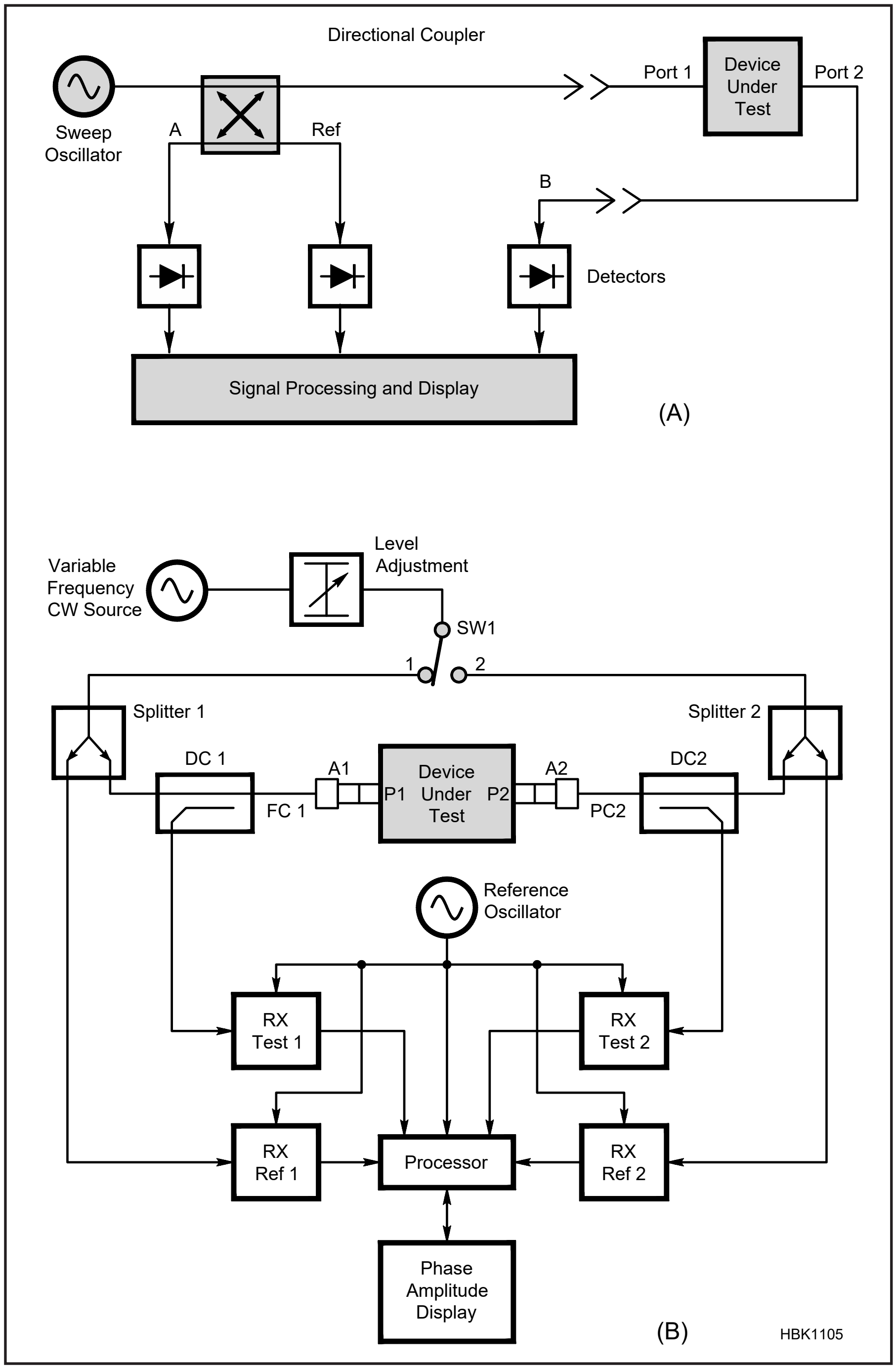


Figure 25.26 — Block diagram showing the elements of a typical scalar (A) and vector (B) antenna analyzers. Figure 25.26B – Created by Chris Angove from original research and study - From English Wikipedia, CC BY-SA 3.0, commons.wikimedia.org/w/index.php?curid=1703095

The detectors mix the reference signal (REF) with each directional coupler output and the input signal. The mixer/detector outputs and the reference signal are all fed to a processor circuit that measures the magnitude and phase of each signal. The measured data are then converted to S parameters as explained previously. From that data, other parameters such as impedance can be calculated.

Usually, the VNA or its host software provides the capability to plot the S parameters on a Smith chart for easy evaluation. S_{11} and S_{21} are often available in a magnitude versus frequency graph, as well.

A *vector impedance analyzer* (VIA) has similar construction to a VNA but only in-

Used Network Analyzers

Used lab-quality network analyzers, both scalar and vector, are readily available on the surplus market. Some require an external sweep oscillator while others include an internal signal source. The tests sets may be internal or external as well. These instruments are generally well-designed and offer excellent quality measurements but are typically bulky and difficult to repair, if needed. The high quality of analyzers available at low prices today makes the older units much less desirable unless needed for a laboratory setup or to replace a specific instrument.

Choosing A Computer Controlled Analyzer

The most capable and useful analyzers available at moderate cost are controlled by a computer. They convert the measurement to baseband audio, which is fed to the computer via an integrated USB interface. Software on the computer controls the analyzer, performs computations to turn the audio into a measured result, and plots the results. These are things to look for:

• It should be a precision instrument, solidly built and with a high-quality bridge, so that it can be reliably calibrated.

• A frequency range of 100 kHz to 600 MHz is a good starting point; 10 kHz to 1.5 GHz is even better.

• The dynamic range of the instrument, including the USB interface, places an upper limit on the range of impedance values (or gain/loss response) that can be displayed in a single measurement. A 12-bit interface (about 88 dB dynamic range) is sufficient for most amateur measurements with a 16-bit interface (about 90 dB dynamic range) desirable for precision work.

• It should be able to make and display measurements at both linear and logarithmic scales; linear sweeps are best for narrow sweeps (a ham band); log sweeps are important for wide sweeps (showing a filter's out-of-band rejection or a choke's impedance and over a wide range).

• An analyzer that can be powered from the USB port simplifies setup both in the field and in the station.

The analyzer's software should include:

• The capability to store and recall calibration data for multiple measurement fixtures or configurations.

• Support the use of correction factors to account for various test fixtures and calibration standards results in greater accuracy for these special setups.

• Perform Time Delay Reflectometry (TDR) from a wide-frequency sweep of a transmission line's S11 (Return Loss). The TDR function should find the line's electrical length and any discontinuities (splices or faults). Cursors should show position in feet or meters with multiple windowing functions for the most useful view of the response.

cludes the 50-Ω generator and bridge circuitry to measure S_{11}, so it can make only single-port measurements such as impedance, SWR, etc. From S_{11}, the analyzer can compute the complex impedance (resistance separately from reactance, including the sign of the reactance to tell us whether it's inductive or capacitive). Most antenna analyzers are VIAs. A two-port VNA can also be configured to do single-port swept frequency measurements.

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.

DUAL-DDS VNA DESIGN

Figure 25.27 shows an alternative VNA design that uses two DDS signal sources with one applied to the antenna system and the other used as a reference. Typically, the clock oscillator is crystal-controlled for high accuracy. One DDS is programmed to output a signal at the actual test frequency and the other DDS is programmed to a slightly higher frequency in the 1 to 10 kHz range, shown in the figure as a 2 kHz offset. See the reference entries for articles by Bob Clunn, W5BIG, Thomas Baier, DG8SAQ, and Michael Knitter, DG5MK, giving a detailed design description of this type of analyzer.

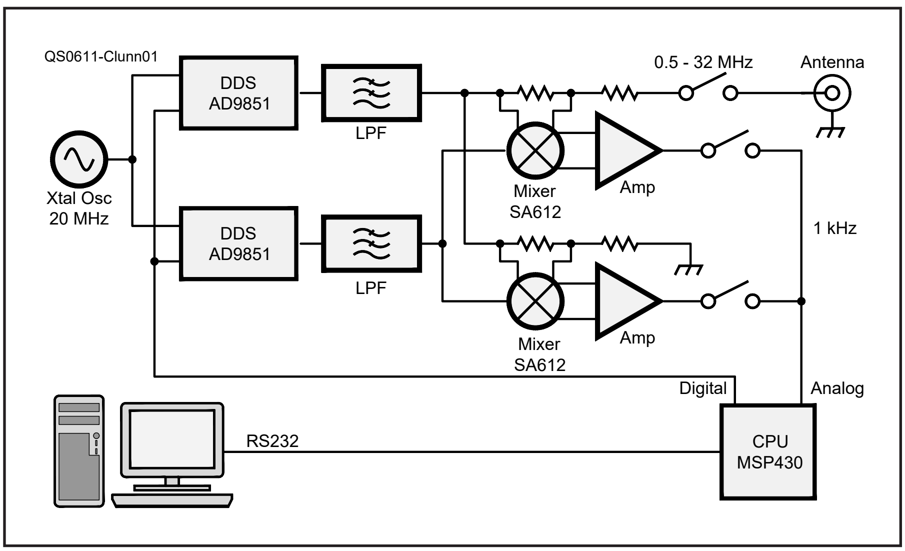


Figure 25.27 — Block diagram of a dual-DDS-based antenna analyzer. By mixing two signal sources that are very close in frequency, with one of the sources applied to the load (antenna), low-frequency signals are generated. The signals contain the necessary information to measure impedance magnitude and phase.

The signals are then mixed to produce a low frequency output that can easily digitized by an inexpensive analog to digital converter (ADC). The measurements are then processed mathematically in a microprocessor or PC to yield full information about the complex impedance being measured. This type of analyzer measures both the magnitude and phase of a test impedance over a wide range of frequencies.

One signal corresponds to the voltage applied to the test port (V1) and the other signal voltage corresponds to the current flowing in the test port (V2). The ratio of these signals (V1 / V2) is the impedance at the port and the difference of their phases is the phase angle of the impedance:

Magnitude (Z) = Magnitude (V1) / Magnitude (V2)

Phase (Z) = Phase (V1) – Phase (V2)

The frequency is swept in small steps across the specified range and a measurement of V1 and V2 is made at each frequency.

25.8.4 Calibration and Measurement Plane

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 table of correction factors that are applied to the raw measurements. The mathematics involved is quite complex, but it is all handled by the instrument software and is invisible to the user. A thorough discussion of VNA error-correction techniques can be found in Agilent Technologies application note 5965-7709. (**www.keysight.com**)

The point at which the measurements are to be performed is called the *measurement plane* or *reference plane*. That point can be at the measurement connector of a VNA, at a test fixture connected to the analyzer, or at the ends of cables. (The end of the cable can even be a long distance from the analyzer at the top of a tower!) Calibration consists of taking measurements of known impedances. The analyzer makes the measurements across the desired frequency range and then applies mathematical corrections at each frequency so that the net result is the exact known values. This set of measurements is saved as a *calibration file* that can be stored and retrieved whenever that measurement plane is used in the future. Most analyzers can save multiple configuration files for different measurements. This process is very quick and easy and the final results of any measurement are much more accurate than from a simple analyzer that does not have a calibration procedure.

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 measurement plane should always be at the end of the cables that are to be used for the test.

Calibration includes the effects of length for any coax between the analyzer and the measurement plane; substituting a different piece of coax requires another calibration. Adding a connector adapter to connect an antenna to the plane that has been calibrated introduces an error that is small enough not to matter for most antenna measurements, but can be significant on the higher HF bands, and much more so at VHF and UHF.

Calibration for S_{21} and S_{12} adds additional calibration measurements with the analyzer output and inputs connected together with cables of the same length (preferably the same cables) for connecting the VNA output to the Device Under Test (DUT) and the output of the DUT to the VNA input, with the two cables mated through a double female adapter (aka a barrel adaptor).

For antenna measurements, the measurement plane might have to move when one or more connector adapters are attached to the same cable. Since the analyzer is then connected at a different spot on the line and has different geometry, it needs its own calibration.

SOLT CALIBRATION

The most popular technique is called a Short-Open-Load-Through (SOLT) calibration because it uses four calibration standards, a short-circuit (0 Ω), an open-circuit (∞ Ω), a 50-Ω load, and a through (direct) connection between the two ports. If the analyzer is a single-port VIA, there is no through connection and only the SOL calibration is performed. A set of standards with those three impedances over a very wide frequency range are provided with the analyzer or available as a calibration test kit.

Figure 25.28 — A simple vector impedance analyzer (VIA) test fixture for measuring component impedance into the VHF range. The fixture is a BNC-to-banana plug adaptor with alligator clips on the plugs. Below the clips are the short-circuit calibration load and a non-inductive 50-Ω resistor to perform the SOL calibration. (Photo courtesy of Jim Brown, K9YC)

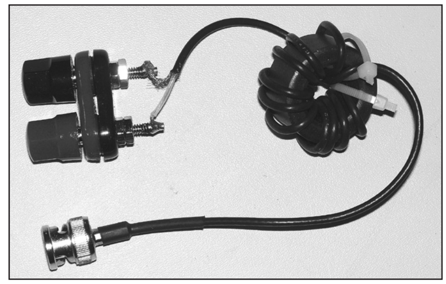


Figure 25.29 — A common-mode choke near the test fixture should be used when making measurements through a length of coaxial cable or measuring two-wire transmission line impedances. This choke is made of 12 turns of RG-193 miniature coax through a 2.4-inch OD, Type 31 core (Fair-Rite 2631803802) and is effective in the upper HF range. The best choke should present at least 5,000 Ω of resistive impedance at the measurement frequency. See the choke section of the Transmission Lines chapter for more information about this type of choke.

Figure 25.28 shows a simple component test fixture consisting of a BNC-to-banana plug adaptor with alligator clips slipped on to the plugs. Under the clips is shown the Short test load (#10 bare copper wire folded back on itself and coated with solder to minimize both resistance and inductance) and a large-format 1% 49.9-Ω chip resistor with short leads soldered to it of the minimum length that allows them to attach to the clips. When calibrated, this fixture can make reliable measurements well into the VHF range. Calibration accuracy is then verified to the highest frequency of interest by measuring high value chip resistors (5–10 kΩ), looking primarily for the effects of parallel capacitance or series inductance. For this test fixture, the measurement plane is at the tips of the alligator clips. The simple fixture like that in the figure is typically used at the end of a convenient length of coaxial cable (RG-400 is a good choice) for S_{11} measurement of chokes and other components.

Most antenna analyzers have a single-ended (unbalanced) output with a coax connector. For making measurements on a two-wire transmission line, such as window line or ladder line, a common-mode choke can be added between the analyzer's RF connector and the input to the transmission line as in **Figure 25.29**. The calibration loads are attached to the output side of this balun.

This calibration procedure can be extended to allow measuring the actual driving point impedance of the antenna. The transmission line is disconnected at the antenna and the calibration loads are attached at the far end instead of the antenna. This shifts the measurement plane to the antenna itself which is handy when designing matching networks.

25.8.5 Analyzer Measurements

Firmware built into vector analyzers, and/or software running on an associated computer, can display impedances computed from S_{11} in many forms. Some of the more common are: 1) Smith Chart; 2) Series Resistance (R_S) and Reactance (X_S); 3) Parallel Resistance (R_P) and Reactance (X_P); the magnitude of the Impedance (Z_{MAG} or $|Z|$) 4) Inductance (L) and parallel capacitance (C_P); 5) Capacitance (C) and series resistance (R_S). Depending on firmware and/or software, the frequency sweep can be made and data presented linearly or logarithmically. (See the section Measuring Components With an Analyzer later in this chapter.)

Using built-in firmware or associated computer software, most modern analyzers can perform *time delay reflectometry* (TDR) to determine the length of a transmission line, to look for splices or faults along the line, even find the impedance of each of multiple sections connected in series. (See the discussion of TDR later in this chapter)

Most modern vector analyzers and/or associated computer software can export a measurement for use by design and analysis software like *SimSmith* (**www.ae6ty.com/smith_charts.html**) and *ZPlots* (**ac6la.com/zplots1.html**) to study feed line loss and design matching networks. (See the sidebar on Touchstone format) These two freeware packages can transform an impedance measurement made anywhere along a feed line to the impedance at the antenna feed point if the length and parameters of the feed line are known. *ZPlots* also offers some design capability for common amateur applications. *SimSmith* allows the design of far more complex two-port networks and transmission line systems.

25.9 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 nearby 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 *PathSim*, a free software HF channel simulator. (**moetronix.com/ae4jy**) A hardware HF channel simulator has been described by Johann B Forrer, KC7WW. (See the Reference for Forrer.)

25.10 Software-Based Instruments

Most amateurs these days own a personal computer with a powerful microprocessor, lot of memory and mass data storage, a large color display, and a sound card with stereo high-bandwidth 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.

Several manufacturers offer hybrid RF instruments that combine a computer with a separate signal interface module to create oscilloscopes, spectrum analyzers, and other types of instruments. The connection between the module and the computer is usually a USB 2.0 or 3.0 interface for high speed.

Audio-frequency instruments can be implemented directly using the computer sound card, which typically has a frequency response from perhaps 50 Hz up to about 20 kHz. While computer sound cards can be quite useful for measurement, the low-cost sound cards built into computers may not have good noise performance. Laptops may not offer a line-level input, only the more sensitive microphone input which is more susceptible to overload. Sound card outputs may distort at levels near their rated output. It is good practice to keep their output level at least 6 dB below rated output to minimize distortion on signal peaks. Both input and output signal levels should also be set so that the lowest amplitude components are at least 10 dB above the noise floor.

Inexpensive standalone sound cards with a USB interface can offer better quality than built-in motherboard sound card interfaces. These work well in applications such as computer-to-radio interfaces for digital modes. Because these products are made for a mass market, models tend to be updated every year or two, but the quality brands such as Numark and Tascam tend to offer stable designs.

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.

One problem with using a sound card is that the low-level signals may be susceptible to picking up low frequency hum and buzz, noise, and radio-frequency interference (RFI). If the computer and the device under test are grounded separately to the ac power system, hum, buzz, and other noise can be generated from currents flowing in the ground connection between the two. Similarly, the necessary loops formed by the shields of multiple cables can pick up 50 or 60 Hz magnetic fields, generating a hum voltage. It is helpful to bond the sound card and PC to whatever device is being tested by using a short, low-resistance connection. 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 bonding help prevent hum and noise pickup.

In general, it is always good practice to bond together the chassis or shielding enclosure of every piece of equipment in a system with short, heavy copper wires (#14 AWG or larger). Failure to do so often results in hum, buzz, and RFI being introduced into the signals and data. For the same reasons, all interconnected equipment should be powered from the same ac outlet or from outlets that share the same 'green wire' (that is, they plug into the same multiple wall outlet box). If the equipment is powered from different outlets, the green wires of those outlet boxes should be bonded together. Good quality cables and attention to proper shielding help prevent hum, buzz, noise, and RFI. (See the discussions of bonding in the **Safe Practices** and **RFI and EMC** chapters, as well as the ARRL book *Grounding and Bonding for the Radio Amateur*.)

25.11 RF and Microwave Test Accessories

This 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 common in the video and television industry.

25.11.1 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 **Figure 25.30**, 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 low SWR over a wide band of frequencies. (This is often specified as *return loss* (RL) which is similar to SWR as described in the **RF Techniques** chapter.) A *feed-through termination*, such as the one in **Figure 25.31**, 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.

25.11.2 Connectors and Adapters

Performing accurate and repeatable tests and measurements demands good quality in every aspect of the test process. This includes coaxial cables and connectors. Test cables should use very high-quality coaxial cable, such as RG-400 or the equivalent, particularly if measurements will be taken above the HF range. To make really accurate measurements of signal levels, the loss of the coaxial cables used in the measurement needs to be accounted for, as well.

Coaxial connectors and adaptors can be a source of problems if poorly made or worn. It is worth the extra expense to obtain a few high-quality connectors and adaptors for making measurements. The manufacturer should be clearly identified on the body of the connector. Silver-plating is desirable and may be required for measurements at extremely low signal levels on in the UHF and microwave range.

Cables, connectors, and adapters can be tested by using an antenna or network analyzer. With the device being tested connected to the analyzer, move the connector around and bend cables, paying particular attention to where the connector is attached to the cable. Watch the analyzer display for any intermittent changes in behavior that indicate a possible poor or loose connection. Check the response of adaptors at the highest frequencies for which they are expected to be used.

25.11.3 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.) The noise source can be combined with a bridge circuit, creating a *noise bridge* which is described in this chapter's section on RF Impedance Measurement. The wideband noise from a noise source can also be used as a test signal to measure frequency response of a filter or amplifier with a spectrum analyzer or tunable receiver.

25.11.4 Dip Meters

This instrument was originally a grid-dip oscillator (GDO), so-called because the indicating "dip" was in the grid current of its vacuum-tube oscillator. Most dip meters today are solid-state but the principle is the same: If you hold the external oscillator coil near a tuned circuit and adjust the dip meter tuning dial to the frequency of the tuned cir-

RF Measurement Test Set

Many low-level RF measurements require similar test setups so why not combine them into one package? This also makes it easier to perform the tests and get consistent results. NOADL and KE4PT show an example in the *QEX* article "Apparatus for RF Measurements" that you can download from the online information.

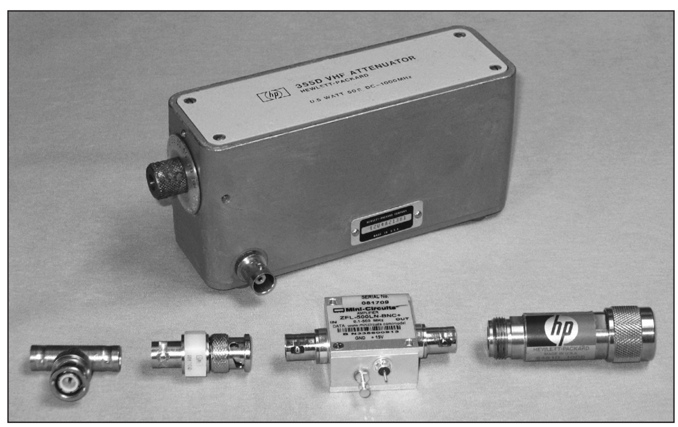


Figure 25.31 — Various RF test accessories. At the rear is a 10 dB step attenuator with a 0-120 dB range. In the foreground from left to right are a BNC "T" connector, a 50-Ω feed-through termination, a 0.1-500 MHz low-noise amplifier, and a 10-dB fixed attenuator.

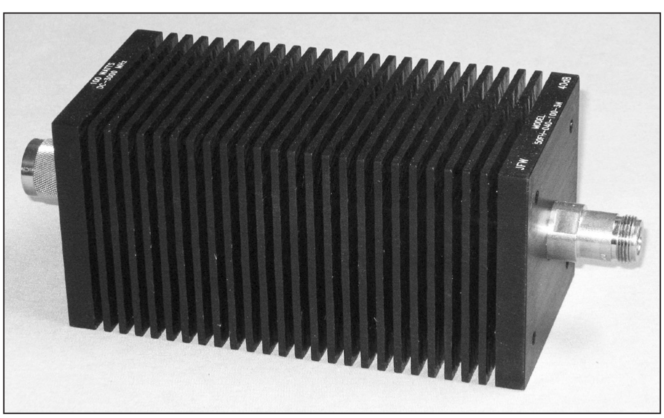


Figure 25.30 — A 100-watt power attenuator rated for dc-3000 MHz. The transmitter is connected to the male type-N connector. The female connector is for the low-power connection.

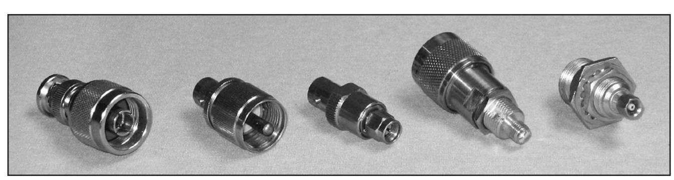


Figure 25.32 — Coaxial connector adapters. From the left are: BNC male to type-N male, BNC female to UHF male, BNC female to SMA male, type-N male to SMA female, and type-N female to SMC male.

cuit, there is a dip in the meter reading as the resonant circuits interact with each other. The coil extends from the end of the instrument and a set of plug-in coils is provided to cover the unit's frequency range.

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. To avoid detuning the circuit being tested, always use the minimum coupling that yields a noticeable indication.

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, 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 above and below the minimum. 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.

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. 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.

25.11.5 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 Keysight 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 and have flanges compatible with your equipment. See the WR-series of waveguide sizes in the **Transmission Lines** chapter.

Adapters such as those shown in **Figure 25.32** are useful for making connections among various pieces of test equipment.

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-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

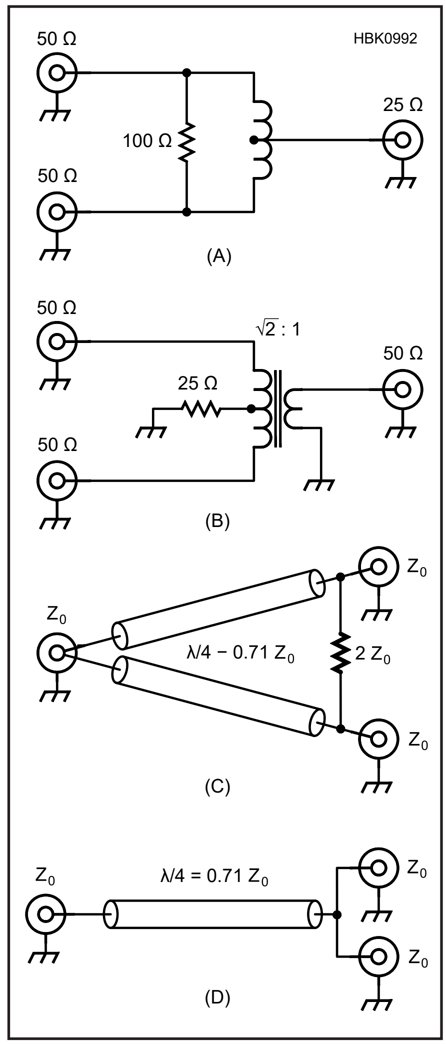


Figure 25.33 — Four forms of the Wilkinson power combiner/splitter. A and B show a lumped-element implementation. The version at C uses a pair of λ/4 transmission lines acting as synchronous transformers. If all three ports are terminated in Z_0, the output ports are isolated from each other. If isolation between the output ports is not required, the divider at D can be used.

comes out each port and half the power is absorbed in the resistors.

A *hybrid combiner*, such as the *Wilkinson combiner* shown in **Figure 25.33**, 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 Figure 25.33B uses a transformer with a turns ratio (2:1 impedance ratio) to obtain 50 Ω on all ports.

The Wilkinson combiner can also be used as a power divider. The transmission line version in Figure 25.33C replaces the transformer of Figure 25.33A with a pair of 1/4-wavelength transmission line transformers (see the **Transmission Lines** chapter) to divide the input signal between the pair of output ports. All three ports must have the same impedance, Z_0. A resistor between the output ports with a value of $2Z_0$ reduces loss and increases isolation. The ¼-wavelength sections must have a characteristic impedance of

$$Z_0 \times \frac{\sqrt{2}}{2} = 0.707\ Z_0$$

Since most amateur equipment and loads have an input of 50 Ω, using RG-59 or RG-11 with an impedance of 75 Ω will result in acceptable performance. If isolation between the output ports is not required and a 35-Ω transmission line can be constructed, the divider in 25.42D can be used. Two pieces of 70-Ω cable or in parallel can also be used. 75-Ω RG-11 can also be used if the resulting slight mismatch is acceptable.

25.12 Making Basic Measurements

25.12.1 Four-wire Measurements

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 **Figure 25.34**. 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.

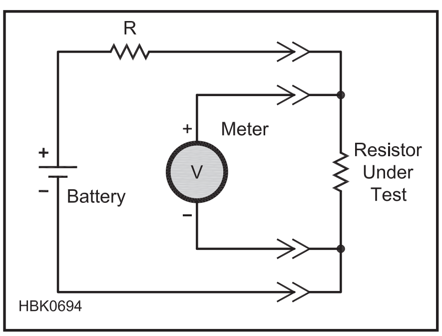


Figure 25.34 — The four-wire technique for measuring low-value resistors. By connecting the current source and the meter separately to the leads of the device under test, the error due to lead resistance is reduced.

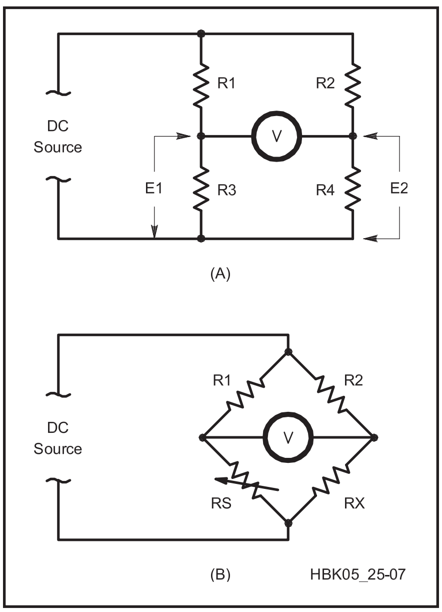


Figure 25.35 — A Wheatstone bridge circuit. A bridge circuit is actually a pair of voltage dividers (A). B shows how bridges are normally drawn.

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 Figure 25.2. The voltmeter 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 as shown in Figure 25.34.

25.12.2 Wheatstone Bridge

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 **Figure 25.35**, 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 Figure 25.35B 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

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

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

25.12.3 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. (Wire tables including resistance in ohms per foot (Ω/ft) are available in the **Antennas** 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 1/4 W resistors in series, each with a value of one-tenth the desired resistance, would be suitable.

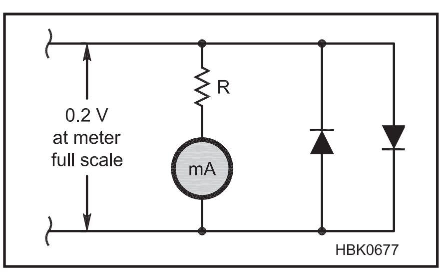


Figure 25.36 — Back-to-back silicon diodes protect the meter by limiting the maximum voltage. See the text for a discussion of how to select the value of R.

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 meter movements 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 **Figure 25.36** 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.

25.12.4 Using Multimeters for AC Measurements

RMS MEASUREMENTS

(See the **Radio Fundamentals** chapter for definitions of values used to measure ac waveforms.) 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.

LIMITATIONS TO ACCURACY

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, especially at frequencies near the upper limit of the multimeter's capabilities. 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.

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)

25.12.5 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. When very close to the station's frequency, watch the variation of the receiver's S-meter instead of relying on the receiver's audio response at very low frequencies. The popular program *WSJT-X* has a frequency calibration tool that used the sound care of a computer connected to the audio output of a receiver.

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. (See the References for Miller and Nash) 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. (See the Reference for Jones) Amateur-level kits are also available or you can build one from scratch. (See the Reference for Sherra)

25.13 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. A table for converting between voltage and power in 50-Ω systems is provided in the supplemental information for this chapter.

25.13.1 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. An RF probe can also be used to detect AM or SSB signals so the modulation can be measured on an ac multimeter. The circuit is typically quite simple, consisting of a diode, a resistor and a couple capacitors, as in the example of **Figure 25.37**. The resistor value is chosen so that the rectified output voltage is approximately

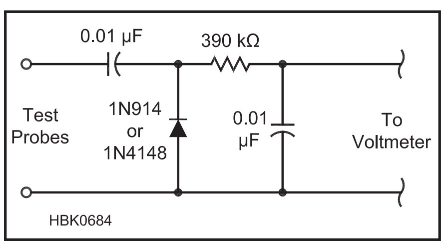


Figure 25.37 — A basic RF probe, used to convert an RF signal into a dc voltage that can be measured by a voltmeter.

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 germanium 1N34 or 1N277 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 with the downloadable supplemental content.

Measuring RF current is a little more difficult. 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. The impedance of the bulb will also be in series with the antenna or feed line, affecting the measurement to some degree.

Another alternative is to construct an RF ammeter using a current transformer that clamps over the wire. (See the Reference for Lau) 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

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

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.

25.13.2 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 (mW) but it is perhaps more common to see units of *dBm*, decibels with respect to one milliwatt. For example, itis 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.)

Most RF systems use 50 Ω as the impedance at various signal interfaces and for the characteristic impedance of most coaxial cables. Since this is so common, it is useful to be able to convert directly between voltage measurements and power, assuming a 50-Ω impedance. A table of voltage-power conversions for 50-Ω systems is provided in this chapter's supplemental information.

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 at the crest of the modulation envelope. (FCC §97.3(b)(6)) 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.

Using 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 $2 \times \sqrt{2}$ 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** online material 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. (See the Reference for Steinbaugh)

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 sensors are fragile and easily damaged. Obtaining replacements for the desired frequency range and power level can 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

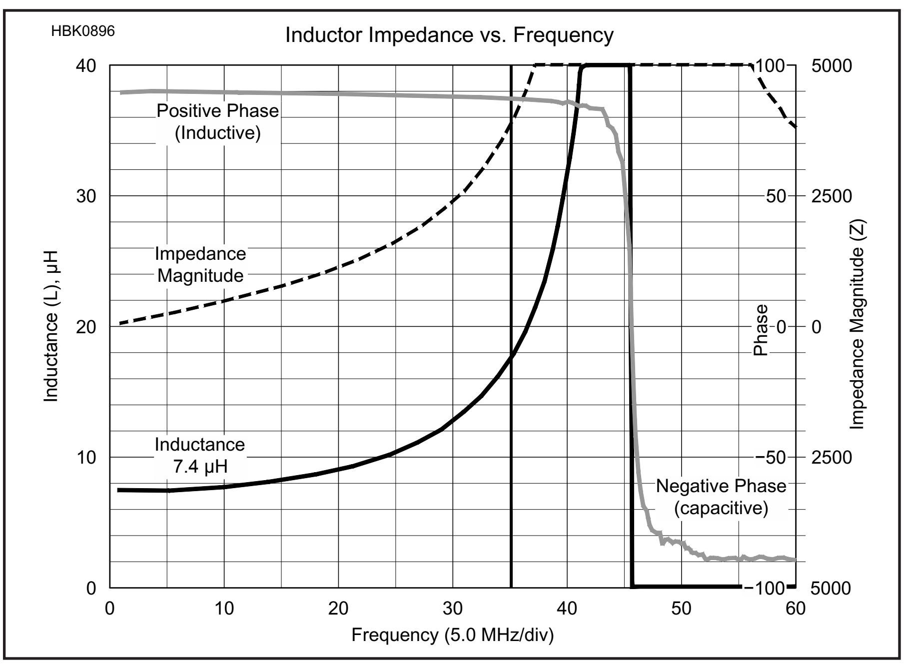


Figure 25.38 — Example of an antenna analyzer (AIM-4170 and companion software) being used to measure an air-core inductor's behavior. The inductor has a nominal value of 7.4 µH and a self-resonant frequency of 45.4 MHz.

dc to 500 MHz with 1-dB accuracy over an 88-dB (nearly 1 billion-to-one) power range.

25.13.3 Measuring Components With an Analyzer

Antenna and network analyzers can be used to measure components at RF and across a wide frequency range. For example, a swept-frequency test can show if an inductor is resonant within the frequency range where it will be used. The **RF Techniques** chapter's section on Effects of Parasitic Characteristics has more information about how components are likely to behave at high frequencies.

Figure 25.38 shows a graph generated from swept-frequency impedance data collected by an antenna analyzer of the type in Figure 25.27. The figure shows the impedance of an air-core inductor with a nominal inductance of 7.4 μH at low frequencies. (The traces are labeled with the measurement they represent.) The self-resonant frequency is 45.4 MHz. This resonance occurs because of the coil's inter-turn capacitance.

Far below the resonant frequency the inductance has a positive reactance and the coil presents its expected value, 7.4 μH. As the test frequency approaches the self-resonant frequency, the parasitic (stray) capacitance that causes self-resonance causes the total reactance to increase, so the inductance appears to be larger. At resonance, the impedance is a resistance of high value. Above the self-resonant frequency, the component increasingly looks like a capacitor, as indicated by the negative phase angle.

The analyzer is also very handy for determining the material of a toroid core. Cores of different ferrite or powdered iron mixes cannot be told apart by their physical appearance, but you can identify the material used by measuring their characteristics over a wide frequency range. A procedure for determining a ferrite component's mix is presented in the **RF Techniques** chapter section on Determining Ferrite Mix.

Capacitors are usually closer to the ideal component than inductors, but they do have some inductance in their leads and electrodes. This is called *equivalent series inductance* (ESL). Eventually they become self-resonant at some high frequency. Above the self-resonant frequency, a capacitor acts like an inductor. This self-resonance should be checked for capacitors that will be used in the VHF/UHF range.

Depending on the dielectric material used, capacitors also have some loss. This loss is most commonly described as *equivalent series resistance* (ESR) which appears in series with its capacitive reactance. The ESR in most capacitors increases with frequency and can be critical in power handling circuits.

Resistors have an effective capacitance in parallel with them as well as inductance in their leads so they are not ideal over a wide frequency range. Physically large power resistors used for dummy loads have larger parasitic components. (Thin-film power resistors in TO-220 packages are available with significantly lower reactance.) Tubular metal and carbon film resistors are often trimmed with a laser to create a spiral track in the deposited film, creating inductance. If the resistor is to be used in an RF circuit, it is prudent to verify its effective frequency range.

25.13.4 Measuring RF Impedance

The most convenient method of measuring RF impedance today is to use a vector impedance analyzer (VIA) or vector network analyzer (VNA). See the preceding sections about those instruments for more information. Regardless of how the measurement is made, at RF there are many factors that can affect an impedance measurement, particularly in excess of a few hundred ohms. Here are some guidelines to minimize those effects:

• Stray capacitance: Even a few pF of stray capacitance can affect an impedance measurement at higher radio frequencies. Keep non-essential cables and surfaces away from the device or circuit being measured or use a test fixture (see below) to control the capacitance.

• Lead inductance: Unless a conductor is part of what is being measured, keep leads as short as practical.

• Skin effect: Minimize the extra resistance created by skin effect. Use silver-plated connectors at VHF and higher frequencies. Use large diameter or width conductors.

• Control test setup: If multiple tests are going to be made, consider making a test jig or fixture so that the layout of the cables and connections and stray capacitance is as consistent as possible between tests.

• Calibration: Make sure your instruments and test accessories are calibrated. Perform a calibration on network analyzers at the intended measurement plane.

• Reference Devices: Before making a measurement on the circuit or system under evaluation, measure a similar device for which the impedance is already known. A significant difference can alert you that there may be flaws or unexpected characteristics in your setup.

• Manufacturer Resources: Particularly for sophisticated analyzers and lab-quality instruments, the user's manual may show good practices for that type of measurement. There may be application or "app" notes available that go into detail.

USING AN IMPEDANCE BRIDGE

Most traditional 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.

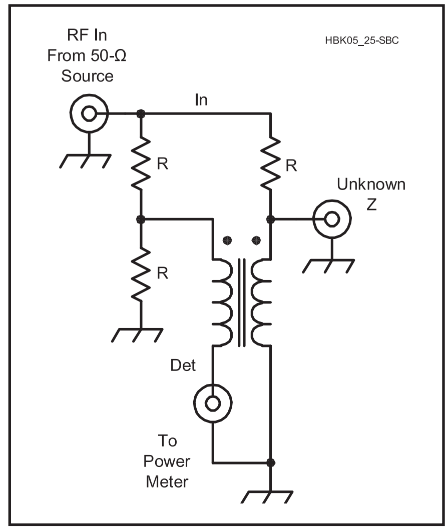


Figure 25.39 — A relatively simple diode detector circuit for measuring impedance that gives adequate results for many applications.

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. (See the Reference for Bird and the **RF Techniques** chapter section on Two-Port Networks for a discussion of return loss.)

The schematic of a simple return-loss bridge is shown in **Figure 25.39**. 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.

USING A NOISE BRIDGE

The noise bridge includes an adjustable bridge circuit similar to that in Figure 25.39. A wide-band noise generator is connected as the source and a conventional receiver is attached to the Power Meter port as a detector. Tune the receiver to the desired frequency and adjust the resistance and reactance controls for minimum noise in the receiver. If the receiver has a panadapter display, the null frequency can be seen on the screen, speeding the adjustment. Noise bridges are rarely used today in favor of the more convenient antenna analyzers. See the References entry for "The Noise Bridge" by Althouse for a more complete description of how a noise bridge works and how it is used.

25.13.5 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.

25.14 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.

25.14.1 Harmonic and Spurious Signal Measurements

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. **Figure 25.40** 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, but look carefully at the specifications for any instrument you are using. Some have a maximum input level that is as low as +10 dBm. For a 1-watt input limit, this means 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. One needs to consider both the maximum-power input rating of the spectrum analyzer and its linearity specifications. At a 0-dBm

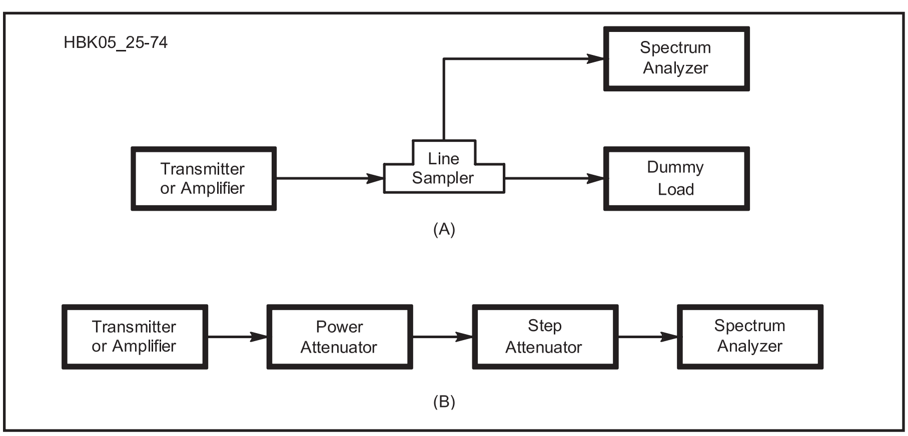


Figure 25.40 — Alternate bench setups for viewing the output of a high-power transmitter or oscillator on a spectrum analyzer. A uses a line sampler to pick off a small amount of the transmitter or amplifier power. In B, most of the transmitter power is dissipated in the power attenuator.

input level, for example, some spectrum analyzers may create harmonics internally that are not really present at its input.

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

It is necessary to ensure that the spectrum analyzer does not generate additional spurious signals that are then attributed to the system under test. Some of the spurious signals that can be generated by a spectrum analyzer are harmonics and IMD. 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 **Figure 25.41**. 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.

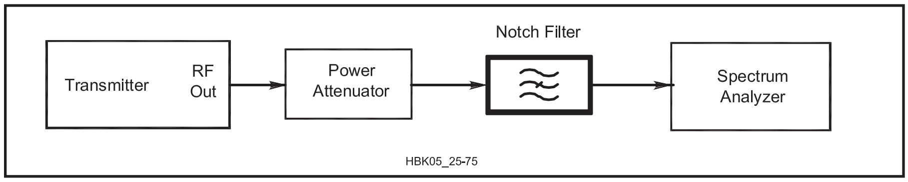


Figure 25.41 — A notch filter is another way to reduce the level of a transmitter's fundamental signal so that the fundamental does not generate harmonics within the analyzer. However, in order to know the amplitude relationship between the fundamental and the transmitter's actual harmonics and spurs, the attenuation of the fundamental in the notch filter must be known.

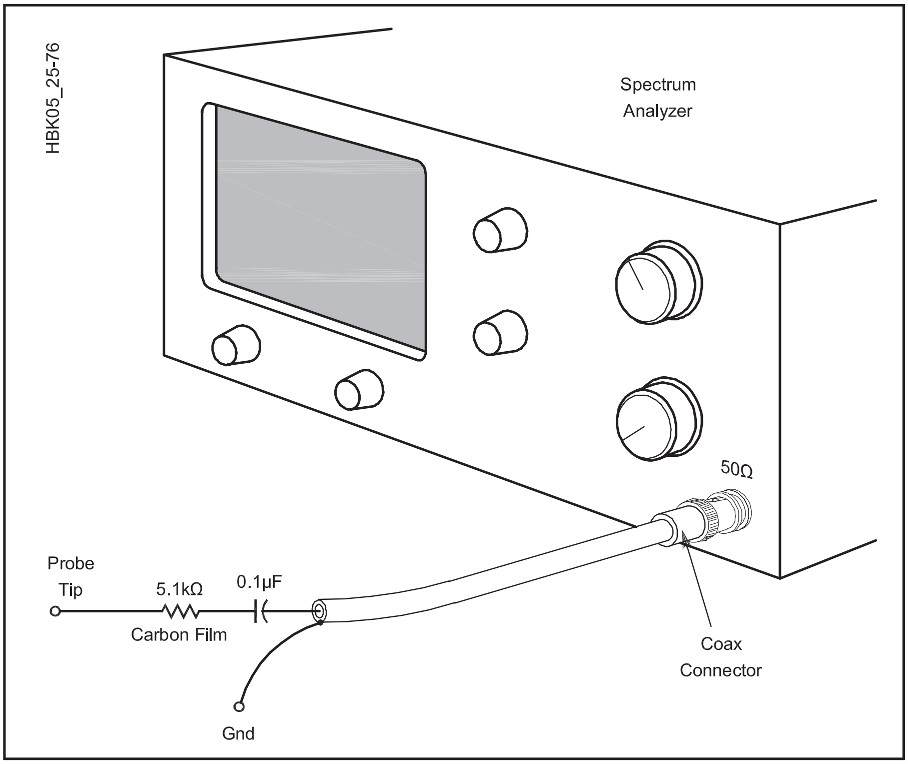


Figure 25.42 — A schematic representation of a voltage probe designed for use with a spectrum analyzer. Keep the probe tip (resistor and capacitor) and ground leads as short as possible.

SIGNAL AND INTERFERENCE TRACING

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 **Figure 25.42** 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. The resistor creates a severe mismatch in the coaxial cable so use the shortest cable practical to keep the low-pass RC-filter created by the resistor and cable's capacitance from degrading accuracy at the highest frequency being measured. Measurements using this setup should be treated as relative or qualitative and not as calibrated levels.

A different type of probe is shown in **Figure 25.43**. 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.

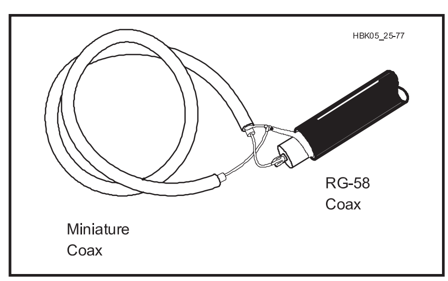


Figure 25.43 — A "sniffer" probe consisting of an inductive pick-up. It has the advantage of not directly contacting the circuit under test.

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.

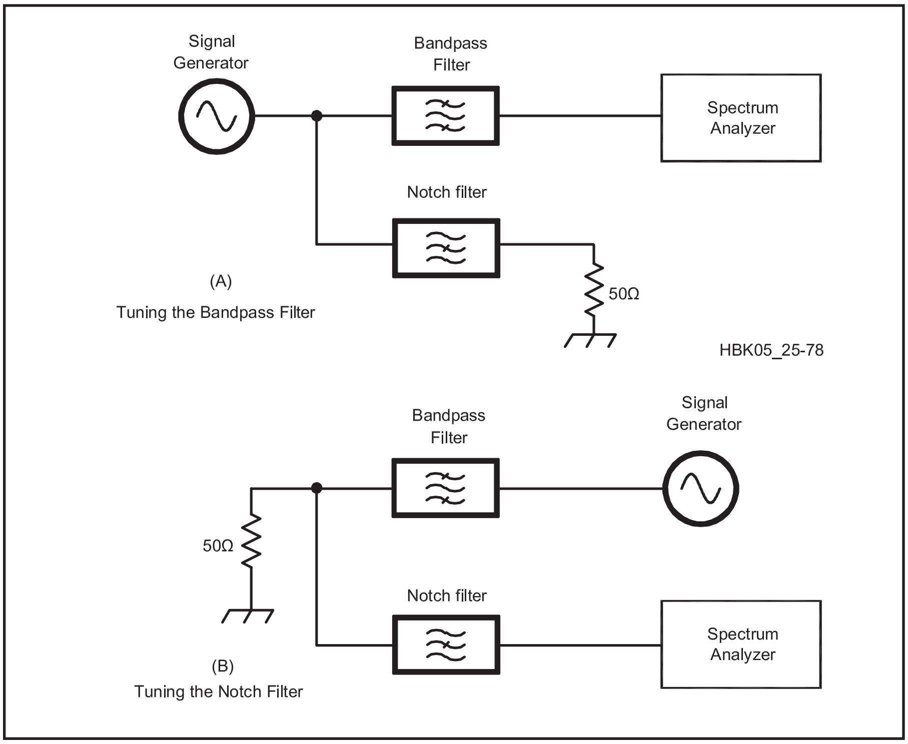


Figure 25.44 — Block diagram of a spectrum analyzer and signal generator being used to tune the band-pass and notch filters of a duplexer. All ports of the duplexer must be properly terminated and good quality coax with intact shielding used to reduce leakage.

25.14.2 Measuring Low-Level Signals

One very powerful characteristic of the spectrum analyzer is the instrument's capability to measure low-level signals. This characteristic is very advantageous when very high levels of attenuation are measured. **Figure 25.44** 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.

25.14.3 Frequency Response Measurement

Some spectrum analyzers can be used in conjunction with a suitable signal generator to measure the frequency response of a circuit. Generators connected in this way create 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. A tracking generator also simplifies many other measurements and adjustments that require involve a signal generator tuned to the same frequency as the spectrum analyzer. Medium-cost spectrum analyzers are available that include a tracking generator integrated with the analyzer for a modest extra cost.

Figure 25.45 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.

You can also 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 audio 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 all filters and audio stages, is shown by the average noise level in the spectrum analyzer window.

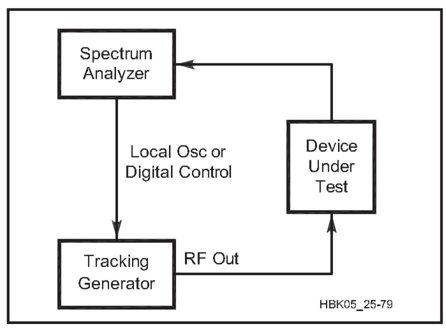


Figure 25.45 — A signal generator (shown in the figure as the "Tracking Generator") locked to the local oscillator of a spectrum analyzer can be used to determine filter response over a range of frequencies.

25.15 Antenna System Measurements

This section discusses measurements of antenna systems. The antenna system typically includes the antenna along with the transmission line and any other accessories such as switches, filters, matching networks, antenna tuners, and so forth. The instruments and measurements in this section assist with construction and installation, performance assessment, and troubleshooting of the entire antenna system. Related information is available in the **Transmission Lines, Antennas,** and **Assembling a Station** chapters.

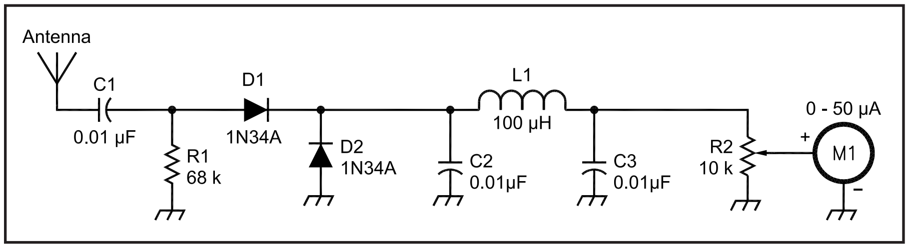


Figure 25.47 — Schematic diagram of the VE7NI field-strength meter.
C1-C3 — 0.01 µF capacitors.
D1, D2 — 1N34A diodes.
L1 — 100 µH inductor.
M1 — Analog meter, 50 µA.
R2 — Sensitivity control potentiometer, 10 kΩ.
Antenna—BNC female chassis mount socket. Antenna selection should match the frequency band for VHF and UHF. A random length of wire might work best for close field measurements on HF to 40 meters. Metal box enclosure is mandatory.

25.15.1 Field Strength Measurements

To determine how well an antenna is actually radiating, a way to detect the signal level at some distance from the antenna is needed. For measurements in the vicinity of the antenna, a *field-strength meter,* generally with a built-in antenna, 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 such as the one shown in **Figure 25.46** 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. **Figure 25.47** shows the schematic of a simple field strength meter. (This article by VE7NI is also included in the book's online material.)

Figure 25.46 — The VE7NI field-strength meter.

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 comparison may not be accurate if the test and reference antenna patterns are significantly different. Even if there are no wires, bodies of water, fences or other conducting objects in the vicinity, the ground reflection can result in an apparent additional gain of up to 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. (See the Reference for Wade) 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 October 2012 (covering the antenna range) and January 2013 (discussing measurements and equipment).]

The most common way to compare antennas is to ask for signal reports on the air. Unfortunately, propagation is so variable on most amateur bands (fading of more than 20 dB in a few seconds is common) 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.

An improved method of assessing antenna performance from distant stations is to use one of the automated receiving systems; the *Reverse Beacon Network* or RBN (**reverse-beacon.net** for CW signals), *PSKReporter* (**pskreporter.info** for PSK signals) and *WSPRNet* (**wsprnet.org** for WSPR signals). These systems consist of a worldwide network of independent receivers that decode signals and report the call sign and signal strength to a central server. The popular FT8 digital mode can also be used, saving decoded transmissions from other stations with objective signal strengths for analysis.

To use the systems for antenna system testing, the server can be queried or multiple decoding cycles can be used for comparative signal strengths after switching between antennas or making adjustments. Recognizing that propagation can change substantially over the short-term, to obtain reliable results, collect a large number of reports from multiple receiving stations over a period of time.

25.15.2 SWR Meters

The analog SWR meter is a descendent of the *QST* project by Lew McCoy, W1ICP which introduced the "Monimatch," to amateurs in 1956, originally created by Walter

Bruene, W5OLY. (**www.collinsradio.org/wp-content/uploads/2015/05/Understanding-the-Bruene-Coupler-Transmission-Line-Bold.pdf**) Directional RF wattmeters, also used to determine SWR, are covered in this chapter's section on Measuring RF Power.

Figure 25.48 shows the schematic of a typical unit. SWR can be computed from forward and reflected power, as shown by the following equation:

$$SWR = \frac{1+\sqrt{P_R / P_F}}{1-\sqrt{P_R / P_F}}$$

If voltages representing the two powers are provided, it's straightforward to create a circuit that computes SWR. This is the reason for the CAL (calibration) control on the meter. The meter is set to FWD (indicating the voltage representing forward power), power is applied, and the CAL control adjusted so that the forward power indication is the full-scale value of 1. By assuming $P_F = 1$ and P_R is a value between 0 and 1, the equation for SWR simplifies to:

$$SWR = \frac{1+\sqrt{P_R}}{1-\sqrt{P_R}}$$

The voltage representing reflected power will always be some fraction of the full-scale voltage, so the meter scale can then be calibrated to read SWR directly, instead of voltage.

THE BRUENE DIRECTIONAL COUPLER

Inductive and capacitive coupling are used to create a *directional coupler* that can provide the independent measurement voltages. The coupling provides samples of forward and reflected voltage and current from the undisturbed center conductor of the coaxial feed line. Figure 25.48 shows how voltages representing reflected power are obtained. As described by W5OLY, "A pickup wire placed parallel to the inner conductor samples the line current by inductive coupling. The voltage e_i induced in the pickup is determined by spacing, length, line current, and frequency. The mechanical dimensions determine the mutual inductance, M. The induced voltage due to line current is:

$e_i = –j2\pi fM$

where f is frequency in Hz and *j* represents a phase shift of 90°. This shows that the higher the frequency, the larger the induced voltage.

"The sample of voltage is picked up by capacitive coupling (C_{CPLG}) from the inner conductor to the pickup wire. A current due to this capacitance flows through R and develops a voltage across it; this voltage also increases with frequency because the reactance of the coupling capacitance goes *down* with frequency. That is:

$e_v = E\ R/ X_C\ R = ER / –j(1/2\pi fC) = j2\pi fERC$

when X_C is much larger than R." D and C form an RF detector that sums e_i and e_v, creating a single voltage proportional to the power in the line.

Since the line current and voltage contain components of both forward and reflected power, the single resulting output voltage also contains components of both. The current and voltage components of reflected power are 180° out-of-phase, compared to those of forward power. So e_i is added to e_v, producing a voltage proportional only to reflected power in the figure.

The different polarities of e_i are obtained by reversing the current sensing pickup. This is accomplished by two identical pickup circuits in the meter; e_v is the same in both circuits, but the e_i pickup direction is reversed from one to the other. One circuit produces a voltage proportional only to forward power and the other proportional only to reflected power. Display or forward or reflected power is controlled by the switch that selects which voltage is applied to the meter.

Continuing with a note from W5OLY, "Since the current and voltage pickups both increase with frequency, their ratio will stay the same. The variation in pickup just means that the sensitivity goes down at lower frequencies." That's why the CAL adjustment is necessary not only at different power levels, but at different frequencies. The meter scale is calibrated according to the equation for SWR when the calibration adjustment places the voltage representing P_F at full-scale.

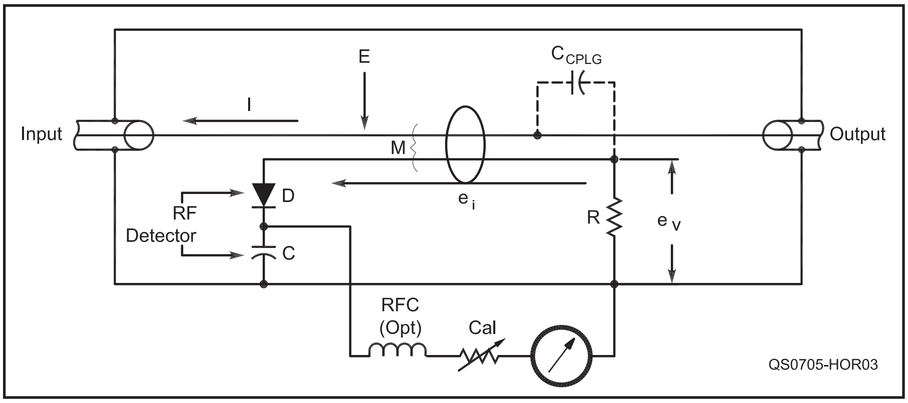


Figure 25.48 — One-half of a Monimatch directional coupler circuit that senses the reflected voltage and current components. The samples of induced voltage e_i and line voltage e_v sum in the RF detector formed by D and C, producing a voltage that drives the meter. The Cal control is adjusted so that a full-scale reading is obtained. The meter is then switched to display the output of an identical circuit oriented in the opposite direction to picks up the forward components. The meter is then calibrated according to the SWR equation in the text.

For best performance from a Monimatch type of directional coupler, the value of R in one of the detector circuits should be adjusted so that the two circuits produce the same value of e_i for any given current. In addition, balanced detector diodes would also provide better performance at low power levels where the detected voltage level is small enough for variations in the diode forward voltage to introduce significant errors. However, for low-cost equipment, using fixed components is generally "good enough."

For SWR meters that have been damaged, repair is very simple. The manufacturer calibrated the meter assuming very similar component values. Replace both detector diodes with the same type of diode, preferably from the same batch of components. The diodes can be matched by using a multimeter with a diode test function that displays the diode's forward voltage. If one of the original diodes is undamaged, choose a pair of diodes with a similar forward drop.

25.15.3 Time-Domain Reflectometry

Time domain reflectometry (TDR) shows what happens to a short, abrupt pulse as it travels through a transmission line. The pulse is reflected by any changes in impedance, such as an open or short (complete reflection) or a change in the line's characteristic impedance (partial reflection). The resulting series of pulses and reflections is displayed as a sequence in time, thus the name of the technique.

In an ideal transmission line terminated by its characteristic impedance, Z_0, the pulse will travel to the far end and be dissipated in the termination, so the trace will be a perfectly flat line. But at any point along the

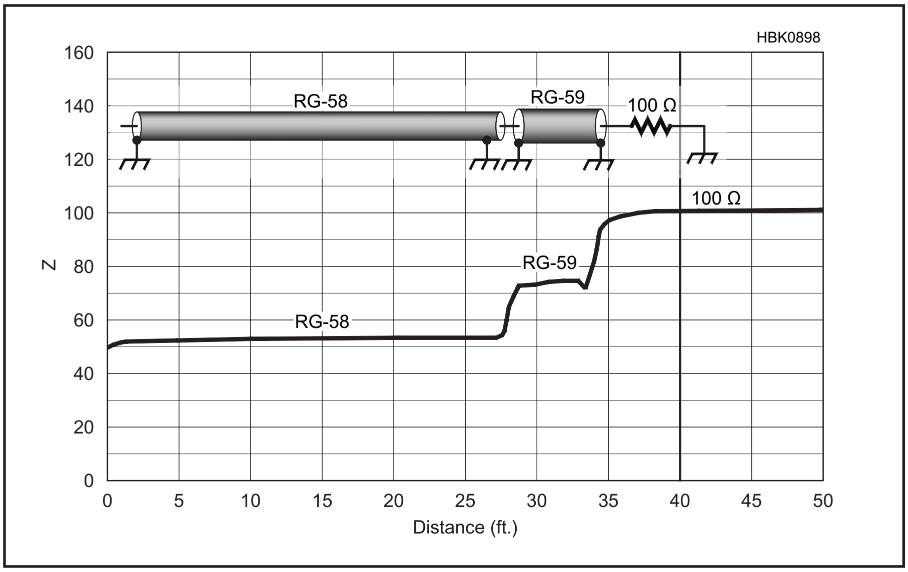


Figure 25.49 — Example of a direct method TDR (see text) connected to a length of RG-58 coaxial cable (50 Ωimpedance) followed by a short length of 75 Ω RG-59 and a 100 Ω load. The discontinuities are clearly shown as the reflections add to the input pulse.

line where the impedance changes (called a *discontinuity*) some of the pulse's energy will be reflected back toward the line's input. The reflected component of the pulse creates an artifact (visually, a "bump") on the otherwise straight line.

The sequence of pulses and their reflections is the *impulse response* of the line. An *impulse* is basically a very short pulse that begins and ends before the system can respond and stabilize. A mathematical impulse is an infinitely-narrow made up of all frequencies from zero to infinity.

While it is not possible to create an ideal (perfect) impulse, a very fast rising edge of a longer pulse is a good enough approximation to measure the line's impulse response and the same information can be measured. The longer pulse is called a *step function*. The ideal step function is an infinitely fast change from one level to another, after which it remains at that level. The response of the line to the longer pulse is called the *step response*. Like the impulse, the infinitely fast change in level also contains all frequencies.

A *time domain reflectometer* is the instrument that generates the pulse and displays the results. A TDR displays amplitude (voltage) on the vertical axis and time on the horizontal axis. The position of each artifact along the TDR trace corresponds to the distance from the transmission line input to the discontinuity that produced it. Large discontinuities occur when a line is open-circuited or short-circuited, or when it is connected to an antenna. (Most antennas are matched to the line at their operating frequency, but at other frequencies they are not. Since the pulse contains all frequencies, an antenna is a large discontinuity.) Small discontinuities occur at splices or when a line is damaged.

The delay between the input pulse's rising edge and the artifact is the round-trip time in the line from the TDR to the discontinuity. If the line's *velocity of propagation* or *velocity factor* (VF) is known, the physical distance from the input to the discontinuity can be calculated. The shape of each reflection can sometimes provide a clue as to the nature of the discontinuity.

In this sense, the TDR is very much like a radar display in which a pulse is transmitted (shown at the center on a radar display) and any echoes from discontinuities in the air (i.e. targets) reflect some of the pulse back toward the transmitter. The farther away the target, the longer it takes for the pulse to travel to the target and back to the receiver. A radar screen shows echoes from all directions. The TDR only shows echoes from one direction, along the line. The larger the echo, the larger the target or discontinuity.

DIRECT METHOD TDR

There are two common TDR implementations, with variations of both. In the "direct method," which is the oldest and simplest, the line is driven by a pulse. This can be a single pulse, or it may be a train of pulses like a square wave.

The pulse and all of the reflections are displayed on an oscilloscope trace that is triggered by the pulse's rising edge. The rise time of the pulse must be much shorter than the round-trip time for the impulse to travel to and from the discontinuity. **Figure 25.49** shows an example of a direct method TDR with the pulse generator and cable attached to the oscilloscope which is displaying the pulse.

Although a digital scope can capture a single pulse and its reflections, making the pulse repetitive means that multiple responses can be averaged to improve the signal-to-noise ratio. The signal-to-noise ratio of a system excited by a single pulse can be rather limited. Repeated pulses also sustain the trace so the operator to see it if the scope is an analog model. The repetition rate of the pulse must be much slower than it takes for the impulse to make a complete round trip time through the line, however. This ensures the line's response dies out completely before the next pulse excites the line again. See the reference article by King about this type of TDR and the *ARRL Antenna Book* also includes material on TDR.

Modern antenna analyzers can use the direct method, as well. For example, the AIM family of vector impedance analyzers (**www.arraysolutions.com**) excite the line with a step function waveform having a very fast rise time. This implementation provides a display of the impedance at every point on the line — it can, for example, show the relative impedance of cables having different Z_0, as well as the position of discontinuities. Figure 25.49 shows a TDR display from an AIM analyzer connected to a length of RG-58 (50 Ω cable), a short length of RG-59 (75 Ω), and a 100 Ω load. The discontinuities at the cable and load transitions are clearly shown. Software converts the raw data from the pulse amplitudes into impedance and the time is converted into distance along the line.

The TDR function can be used to determine if the line has been degraded, for example, by water leaking into the coax or if the line has been shorted or cut somewhere between the transmitter and the antenna. Damage or defects can be located within a few inches and this reduces the effort required to repair the line. Defective connectors can also be indicated by short glitches in the trace corresponding to the location of the connectors.

TRANSFORM METHOD TDR

The other common implementation could be described as the "transform method." Instead of determining the impulse response of the cable with a pulse, the excitation is a sine wave swept over a range of frequencies and the analyzer captures the *frequency response*. An inverse Fourier Transform (see the chapter on **DSP and SDR Fundamentals**) is performed on that frequency response, producing the time-domain response. (Frequency and time are the inverse of each other; the complete frequency response of a system contains its time response and the response to an ideal impulse contains the frequency response. A Fourier Transform (FFT) of the time response provides the frequency response, an Inverse FFT of the frequency response provides the time response.)

Transforming Analyzer Data

Like the trace on a simple oscilloscope, the TDR plot of the impulse response contains all frequencies (or the range of frequencies if it is transformed from a sweep). The scope and the TDR plot are "frequency blind" — that is, they display information only about the time response, and no information about the frequency response. An impulse or sweep measurement can, theoretically, be manipulated mathematically to compute the impedance at every point in the line over the same range of frequencies. The precision of that computation and whether it is practical, depends on how the data is gathered (sweep rate, sweep range, spacing of data points) and the software tools available. Frequency sweep ranges chosen for TDR may be inappropriate for examination of other line properties. Free software such as *SimSmith* (**www.ae6ty.com/smith_charts.html**) and *ZPlots* (**ac6la.com/zplots.html**) can accept swept measurements made at discrete frequencies to compute and plot the impedance at any point on a line if the characteristics of the line are known. Data is interchanged between a measurement device and software programs (and between one software program and another) by means of a plain text file. These files, defined by the Touchstone format (see previous sidebar) can take several forms that are defined by the first line(s), called a "header." The filename extension indicates the type of measurements:.s1p files describe single port measurements, like impedance or a time response and .s2p files describe two-port measurements such as the S21 (gain) transfer function produced by a vector network analyzer.

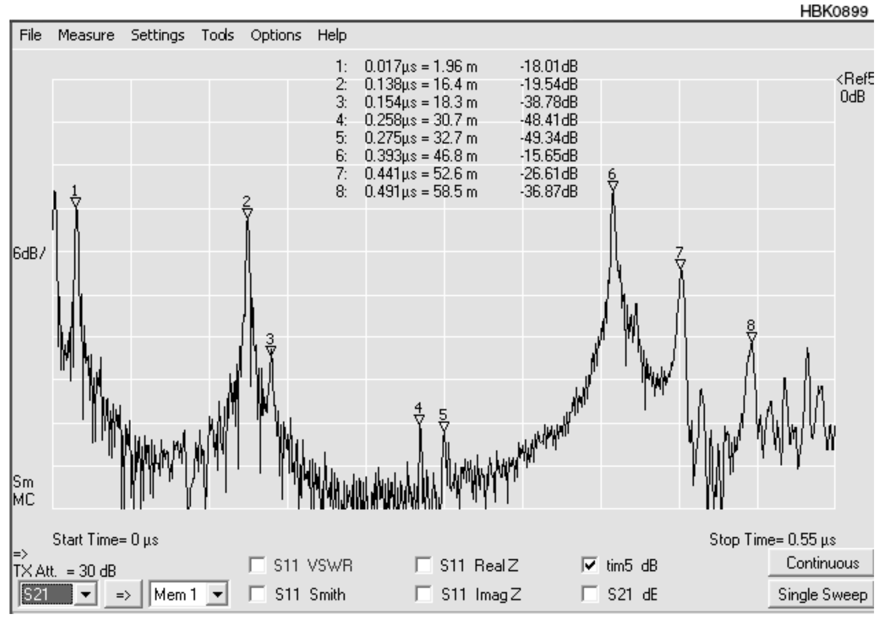


Figure 25.50 — The impulse response of a feed line attached to a 30 meter dipole. The system is swept from 50 to 500 MHz. See text for an explanation of the markers.

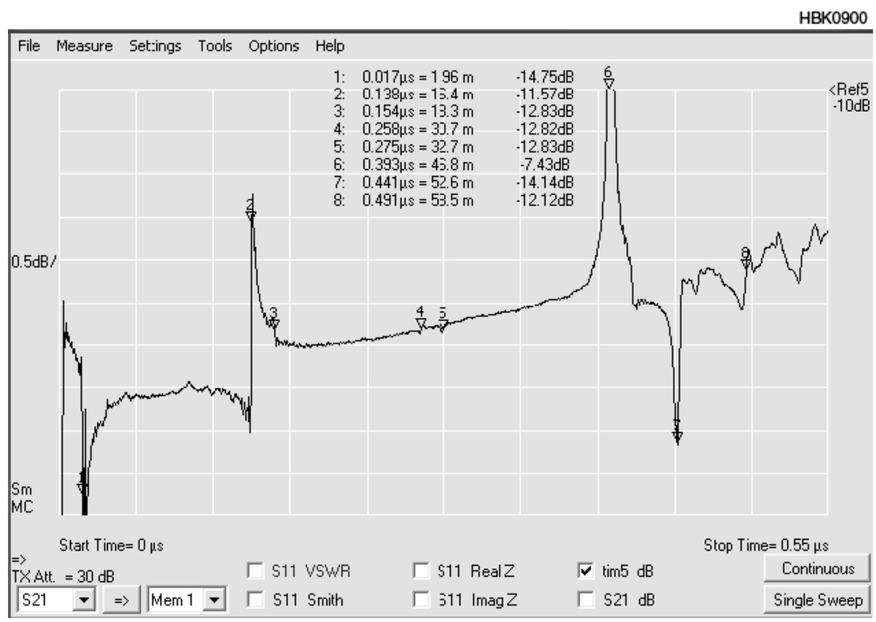


Figure 25.51 — The step response similar to the direct method for the same antenna system in Figure 25.10. See text for an explanation of the graph.

Some antenna analyzers and vector network analyzers use this method. One example is the handheld SARK-110 vector impedance antenna analyzer (**www.sark110.com**) The sine wave exciting the cable need not be a continuous sweep — rather, it can be stepped over a wide range of frequencies and the frequency response is computed from that data. Sweep range, spacing between data points, and the settling time at each data point are set by the user.

An inverse Fourier Transform produces spurious artifacts which must be removed by applying a mathematical windowing function to the transformed data. Several mathematically different windowing functions are commonly used, and which of the windows provides the most useful display depends on the shape of the impulse response.

The frequency content of the excitation strongly influences the degree of detail that the measurement can reveal. When the excitation is an impulse, a very fast rise time reveals greater detail. When the excitation is a swept sine wave, a wider frequency range reveals the greatest detail. Currently available analyzers can sweep from 1 kHz to more than 1 GHz.

For TDR studies using the sweep method, a sweep from 5 to 500 MHz (or from 500 MHz to 1 GHz) will clearly show detail that would be missed with a sweep to only 100 MHz, while a 5 MHz to 1 GHz sweep may provide too much detail (or show discontinuities that don't matter below 50 MHz). Beginning the sweep in the HF range avoids smearing of the data due to the variation of VF with frequency.

Some analyzers also have a feature called *Distance to Fault*. This measures the length of a transmission line to any point(s) along the line where the impedance differs from its characteristic impedance (Z_0). The "different impedance" can be an open or short circuit, the point at which an antenna is connected, or it can be some small change in the impedance such as that which occurs at a connector, or is the result of damage to the cable. With the best TDR systems, the line can be accurately measured with or without an antenna connected.

If the line is disconnected from the antenna, the distance to fault is the total length of the line. If the line has been damaged somewhere this measurement gives you an idea where the damage is, which is very handy when the line is buried or otherwise requires special access. TDR can also locate impedance discontinuities or other changes in the line's characteristic impedance.

EXAMPLES OF THE TRANSFORM METHOD

Figure 25.50 is the impulse response of the feed line for a 30 meter half-wave dipole at a height of 100 feet, computed from a sweep over the range of 50 to 500 MHz. Marker 1 shows the effect of lightning protectors at the station entry bulkhead. Marker 2 is a coax splice (two PL-259s and a PL-258 double-receptacle). Markers 3, 4, and 5 are coax defects. Marker 6 is the antenna feed point. Marker 7 is the end of the antenna (displayed distances are for the feed line, so for the antenna are divided by 0.795) the peak at Marker 8 is unexplained, but most TDR sweeps show multiple reflections after the antenna.

Figure 25.51 is the step response (similar to that from the direct method shown in Figure 25.49) computed from the same sweep as that used for Figure 25.50. The data revealed (and a visual inspection confirmed) that the cable inside the station and from the bulkhead to the coax splice is 50 Ω, but that the cable from there to the antenna is 75 Ω. A list of discontinuities is provided at the top of the graph with time delays and computed distances.

TDR measures time; to convert that measurement to physical distance, we must provide the velocity factor. The 75 Ω cable is Belden 8213, this sample of which has a measured VF of 0.795 at VHF. The 50 Ω cable has a measured VF of 0.8425 at VHF. In the setup screen for this TDR measurement, VF was set at 0.795, so distance measurements will be correct for the 75 Ω cable, but wrong for the 50 Ω cable. Computed results could be made correct for the 50 Ω cable by changing VF in that setup screen, or by leaving VF at 0.795 and applying a correction factor of (0.8425/0.795) to the dimensions of the 50 Ω cable.

25.16 Receiver Measurements

25.16.1 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 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 **Figure 25.52**. 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 may need to 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 1/5 the maximum output level to prevent the peaks from clipping.

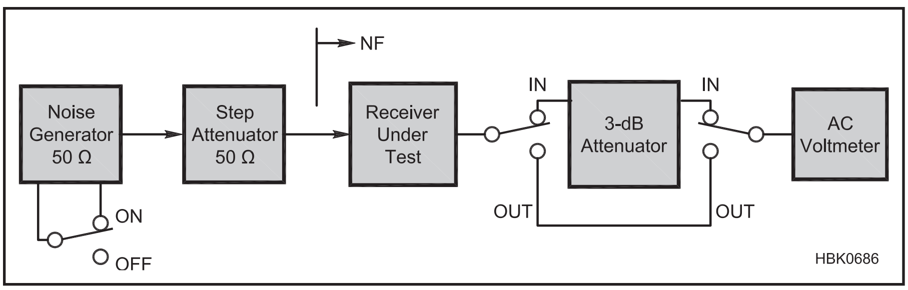


Figure 25.52 — Setup for measuring receiver noise figure.

ARRL Lab Test Procedures Manual

All of the test procedures used by the ARRL Lab in evaluating equipment for *QST* Product Reviews are published in the ARRL Lab's *Test Procedures Manual*. This manual is published online for public use at **arrl.org/product-review**. While the procedures refer to specific pieces of test equipment, they are easily adapted or used as guidelines for developing your own tests.

NOISE FIGURE OF RF AMPLIFIERS

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.

NOISE FIGURE METERS

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. A low-level 5 to 7 dB ENR source such as the HP 346A is best for testing high-performance preamplifiers. A higher level 15 dB ENR source is also common since it can be used for self-calibration of the meter.

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. Rohde & Schwarz (1MA178) and Keysight (5952-8255) both publish good application notes on noise figure measurement that are available online and are listed in this chapter's References section.

25.16.2 Receiver Sensitivity (MDS)

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." (Many people can hear a signal below the MDS due to the additional processing power of the auditory system.)

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 **Figure 25.53** 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.

Typical Test Conditions:

Filter width: SSB 2.4 kHz, CW 500 Hz (sharp passband shape, if available)

Attenuation OFF; Preamplifier OFF

Noise Blanker OFF; Noise Reduction OFF, Notch filter OFF

RF Gain MAX; AGC Medium time constant

In the hypothetical example of Figure 25.53, 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 (e.g. an HP-3400B, Ballantine 323, or Rohde & Schwarz URE3 are often used) 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 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 (S+N/N) 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

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

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

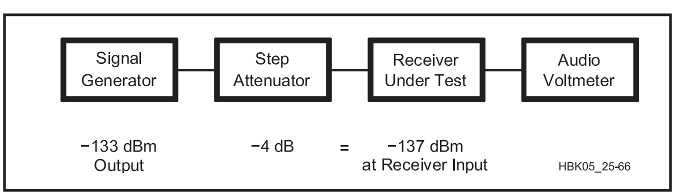


Figure 25.53 — A general test setup for measuring receiver MDS, or noise floor. Signal levels shown are for an example discussed in the text.

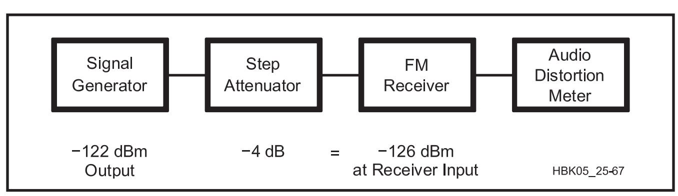


Figure 25.54 — FM SINAD test setup.

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 (e.g. HP-339A or Rohde & Schwarz UPV) is needed as shown in **Figure 25.54**. 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.

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

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

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.

25.16.3 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.

It is important to note that analog superheterodyne receivers and digital SDR receivers respond to strong signals differently. While both may produce distortion products, the signal levels at which the products appear and how the levels change are different in the two types of receivers. Hybrid receivers in which analog circuits are used to convert signals to an IF where they are digitized and processed using DSP techniques, sometimes referred to as "DSP-IF," respond differently than either pure analog or fully-digital receivers. Be cautious in comparing IMD measurements between the different receiver architectures. See the **Receiving** chapter's section on Receiver Dynamic Range for more information.

BLOCKING DYNAMIC RANGE (BDR)

Blocking dynamic range (BDR) or *blocking gain compressions* 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.

It is useful to measure BDR for receivers using both analog superheterodyne and digital SDR architectures, including hybrid analog/digital designs and even direct-sampling receivers. Any analog circuitry in the signal path, including signal conditioning circuits in an ADC or DAC are susceptible to overload. Making the same measurement on all receivers gives a better picture of comparative performance.

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 **Figure 25.55**. 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 Figure 25.55, the blocking level is the level from the signal generator minus the loss of the hybrid combiner and attenuator,

BL = –7 – 3 – 10 = –20 dBm

The blocking dynamic range is given by BDR = BL – MDS = –20 – (–137) = 117 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 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.

In a direct-sampling SDR receiver, no blocking occurs until the ADC is driven into saturation (clipping). This assumes correct design of the active stages preceding the ADC. (Circuitry ahead of the ADC, such as an RF amplifier, can exhibit gain compression, for example.) Thus, for SDRs, the clipping level is measured and stated in dBm, rather than blocking dynamic range (BDR) in dB.

RECIPROCAL MIXING DYNAMIC RANGE (RMDR)

Reciprocal mixing is the name for the mixing of a nearby interfering signal with the phase noise of the receiver's local oscillator or phase noise sidebands of the digitizing ADC's clock signal. The noise signals mix with strong signals close in frequency to the desired signal, producing unwanted noise in the audio output or detection channel. This degrades receiver sensitivity.

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

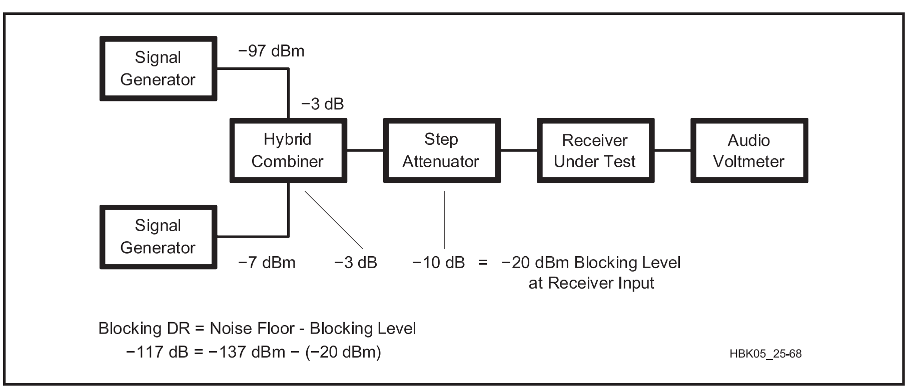


Figure 25.55 — Receiver blocking dynamic range is measured with this equipment and arrangement. Signal levels shown are for the example discussed in the text.

suite includes a separate measurement for reciprocal mixing. The test setup is the same as for MDS except that the signal generator in Figure 25.53 is replaced with a low-phase-noise crystal oscillator (or OCXO) with an output power level of +15 dBm. The OCXO operates at a fixed offset from the desired signal. The oscillator signal level is increased until the receiver noise floor increases by 3 dB as measured at the audio output.

Since the output of the OCXO is usually fixed, a variable attenuator is used to reduce the output level in controlled steps. 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 and 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 dynamic range* is expressed as:

RMDR (in dB) = (OCXO – A) – MDS

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

The same test can also measure phase noise at the particular receiver offset from the oscillator.

Phase noise (in dBc/Hz) = (RMDR + 10 log BW)

Where BW is the receiver's filter bandwidth. If BW = 500 Hz, 10 log BW = 27)

Typical Test Conditions:

Filter width: CW 500 Hz (sharp passband shape, if available)

Attenuation OFF; Preamplifier OFF

Noise Blanker OFF; Noise Reduction OFF, Notch filter OFF

RF Gain MAX; AGC Medium time constant

In general, to assess the impact of strong off-channel noise on weak signals, one should look at the worst-case of blocking and reciprocal-mixing dynamic range.

Dynamic Range Metrics: Superheterodyne vs. SDR Receivers

Reciprocal Mixing (RM) is the noise generated in a superheterodyne receiver when noise from the local oscillator (LO) mixes with strong, adjacent signals. All oscillators have noise sidebands, and some types and designs have more than others. The noise sidebands mix with the strong adjacent off-channel signal, creating noise products in the mixer's output. This noise can degrade the sensitivity of the receiver and is most notable when a strong signal is just outside the IF passband. Reciprocal mixing is worse with a single strong signal 2 kHz away from the tuned frequency than it is 5 or 20 kHz away.

It's interesting to compare to the two-tone third-order IMD dynamic range (3IMD DR) to the reciprocal mixing dynamic range (RMDR). In the case of 3IMD, two strong adjacent signals add up to cause an unwanted effect, but in the case of RM, a *single* strong adjacent station 5 or 2 kHz away from the desired signal has more of an impact on the ability to hear a desired weak signal. For most superheterodyne receivers, RMDR is the worst dynamic range figure at 2 kHz spacing. The major source of phase noise in a direct-sampling SDR receiver is the ADC clock subsystem. Thus, it is essential to ensure that the clock source and distribution chain have as low phase noise as practicable.

It has been observed in the ARRL Laboratory that many direct-sampling SDR receivers exhibit little, if any reciprocal mixing up to the point of analog-to-digital converter (ADC) overload. These receivers have no front-end mixer and local oscillator, thus, no sideband noise of an oscillator to mix with adjacent signals.

We perform the reciprocal mixing test at 14.025 MHz, using a very low-noise Wenzel test oscillator with a measured output of +14 dBm to generate the off-channel, strong signal. A test oscillator must have the lowest sideband noise as possible — considerably lower than the reciprocal mixing being measured. An excellent receiver will exhibit 100 dB of RMDR at 2 kHz spacing.

Blocking — an apparent reduction in receiver gain due to the presence of a strong signal — was problematic in years past. With the AGC off, the desired signal, especially if weak, would disappear from the receiver audio if a strong, off-channel signal was present. The ability to reject adjacent signals has improved greatly in recent years, so much so that blocking dynamic range (BDR) greatly exceeds the other dynamic ranges. During Lab tests, reciprocal mixing in a receiver can raise the audio output noise enough to drown out the desired signal! Some of our laboratory consultants have expressed that the BDR figure is a bit useless because of this behavior. Agreed! Still, some still wish for us to report blocking, since we have the use of signal analyzers set to narrow audio bandwidths (2 or 5 Hz).

Many SDR receivers measured for *QST* Product Review exhibit little or no blocking effects up to the point of analog-to-digital converter (ADC) overload. This behavior, along with an SDR's reciprocal mixing behavior is most desirable. Typical BDR of modern receivers at 2 kHz spacing is 110 dB or greater.

Two-tone Third-Order Intermodulation Distortion (3IMD) happens when two strong signals (2 and 4 kHz away from the tuned frequency, for example) appear at the antenna jack, with a resulting unwanted "phantom" signal heard at the tuned frequency. With a superheterodyne receiver, IMD is generated at the input stages to the first IF, with the undesired mixing products passing through the rest of the receiver stages. In an SDR receiver, the ADC clock quality and the cumulative signal level input to the ADC determines the level of IMD at the speaker (see other sidebar).

Due to design improvements of both superheterodyne and SDR receivers, typical 3IMD DR at 2 kHz spacing is 90 dB or greater. Due to changes in technology, most modern receivers do not exhibit a 3:1 relationship between the IMD signal level and the IMD input level. The ratio can be significantly greater or less than 3:1. Since the third order intercept point (IP3) figure is calculated based on the assumption of a 3:1 ratio, this figure is meaningless with today's receivers. Emphasis must be placed on all three dynamic ranges. The lowest dynamic range at 2 kHz of the three dynamic ranges we publish is the dynamic range of the receiver. The benchmark for excellent performance is 100 dB.

INTERMODULATION DISTORTION DYNAMIC RANGE (DR2 AND DR3)

Intermodulation distortion (IMD) means the creation of unwanted signals at new frequencies because of two or more strong interfering signals modulating each other. This happens due to non-linearities of circuits, active and passive, in the path of the strong signals.

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 **Figure 25.56** is the same for second and third-order IMD. For receivers with high dynamic range, to obtain sufficient output power and isolation of the two signal generators, it may be necessary to follow each one with a wideband 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

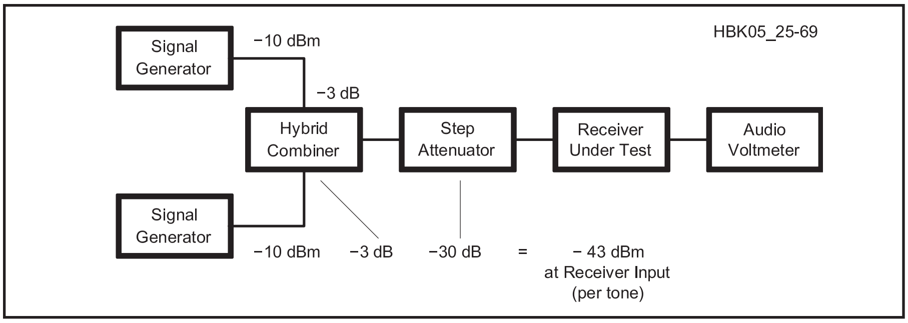


Figure 25.56 — The test setup for receiver intermodulation distortion dynamic range. Signal levels shown are for the example discussed in the text.

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.

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

IMD = –10 – 3 – 30 = –43 dBm

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

IMD_DR = IMD – MDS = –43 – (–137) = 94 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. (This is the *subtractive method* discussed in ITU-R Recommendation SM.1837-1, Section 2.)

If the subtractive method is used for DR3 measurement, RMDR must be measured in the course of the same test suite to allow the test engineer to determine whether the DUT is IMD-limited or phase-noise-limited.

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.

An alternative method for measuring IMD is presented in the supplemental paper "Testing and Calculating Intermodulation Distortion in Receivers," by Ulrich Rohde, N1UL. Avoiding potential errors of measuring noise power and audio response characteristics of the receiver, Rohde's method measures IMD product levels at a repeatable S meter reading. An adjustable attenuator is used to control the IMD product levels, with the difference in attenuator readings being much less error-prone than noise power measurements.

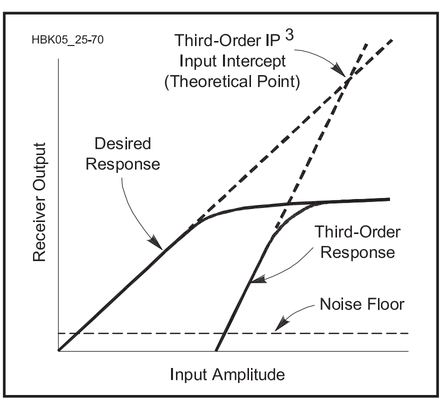


Figure 25.57 — The third-order intercept point can be determined by extending the lines representing the interfering signal level and the third-order intermodulation products on a plot of the signal levels in dB.

THIRD-ORDER INTERCEPT POINT (IP3)

The third-order intercept (IP3) is generally not a valid concept for direct-sampling SDR receivers that do not use an analog front end. 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 IP3 measurements of an SDR and a conventional analog receiver is likely to give misleading results.

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 **Figure 25.57**. 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 × IMD_DR 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.

Note that IP3 (3rd-order intercept) is not a valid test metric for a direct-sampling SDR, as IMD in an ADC follows a quasi-1st-order rather than a 3rd-order law. The transfer and IMD curves diverge and never intersect. In a conventional receiver, IP3 is the convergence point of the transfer and IMD curves.

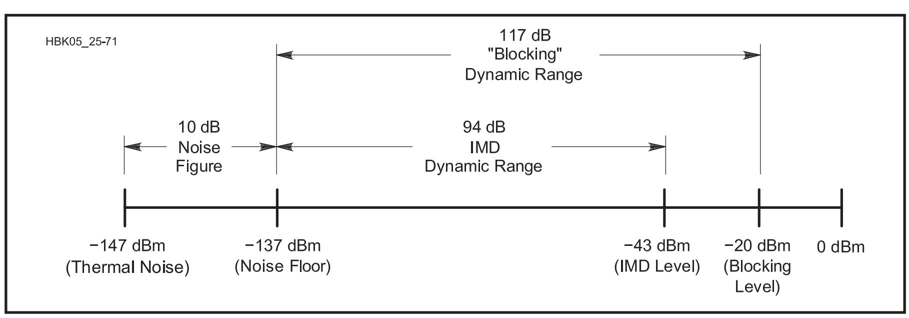


Figure 25.58 — Performance plot of the example receiver discussed in the text.

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

IP2 = MDS + 2 × 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.

Figure 25.58 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.

Third-order IMD dynamic range may also be measured on an FM receiver. The test setup is the same as Figure 25.56 except that the audio voltmeter is replaced with an audio distortion meter, as in Figure 25.54. 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 dB

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

SDR Behavior with Two-tone Third OrderIntermodulation Dynamic Range: The Third Signal

The ARRL Laboratory determined that our normal two-tone third order IMD dynamic range (3IMD DR) tests did not always apply to the way software defined radio (SDR) receivers work in the real world.* With an SDR receiver, all of the demodulation is done by software. An SDR converts a block of analog RF spectrum to a digital stream of data, then uses software to select and demodulate a desired signal from that data. The digitization of an analog signal results in a sequence of small steps in signal level. These small steps are a type of nonlinearity that forms intermodulation (IMD) products, similar to those observed in analog receivers. (See reference entry for Allison.)

For analog receivers, once intermodulation occurs, it gets strong quickly as the off-channel signals used for testing are increased. With an SDR, however, because the small steps are the same for all levels of the signal being digitized, intermodulation at low levels does not vary significantly with the level of test signals. For the steady-state sine waves seen from signal generators in a laboratory environment, the steps are the same for each cycle of RF. Any intermodulation that is created adds coherently in the receiver output.

In some SDR designs, random noise is added to the RF input to randomize the digitization process, preventing intermodulation products from adding coherently. This technique is called dithering. When a receiver is connected to the antenna, band noise present at the input to the analog to digital converter (ADC) can, in most cases, serve to randomize the digitization. The net effect is the same — low-level intermodulation seen in a laboratory test environment will not be present when that receiver is connected to an antenna and used to receive on-the-air signals. This effect was verified by measurements in the Lab by adding a receive antenna (acting as our dithering source) to our two-tone IMD test fixture; the level of band noise varied with local conditions and propagation, thus, the amount of dithering present also varied, causing IMD products to vary as well.

With such observations, we had to find out if the IMD products varied as a function of noise or as a function of level. We substituted a random noise generator as our dithering source. The result: low-level IMD products were masked by the raised noise level with little change in the level of the IMD products. Varying the level of the noise had no effect, except the unwanted effect of a raised noise floor. Sherwood & Farson have confirmed via lab testing that noise added to the 2-tone test signal at the DUT RF input did not reduce IMD product amplitude, but merely appeared as noise in the DUT audio output. For noise dithering to be effective, ADC on-chip dithering is required in which added digitized noise is digitally subtracted from the ADC output bitstream. (See reference entry for Kester.)

Since noise or its level had no effect on the level of the IMD products, the only other dithering source received on an antenna must be discrete signals! We soon discovered that by using a strong, off frequency, single signal as our dithering source (the third signal), the IMD product level could be reduced. By varying the level of the third signal (generator), the IMD product level also varied. Thus, when connected to an antenna, the levels of all the signals at the input to the ADC vary considerably with propagation and mode, and thus the dynamic range varies with propagation and mode. This significant discovery determined that adding noise (dithering) does not necessarily improve an SDR's 3IMD DR, but that a single, strong, constant, off-frequency signal does. By experimenting, we found that by varying both the third signal's level and frequency, a "sweet spot" could easily be found that maximizes the 3IMD DR. A third signal at a level of –43 dBm, 200 kHz away from the tuned frequency is a good example of a "sweet spot."

This single additional signal, at a specific level and frequency, hardly represents actual on-air conditions. In reality, the 3IMD DR can be a moving target, varying by 30 dB or more depending on conditions! With SDRs that exhibit improved 3IMD DR, *QST* Product Reviews report "up to XX dB," which represents the absolute best-case dynamic range. A user of such devices can expect much lower dynamic range under average band conditions, with an effort made to mention the expected dynamic range during "quiet" conditions.

It is interesting to note that by design, some manufacturers of SDRs now employ a "third signal" to improve 3IMD DR; the third signal is internally generated and tracks with the tuned signal, at a specific level, which minimizes IMD products and maximizes 3IMD DR.

Please keep in mind that when an antenna is attached to a receiver, atmospheric, solar and man-made noise will mask much of the low-level IMD products that are measured in a laboratory environment. That is, the dynamic range can be higher under laboratory "quiet" conditions, where a receiver's noise floor is measurable.

INTERFERENCE-FREE SIGNAL STRENGTH (IFSS)

Since a direct-sampling SDR does not experience IMD in the same way as an analog superheterodyne, a different test is required to measure the spurious products generated by strong signals. Using the same test configuration as for single-frequency DR3 tests described above, the IFSS test creates a chart of absolute IMD product power level vs. test signal input power level, measured at 1 dB intervals over the range from ADC clip level to the noise floor; with the ITU-R P.372 site band-noise levels at the test frequency as datum lines. The same instrumentation is used as for the "classical" single-point DR3 test.

ADC CLIPPING

Also known as *overflow*, *overload*, and *saturation*, *ADC overload* occurs when the signal level into an ADC exceeds the maximum encoded level for the ADC. At and beyond this point, the receiver will exhibit spurious and non-linear responses. Even if the overload condition is removed, there may be persistent effects in DSP functions such as filtering or time-sensitive functions such as AGC or noise reduction. Many receivers provide an indicator that lights when the overload condition is present.

In the ADC Overload test, the receiver is offset +25 kHz above the test signal frequency and the input level required to light the overflow indication or screen icon is noted. The overload indication should only occur when a strong out-of-channel signal is present. In-channel signals stimulate AGC action to reduce signal levels at the ADC input.

Typical Test Conditions:

Receiver tuned to 14.1 MHz, Test signal frequency 14.125 MHz

Selectivity filter width: CW 500 Hz (sharp passband shape, if available)

Attenuation OFF; Preamplifier OFF

Noise Blanker OFF; Noise Reduction OFF, Notch filter OFF

RF Gain MAX; AGC Medium time constant;

NOISE POWER RATIO (NPR)

Noise Power Ratio is a very effective test of overall receiver dynamic performance on a band crowded with extremely strong signals. In other words, exactly the conditions in which a receiver's strong-signal performance is most likely to be tested on the air.

The NPR test is performed by applying a band of white noise (called *noise loading*) to the receiver with a narrow notch filter. (A bandpass filter is used to limit the bandwidth of the applied white noise.) The receiver is configured to have a bandwidth narrower than the notch, then tuned to the center frequency of the notch. The output of the receiver is called *idle channel noise (ICN)*. The test is performed with and without the notch. When testing an SDR receiver, increase the noise level until the onset of clipping is reached, then reduce the noise level until no clipping indication occurs for at least 10 seconds (typically 1dB below clip level).

The theory behind the NPR test is that the noise outside the notch will cause reciprocal mixing noise and multiple IMD products, which will then appear in the receiver output (the idle channel) and raise the ICN. The test requires the receiver IF or DSP filter bandwidth to be no wider than the bottom of the notch; the filter bandwidth must not be wide enough to allow noise outside the notch to spill over into the receiver's passband.

NPR for a given noise bandwidth is the reciprocal of the ratio of the noise power in the notched band to the power in an equal bandwidth adjacent to the notch. For a given noise bandwidth:

$$NPR = P_{TOT} - BWR - MDS$$

where

P_{TOT} is the total noise power in dBm in the noise bandwidth (*noise loading level*)

$BWR = 10 \log (B_{RF} / B_{IF})$

B_{RF} = RF bandwidth or noise bandwidth in Hz of the filter applied to band-limit the noise

B_{IF} = Receiver selectivity filter bandwidth in Hz (IF for superheterodyne, DSP filter for SDR)

This can also be expressed as:

$$NPR = D_N + 10 \log B_{IF} - MDS$$

Where D_N is the noise density in dBm/Hz = $P_{TOT} - 10 \log B_{RF}$

The band-limited filter should be wider than the receiver's front-end bandwidth to ensure that the test subjects all stages of the receiver to noise loading, including any front-end filter or preselector. That ensures any effects of the noise in the front end of the receiver will be taken into account in the measurement.

In a superheterodyne receiver, the effect of the high noise power outside the notch most strongly affects the first and second mixer. The noise also mixes with the noise sidebands of the LO to cause reciprocal mixing. The noise components also mix with each other and any internally generated noise or spurious products to produce a very large number of IMD products. Some of both the reciprocal mixing noise and IMD products will fall into the receiver's passband, degrading ICN.

See the paper on Noise Power Ratio (NPR) Testing of HF Receivers by Farson in the References for a detailed analysis of NPR testing.

Typical Test Conditions:

Receiver tuned to notch filter center frequency, ±1.5 kHz

Selectivity filter width: SSB 2.4 kHz

Attenuation OFF; Preamplifier OFF

Noise Blanker OFF; Noise Reduction OFF, Notch filter OFF

RF Gain MAX; AGC Slow time constant;

25.16.4 Standard ARRL Receiver Measurements

LATENCY

Latency is the time interval between arrival of an impulsive event at the receiver RF input and the corresponding output event at the receiver audio output. Latency is measured at various detection bandwidth settings. The RF test signal is generated by a pulse-modulated or "bursty" RF signal source. A dual-channel oscilloscope is used to measure the time interval.

Various aspects of receiver design exert influence on latency. Among these are DSP processing speed and group delay across selectivity filters. Since the DSP processing speed is fixed, latency is measured for various filter configurations of bandwidth and shape factor.

To measure latency, a pulse or pulse-modulated signal generator feeds repetitive pulses via a hybrid splitter to the receiver antenna input and to Channel 1 of a dual-trace oscilloscope. Channel 2 is connected to the receiver AF output. The scope is triggered from the generator's trigger output. The time interval between the pulse displayed on Channels 1 and 2 is the receiver's latency as in **Figure 25.59**. This is *not* the same as latency of a receiver operated by remote control where the audio channel includes network delays.

Typical Test Conditions:

Receiver tuned to 3.6 MHz

Attenuation OFF; Preamplifier OFF

Noise Blanker OFF; Noise Reduction OFF, Notch filter OFF

RF Gain MAX; AGC Fast time constant

Pulse generator (if used) output 16 V_{P-P} with 60 dB attenuation between pulse generator output and receiver input, 60 ns duration, 0.2 s period

Signal generator (if used):

RF output –70 dBm

Pulse duration 200 µ

Pulse period 200 µ

FIRST IF AND IMAGE REJECTION

The first *IF rejection* and *image rejection* tests are performed on superheterodyne receivers. These tests measure 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 Figure 25.53. 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

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 Figure 25.56 except that a distortion meter is substituted for the audio voltmeter, as was done for the FM SINAD test illustrated in Figure 25.54. 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.

SQUELCH SENSITIVITY

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 OUTPUT POWER

Audio power output is tested with the same setup as the FM SINAD test illustrated in Figure 25.54. The receiver under test is set for SSB mode with the widest bandwidth

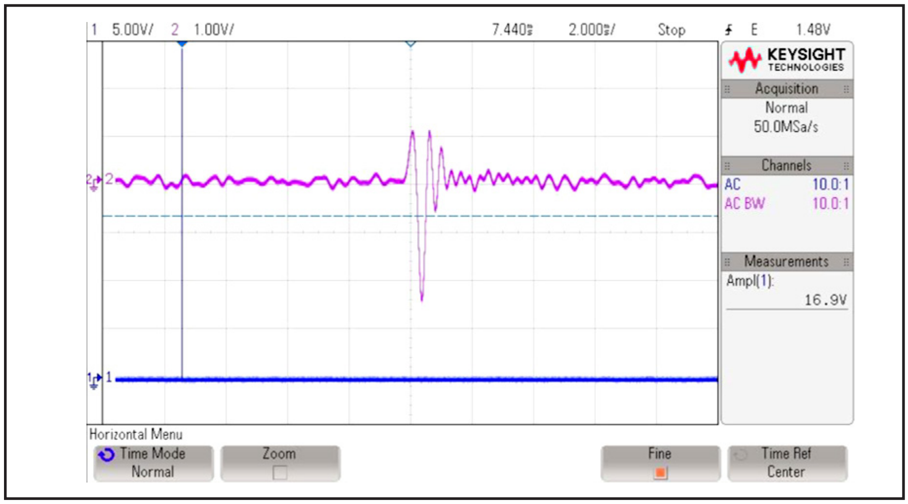


Figure 25.59 — Receiver latency is measured by inputting a very narrow pulse to the receiver (vertical line on the bottom trace with the marker on the top) and measuring the time until the response is observed in the receiver output (top trace). (Image courtesy of Adam Farson, AB4OJ/VA7OJ)

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 (1/2 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.

MISCELLANEOUS TESTS

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 Figure 25.56 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.

25.16.5 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 of 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.

25.17 Transmitter Measurements

25.17.1 RF Power Output

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. PEP can also be measured with an oscilloscope previously calibrated against a wattmeter in CW mode.

25.17.2 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 transmitter in CW mode using the test setup shown in **Figure 25.60**. 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.

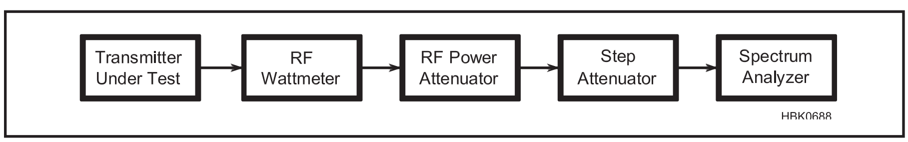


Figure 25.60 — The transmitter spectral purity test uses a spectrum analyzer to display the spurious frequencies.

Representative spectrum analyzer plots are shown in **Figure 25.61**. 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

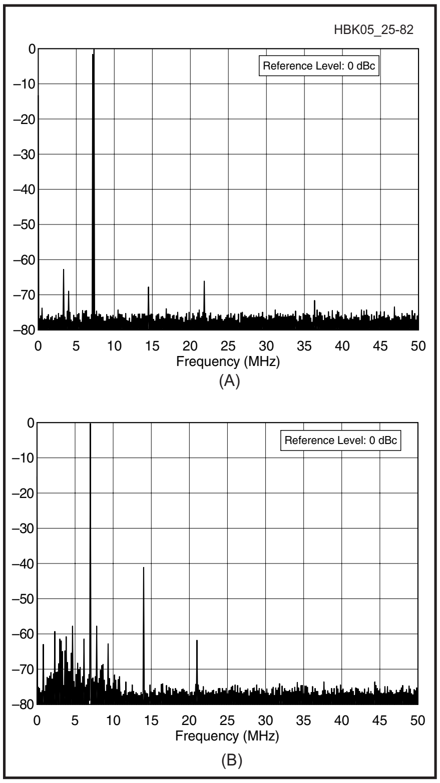


Figure 25.61 — Comparison of the spurious signal levels of two 100-watt transmitters, as shown on the spectrum analyzer display. The display on the top shows about 63 dB worst-case spurious signal suppression while the one on the bottom has a second harmonic suppressed approximately 42 dB. For transmitters below 30 MHz installed after January 1, 2003, the worst-case spurious emission must be at least 43 dB below the carrier power.

equipment and special measuring techniques as described in the Reference for Pontius. The ARRL Lab uses a special low-noise oscillator and a Hewlett-Packard (now Keysight 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 **Figure 25.62**.

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 a signal that is wider than the bandwidth of the original modulation and causes interference to other stations on nearby frequencies. For this test, the same test setup is used as for the harmonics and spurious frequencies test shown in Figure 25.60 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

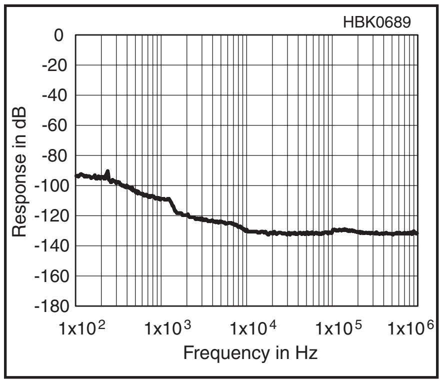


Figure 25.62 — Spectral display of an amateur transmitter output during composite-noise testing in the ARRL Lab. Power output was 200 W on the 14 MHz band. The carrier, off the left edge of the plot, is not shown. This plot shows composite transmitted noise 100 Hz to 1 MHz from the carrier on a logarithmic scale. The vertical scale is in dB with respect to the carrier.

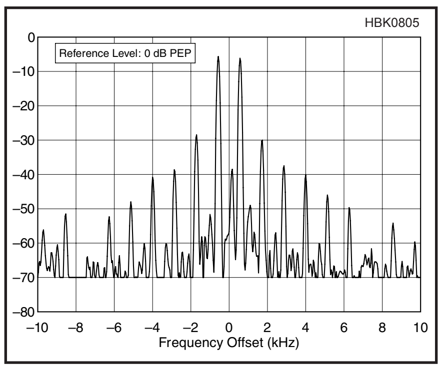


Figure 25.63 — An SSB transmitter two-tone test as seen on a spectrum analyzer. Each horizontal division represents 2 kHz and each vertical division is 10 dB. The third-order products are 30 dB below the PEP (top line), the fifth-order products are down 38 dB and seventh-order products are down 40 dB.

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 amplitudes 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 **Figure 25.63**, the IMD distortion may be read out directly. For the signal shown, the third-order products are at –30 dB from PEP.

Clean Signal Initiative

Recognizing concerns about potential interference from new amateur transmitting technology and modes, the ARRL created the Clean Signal Initiative in 2021. The goal is to provide guidance and training materials to avoid transmitting poor-quality signals that create interference to other amateurs. The initiative will engage ARRL Lab staff and volunteer technical experts to establish formalized technical standards for transmitted signal purity of amateur radio equipment, certify new equipment to these standards, and work with manufacturers to ensure compliance with these standards for the good of amateur radio worldwide. It also requires that the ARRL address the educational aspect of transmitted signal purity by developing materials to train amateurs in the proper operation of amateur radio transmitting equipment. The ARRL Lab is leading a CSI Working Group comprised of knowledgeable amateurs and industry professionals to develop these principles into an industry standard that will help improve amateur radio technology and on-the-air practices.

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.

25.17.3 Transmitted Composite Noise

As discussed in the **Transmitting** chapter, transmitters output noise along with the desired signal. *Composite noise* is a combination of phase noise and amplitude noise. (Both are discussed in the **Oscillators and Synthesizers** chapter.) The transmitted noise is a source of interference to stations operating on the same band, particularly those operating on adjacent channels. See the discussion on RMDR in the Receiver Measurements section. Wideband noise across multiple bands is also encountered.

Transmitted noise is measured using a spectrum analyzer with very narrow filter that is swept across the frequency range of the measurement and the result converted to a dBc/Hz value. If a signal source analyzer is available (such as the Rohde & Schwarz

FSUP), that is a much more convenient option, since the conversion is performed automatically. The analyzer captures phase noise as a plot of phase noise in dBc/Hz (dB below carrier in a 1 Hz bandwidth) vs. offset. A typical wideband measurement is shown in **Figure 25.64**.

If amplitude noise is to be measured, a diode detector is used. An alternative is a signal-source analyzer with phase and amplitude noise measurement capability such as the FSUP analyzer referenced in the previous paragraph. The test is made across the same spectrum as composite noise. To compute composite noise, convert both levels to absolute power/Hz, add them together, and convert back to dBc/Hz.

The transmitter output is connected to the spectrum analyzer directly through a high-power attenuator so the transmitter can be operated at full power output. An alternative is to use a dummy load and a directional coupler. Next, a swept-frequency measurement is made across the frequency range of interest. 10 Hz – 500 kHz or 10 Hz – 1 MHz offsets from the carrier frequency are typical.

Typical Test Configuration:
Frequency – each band of interest
Modulation – RTTY or another 100% duty cycle FSK/AFSK mode
Power output — full power

The plots of phase or composite noise are useful for comparing different radios and can also be used to calculate the amount of interference that may be received from a nearby transmitter. An approximation is given by

$A_{QRM} = NL + 10 \times \log(BW)$

where

A_{QRM} is the interfering signal level in dBc

NL is the noise level on the receiver frequency in dBc (from the noise plot)

BW = receiver IF bandwidth in Hz (for SDR receivers, use the receive filter bandwidth)

For instance, to find receiver interference level 20 kHz from the transmitter frequency, read the noise level (NL) on the noise plot at 20 kHz from the carrier. If NL is -90 dBc at 20 kHz from the carrier and the receiver is using a 2.5 kHz SSB filter, the interference level will be -56 dBc. If the transmitted signal is 20 dB over S9 and each S unit represents 6 dB, the interference will be (S9+20) – 56 = S9-36 dB = S3. In other words, the interference will be as strong as an S3 signal.

As of April 2022, only phase noise is measured during ARRL Lab testing for *QST* Product Reviews. The specific measurements for transmitted AM noise are being developed.

25.17.4 Latency

As in the receiver test, latency is the time interval between arrival of an event at the transmitter audio or keying input and the corresponding output event at the transmitter's RF output. Latency is measured at various transmit occupied bandwidth settings.

The audio test signal is applied from a pulse or function generator to a line input, if available, and a speech input, if not. A dual-channel oscilloscope is used to measure the time interval. The transmitter's output is connector to the oscilloscope through a high-power attenuator or via a directional coupler and dummy load. Latency should be measured at all available transmit bandwidth settings. (See the Receiver Measurements section for more information.) **Figure 25.65** shows a typical latency test result.

25.17.5 CW Keying Tests

KEYING SIDEBANDS

Keying sidebands that include both noise and components resulting from imperfections in the keying waveform can spread out for sev-

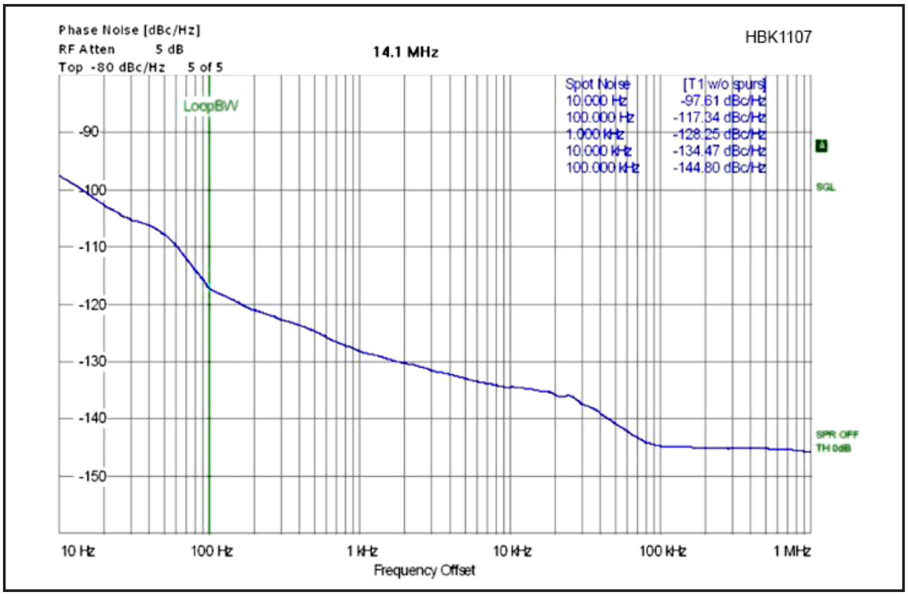


Figure 25.64 — The spectrum of transmitted phase noise from a modern transceiver operating on 20 meters. Noise is measured up to 1 MHz from the carrier signal. (Image courtesy of Adam Farson, AB4OJ/VA7OJ)

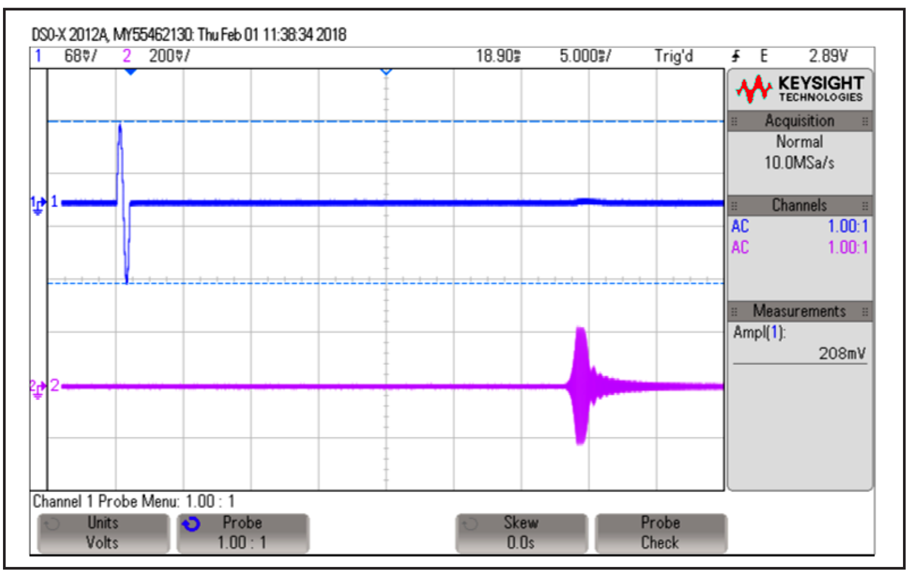


Figure 25.65 — Transmit latency is measured by inputting a very narrow pulse to the transmitter audio input (pulse waveform on the top trace with the marker on the top) and measuring the time until the response is observed in the transmitter's RF output (bottom trace). (Image courtesy of Adam Farson, AB4OJ/VA7OJ)

eral kHz around the carrier frequency. Usually referred to as *key clicks*, these sidebands can be a source of strong interference to adjacent stations.

Noise can be keyed with the CW signal as well as be present whenever the transmitter is enabled, whether it is keyed or not. For more information about bandwidth of CW signals, see the section Transmitting CW in the **Transmitting** chapter.

The spectrum of the keyed signal is measured with the same setup as for Transmitted Phase Noise. Note that the waveform rise/fall times (configurable on virtually all modern transceivers) are important parameters for the test. A typical CW keying sideband measurement is shown in **Figure 25.66**. Tests may be run with different combinations of rise and fall times, if desired.

The keying speed is set to a high speed so that the sidebands can be fully captured by the analyzer's slow sweep across the band, usually ±5 kHz from the transmitted signal using a very narrow RBW (resolution bandwidth) filter. The keying waveform is generally a continuous stream of "dits" with a 50% duty cycle.

Typical Test Configuration
Frequency — each band of interest
Modulation — CW dits (50% duty cycle) at a maximum 50 WPM
Power output — full power

Note that as keying speed begins to exceed 45 WPM, the keying waveform's necessary sidebands will begin to affect the test results. For this reason, keying speeds in excess of 50 WPM are not recommended.

KEYING WAVEFORM AND DELAY

A typical setup for measuring CW keying waveform and time delay is shown in **Figure 25.67**. 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. **Figure 25.68** 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 **Transmitting** and **Transceiver Design Topics** chapters for further discussion of CW keying issues.

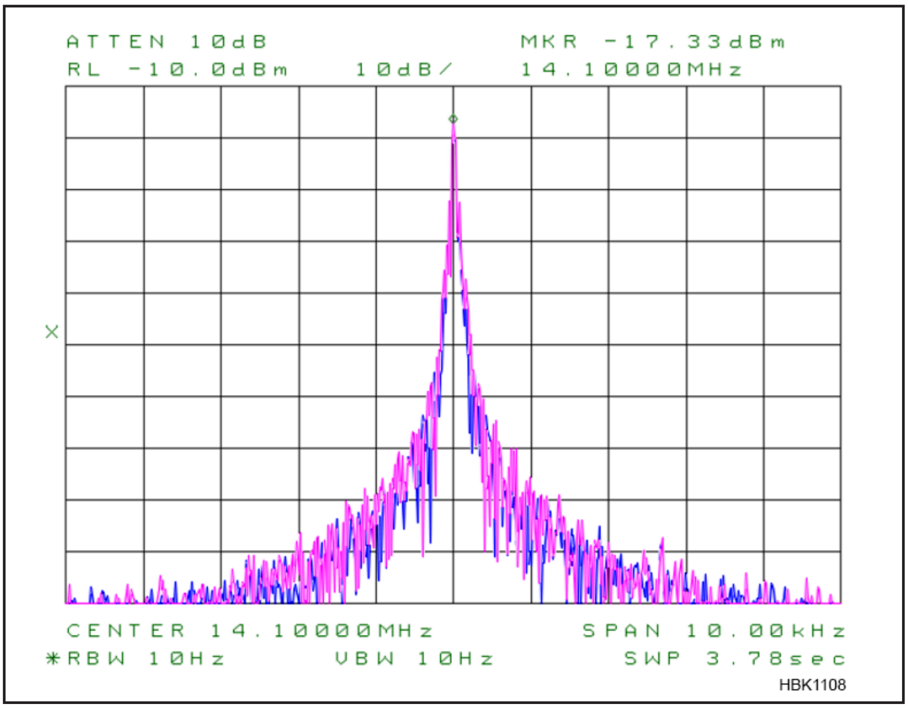


Figure 25.66 — The spectrum of a 48 WPM CW signal from a modern transceiver operating on 20 meter. The spectrum is measured up to 5 kHz above and below the carrier signal. (Image courtesy of Adam Farson, AB4OJ/VA7OJ.)

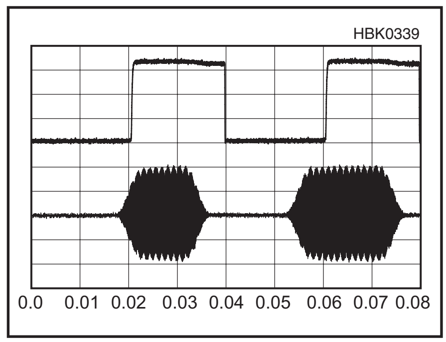


Figure 25.68 — Typical CW keying waveform for a modern amateur radio transceiver during testing in the ARRL Lab. This plot shows the first two dits in full break-in (QSK) mode using external keying. Equivalent keying speed is 60 WPM. The upper trace is the actual key closure; the lower trace is the RF 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.

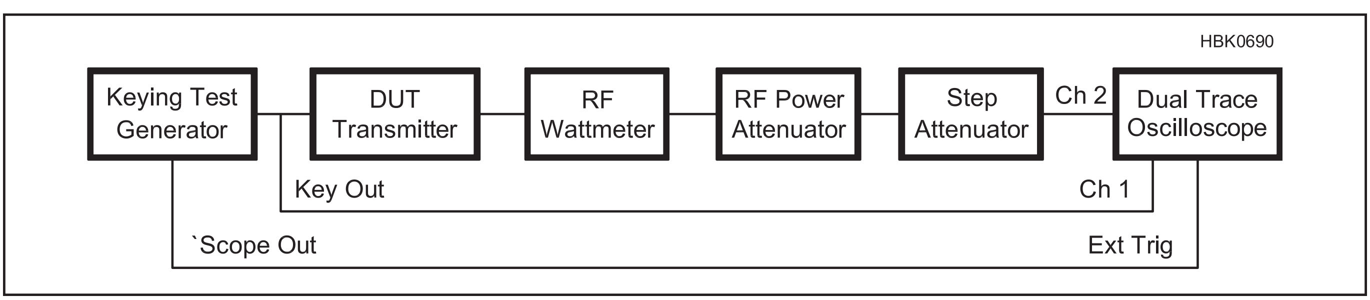


Figure 25.67 — CW keying waveform test setup.

MAXIMUM FULL BREAK-IN KEYING SPEED

Maximum QSK (CW full-break-in) keying speed is a measure of the maximum speed at which the receiver can still be heard between code elements in full-QSK CW mode. This is also a good indication of *receiver recovery time* which can be important when using fast ARQ digital modes that require messages to be acknowledged with a minimum of delay.

The test is performed by injecting a CW test signal offset from the carrier by 800 Hz into the transceiver's RF port via the coupling port of a directional coupler. This produces a continuous 800 Hz tone in the transceiver's audio output. A series of "dits" (50% duty cycle) with the standard 600 Hz offset at low power (< 10W) is transmitted into the coupler's input port. The coupler's output port is terminated in 50 Ω. The keying speed is increased until the 800 Hz tone can still just be heard in the receiver. This is the maximum usable QSK speed.

25.17.6 Transmit-Receive Turnaround Time

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 **Figure 25.69**. 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 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 *QST* Product Review tests at the ARRL Lab.

25.17.7 PTT-To-RF Output

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 **Figure 25.70** 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.

25.17.8 Other Common Transmitter Measurements

AM MODULATION PERCENTAGE

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.

FM DEVIATION

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.

The Bessel Null method of measuring FM deviation uses 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.

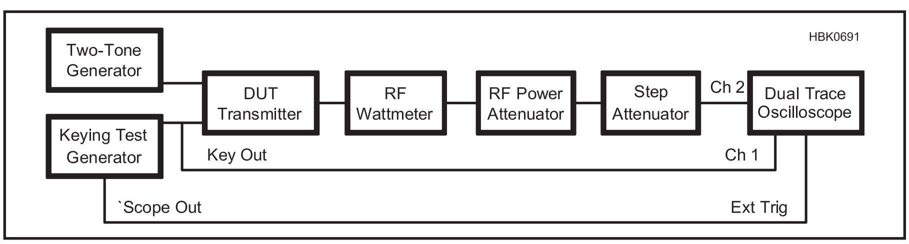


Figure 25.69 — Transmit-receive turnaround time test setup.

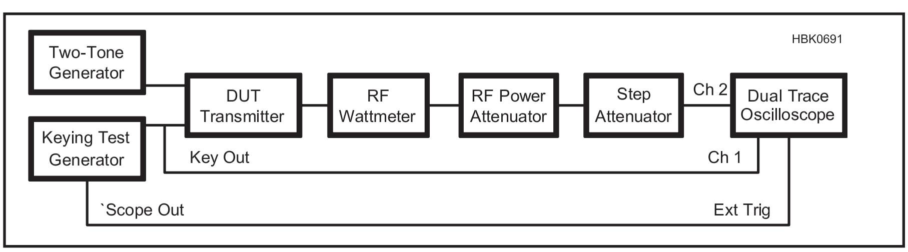


Figure 25.70 — PTT-to-RF-output test setup for voice-mode transmitters.

25.18 Construction Projects

25.18.1 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.

A circuit of a simple RC oscillator that is useful for general testing is given in **Figure 25.71**. 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 degrees. 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.

25.18.2 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 **Figure 25.72**). 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*.

25.18.3 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.

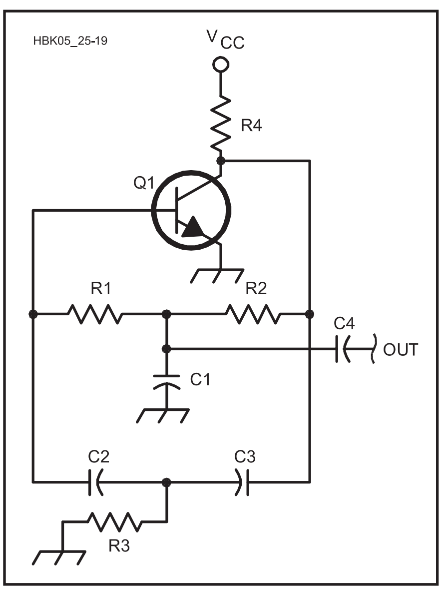


Figure 25.71 — Values for the twin-T audio oscillator circuit range from 18 kΩ for R1-R2 and 0.05 µF for C1 (750 Hz) to 15 kΩ and 0.02 µF for 1800 Hz. For the same frequency range, R3 and C2-C3 vary from 1,800 Ω and 0.02 µF to 1,500 Ω and 0.01 µF. R4 is 3,300 Ω and C4, the output coupling capacitor, can be 0.05 µF for high-impedance loads.

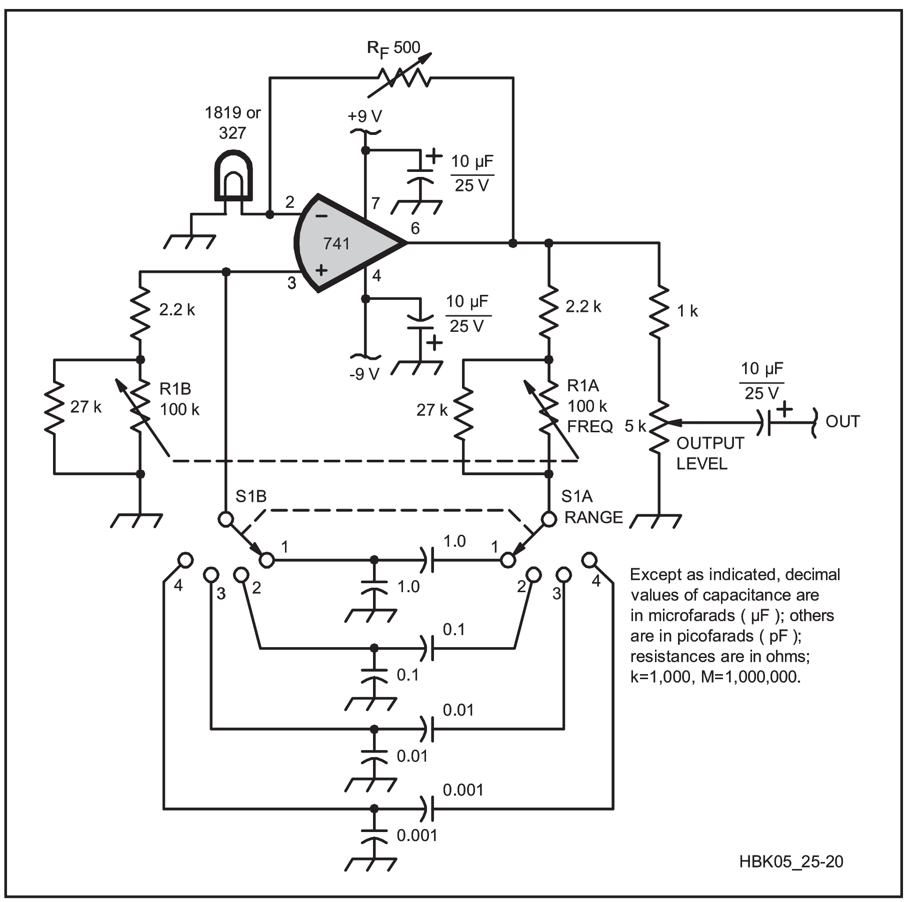


Figure 25.72 — An audio oscillator based on a single IC. The 741 op-amp is shown but most op-amps will work in this application. The frequency range is set by switch S1.

CIRCUIT DETAILS

Each of the two tones is generated by a separate Wien bridge oscillator, U1B and U2B. (See **Figure 25.73**.) 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 with the online content.) 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 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.

25.18.4 RF Current Meter

The following project was designed by Tom Rauch, W8JI (**http://w8ji.com/building_a_current_meter.htm**). The circuit of **Figure 25.74** 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 **Figure 25.75** 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 1/2-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.

25.18.5 RF Ammeters

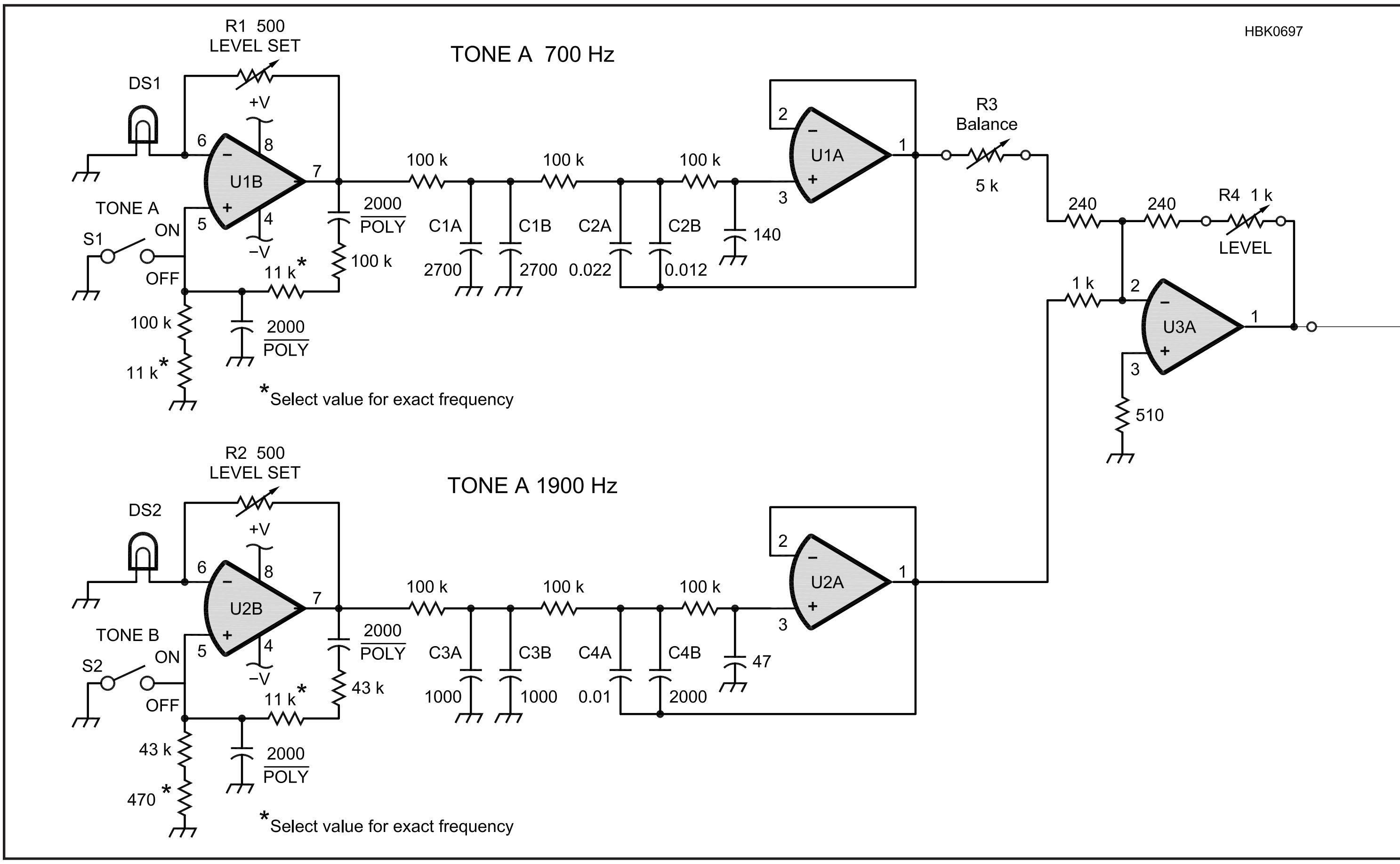
When it comes to getting your own RF ammeter, there's good news and bad news as related by John Stanley, K4ERO. First, the bad news. New RF ammeters are expensive, and even surplus pricing can vary widely between $10 and $100 in today's market. AM radio stations are the main users of new units. The FCC 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 (Sutter, F., "What, No Meters?" *QST*, Oct. 1938, p. 49) or even forward and reflected power (Wright, C., "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 (Lau, Z., "A Relative RF Ammeter for Open-Wire Lines," *QST*, Oct. 1988, pp. 15 – 17).

25.18.6 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 25.3**. 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 **Figure 25.76**. 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 25.4** 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 25.3
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.

CONSTRUCTION

The enclosure can be built using only a nibbling tool, drill press, metal shears, and a soldering gun or heavy soldering iron. (Use

Table 25.4
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

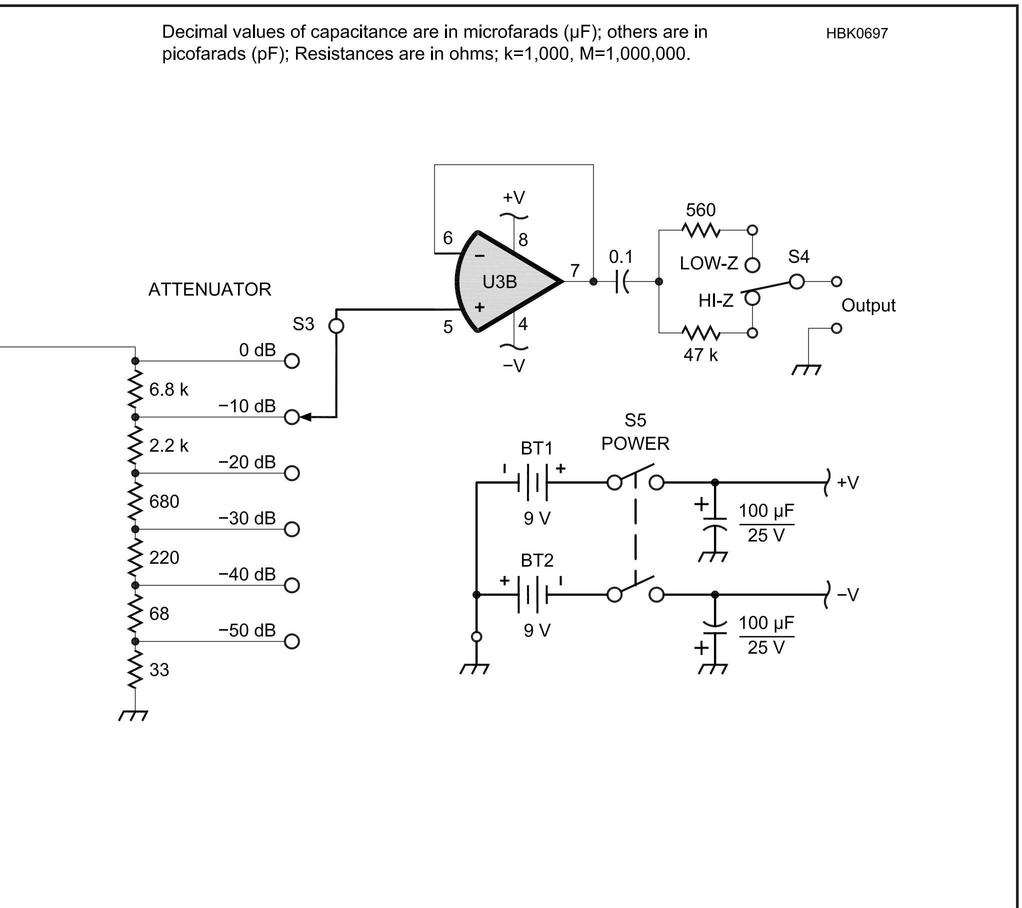


Figure 27.73 — Two-tone audio generator schematic.

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.

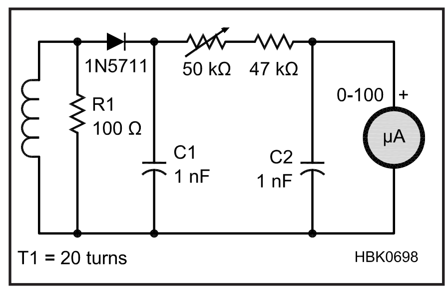


Figure 25.74 — The schematic of the RF current probe. See text for component information.

Figure 25.75 — 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.

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 1/4 × 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 **Figure 25.77**). Drill holes for the #4-40 mounting screws and nibble or punch rectangular holes for the bodies of the slide switches.

Before mounting any parts, solder in place one of the 1-inch-wide chassis side pieces to make the assembly more rigid. Solder the side piece to the edge of the top plate that faces the "through" side of the switches; this makes later assembly easier (see **Figure 25.78**). Although the BNC input and output connectors are shown mounted on the top (front) panel, better lead dress and high-frequency performance may result from mounting the connectors at the ends of the enclosure.

DPDT slide switches designed for sub-panel mounting often have mounting holes tapped for #4-40 screws. Enlarge the holes to allow a #4-40 screw to slide through. Before mounting the switches, make the "through" switch connection (see Figure 25.76) 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

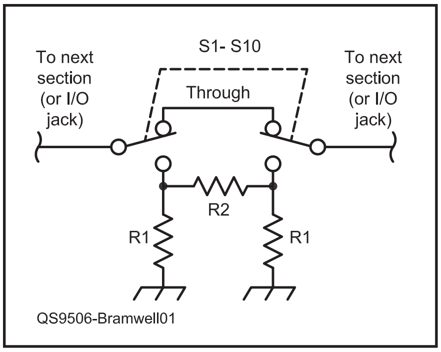


Figure 25.76 — Schematic of one section of the attenuator. All resistors are ¼-W, 1%-tolerance metal-film units. See Table 25.4 for the resistor values required for each attenuator section. There are six 10 dB sections and one each of 1, 2, 3, and 5 dB.

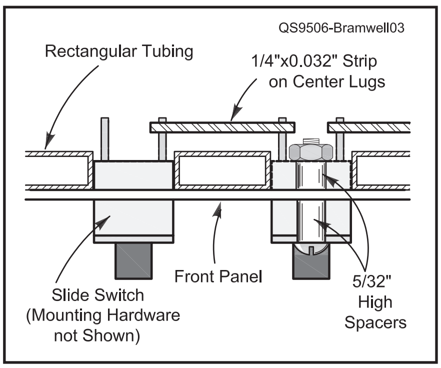


Figure 25.77 — Key to obtaining acceptable insertion loss in the "through" position is to make the whole device look as much as possible like 50-Ω coax. The rectangular tubing and the ¼ × 0.032-inch brass strip between the switch sections form a 50-Ω stripline.

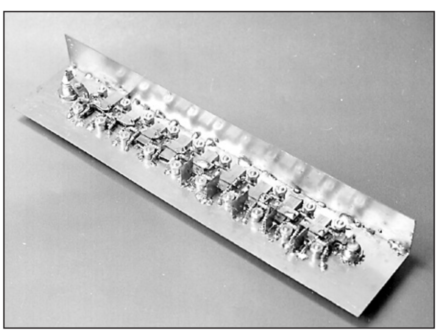


Figure 25.78 — Solder one of the 1-inch-wide chassis side pieces in place to make the assembly more rigid during construction. Solder the side piece to the edge of the top plate that faces the "through" side of the switches; this makes the rest of the assembly easier.

the ¼ × 0.032-inch brass strip come into the picture (see Figure 25.77); 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 brass shield between each 10-dB section to ensure that signals don't couple around the sections at higher frequencies.

Use parallel ¼-inch strips of 0.005-inch-thick brass spaced 0.033 inch apart to form 50-Ω feed lines from the BNC connectors to the switch contacts at each end of the stripline as shown in Figure 25.86. (Use the undrilled piece of 0.032-inch-thick brass to insure the proper line spacing.) The attenuator with all switches and shields in place is shown ready for final mechanical assembly in **Figure 25.79**.

Finally, solder in place the remaining enclosure side, cut and solder the end pieces, and solder brass #4-40 nuts to the inside walls of the case to hold the rear (or bottom) panel. Drill and attach the rear panel and round off the sharp corners to prevent scratching or cutting anyone or anything. Add stick-on feet and labels and your step attenuator of **Figure 25.80** 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.

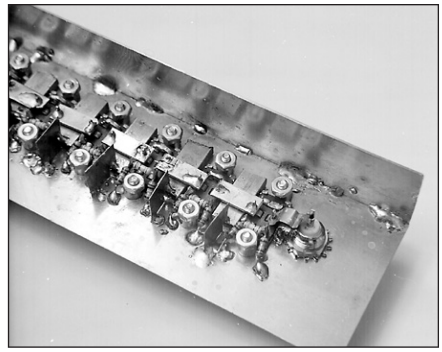


Figure 25.79 — The attenuator before final mechanical assembly. The ¼-inch strips are spaced 0.033 inch apart to form a 50-Ω connection from the BNC connector to the stripline. There are ½-inch square shields between 10-dB sections. The square shields have a notch in one corner to accommodate the end of the rectangular tubing.

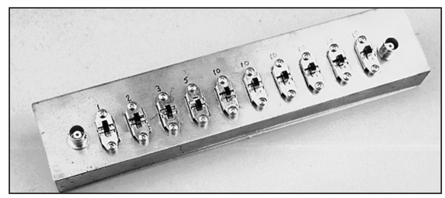


Figure 25.80 — The completed step attenuator in the enclosure of brass sheet. The BNC connectors may be mounted on the front panel at the end of

25.18.7 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

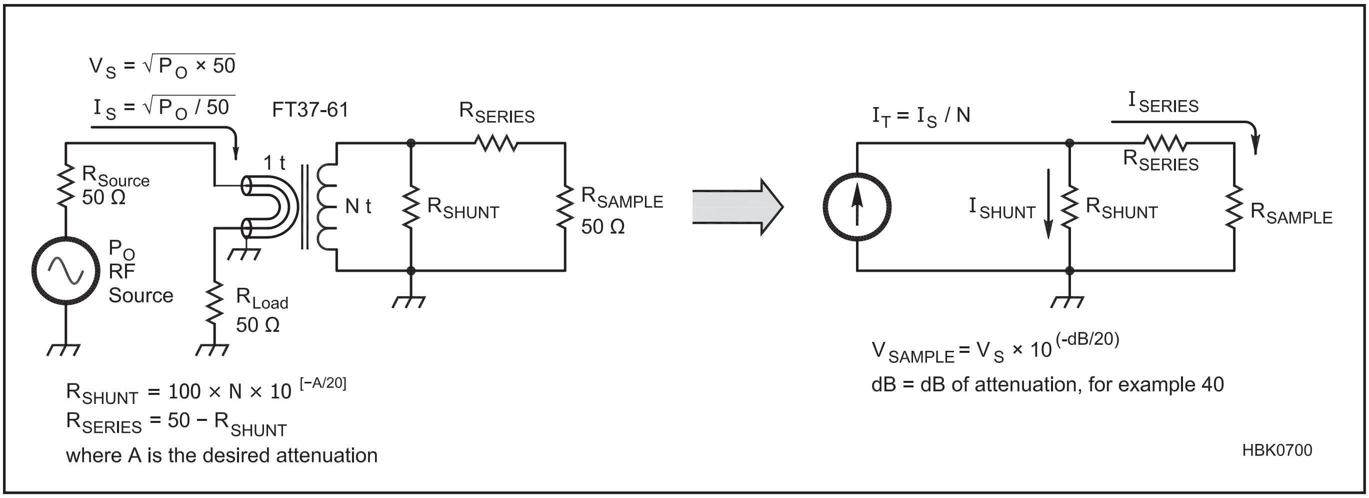


Figure 25.81 — RF sampler circuit diagram and equivalent circuit showing calculations.

conductor (usually the insulated center conductor from a piece of coaxial cable) from the transmitter or amplifier to the dummy load through a toroidal inductor forming a transformer with a single turn primary. The secondary of the transformer is connected to a resistor network and then to the test equipment as shown in **Figure 25.81**. 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. Figure 25.81 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.

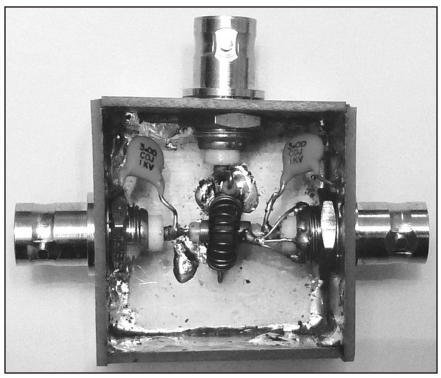


Figure 25.82 — RF sampler using box construction.

Figure 25.82 shows a photo of a sampler built into a 1.3 × 1.3 × 1 inch (inside dimensions) box constructed from single-sided circuit board material. The through-line connection is made with a short piece of UT-141 semi-rigid coax with the shield grounded only on one side to provide electrostatic shielding between the toroid and the center conductor of the coax. (Do not ground both ends of the shield or a shorted turn is created.) R_{SHUNT} is hidden under the toroid, and R_{SERIES} is shown connected to the sample port. This construction technique looks like a short piece of 200-Ω transmission line in the through-line which affects the SWR at higher frequencies. This can be corrected by compensating with two 3 pF capacitors connected to the through-line input and output connectors as shown in the photo. The through-line SWR was reduced from 1.43:1 to 1.09:1 at 180 MHz by adding the capacitors. This compensation, however, causes the attenuation to differ at high frequencies depending on the direction of the through-line connection. A sampler constructed using the box technique is useable from below 1 MHz through 30 MHz.

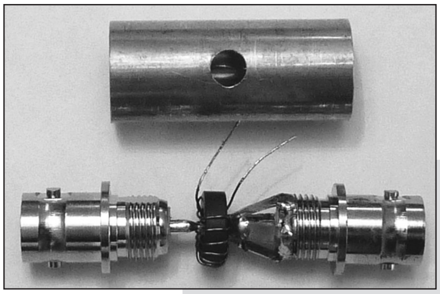
Figure 25.83 shows a different approach using 9⁄16 inch diameter, 0.014 inch wall thickness, hobby brass tubing. This lowers the impedance of the through-line so that no compensation is needed. The through-line SWR for the tube sampler is 1.08:1 at 180 MHz which is as good as the box sampler and the sensitivity to through-line direction is reduced. Although the high frequency attenuation is not as good as the box sampler, the construction technique provides a more consistent result. A sampler constructed using the tube technique should be 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, noninductive metal-oxide resistor and R_{SERIES} is a 34.8 Ω, ¼ W, 1% noninductive metal-film resistor. The power dissipation of the resistors and the flux handling capability of the ferrite core are adequate for sampling a 1,500-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 with the online content.

25.18.8 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



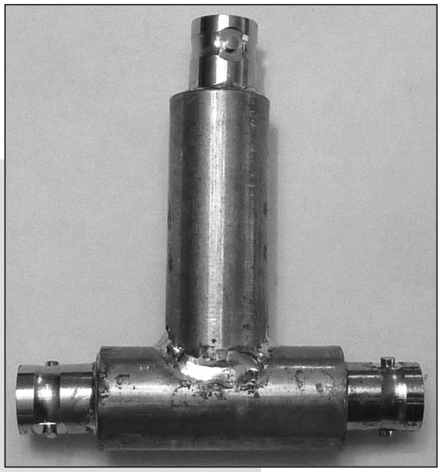


Figure 25.83 — RF sampler using tube construction.

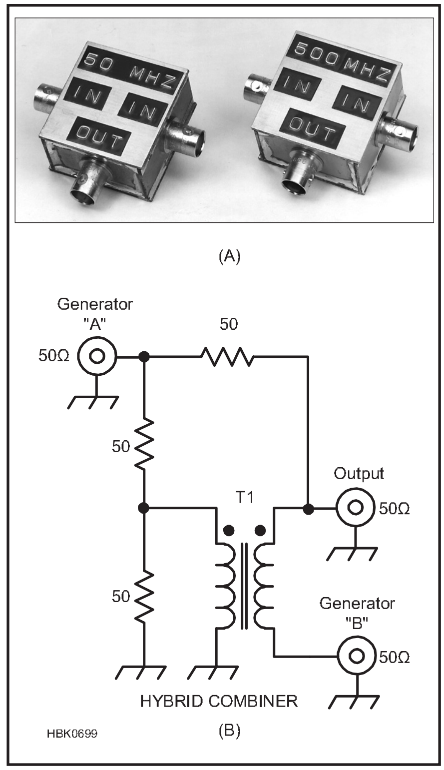


Figure 25.84 —The hybrid combiner on the left of A is designed to cover the 1 to 50-MHz range; the one on the right 50 to 500 MHz. B shows the circuit diagram of the hybrid combiner. Transformer T1 is wound with 10 bifilar turns of #30 AWG enameled wire. For the 1 to 50-MHz model, T1 is an FT-23-77 ferrite 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.

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 examples 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.

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.

25.18.9 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 **Figure 25.84**) 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 Figure 25.84A. 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 Figure 25.84B

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 with the online content. Return loss is discussed in the **RF Techniques** chapter.

25.19 References

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