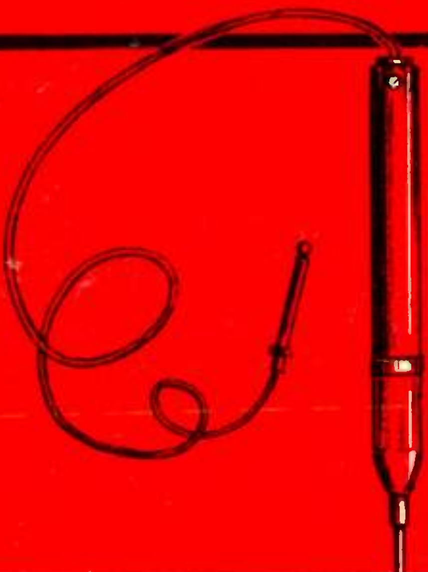
AMATEUR TESTS
and
MEASUREMENTS



Louis M. Dezettel, W5REZ

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Preface

What does the amateur need to know about his equipment? He needs to know whether or not he is within the law as to power input to the final of his transmitter, and that he is operating within the frequency limits of the bands assigned to amateur operation. To do so requires the use of a few basic instruments and a fundamental knowledge of checking transmitter performance. The knowledge is probably no greater than that needed to pass the license exams in the first place.

What does the amateur want to know? He wants to know what the sensitivity and selectivity of his receiver are, and if they are as good as can be obtained. He wants to know how to align his receiver for top performance. He wants to know how to get the most out of his transmitter, what the modulation capabilities are, and how to load and modulate his ssb rig for best results. He wants to know if his antenna is resonant, and how to measure for optimum match between his transmitter and the antenna.

The QSO's on the air are proof of the average amateur's pride in the quality of performance of his rig. This book is devoted to contributing to that pride, in showing amateurs how to measure the performance of their rigs, and how to get the most out of them, with the highest quality of performance.

Stress is placed on making measurements and tests with inexpensive equipment. Most of the equipment described in this book should be a part of every progressive amateur's station list. The pieces used are probably already parts of more than half the amateur stations now in existence. Many of them are easily home-built, and construction of these is described.

This book begins with fundamental a-c and d-c measurements, and includes descriptions of the basic instruments used to make the measurements. One feature is how to check tubes and transistors while they are in the circuit, based on current and voltage measurements during actual operation. The section on transistor checking is followed by a description of a transistor checker you can build yourself.

The chapters on receiver and transmitter alignment and calibration are fairly complete, but stop short of requiring very expensive laboratory-type equipment. Oscilloscope patterns are shown in more than one place in the book for checking the performance of a transmitter, especially an ssb transmitter.

Adjusting the antenna to resonance, and matching its impedance, is probably the most important thing an amateur can do to get increased performance from his transmitter. The antenna chapter shows you how to do these, both at the antenna and inside the ham shack. This book is worth owning even if only for the chapter on antenna adjustments using simple equipment.

LOUIS M. DEZETTEL, W5REZ

Contents

CHAPTER 1

D-C MEASUREMENTS 7

Sensitivity—VOM Versus VTVM—Solid-State VTVM—Instrument Accuracy—Loading Effect—Practical D-C Measurements—Vacuum-Tube Amplifiers—Transistor Circuits—Transistor Oscillators—Batteries—Power Supply

CHAPTER 2

A-C MEASUREMENTS 23

Meter Rectifiers—Oscilloscopes—Measuring Frequency—The Grid-Dip Oscillator—Alternating-Current Measurements—The Signal Generator—The Audio Generator—Checking Capacitors and Inductors—Power Transformers—Audio Transformers—Measuring Capacitors—Measuring Inductances—Headphone and Speaker Impedance—Measuring *Q*—R-F and I-F Transformers—Frequency Measurements—Filters

CHAPTER 3

TUBE AND TRANSISTOR TESTING 57

Transconductance Tube Checkers—In-Circuit Tube Testing—Semiconductor Testing—Checking Diodes and Transistors Out of Circuit—Transistors—Diode Measurements—Zener- and Signal-Diode Measurements—Transistor Measurements—Collector Characteristics Curve—Field-Effect Transistors—Handling FET's—A Transistor Tester—Using The Tester—Transistor Manuals

CHAPTER 4

RECEIVER MEASUREMENTS 78

Measuring Instruments—Measuring Sensitivity—Sensitivity Hookup—Stage-by-Stage Gain—Measuring Selectivity—The Selectivity Hookup The Sweep Generator for Selectivity—Reading The Scope—The

Sweep Hookup—Adding Markers—Measuring Image Response—Noise Figure—Input Impedance—Receiver Alignment—A-M and C-W Receivers—SSB Receivers—Sweep Generator—BFO Adjustment—SSB Carrier-Insertion Adjustment—Receiver Calibration—Front-End Alignment—Using The Noise Generator—High-Priced Receivers

CHAPTER 5

TRANSMITTER MEASUREMENTS 117

Definitions of Power—A-C Power Out—Connecting an Oscilloscope What You See—Power Efficiency—Measuring Power Out—The Dummy Load—Loading The Transmitter—Modulation Measurements—C-W Keying—SSB Modulation Percent—Transmitter Alignment—VFO Calibration—SSB Carrier Adjustment—Neutralizing—Parasitics—Class-C Operation—Class-B Linearity

CHAPTER 6

ANTENNAS AND FEEDERS 148

Tuning The Antenna—The Half-Wave Feeder—How To Cut Half-Wave Line—Measuring Antenna Impedance—About SWR—The Value of Low SWR—Tuned Lines—Tuning A Vertical—About Vertical Antenna Ground—The SWR Meter—Using The Sweep Generator for Antenna Resonance—Front-to-Back Antenna Ratio—Matching Transmission Line to Antenna—The Gamma Match—The Q-Bar—Matching Stubs—Antenna Traps

CHAPTER 7

INSTRUMENTS EVERY AMATEUR SHOULD HAVE 178

VOM—VTVM—SWR Meter—Building The SWR Meter—Accuracy—Using The SWR Meter—Impedance Bridge—Construction—Using The Impedance Bridge—100-kHz Crystal Calibrator—Grid-Dip Oscillator—Semiconductor GDO's—Monitoring Scope—Coupling RF—Audio Phase—Observing Patterns—SSB Patterns

INDEX 207

CHAPTER 1

D-C Measurements

Measurements of direct current (dc) are made with a VOM or a VTVM. VOM stands for volt-ohm-milliammeter, an instrument which performs the function of measuring volts (both d-c and a-c), resistance, and current. A switch not only selects the functions of the VOM, but selects the ranges of each function (Fig. 1-1). The basic meter movement deflects a needle mounted on a coil, the needle moving a distance proportional to the current passing through the coil. High-value resistors in series with a basic meter movement establish different ranges for reading volts. Shunt resistors bypass some of the current so that the instrument can indicate more current than the basic meter movement can take. Batteries and resistors in series set up the circuit for reading the value of an external resistor.

SENSITIVITY

When making voltage measurements in high-resistance circuits, such as at the plate of a vacuum-tube audio amplifier, it is important to draw the least amount of current to deflect the needle on the meter. Otherwise, the voltage read will be inaccurate due to the extra load.

If a 1-mA basic meter movement is used in a VOM, a 100,000-ohm multiplier (series) resistor will cause the meter to be deflected to full scale with 100 volts applied to the VOM test leads (OHM's law tells us that 100 volts through a 100,000-ohm resistor produces a current of 1 milliampere). An instrument with a 1-mA meter movement is called a 1000 ohms-per-volt VOM. Better VOM's are 5000 ohms/volt and 20,000 ohms/volt. To keep the loading effect of the instrument down, and for working with transistors, the amateur should seriously consider buying the 20,000 ohms/volt VOM es-

Courtesy Triplett Electrical Instrument Co.

Fig. 1-1. The Triplett Model 630 VOM is typical of many of the better 20,000 ohms/volt VOM's.

pecially since some inexpensive imports are now available with this sensitivity.

VOM VERSUS VTVM

Where high sensitivity is important the logical instrument is the VTVM (and the solid-state equivalent). VTVM stands for vacuum-tube voltmeter. Most of them have a constant input resistance of 11 megohms on all ranges. There is hardly a circuit in which an 11 megohm load will alter the readings to any appreciable degree (Fig. 1-2).

The VTVM uses two triodes in a bridge circuit with the meter movement connected between the cathodes of the two triodes. The grid of one triode is connected to ground, and the grid of the other triode connects to a bank of voltage-divider resistors at the input. The string of voltage-divider resistors in series totals 11 megohms. The range switch selects a portion of the voltage on the string, the position depending on the range. Any voltage above ground results in the cathode current in one triode being different from the cathode current in the other triode, with a consequent voltage difference between the two cathodes. This upsets the balance and causes the meter to indicate.

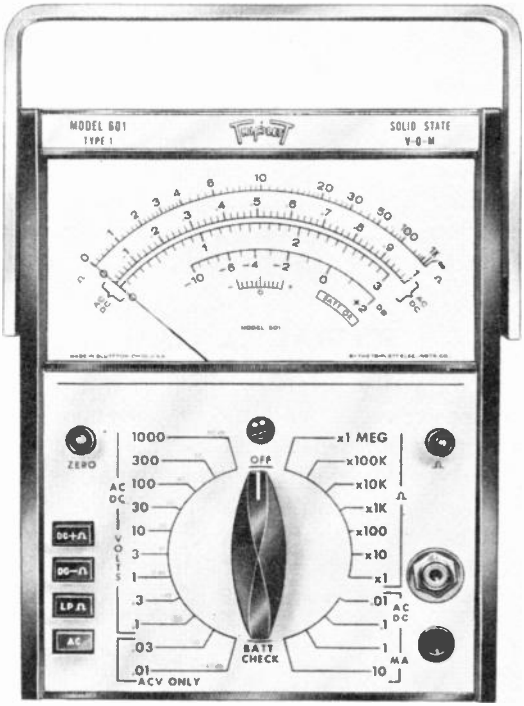
The resistor bank in the instrument actually totals 10 megohms. There is a 1-megohm resistor in the tip of the d-c volts probe to make a total of 11 megohms. The 1-megohm resistor in the probe isolates the test-probe cable capacitance (a shielded cable is used to keep stray ac from affecting the reading) and the instrument itself from the circuit under test. A VTVM has no current-reading function.

SOLID-STATE VTVM

The standard VTVM depends on the 120-volt a-c supply for its primary operating voltage. With the improvements in transistors (especially the high-impedance FET's), the selection of battery-operated VTVM-type instruments is increasing every day. Most have a VTVM 11-megohm input resistance, like that of the a-c operated VTVM, and other functions are very similar (Fig. 1-3). The solid-state instruments employ two or four transistors in a bridge circuit. Portability is the principal advantage of the solid-state instrument.

Courtesy RCA

Fig. 1-2. One of the most popular of standard VTVM's is the RCA *Volt-Ohmyst*®. This Model WV-77E is also available in kit form.



Courtesy Triplett Electrical Instrument Co.

Fig. 1-3. The Triplett Model 601 has functions and sensitivity like a standard a-c VTVM, but is battery powered. Because it also includes a current-reading function, it is called a solid-state VOM.

INSTRUMENT ACCURACY

Accuracy depends on two things: the quality of the meter movement, and the tolerance of the resistors. Typical accuracy specifications on the better instruments are 2 percent on dc and 3 percent on ac. The 2-percent accuracy on dc will usually mean the manufacturer is using 1-percent resistors and a 1-percent meter movement. Incidentally, accuracy figures on meter movements means *accuracy of full scale*. That is, the 500-volt scale may have a 5-volt error at any part of the scale. You can see that the same 5 volts at the low end of the scale can be a high percentage of the figure being read. The change in the accuracy figure of a-c means the rectifier has some tolerance of its own and adds to instrument inaccuracy.

LOADING EFFECT

If the source of the voltage being measured has high resistance in it, a VOM can show an error in its reading because it represents an additional load at the point of measurement. Fig. 1-4 is the schematic

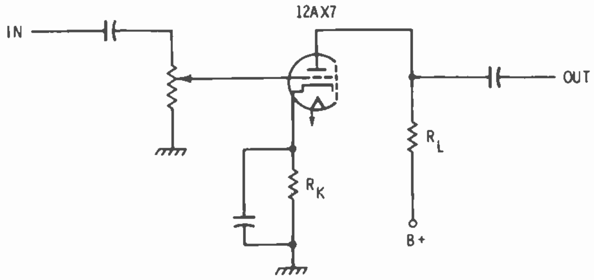


Fig. 1-4. A standard audio-voltage amplifier using one triode of a 12AX7.

of a triode audio-voltage amplifier. The plate-load resistor (R_L) is 100,000 ohms. A 1000-ohms-per-volt VOM set for 100 volts is connected to the plate for a voltage measurement. On the 100-volt range, the total series resistance in the instrument is 100,000 ohms, as explained previously. Assume a 100-volt power-supply source. If the tube were pulled out, there would be no current drawn through the plate-load resistor, and the plate side should have a voltage of 100 the same as the supply voltage. But if you connect the VOM to the plate terminal, it will read only 50 volts, because its internal 100,000 ohms is in series with the 100,000 ohms of the plate-load resistor, and at the test point the voltage is one half because there is now current drawn. The current through the meter is now 100 volts divided by 200,000 ohms (100,000 + 100,000) or 0.5 mA. This current through the plate-load resistor equals 50 volts and that is what your meter will indicate. Now, reinsert the tube. It, too, draws current and will look like a resistor, probably about 100,000 ohms. With the tube in the circuit the plate voltage should read 50 volts, but as soon as you touch the VOM probe to the plate you have shunted the tube with an additional 100,000 ohms, so it will look like a 50,000 ohm resistance (two 100,000- ohm resistors in parallel) in series with the plate-load resistor. The total resistance is now 150,000 ohms and the voltage across the 100,000-ohm plate resistor will indicate only ⅔ of the 50 volts actually there, and this is a 33⅓-percent error.

A VTVM, on the other hand, is an 11-megohm load, which reduces the effective resistance by only about 1-percent and results in a 1-percent error in measurement at this point. Go through the mathematics as we did for the VOM, and you will see the difference.

Wherever high-value resistors are used in circuits, a high-impedance VTVM will reduce the error in the measurement of direct current. In addition to the plate circuit of a vacuum-tube audio voltage amplifier, there may be occasions for measuring at the grid. Some audio amplifiers, although not usually in ham equipment, use what is called "contact-potential" biasing for the vacuum-tube amplifier (Fig. 1-5). In

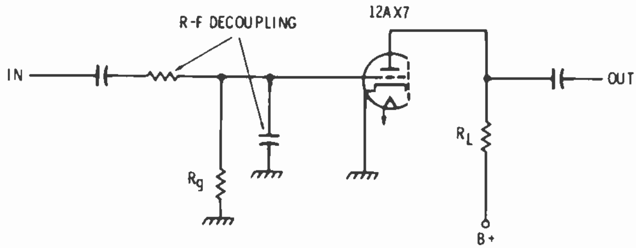


Fig. 1-5. Audio amplifier using contact-potential biasing, created by the space charge between grid and cathode. A very high value for R_G is required.

these circuits the cathode is grounded and the grid is returned to ground through a 5- or 10-megohm resistor. Electron space charge in the tube creates a bias. Even the VTVM does not measure this voltage accurately, but still will measure it much more accurately than will a VOM, which becomes practically a short circuit for this bias arrangement.

PRACTICAL D-C MEASUREMENTS

Ham equipment usually comes with an operating manual which also includes a chart of voltages for various terminals of the tubes, if tubes are used. The sensitivity of the measuring instrument is usually indicated. Generally it is assumed a 20,000-ohms/volt VOM is used. The readings given are for voltages with respect to ground, and may be positive (+) or negative (−) dc with a-c readings for the heaters, and for the plates of a tube rectifier if one is used. If your voltage readings agree with the chart, it is assumed that all tubes are functioning properly. This is one of the best methods of servicing a receiver or transmitter.

Solid-state equipment is more difficult to check because of the sizes of components. If the transistors are wired directly into a printed-circuit board, as they frequently are, the job of tracing transistor terminals on the underside of the board is very tedious. Many pieces of solid-state equipment include test points along the circuit. These are short stubs of heavy wire sticking through the topside of the printed-circuit board identified by printed letters on the board. Many of the test points are there for alignment purposes as well as for voltage check points.

VACUUM-TUBE AMPLIFIERS

Even without a voltage chart it is easy to determine the performance of the active components of ham equipment (active components are tubes, transistors, and relays; passive components are resistors, capacitors, etc.). Comparing voltage readings with specs in a tube or transistor manual can tell you whether such active components are performing as they should.

Fig. 1-6 is the schematic for a popular output stage in a receiver. It uses a single 6K6GT tube, capable of an output of about 3 to 4 watts. A typical voltage supply value to the tube is 250 volts. This would be read at point #2. R_3 is a cathode-biasing resistor. The voltage drop across it is the bias voltage between cathode and grid. A typical value for this resistor is between 500 and 600 ohms, probably around 550 ohms. With the black lead connected to the chassis, touch the red (+) lead to terminal 8 of the tube socket. A reading of + 18 volts is about right. Ohm's law tells us a current of about 33 mA will

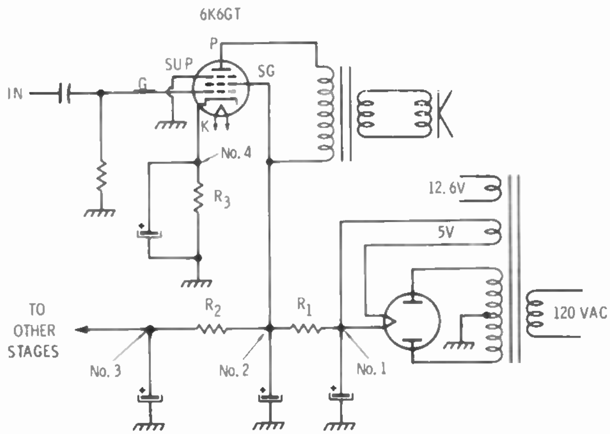


Fig. 1-6. Typical output stage and power supply as used in some amateur receivers. Symbols and test-point numbers are described in the text.

produce an 18-volt reading across a 550-ohm resistor. According to the tube manual this is about right. The tube is working, that is, and there is emission of the prescribed amount for this tube in this circuit. It is assumed that you know the current in the cathode circuit equals the current in the plate and screen circuits combined. If no voltage was indicated across the cathode resistor, it would mean that the tube had lost its emission. If less than the right amount is indicated, the tube is getting soft and should be replaced. Other information from this circuit is the total current drawn by all stages in the receiver. All the current goes through the filter resistor (R_1). The manual will give you the value, or you read the color code, or measure the resistance with the ohmmeter. Measure the voltage across the resistor, either by placing the positive lead of the voltmeter at terminal #1 and the negative lead at terminal #2, or by grounding the negative lead and measuring the voltage at both terminals #1 and #2 and take the difference. The voltage divided by the resistance is the current. As a rule the highest current is drawn by the output stage. To measure for the current to all other stages, measure the voltage across R_2, and divide by the resistance.

Another version of the output-stage circuit is shown in Fig. 1-7. This is sometimes seen in transceivers because of the availability of bias for the transmitter tubes. Plate current can be measured by measuring the voltage *across* the primary of the output transformer with power on and then measuring the resistance with the power off. Both the voltage and resistance will be very low because you are measuring the resistance of the wire in the primary. While it is much more accurate to measure across the primary rather than between each side and ground, be careful to place the VOM out of the way so you will not touch any part of exposed metal such as other terminals. They may be "hot" with respect to the case of the receiver or transformer. Since

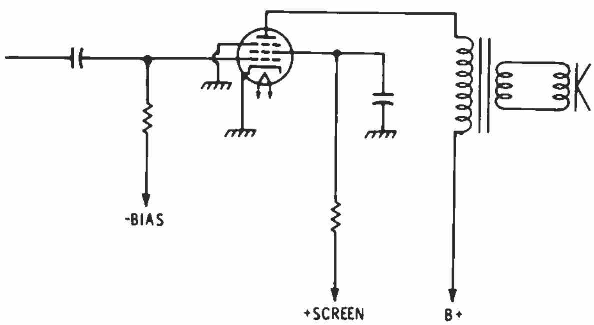


Fig. 1-7. An audio output stage using fixed bias.

fixed bias is used, bias measurement is directly from grid to chassis ground. A bias measurement will be the same whether the tube is OK or not. You will have to depend on the plate-current measurement described for a check on performance. If you are checking the receive functions of the tubes in a transceiver, remember to have the transceiver in the *receive* mode. When the transceiver is in the *transmit* mode, high bias is applied to some tubes in the receiver to kill their amplification while transmitting.

There are two popular forms of low-level audio amplifiers. The most used is the one shown in the schematic of Fig. 1-4. The high-μ twin triode 12AX7 is frequently the tube seen in this circuit. The tube manual will probably refer to the 6AV6 for characteristics because the triode in this triode twin-diode tube is the same as the two triodes in the 12AX7. The plate supply voltage to the tube is most likely 100 volts after the original voltage is dropped by a decoupling resistor. At 100 volts, the right grid bias is -1 volt. Measure across the cathode resistor again (R_k), in the same way as the first example of the 6K6GT. And from this you can calculate current in the plate circuit by dividing the voltage by the resistance. The tube manual says it should be 0.5 mA. Measuring the voltage drop across the plate-load resistor and dividing by the resistance should give you the same answer, but here is where you must be careful as to the sensitivity of the instrument. It takes a VTVM or a FET voltmeter to give you any reasonable accuracy. A 20,000 ohms/volt VOM on the 100-volt scale

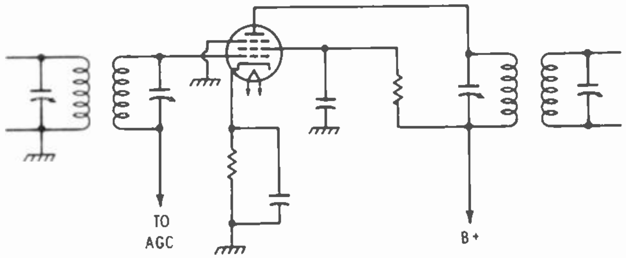


Fig. 1-8. R-f and i-f amplifiers are like audio amplifiers except for the tuned circuits used at input and output. Fixed bias is from the cathode resistor. Additional bias from a diode detector varies the gain for agc.

will have a total resistance of 1 megohm. If the plate-load resistor is 100,000 ohms, as it sometimes is, the error introduced by the instrument is 10 percent.

Fig. 1-5 is the schematic of an audio amplifier often used as the microphone first stage amplifier in a transmitter, because the grounded cathode reduces hum and r-f pickup. Bias is obtained by the space charge of electrons gathering around the cathode and grid. The *only* way to measure bias is with a VTVM between grid and ground, and even then there will be about a 20-percent error It will be more accurate to check for proper plate current by measuring the voltage across the plate-load resistor, and dividing by the value of the resistor. If the answer is about 0.5 mA, you will know the bias is right.

R-f and i-f vacuum-tube amplifiers (Fig. 1-8) will also usually have cathode resistors for fixed bias. Their performance is checked by measuring the voltage across the cathode resistor as in the case of the audio amplifier. Actual bias values in i-f amplifiers are usually not the same as they are for audio applications, since there is no attempt to operate the tubes as class-A amplifiers because there is no need to. Almost any low value of bias voltage is considered OK. When checking the bias on r-f or i-f tubes, the r-f gain control must be at maximum, and there must be no signal received, otherwise the agc voltage will reduce the plate current much below that indicated by the tube manuals.

Buffer and driver stages in a transmitter are measured in the same way as r-f and i-f amplifiers in a receiver, except they often have a fixed bias on the grid. Also power-type tubes are used, so currents will run more like the audio output stage mentioned earlier. Sometimes they have a cathode resistor for protective bias and sometimes they do not. About the only way some can be checked for emission while in the chassis is to measure across the screen dropping resistor, (which most will have) and check with the tube manuals to compare screen current. A typical example is a 100-ohm decoupling resistor in the screen of a 6GK6 driver tube fed from a 275-volt supply. The voltage across this resistor measured only 0.5 volt, but Ohm's law gives us 5 mA of current which is about right, and the tube is drawing about the current it should.

An oscillator is oscillating when there is d-c voltage between grid and ground. The value varies considerably and will usually be somewhere between −5 and −10 volts. This is grid bias due to the oscillation. If there is no bias there is no oscillation.

Nearly all output r-f power stages have self-bias as well as fixed bias. In case of excitation and fixed-bias failure, the self-bias resistor (R_k) protects the tube or tubes from excessive current. Fig. 1-9 is similar to many used in medium-power transceivers today. Because of high voltage on the output-stage tubes, they are usually enclosed in a separate compartment, and this means *CAUTION* to you. When measuring voltage at the plate or plate supply, use only good-condition leads that are dry and clean of grease. Ground the negative lead

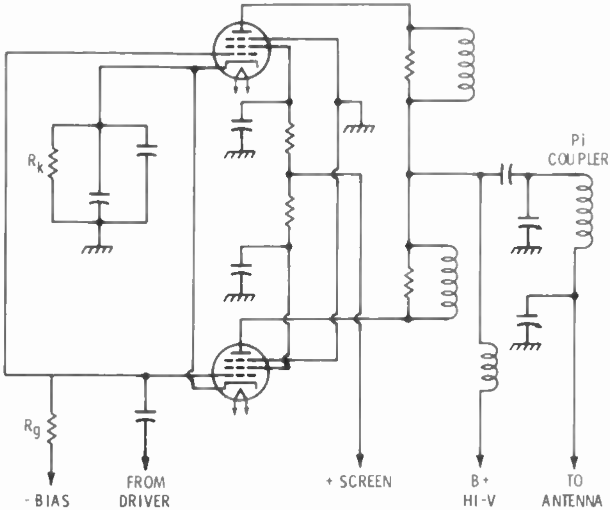


Fig. 1-9. Two tubes in parallel in an r-f output stage. R_K is a cathode resistor for self-bias to limit plate current.

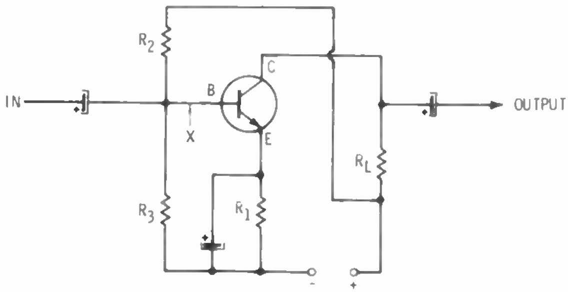


Fig. 1-10. Most frequently used transistor amplifier circuit as a low-level stage. The resistor symbols are explained in the text.

to the case or chassis, and *place one hand behind your back* when you apply the positive probe to the plate or plate supply. Be sure your meter is set on a range high enough to measure the expected voltage. Again current is measured by measuring the voltage across the cathode resistor and using Ohm's law. To check for balance between the two tubes in a parallel-connected output stage (or push-pull) measure the voltage drop across each screen-decoupling resistor.

TRANSISTOR CIRCUITS

There are two basic differences between vacuum tubes and transistors, and these differences affect the measurements made on transistors. Transistors are *current-amplifying* devices; the current in the collector circuit is controlled by the current in the base circuit. The second difference is that *no current flows in the collector until the base is forward-biased.* There is an important exception, and that is the field-effect transistor (FET). This transistor has all the attributes of a vacuum tube.

In addition to the above, the conventional transistor is a low-impedance device. A low, low-voltage full-scale range is desirable. High sensitivity as a milliammeter is highly desirable since base currents in a transistor are frequently in the microampere range.

The schematic of Fig. 1-10 shows the basic circuit of most amplifiers using a conventional transistor. While an audio amplifier is shown, the same method of biasing is often used for r-f and i-f amplifiers. In an r-f or i-f amplifier, the load resistor (R_L) would be replaced by the primary of a transformer, and the capacitors would not be electrolytics.

Note the use of both fixed and automatic bias. R_2 and R_3 form a voltage divider for fixed bias to the base of the transistor. R_1 carries current from both the collector and base, and so serves to create automatic bias, the same as a cathode resistor on a vacuum tube.

It might be well to review for a moment the reason for the double method of biasing a transistor. The voltage divider supplies the forward bias needed to establish the right amount of current flow in the collector circuit. In this circuit the voltage at point X (the base) will be positive (+) with respect to the emitter. When current flows in a transistor, heat is developed internally. As the transistor warms up, current increases. This could continue until the transistor is destroyed if some method to stop it were not used. This is called *thermal runaway.* Resistor R_1 in the emitter circuit develops a "bucking" bias voltage. It is in opposition to the fixed bias of the voltage divider. Since current tends to increase due to heat within the transistor, the bucking bias increases and reduces the forward bias from the voltage divider, which, in turn, tends to reduce current through the transistor. When the right values for R_1, R_2, and R_3 are selected, the current through the transistor is stabilized at the correct value.

Collector *current* is measured by measuring the voltage across load resistor R_L and dividing the voltage by the resistance. Bias *voltage* is measured between base and emitter. The specs in a transistor manual give base current, which can best be measured by breaking the connection to the base at point X and inserting the milliammeter function of the VOM in series with the base. This is almost impossible to do in manufactured equipment. This method should be used in experimental breadboard setups for determining the right values of resistors for biasing. For measuring bias current in commercial equipment, find the collector current and emitter current from the voltages across R_L and R_L. The difference between the two is base current.

TRANSISTOR OSCILLATORS

Oscillation is easy to achieve in a transistor. Almost any input and output circuit resonant to the same frequency will oscillate. A typical crystal oscillator is shown in the schematic of Fig. 1-11. Note, again, the similarity of biasing to that of the amplifier. Fig. 1-12 is a vfo as used in the Swan 500 and uses the common-base configuration. If you are using a VOM, wrap one pigtail of a 10,000-ohm resistor around the test probe. Ground the other probe. Touch the other resistor pig-

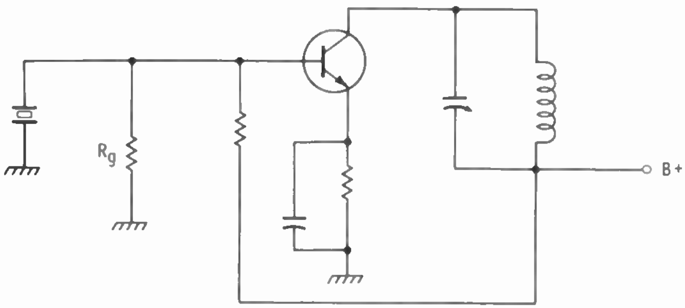


Fig. 1-11. A transistor crystal oscillator. Biasing is similar to the amplifier of Fig. 1-10. D-c voltage across R_G will be different between the oscillating and non-oscillating conditions.

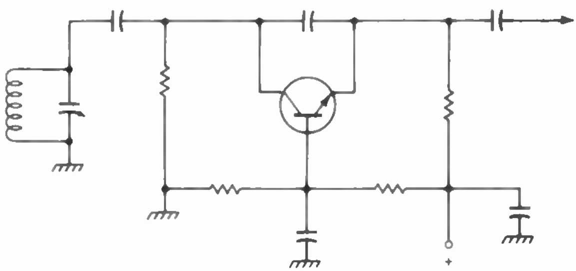


Fig. 1-12. A common-base configuration of a transistor in a vfo circuit.

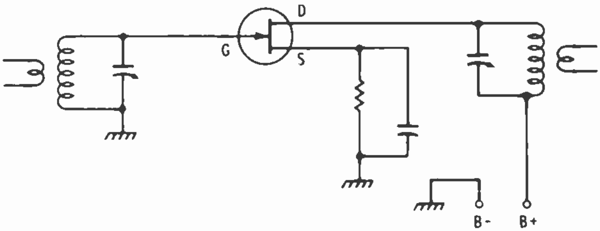


Fig. 1-13. A FET transistor as an r-f amplifier acts like a vacuum tube.

tail to the emitter. Alternately place a short across the resonant circuit, and lift the short. A difference in voltage read means that the circuit is oscillating. The 10,000-ohm resistor isolates the VOM from affecting the r-f circuit. A VTVM already has an isolating resistor built into the probe.

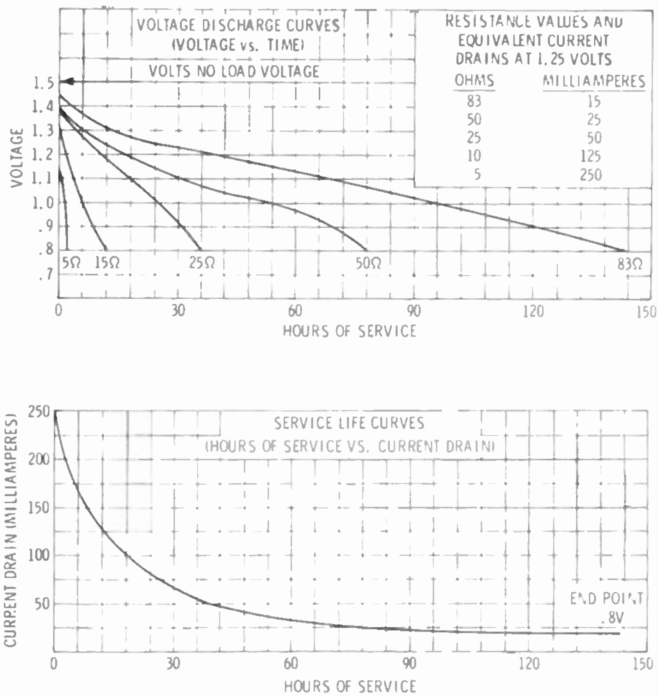
Fig. 1-13 is a FET transistor amplifier. It is almost always an r-f amplifier because its high impedance does not load down the resonant circuits. Note the similarity to vacuum-tube circuitry. The elements are called *gate*, *drain*, and *source* instead of base, collector, and emitter. The voltage across the source resistor is the bias value, and dividing the voltage by the value of the resistor will give you the current through the drain-source circuit.

BATTERIES

The open-circuit voltage reading on a new battery should be the rated voltage of the battery. As soon as a battery is connected into a circuit a load is applied, and the open-circuit voltage value will drop a small amount. This is due to the internal resistance of the battery. As batteries deliver power and the internal chemical action begins to be used, the internal resistance increases still more, resulting in a still further drop in voltage delivered to the load. The end-of-life of a 1.5-volt battery is considered to be at 0.9 volts. The actual number of hours a battery will last depends on its size, the current drawn, and the duty cycle—that is the ratio of on time to off time. Zinc-carbon batteries cannot take continuous drain too well, but do have pretty good recovery when off for a while. Other types are able to operate continuously with good service.

The curves of Fig. 1-14 show the hours of service that may be expected of the AA size manganese-alkaline battery. Note the rather fast initial drop in voltage in the top curve. The AA size is a popular portable-radio battery for moderate drains out of a small package, and is excellent for ham experiments with transistors. It has a long shelf life, compared with the cheaper zinc-carbon, retaining about 80 percent of its life up to about 5 years. Measuring the voltage and current will tell you how much life is left in the battery. The industry NEDA number is 15A. Mallory's number is MN-1500, Burgess's AL-9, Eveready's E91BP.

Table 1-1 lists popular sizes and kinds of batteries, and the recommended load resistor for measuring. The resistances were selected to draw the current that would result in a total life of 100 hours for each type. The end voltage should be about 0.9 volts, except for the 9-volt transistor radio battery, for which an end life of 6 volts was figured. Duty cycle was based on 2 hours per day for the zinc-carbon batteries,



NOTES: CURRENT DRAIN VALUES DETERMINED AT 1.25 VOLTS
INDICATED PERFORMANCE AT 70° F
ALL CURVES REPRESENT CONTINUOUS LOAD CONDITIONS

Courtesy Mallory Battery Co.

Fig. 1-14. Voltage discharge of the manganese-alkaline size AA battery under different loads is shown in the upper curve. The lower curve shows hours of service under different current drains.

and continuous duty for the others. All open-circuit voltages are 1.5 volts, except for mercury which is 1.4 volts, for new batteries.

POWER SUPPLY

The regulation of a power supply is the ratio of the difference in voltage between no load and full load, divided by the full-load voltage. Example: At no load the output measures 500 volts. With the load connected, the voltage reads 400.

$$\frac{E_{\text{no load}} - E_{\text{full load}}}{E_{\text{full load}}} = \frac{500 - 400}{400} = .25, \text{ or } 25\%$$

The internal resistance of the power supply is also easy to figure. It is the difference between the no-load voltage and full-load voltage divided by the current. Example, with the same voltages given above:

$$\frac{E_{\text{no load}} - E_{\text{full load}}}{I} = \frac{500 - 400}{200 \text{ mA}} = \frac{100}{0.2\text{A}} = 500 \text{ ohms}$$

TABLE 1-1. Popular Batteries Showing Probable Life in Milliampere Hours, With Resistive Load and Current Shown

Zinc 1.5V	mAh*	Load Ohms	I mA	Duty Hrs/24 .9V End
N	370	350	3.5	2
AAA	430	300	4.5	2
AA	1100	110	11	2
C	2400	50	24	2
D	4500	25	50	4
9V	300	2500	3	2 (6V end)
Manganese Alkaline				
N	520	250	5	24
AAA	650	200	7	24
AA	1500	100	15	24
C	4000	30	40	24
D	8000	15	80	24
Mercury				
AA	2300	50	25	24
D	14,000	10	140	24

*Milliampere hours.

CHAPTER 2

A-C Measurements

VOM's and VTVM's are designed to measure a-c voltage, but their accuracy of measurement will depend on the a-c frequency. VOM's have built in rectifiers to convert the alternating current to direct current since direct current is required to operate the meter movement. They are calibrated to read with accuracy (usually ±3%) at 60 Hz, with a drop-off in accuracy at higher frequencies. The manufacturer's specs will state the range of frequencies, which sometimes includes the entire audio range (beyond 15,000 Hz). You are not likely to find a VOM which will measure radio frequencies.

VTVM's (a-c operated or solid state) have internal or external rectifiers. Those with internal rectifiers include frequency-compensation circuits to make accurate readings up to about 3 MHz. For these you can usually buy an external r-f probe that fits over the regular probe. With the external r-f probe (and with those instruments with the built-on external probe), readings to 250 MHz can be made. External probes convert the r-f current to direct current, and are read on the d-c scales.

METER RECTIFIERS

Internal rectifiers follow the series of voltage-multiplier resistors, so value readings can be made as high as the instrument's d-c reading capability. However, external probes are limited as to the highest voltage read. The limit is set by the reverse-breakdown voltage of the silicon diodes, or heater-to-cathode breakdown voltage in the case of vacuum-tube diodes. This limit is often around 100 volts. This means that you cannot measure the r-f voltage at the top of the final tank circuit in a transmitter, but 95 percent of r-f voltage measurements can be made.

Because of the loss of some voltage through the rectifier and the need for extra circuitry for frequency compensation, as well as adjustment to make the a-c scales match the d-c scales, sensitivity on a-c is usually lower than on dc. This may be a drawback in some high-impedance circuit measurements, such as the high-impedance input of an audio amplifier. Most a-c measurements are made at low-impedance points, however.

Rectifiers have low conduction at very low currents through them, and for this reason, the instrument will have a separate scale for low-range a-c readings. At higher ranges the a-c readings are fairly linear, and high-range a-c scales are the same as the d-c meter scales.

Meter rectifiers for alternating current are very much like rectifier systems in power supplies. They may be half-wave, full-wave, or bridge types (Fig. 2-1). Each type provides successively higher sensitivity.

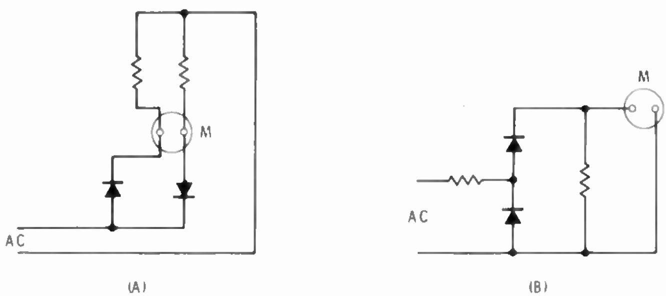
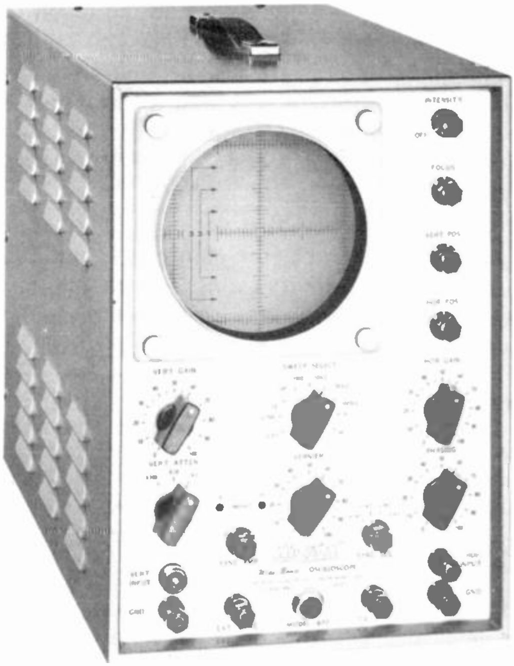


Fig. 2-1. At A two diodes are used in a full-wave meter rectifier circuit. B shows a half-wave rectifier. The lower diode conducts on the other half cycle to present a constant load to the a-c source.

Sometimes a voltage-doubler system is used for an even higher sensitivity. As in power supplies, capacitors are frequently used, and when used, charge up to the peak value of the a-c waveform. Thus, with VTVM's, the meter actually reads the peak or the peak-to-peak value. This is important in TV servicing because of the complex waveforms encountered. A half-wave rectifier passes current in one direction of the a-c cycle only, so it is a peak-reading meter. A full-wave rectifier and bridge rectifier rectifies both halves of the waveform, and thus is a peak-to-peak reading a-c meter. However, the scales are marked in rms values (as well as peak-to-peak), which is the peak-to-peak value times .3535. Keep in mind the rms readings are only correct for sine-wave voltages. Complex waveforms with high peaks will show a higher reading than average alternating current. This is important only when reading voice or random audio-frequency signals. Any meaningful tests of audio amplifiers or modulators should be made with a sine-wave generator supplying the signal.

OSCILLOSCOPES

Because the meter movement in a VOM or VTVM is mechanical it has mass, and therefore has inertia. The needle cannot follow the rapid changes of the a-c cycle. An oscilloscope, on the other hand, is electronic. The electron beam scanning the face of the scope tube has no inertia and will travel as fast as internal circuitry will permit. The better the instrument is engineered (and the higher the price), the higher the frequency it can display (Fig. 2-2). The vertical-



Courtesy Hickok Electrical Instrument Co

Fig. 2-2. A wideband, high-gain oscilloscope. Vertical amplifier response is within 3 dB to 4.5 MHz. Sensitivity is 40 mV per inch of deflection.

deflection plates in the cathode-ray tube carry the amplified (or direct) signal voltage to be observed. The horizontal deflection plates carry a repeating linear voltage to sweep the beam across the tube at a fixed time rate. The time rate is selectable from the front panel. As the beam sweeps from left to right at a fixed rate, it also moves up and down following the voltage changes of the signal under observation. Thus, you *see* the exact appearance of the wave, and all its complexities if there are any. If the signal is a nonchanging repeating frequency, and the horizontal sweep is synchronized with it, the signal appears to stand still on the face of the CRT for easy study.

The value of the scope to the amateur is almost exclusively based on the vertical amplifier. Low-priced scopes have a frequency response that is more than adequate to observe any audio frequency, but not radio frequencies. Medium-priced scopes have a response to about 4.5 or 5 MHz, and this will allow direct observation of r-f waveforms in the 3.5-MHz band. Lab-type scopes go much higher in response, but they become very expensive and outside the reach of the average ham. However, the internal vertical amplifier can be bypassed and a direct connection made to the vertical-deflection plate terminals of the CRT. This lifts the limit of frequency, but a fairly high signal level is needed to get any deflection. Some instruments do not provide for direct connection to the CRT deflection plates, but any ham can make his own easily.

Vertical amplifier sensitivity also varies with different scopes. As a rule, the higher the sensitivity (the gain of the amplifier) the lower the frequency response. Sensitivity ratings are in volts or millivolts per inch of deflection (sometimes in centimeters of deflection). For ham use a sensitivity of $\frac{1}{10}$ volt per inch is usually sufficient, although quite low. If you wish to check on low-level speech circuits (such as clippers), you will need a much higher input sensitivity—more like 10 millivolts per inch.

MEASURING FREQUENCY

The instruments mentioned above are for measuring or observing a-c voltages. Since frequency is also involved it is frequently necessary to measure the frequency of an unknown a-c voltage. Frequencies that hams are interested in measuring are usually in the r-f range. Instruments for measuring the frequency of an r-f a-c voltage vary in complexity and cost, depending on the accuracy of the measurement.

The most inexpensive instrument for measuring radio frequencies is the *absorption wavemeter* (Fig. 2-3). It is simply a tuned circuit to which a rectifier and meter is connected. When the coil is coupled to the r-f source the indicator will show a deflection on the needle when the capacitor is tuned to resonance at the frequency being measured. The capacitor dial is calibrated in frequency with the coil in use (the coils are usually plug-in). Since only one tuned circuit is involved, the accuracy is low—just about good enough to tell you in what part of the amateur band the radio frequency is.

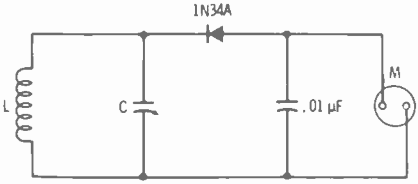


Fig. 2-3. A simple wavemeter for measuring frequency. L is one of several plug-in coils for covering all amateur hf and vhf bands. C has a maximum capacity of about 50 picofarads.

Accurate frequency measurements on transmitters are made with a *heterodyne frequency meter* (Fig. 2-4). This instrument includes a precision calibrated oscillator and a detector. The frequency to be measured beats with the internal oscillator and is heard in headphones, and sometimes is shown on a meter. Better instruments include

Courtesy Lampkin Laboratories, Inc.

Fig. 2-4. A high-quality frequency meter for making measurements of transmitter frequency between 100 kHZ and 175 MHz. A variable oscillator beats against the incoming signal. A built-in crystal oscillator is a calibration standard.

crystal oscillators for use in setting the local adjustable oscillator to exact frequency. These instruments are sometimes called *secondary frequency standards,* in that the crystal oscillator can be adjusted precisely by beating it against WWV, the National Bureau of Standards precision transmitter in Ft. Collins, Colorado.

THE GRID-DIP OSCILLATOR

The handiest instrument for the amateur (and inexpensive too) is the grid-dip oscillator (Fig. 2-5). It is a resonant circuit like the wavemeter but includes the resonant circuit in an oscillator using a vacuum tube, transistor, or tunnel diode. An indicating meter is connected to the grid of the vacuum tube, to the base of a transistor, or to the load resistor in series with the tunnel diode. The GDO will measure frequency in your transmitter with its power off. When the GDO is coupled to a coil-capacitor circuit and made to resonate with it, energy is absorbed from the GDO by the coupled circuit. When this happens the d-c voltage at the grid drops, and the meter needle shows a lower reading. The name of the instrument comes from this action.

Fig. 2-5A. A commercial high-quality grid-dip oscillator. Its accuracy is much higher than is normally available from a kit or home-built unit.

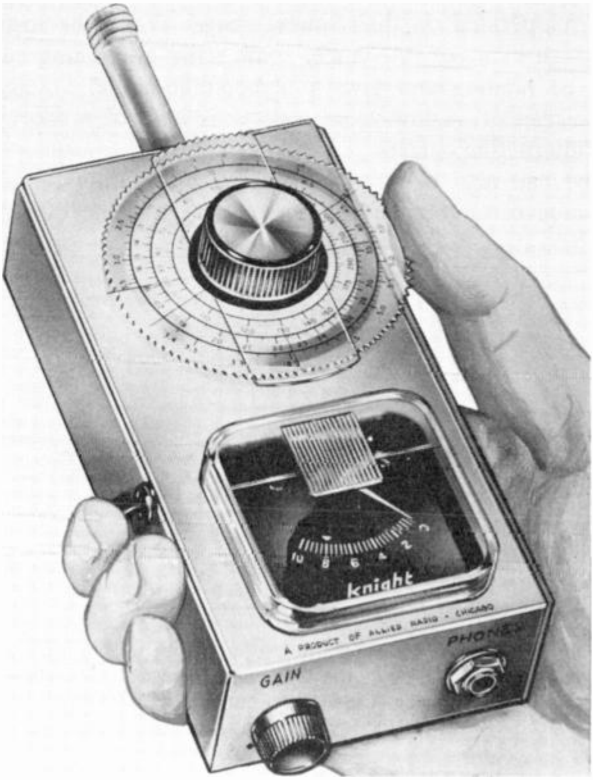
Courtesy James Millen Mfg. Co., Inc.

The needle takes a dip when the GDO is resonant to the tuned circuit to which it is coupled. Most GDO's include a phone jack. With headphones plugged in, the instrument acts like a regenerative detector and a beat note will be heard when the GDO is resonant with an r-f voltage. This action is like that of the heterodyne frequency meter, but, as with the absorption wavemeter, accuracy is comparatively low. Its greatest value is in experimental work for matching coils and capacitors to resonate at a desired frequency.

The GDO is also a signal generator, although it lacks the refinements of a good one. It can also be used to feed an impedance bridge for checking antenna feeders—more about this in a later chapter.

ALTERNATING-CURRENT MEASUREMENTS

While the VOM measures direct current, and all of the instruments mentioned at the beginning of this chapter measure both dc and ac volts, none measure alternating current directly. Alternating current



Courtesy Allied Radio Corp.

Fig. 2-5B. Low cost grid-dip oscillators are available in kit form. The *Knight-Kit* shown here is one. By listening to its signal in a calibrated receiver, the adjustable cursor can be set to give fairly good accuracy.

can be measured by measuring the voltage across a resistor in series with the current (Fig. 2-6), and using Ohm's law. Ohm's law says:

$$I = \frac{E}{R}$$

The voltage measured across a resistor divided by the value of the resistor is the current through it. The instrument should be set for the lowest a-c voltage range, and the resistor should be as low a value as possible. When measuring r-f current, the resistor must be noninductive.

When current values are low and it is necessary to use a high value of resistance to get a decent reading, another problem arises. The very high-impedance inputs of VTVM's will pick up stray alternating currents when operated on their lowest range. When the instrument is across a high value of resistance, some stray alternating current can increase the reading and give a higher-than-actual voltage reading across the resistor. A low-value resistor is kind of a short circuit to the stray alternating current.

A scope can also be used to observe alternating-current action, but the same precaution of use across a high value of resistance is needed.

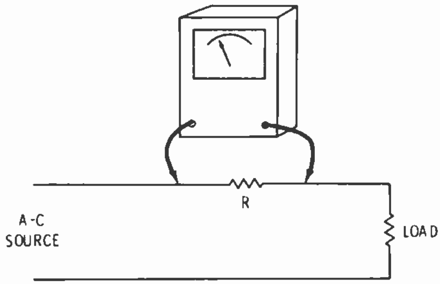


Fig. 2-6. To measure alternating current, connect a resistor in series with the load, and measure the voltage across the resistor. Convert to current with Ohm's law. Always use the lowest possible resistor value.

THE SIGNAL GENERATOR

Although not a measuring instrument, the signal generator is an important instrument for producing r-f signals for making it possible to measure components and equipment. It is a miniature transmitter which is tunable over a very wide range of radio frequencies (Fig. 2-7).

A signal generator contains a stable variable-frequency oscillator. The oscillator is bandswitched, and each band is tuned over its range with a variable capacitor, like a general-coverage receiver. The oscillator is sometimes followed by a buffer stage to isolate the oscillator from the output. The buffer is followed by an attenuator, usually in two parts. One is a step attenuator with steps of either 10 dB between each step or 10:1 voltage ratio between steps. The other attenuator part is a variable potentiometer. An output meter is part of better signal generators. The meter usually precedes the attenuator and is used to set the output to a predetermined level, say 0.1 volt. The step

Fig. 2-7. A quality signal generator available in kit form.
It covers 100 kHz to 9.5 MHz in four bands. The
output meter provides calibrated r-f output

attenuator then reduces the output in fixed increments, down into the microvolt range.

For use where a modulated r-f output is necessary, signal generators include internal audio modulation. Most use a single 400-Hz audio signal, but some include a variable-frequency audio signal. The amount of audio is adjustable.

THE AUDIO GENERATOR

Of less importance to the amateur, but a generally useful instrument to have is the audio generator (Fig. 2-8). If high fidelity is high on your list of activities, an audio generator and a scope are two instruments worth owning. An audio generator is also valuable in ssb transmitter adjustments.

Like the signal generator, the audio generator has an oscillator, a buffer, and an attenuator. The oscillator is of the RC type, usually a Wien bridge. It is bandswitched by switching precision resistors in the bridge circuit. The range of audio frequencies in each band is covered by a dual-ganged variable capacitor which is part of the bridge circuit. The output also has a stepped attenuator marked in dB or in 10:1 voltage steps.

The output is a nearly pure sine wave with a minimum of distortion. The better the instrument is the lower is the distortion. For ham use, the very low distortion is not important. Many audio generators include a square-wave output in addition to the sine wave. A switch cuts in a positive- and negative-peak clipping circuit. When a good part of the peaks of a sine wave are clipped off, the remaining sides are straight enough to make fairly good square waves. A square wave through an

Courtesy Heath Co.

Fig. 2-8. A sine-/ square-wave audio oscillator. It covers the range from 20 Hz to 200 kHz. Output is a choice of pure sine wave or a square wave, which has value in hi-fi measurements.

audio amplifier will show up transients as spurious spikes when viewed on a scope.

CHECKING CAPACITORS AND INDUCTORS

The simplest check of a component is, of course, that for continuity. If an ohmmeter shows infinite resistance through a choke or a winding of a transformer, that circuit is open. If the ohmmeter reads zero resistance across the leads of a capacitor it is shorted.

But an ohmmeter can do more than that on a capacitor. It can show that there is capacitance, and it can measure leakage, which is an important parameter of a capacitor used to couple between stages of vacuum-tube amplifiers. The presence of capacitance is shown by a kick of the needle when the leads are first applied. When you first touch the leads of the ohmmeter to the capacitor it takes a charge from the ohmmeter battery. During the short interval that current from the battery is entering the capacitor the current drawn is high enough to cause the meter needle to kick up-scale in the case of a VOM, down-scale in the case of a VTVM. As the capacitor becomes charged, the current decreases and the needle returns to a high-resistance reading. The amount of time it takes the needle to reach its highest-resistance reading is an indication of the amount of capacitance in a capacitor. There is a formula for figuring the RC time constant of the charge of a capacitor, but it involves time in fractions of a second which makes it impractical to measure. It is sufficient to know that the higher the capacitance is, the longer it takes the meter to settle down to a high-resistance reading.

A good capacitor should have a resistance of over 50 megohms after the needle has settled down. This is the leakage resistance. Lower resistances (except for electrolytics) mean the capacitor is old and has absorbed some moisture, or is of poor construction. When electrolytic capacitors are checked, the kick will be to a lower resistance reading and the needle will take longer to settle down to a high reading, if the polarity of your ohmmeter is correct. The positive (+) of the internal battery must be connected to the positive (+) terminal of the electrolytic. Any leakage resistance reading of 100,000 ohms or more is considered satisfactory for electrolytic capacitors, because of the electrolyte used in them for a dielectric.

The presence of inductance in an iron-core inductor, such as a filter choke or audio or power transformer, is checked in the same way. The effect is the opposite, however. The needle on the ohmmeter will move from a high resistance value slowly down to the d-c resistance of the winding. The higher the inductance is, the longer it takes for the magnetic field to build up as a result of the charge from the ohmmeter battery. A quick movement of the needle (only slightly more than the movement occurring when measuring a resistor) means that possibly some of the turns of wire are shorted, thus affecting the inductance.

POWER TRANSFORMERS

The leads on a power transformer generally follow a standard color-code system. Fig. 2-9 shows the standard colors used on a typical power transformer. Assume you have a transformer of unknown voltages and you want to measure the voltages. Separate the leads and

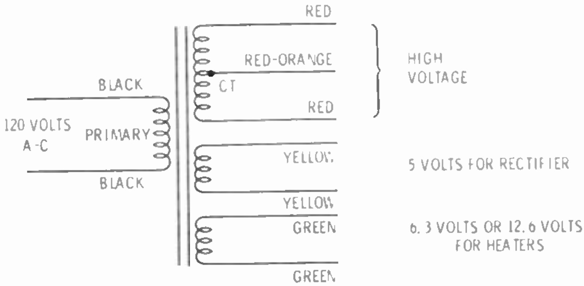


Fig. 2-9. The leads out of power transformers are color-coded like this.

study the colors of the insulation on them. If you are able to find agreement with the sketch of Fig. 2-9, select the black leads and connect them to the 120-volt house supply. Be careful in this operation; you are dealing with high voltages. Twist the black leads to the ends of a line cord, and cover them with tape, or with wiring nuts used to make electrical connections. Place your voltmeter on the highest a-c range and measure between a red lead and a red-orange (the center tap of the high-voltage secondary) lead. The voltage should be quite high, depending on the transformer, and be the same between either red lead and the red-orange lead. The balance of the leads should be from heater or filament secondaries. You should measure 5 volts ac across the yellow leads, and either 6.3 or 12.6 volts between the green leads.

You can also measure the approximate power rating on the basis of temperature rise. The same length of copper wire will measure a different resistance when hot or cold; higher when hot. Assume you have a power transformer with a 500-volt center-tapped high-voltage secondary. You are building a piece of equipment in which you want to draw 50 mA in the secondary. Measure the d-c resistance of the secondary winding while cold. Connect a 10,000-ohm, 25-watt resistor across the secondary and plug the primary into the 120-volt a-c socket. At 500 volts, the resistor will draw 50 mA. Leave it connected for about an hour thus allowing the transformer to warm up. Remove the transformer from the power, remove the resistor and again measure the d-c resistance of the high-voltage secondary. It will now indicate a higher value. The heat rise is figured from the following formula:

$$t_2 = \frac{R_2}{R_1} \times (234.5 + t_1) - 234.5$$

where,

t_1 is cold temperature in °C (room temp),
t_2 is hot temperature,
R_1 is resistance at t_1,
R_2 is resistance at t_2.

A heat rise of 40° C is considered within limits. At any figure substantially below that, the transformer will run comparatively cool for your application. But what about the power from the filament windings? Our measurement was made on the basis of a 50-mA continuous current from the entire high-voltage secondary. A center-tapped secondary in a full-wave rectifier circuit is called on to deliver current out of each half of the secondary winding only on alternate cycles. Our checking system gave a 100 percent safety factor, and some of this will be used for filaments. Unless you figure on a long string of tubes in the circuit, you will still be well within the heat-rise figure of the transformer.

AUDIO TRANSFORMERS

Audio transformers are used to match impedances in audio circuits, whether it is to match the output of a power amplifier to a speaker or the collector of a transistor to the base of another following it. The impedance ratio between primary and secondary of a transformer is easy to figure when you know the turns ratio. The impedance ratio is equal to the square of the turns ratio. The turns ratio is equal to the voltage ratio. Using voltage as our measurement we merely square the figures to get impedance ratio. Remember we are concerned with a ratio, so the actual voltage is not important.

For example, we are building our own receiver and wish to use a 6K6GT in the output to drive a 3.4-ohm speaker. The tube manual says a 7600-ohm load is recommended. Here is a small output transformer whose instructions have been lost. It has a red lead, a blue lead, and two black leads. The red and blue leads are the primary, and black leads are the speaker secondary. This is standard color coding for output transformers. If instead of two black leads it had a green and a black lead, it would be an interstage transformer. However, a d-c ohmmeter check would verify the type of transformer. The resistance between the red and blue leads would be between 25 and 100 ohms, and between the two black leads, less than one ohm. If it were an interstage transformer the resistance between the secondary leads would be greater than the resistance on the primary. Since we have determined it is an output transformer, wire up a line cord to the primary, in the same way we did for the power transformer before. Measure the voltage across the primary. It should, of course, be the line voltage. Measure the voltage across the secondary. As near as we can read on our meter it is 2.4 volts. The ratio of 120 (the line voltage) to 2.4 is approximately 50 to 1. The square of 50 is about 2500. The impedance ratio is, therefore, 2500 to 1. Since it is a ratio, we can also say it is 7600 to 3.13. That is pretty close to the ideal of 7600 ohms for the plate and 3.4 for the speaker. The impedance of 3.4 ohms for a speaker is a very loose figure anyway. Your transformer could be as much as 20-percent off the right ratio and still work quite well.

The primary of a small transistor transformer should never be connected to the 120-volt line. The impedance of the primary is much too low for so high a voltage. Instead, find an old power or filament transformer and use 12.6 volts for the primary. Because transistors are low-impedance devices, the required impedance ratio is more like 300 to 3.2 for a speaker matching transformer. Check the specifications for the transistor you plan on using.

MEASURING CAPACITORS

There are two methods for measuring the capacitance of a capacitor. One is to measure the a-c voltage across the capacitor, and across a resistor in series, adjusting the value of the resistor until the two voltages are equal, and applying a formula. The other method is to measure the resonant frequency of a coil and capacitor of known value; then substitute the unknown capacitor and measure resonance. The first method is good only for high values of capacitance, because the handiest frequency available for such measurements is the 60 Hz of the house supply. With an audio generator set at higher frequencies you can measure lower values of capacitors. The second method is for low values of capacitance. Neither method is highly accurate with the use of equipment the amateur is likely to have, but either is good enough for practical use.

Accurate measurement of capacitors and inductors can only be done on a laboratory-type bridge. This requires an investment it will never pay the amateur to make.

Fig. 2-10 is the schematic setup for measuring capacitors by the method mentioned first.

Connect a 50,000-ohm variable resistor in series with the capacitor and a source of 60-Hz voltage. A 12.6-volt filament winding on a transformer is safe to use. Make a-c voltage measurements alternately across the capacitor and resistor, at the same time adjusting the re-

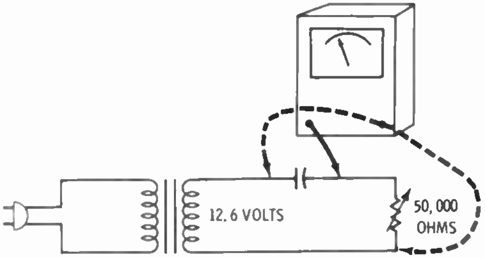


Fig. 2-10. When a voltage across a resistor and a capacitor are equal, the capacitor's reactance equals the value of the resistor.

sistor until the voltage across both are the same. When the two voltages are equal, the capacitor has a capacitive reactance equal to the resistor. Measure the resistance of the variable resistor after the adjustment. The standard formula for capacitive reactance is:

$$X_c = \frac{1}{2\pi f C}$$

Transposed for C as the unknown it becomes:

$$C = \frac{1}{2\pi f X_c}$$

where,

C is capacitance in farads,
π is 3.1416,
f is frequency in Hz,
X_c is capacitive reactance in ohms.

To change to C in microfarads (μF), raise the numerator 1 to 10^6, thus:

$$C = \frac{10^6}{2\pi f X_c}$$

Here is an example of how it works: An output of 14.8 a-c volts was available. The unknown capacitor and a variable resistor was connected in series, across this voltage. The resistor was adjusted until 9.8 volts was measured across both the resistor and capacitor. The resistor measured 12,450 ohms at this balance. Substituting in the above formula:

$$C = \frac{10^6}{2 \times 3.1416 \times 60 \times 12{,}450}$$

$$C = \frac{10^6}{4700 \times 10^3}$$

$$C = \frac{1000}{4700}$$

$$C = .215\ \mu\text{F} \quad \text{(approximately)}$$

The capacitor in this case was marked .22 μF with probably a broad tolerance, so we are pretty close.

It is interesting to note that the sum of the two voltages measured seems to be greater than the supply voltage. This is because the circuit of a capacitor in series with a resistor introduces a phase lag (the voltage lags behind the current).

Using a 50,000-ohm variable resistor permits measuring capacitors from about .05 μF up. A higher resistor value might introduce stray a-c pickup into the reading and result in considerable error. If your signal source is an audio generator, increased frequency will permit reading smaller capacitances. For instance, using 600 Hz places the lower limit of capacitance measurement at about .005 μF. You will

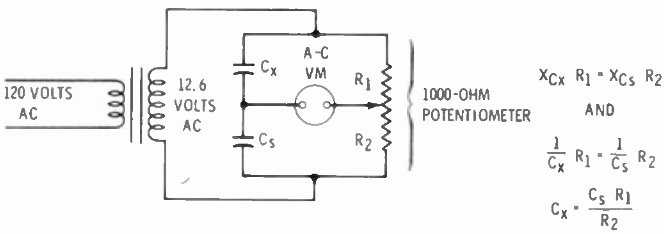


Fig. 2-11. A simple bridge for measuring capacitance. A capacitor (C_s) of known value must be used as a standard.

recognize this as half of a bridge circuit, with the resistor adjusting the bridge for equal voltages each side of the two arms.

With only a little more effort you can make a homemade (or should we say a ham-made) bridge, which checks unknown values of capacitors against one of a known value. Fig. 2-11 is the schematic of a simple bridge. A variable 1000-ohm potentiometer represents two arms of the bridge. The unknown capacitor, and the known capacitor are the other two arms. Set your meter on the 15-volt range. Adjust the potentiometer for minimum reading on the meter; it should come pretty close to zero. Now disconnect the voltage source, switch your instrument to ohms and measure the resistance each side of the center terminal of the potentiometer. The bridge is in balance when the product of the reactance of C_x and resistance of R_1 equals the product of the reactance of C_s and the resistance of R_2. This can be written as follows:

$$X_{CX} R_1 = X_{CS} R_2$$

Capacitance is inversely proportional to capacitive reactance. The capacitance formula can be written as follows:

$$\frac{1}{C_x} R_1 = \frac{1}{C_s} R_2$$

or

$$C_s R_1 = C_x R_2$$

and

$$C_x = \frac{C_s R_1}{R_2}$$

Example: A capacitor marked .047 μF was connected into the circuit for C_s. Another capacitor was connected in the C_x spot. The potentiometer arm was balanced for a null on the meter, and when measured for resistance, the R_1 part measured 800 ohms, and the R_2 part measured 200 ohms. Substituting in the above formula:

$$C_x = \frac{.047 \times 800}{200}$$

$$C_x = .188\ \mu\text{F}$$

The capacitor C_x was actually marked .22 μF. The difference between the calculated value and marked value is the tolerance of both the standard capacitor and the one under measurement. The obvious conclusion is that the capacitor used as a standard should be a close-tolerance capacitor, or you must expect a $\pm 20\%$ variation.

Lab-type bridges also include a variable resistor in series with C_s for balancing out the leakage resistance of capacitors. For practical purposes for the ham this can be overlooked. If a leakage test with the VTVM shows less than 1 megohm of leakage of any kind but an electrolytic, the capacitor should be thrown away.

Electrolytic capacitors can be measured in the same way but it takes a quality, high-value capacitor for the standard, and the addition of the variable resistor in series with the standard. Also required is a polarizing d-c voltage in series with the a-c source.

Fig. 2-12 is the circuit. The R_1R_2 balance control is adjusted simultaneously with the leakage resistor control (R_3) for minimum reading. R_3 balances out the leakage resistance of the electrolytic capacitor, which resistance is considered to be in series with the electrolytic

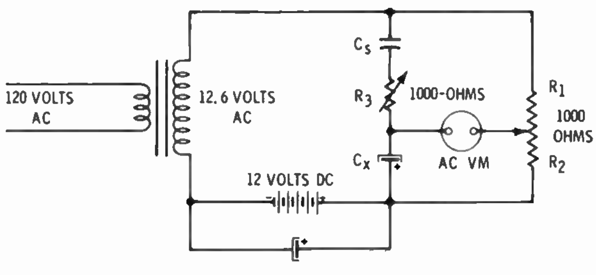


Fig. 2-12. A bridge circuit for measuring electrolytic capacitors. C_s must be an oil filled nonelectrolytic type. Direct current is added to polarize the electrolytic capacitor under test.

capacitor. Once balance is attained, the solution is the same as for other capacitors.

Small values of capacitors are best measured in a resonant circuit. This involves the use of grid-dip oscillator, or a signal generator. Every ham should have a GDO, and the method to be described uses the GDO. You will also need a coil and a small capacitor of known value. The coil may be any type that you think should resonate with the capacitor within the ranges of the GDO (Fig. 2-13).

Connect the capacitor across the coil ends. Couple the GDO to the coil and measure the resonant frequency. Start with the highest-

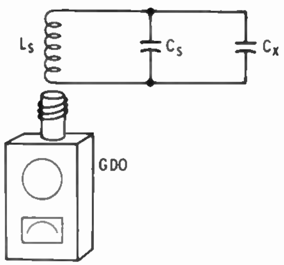


Fig. 2-13. Using standard capacitors and inductances in a resonant circuit, any small value of unknown capacitor can be measured by the change in resonant frequency.

inductance plug-in coil you have for the GDO. Couple and tune the GDO for a dip on the needle. Move the GDO coil slowly out of the field of the coil under test until the needle barely dips. The looser the coupling is, the more accurate the reading will be. Note the resonant-frequency reading. Connect the unknown capacitor in parallel with the known-value capacitor and measure for resonance again. The solution requires solving for the inductance of the coil with the known capacitor and frequency; then knowing the inductance, solving for the value of the unknown capacitor from the inductance and the new frequency. It is based on the formula:

$$f = \frac{1}{2\pi\sqrt{LC}}$$

Squaring, we get:

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

and

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

and

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

where,

f equals frequency in Hz,
L equals inductance in henrys,
C equals capacitance in farads.

If all values are in MHz, microfarads(μF) and microhenrys(μH), the same formulas apply.

Example: Starting with a known capacitor of 100 pF (picofarads, or micromicrofarads) connected across a coil, the resonant frequency was measured to be 7.8 MHz. Solving for L in the above formula (and using a slide rule for approximate calculations) we get:

$$L = \frac{1}{4 \times 9.8 \times 60.8 \times .0001} = \frac{1}{.239}$$

$$L = 4.19\ \mu\text{H}$$

Connect the unknown capacitor across the 100-pF capacitor. The new frequency measures 5.48 MHz. Having L we now solve for C.

$$C = \frac{1}{4 \times 9.8 \times 30.3 \times 4.19} = \frac{1}{4987}$$

$$C = .0002 \text{ (approximately)}$$

Subtract the known value of .0001 μF from the calculated value of .0002 μF (.0002 — .0001 = .0001). The unknown value is probably a 100-pF capacitor, but, again, tolerances must be considered.

If you have many occasions to measure very small capacitors, it will be worth your while to select a coil to keep as a standard for L. Put it away with a chart of capacitance versus the inverse of frequency squared (capacitance increases linearly with the inverse of the frequency squared), like that shown in Fig. 2-14. Copy the figures of the horizontal capacitances and vertical inverse frequency-squared ratio as shown. Draw your own diagonal line for the coil you have. Any small r-f coil will do. You will need a couple of 1-percent capacitors, say 10 pF and 100 pF. Place marks on the graph where resonance occurs for each of the capacitors, as measured by coupling a GDO to the resonant circuit. Draw a straight line through both marks. The lower left end of the line will extend beyond the zero-

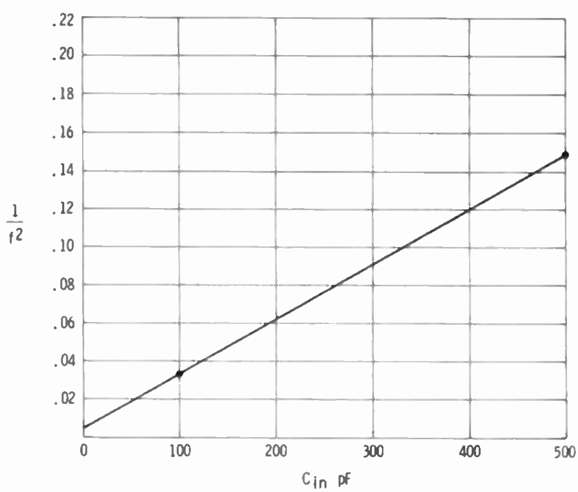


Fig. 2-14. How a chart can be made to find the value of small capacitors by measuring the resonant frequency with a standard coil. The diagonal line in this sample was based on a 100-pF and a 500-pF capacitor.

capacitance figure. This is the distributed capacitances of the coil. Thereafter, for any unknown value of small capacitor, find the resonant frequency with the standard coil, go across to the diagonal line, and down to capacitor value.

MEASURING INDUCTANCES

An inductance may be measured by the method described for capacitors. For high values of iron-core inductors, use one of the first two methods. For low values of inductor use the GDO resonance method, with a known value of small capacitor.

One difference must be considered in measuring iron-core or high-value inductors. They all have resistance, which is the resistance of the wire with which they are wound. In the case of r-f coils, the resistance is so low it may be disregarded.

In the first method (that of an inductor in series with a resistor and adjusting the resistor for equal voltage across it and the inductor) the answer will be the impedance of the inductor. Impedance is a combination of reactance and resistance.

Remove the power and measure the resistance of the inductor with the ohmmeter. Find true X_L from the following formula:

$$Z = \sqrt{R^2+X^2}$$

or

$$Z^2 = R^2+X^2$$

therefore

$$X^2 = Z^2-R^2$$

where,

Z is the apparent impedance as measured,
R is the d-c resistance of the wire in the inductor.

Having reactance (X) we find the inductance from the following formula:

$$L = \frac{X}{2\pi f}$$

If you have an inductor of known value you can use it as the standard in the bridge method described for capacitors. If the d-c resistance of the two inductors (the known and the unknown) are different, you will need to add resistance in series with the one with less resistance to make the two equal. Assuming the standard inductor has the lesser d-c resistance, the circuit will look like Fig. 2-15. If the two resistances are not equalized the balance on the R_1 R_2 arm will be incorrect.

For small air-core inductors as used to resonate in the ham bands, the GDO method described for measuring small capacitors with a standard value capacitor applies. On the basis of the resonant frequency with a capacitor of known value you would solve for inductance, as was done earlier for capacitors, using the setup of Fig. 2-13. This does not take into account the effect of distributed capacity in the coil, but in the small inductors used in the ham bands distributed capacitance has little effect. Besides, the wide tolerances of capacitors

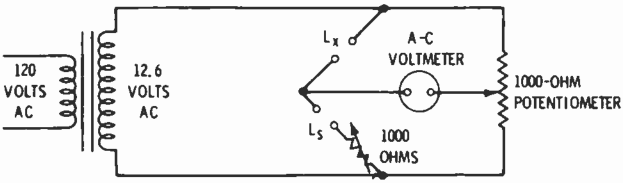


Fig. 2-15. A bridge for measuring iron-core inductors. The standard inductor (L_s) must have a known value. If its d-c resistance is higher than the inductor being measured (L_x), switch the positions of the two. The series resistor should be in series with the lower-resistance inductor.

that may be at hand and the lack of calibration accuracy of the GDO readings would result in wide errors of distributed capacitance. Distributed capacitance on ham-band coils usually is less than 1 pF on large air-wound coils, and perhaps as much as 10 pF on close-spaced r-f coils.

Distributed capacity is important only in r-f chokes where it should result in self-resonance in the ham bands for the greatest effect. The real importance is for the r-f choke to offer the highest r-f impedance in a circuit, and it does so when it is resonant at the frequency of concern. But its self-resonance is as much affected by parallel capacities of the associated circuitry as the distributed capacitance of the coils of the choke. The measurement, therefore, should be made with the choke connected into the circuit in which it is used or is to be used.

If one knows the inductance of a coil from its published specifications, the distributed capacitance can be determined. Couple the GDO to the coil and measure its self-resonant frequency. Start with the lowest-frequency coil in your GDO, and work up. From frequency, the distributed capacitance is solved from:

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

HEADPHONE AND SPEAKER IMPEDANCE

The impedances of headphones and speakers are specified for frequencies other than 60 Hz; for headphones it is usually 1000 Hz, and for speakers, 400 Hz. This obviously calls for the use of a signal source of these frequencies, the audio generator being the best source. As in the case of C and L measurements, the simplest method is the equal-voltage method; that is, equal voltages across the headphone or speaker, and a resistor in series.

For head phones set the audio generator for 1000 Hz and connect a 5000-ohm variable resistor in series with the headphone and across the output of the generator. Make alternate a-c voltage measurements across the headphones and across the resistor. Adjust the variable resistor until the voltages are equal. The impedance of the headphones then is equal to the resistance of the resistor setting at 1000 Hz.

The a-c impedance of loudspeakers is very close to the d-c resistance of the voice coil. Begin by measuring the d-c resistance of the voice coil. Connect a variable resistor of about twice that value in series with the speaker. A 20-ohm variable resistor will be high enough to measure any speaker, including 16-ohm high-fidelity speakers. With the audio generator set for 400 Hz, adjust the resistor for equal voltages. Most speakers used on ham receivers are the same as used in standard radio sets, and will probably measure 3.4 ohms. These speakers are called "replacement" speakers in the radio parts catalogs.

MEASURING *Q*

The Q of a circuit is the figure of merit of selectivity of the circuit. It is an indication of how sharp the resonant point is. The Q is affected by the winding factor of the coil and by the resistance in the circuit. The resistance may be the resistance of the wires of the coil, the resistance in the capacitor (usually negligible), and principally by the resistance of the circuit associated with the resonant circuit—that is, the rest of the circuit in which it is used.

With a GDO the relative Q is reflected by the sharpness with which the needle on the GDO dips when tuning through resonance. A quick flick of the needle indicates a high Q. A broad dip indicates a low Q. To see this, note the effect on the dip when measuring a coil-capacitor resonant circuit both with and without a 1000-ohm resistor across the circuit. Resonance will show up as a broad dip with the resistor across the circuit.

The actual measurement of Q requires a signal source and a VTVM with an r-f probe. Connect the probe of the VTVM across the coil-and-capacitor circuit, and apply an r-f signal to the coil by coupling the GDO to the coil. Measure the voltage at resonance. Adjust the GDO frequency to either side of resonance until the meter reads .707 of the resonance value. Adjust the GDO to the other side of resonance until the meter again reads .707 of the resonance value. Q is given by the formula.

$$Q = \frac{f_0}{f_2 - f_1}$$

where,

f_0 is the meter frequency at resonance,
f_1 is the frequency at .707 below resonance,
f_2 is the frequency at .707 above resonance.

The Q of a tuned circuit is easily measured with an r-f signal generator. The setup is like that of Fig. 2-16. Note the resistor in series with the signal generator and tuned circuit. The value of the resistor

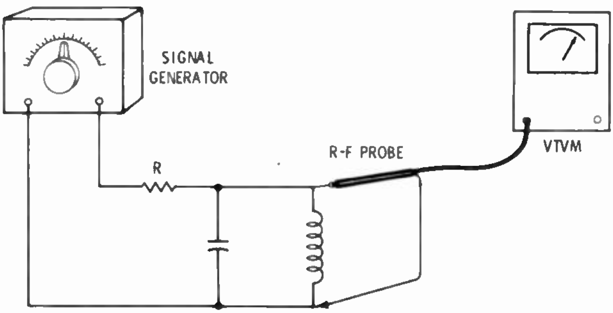


Fig. 2-16. Setup for measuring Q of a resonant circuit. The resistor (R) should have a value equal to the normal associated circuitry—for example, the plate resistance of the tube preceding the resonant circuit.

should be the same as the plate resistance of the tube or circuit to which the tuned circuit is normally connected. The information can be obtained from the tube manuals.

R-F AND I-F TRANSFORMERS

The build-it-yourself ham can predetermine the selectivity of transformers he may want to use. The setup of Fig. 2-17 is similar to that of Fig. 2-16, but a resistor has been added to the output to simulate the grid loading of a following stage. Selectivity measurements are made like that for the Q of a circuit mentioned before, except a number of frequency-versus-output voltages are plotted. The information is converted to a curve like that of Fig. 2-18. Changing the voltage-ratio figures to dB results in a selectivity curve for that transformer. (dB is 20 times the logarithm of the voltage ratio).

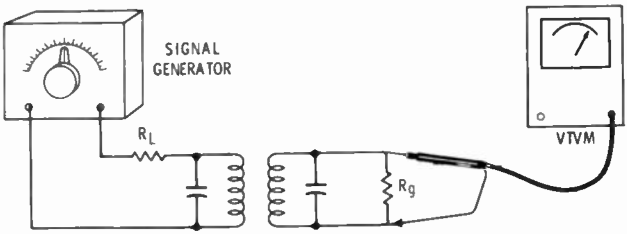


Fig. 2-17. The Q and selectivity of an r-f or i-f transformer can be measured with this setup. R_L is equal to the plate resistance of the preceding stage. R_g is equal to the grid load of the following stage.

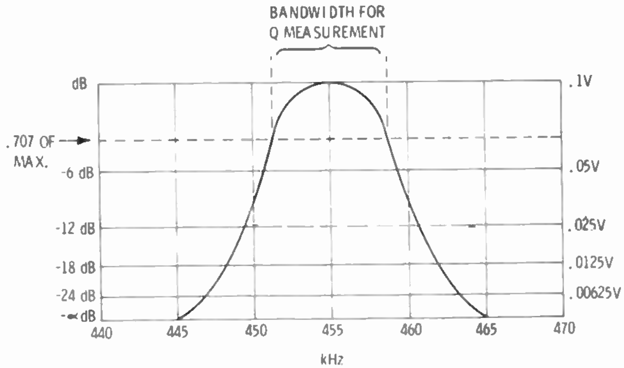


Fig. 2-18. By charting volts versus frequency and converting volts to dB, a selectivity curve like this can be drawn. Bandwidth shown in specs is at the −6dB points.

Transformers are designed to cover the range from high gain and sharp selectivity to those with a broad passband and low gain. Sometimes the latter type even have a double hump with a small dip at resonance.

FREQUENCY MEASUREMENTS

It obviously follows that the method described above for *Q* and transformer selectivity measurements may also be used for measuring the resonant frequency of an unknown coil-and-capacitor combination, or r-f or i-f transformers. Tuning the signal generator to the lowest frequency that gives a peak reading on the VTVM, the generator dial will indicate the resonant frequency.

The handiest instrument for measuring frequency is the GDO. What it lacks in accuracy it makes up in versatility. Resonant circuits wired into a piece of gear may be measured for frequency, if they are not shielded to prevent coupling the GDO to it. Inductively couple a GDO to the coil of any resonant circuit. The meter on the GDO will dip down scale when the GDO and the resonant circuit are at the same frequency.

How to Couple a GDO

Care must be used in coupling a GDO to a resonant circuit. The accuracy of the GDO is affected by the external load of the circuit to which it is coupled. For minimum effect the coupling should be as loose as possible, while still getting a dip on the meter.

When first coupling to a tuned circuit, place the coil end against the end of the coil of the resonant circuit. This puts the magnetic field in line with the field path of the resonant circuit. You will note a pronounced dip on the needle of the GDO. You will also note the point of greatest dip is different depending on the direction from which you turned the tuning dial on the GDO. If this occurs you are definitely overcoupled. The object is to draw the GDO coil away from the resonant circuit until there is a barely perceptible dip on the needle. This is the point of least effect of the external tuned circuit and greatest accuracy. The higher the *Q* of the resonant circuit the farther away from it you will get a dip, and the sharper will be the dip.

The GDO may also be coupled to a piece of wire for an indication, if the wire is part of a tuned circuit. The connecting lead between a coil and capacitor is part of the resonant circuit. Placing the coil of the GDO against the wire is coupling to the circuit. Because the magnetic field around a wire is much less than around the coil of a resonant circuit, it may be necessary to place the coil of the GDO right up against the wire. Unless heavily swamped with resistance, any piece of wire will show some resonance. Try this in a transmitter (turned off, of course) at any point. You may discover other resonant frequencies than the one the transmitter should be operating on. In fact, this is one way of checking for spurious oscillation, and will be covered more fully in a later chapter. A straight piece of wire connected to nothing is a resonant circuit—like an antenna. By coupling your GDO to the center of the wire you can measure the resonant frequency of that wire like a half-wave dipole antenna.

Coupling to a circuit under test must be inductive. It is possible to get a dip by capacity effect between the coil and a large surface, but most such attempts usually fail, and the capacitive-coupling method should not be considered.

Measuring Crystal Frequency

A quartz crystal is a mechanical resonant circuit of high stability and Q. Its approximate frequency can be measured with a GDO by coupling to it as you would any resonant circuit. Wind about two turns of wire around the coil of the GDO, and connect the stripped ends to the pins of the crystal. The GDO will dip at resonance. Crystals can have more than one mode of vibration, and some are especially designed as overtone oscillators, especially the ones for use in the Citizens Band. Start with the lowest-frequency coil plugged into the GDO, and work up in frequency until the first dip is found. This is the approximate frequency of the crystal fundamental.

In some GDOs the coil jack's pin spacing and the circuit is such that FT style crystals may be plugged into the GDO in place of a coil, and the instrument will oscillate at the crystal frequency. The *Knight-Kit* GDO is an example of such a unit. In GDO's using tunnel diodes as the oscillator, the tunnel-diode current flows through the coil. Since crystals have no d-c continuity they cannot be substituted for coils in tunnel-diode GDO's.

A crystal plugged into a GDO can be heard in a receiver. Its frequency would be determined from the dial reading of the receiver.

A crystal offers very low impedance to the flow of current at its resonant frequency. When connected in series with a signal generator and a VTVM with an r-f probe, the frequency at which the VTVM indicates the highest is the frequency of crystal resonance.

FILTERS

Filters for attenuating frequencies above or below a specified frequency (called low-pass and high-pass) are made up basically of L sections as in Fig. 2-19. This type of filter is called a *constant-k* filter. The product of the arithmetic impedance values of each leg is a constant ($Z_C \times Z_L = k^2$), and is true for all frequencies. Constant-k filters have excellent attenuation above or below the cutoff frequency. The same filter will often appear as in Fig. 2-20 called a π-section, low-pass filter, or as in Fig. 2-21 called a T-section, low-pass filter. Adding sections increases the attenuation rate above or below the

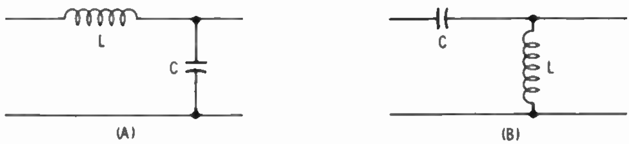


Fig. 2-19. Constant-*k* filter sections. A is a low-pass, and B a high-pass filter.

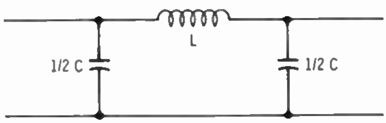


Fig. 2-20. A pi-section constant-*k* low-pass filter network. The capacitor value is split between the input and output.

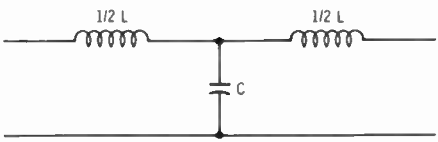


Fig. 2-21. A T-section constant-*k* low-pass filter.

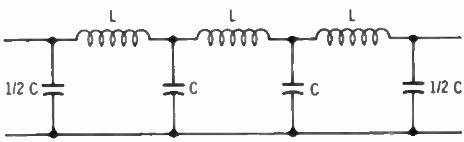


Fig. 2-22. A multisection constant-*k* low-pass filter.

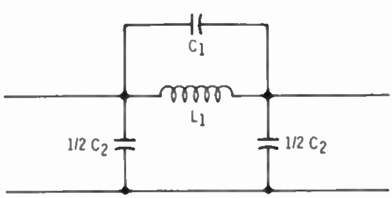


Fig. 2-23. An *m*-derived low-pass filter section. The parallel coil and capacitor form a resonant circuit.

cutoff frequency. Fig. 2-22 is a multisection constant-k filter for reducing TVI from a transmitter output.

Another configuration, called m-derived, uses resonant circuits. This type puts a deeper dip in the curve at some remote frequency beyond cutoff, in the case of a low-pass filter. A single m-derived π-section low-pass filter looks like Fig. 2-23.

Constant-k low-pass filters have better overall attenuation beyond the cutoff frequency, while m-derived filters have a sharper initial drop

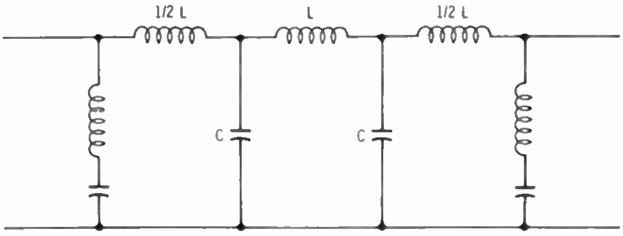


Fig. 2-24. The circuit of a commercial low-pass filter used at the output of a transmitter. The end sections are m-derived, while the center sections are constant-k.

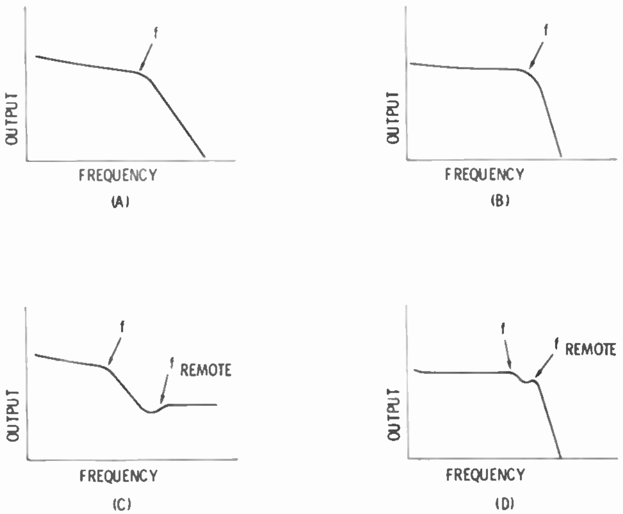


Fig. 2-25. Approximate curves for various filter types. A is that of a single-section constant-k. B shows the sharper drop of a 3-section constant-k. C is a single-section m-derived. D is the actual measurement of a commercial TVI low-pass filter for amateur transmitters. The sharp drop is at approximately 45 MHz. All of these are low-pass filters. The letter f indicates the cutoff frequency; f_{remote} is the dip of the resonant circuit in m-derived filters.

between the cutoff frequency and some remote frequency. Commercial TVI filters for use at the output of a transmitter combine both types (Fig. 2-24).

Fig. 2-25 shows approximate curves of typical filters. The one at D is an actual measurement of a commercial TVI high-pass filter measured in the manner about to be described. The response shown is true only when the filter is terminated with the proper load resistor. The value of the load resistor plays a prominent part in the formulas. If you use a TVI filter in the output of your rig it is most important that it be followed by a perfect 52-ohm resistive load by the antenna feeder. This means a perfect match, or a 1:1 SWR. Otherwise you might not be suppressing higher harmonics, and may even be suppressing some of your operating-frequency signal power.

Filter Formulas

The formulas for constant-*k* filters are:

$$L = \frac{R}{\pi f_C} \text{ and } C = \frac{1}{\pi f_C R}$$

where,

L is in henrys,
C is in farads,
f_C is the cutoff frequency in Hz,
R is in ohms.

The cutoff frequency is represented by f_C, and R is the load resistance. Furthermore, if your feeder is reflecting a high SWR it is also reflecting some reactance, and this also upsets the filter values.

Resonant *m*-derived filters are filters whose LC ratio is related to the load, (R). The factor *m* becomes part of the precise formula. It is a figure somewhere between 0 and 1. The usual value for *m* in r-f filters is 0.6, and is already calculated in the formulas appearing in Fig. 2-26.

These *m*-derived filters are frequently used at each end of a constant-*k* filter system. The two *m*-derived configurations are shown in Fig. 2-26. B and D are complete filters using the *m*-derived sections at the ends.

There are many other configurations for both the constant-*k* and *m*-derived type filters. There is also a whole series for high-pass purposes.

Those shown here are applicable to transmitter output filters for reduction of harmonics above the highest output band (commercial units are usually designed for passing frequencies through the 10-meter

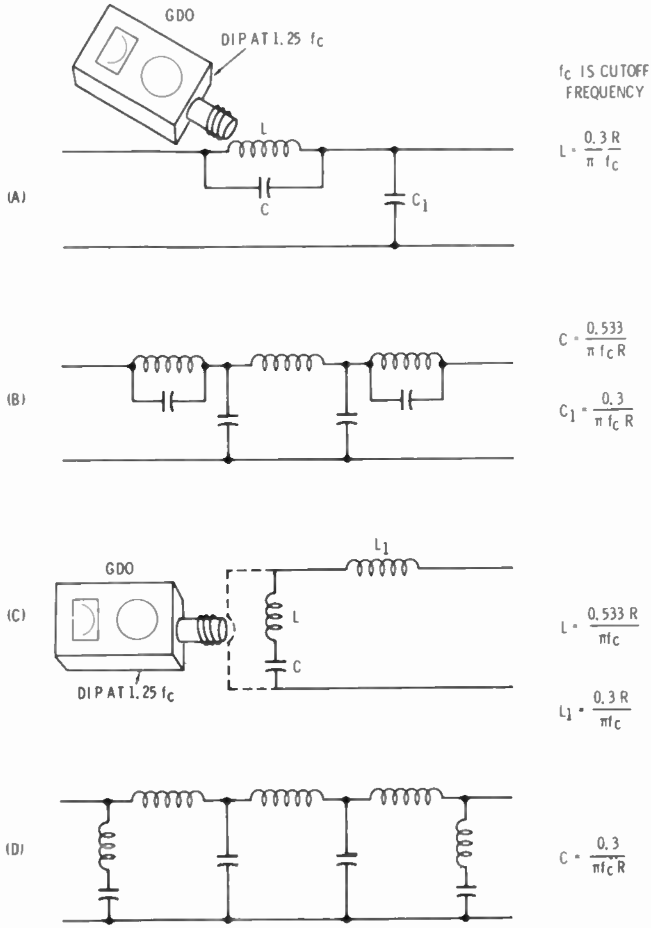


Fig. 2-26. Filters using m-derived sections. At A is a single pi section and the formula for the constants. B is a complete low-pass filter using *m*-derived pi sections on the ends. C and D are like A and B but with T sections. Also shown is how to couple a GDO for measuring *m*-derived sections.

band, and attenuating sharply above 30 MHz) to reduce TVI. The information here will make it possible to measure commercial filters or to design your own.

Audio Filters

Low-pass audio-frequency filters are sometimes used in the speech equipment of transmitters to cut off frequencies above about 3000 Hz. They reduce splatter caused by harmonics of the speech frequencies, especially following a speech clipper which, by its clipping action, is a producer of undesirable harmonics.

Audio filters generally follow the constant-*k* design but seldom adhere precisely to formulas. They are practical compromises because input and output load resistance is seldom the same, and exact values of *L* and *C* are hard to come by at audio frequencies.

Filter Measurements

All you need to measure an r-f filter is a GDO or signal generator and a VTVM with an external r-f probe having high frequency-response limits. The setup is as shown in Fig. 2-27. Two or three turns of wire around the GDO coil is connected in series with a 52-ohm noninductive resistor (any carbon resistor will do), and connected to the input of the filter (filters are symmetrical and either end may be used for input or output). A 52-ohm noninductive resistor is connected across the output.

Use the coil in the GDO with which you observe a pronounced drop in output beyond some frequency. For a transmitter TVI filter this will be at about 35 to 45 MHz. Make a chart of frequency and voltage input at each major division on the GDO dial with the VTVM r-f probe connected to the input of the filter. Make sure the coupling to the GDO is not disturbed and run through the same frequencies charting the output voltages. You will probably use the 1- and 5-volt

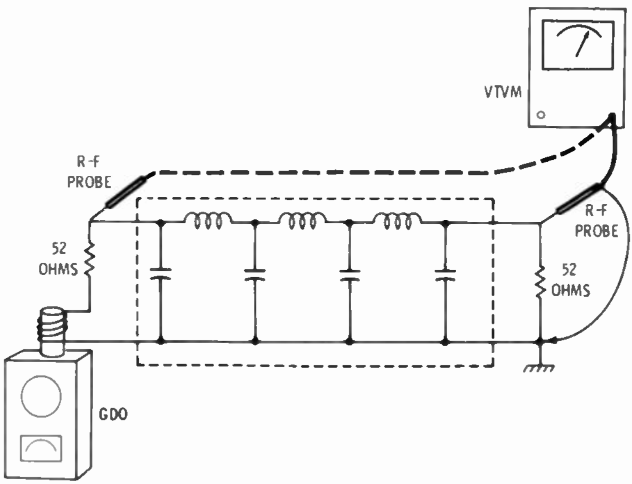


Fig. 2-27. The setup used to measure the attenuation of a filter. Compare the output to input voltages at various frequencies, and draw a curve.

ranges on the VTVM. With a slide rule, chart the ratio between output and input voltages. Plot the ratio figures on log graph paper versus frequency and you will have the voltage attenuation curve for your filter. For converting to dB remember that dB is 20 times the log of voltage ratios. This can be read directly on a slide rule with an L scale. Convert the voltage ratios from the slide-rule D scale to the L scale, and multiply by 20. If you prefer to plot attenuation in terms of power, square the voltage ratios.

Fig. 2-28 is the schematic of a simple, balanced high-pass filter. This, and variations of it, are used at a TV set input to reduce swamping by adjacent ham transmitters. They are designed for 300-ohm lines, the usual TV lead-in. These can be checked as described above, but with 300-ohm resistors at the input and output.

An easy method of checking the filter of Fig. 2-28 is shown in Fig. 2-29. Short the input and measure the resonant frequency by coupling the GDO coil to the short. The resonant frequency is the cutoff frequency (f_c). No resistors are used. This method is good only for single-section filters. In multiple section filters the other coils and capacitors upset the measured resonant frequency.

The method described for multisection filters is used to plot the attenuation of band-pass, or band-attenuation filters, or filters used at audio frequencies.

Fig. 2-28. A simple high-pass filter as used at the input of TV sets. The sketch at the right shows approximate response.

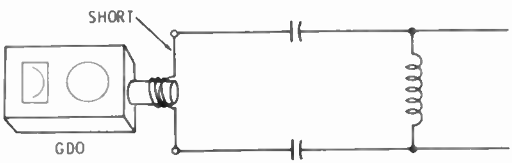


Fig. 2-29. The method of coupling a GDO to a simple high-pass filter. The resonant frequency measured is the cutoff frequency of the filter.

For measuring audio-frequency filters you will need an audio generator for a signal source. Any VTVM and most VOM's will have good audio frequency response on the a-c volts function to serve for measuring, without the need for special probes.

When making measurements of filters at radio frequencies, external leads must be kept short. Connect the load resistor directly from the output to ground. A banana plug makes a good connector for the center terminal of coaxial male connectors on commercial filters. Use only two or three turns around the GDO coil on the input and keep the rest of the connection as short as possible. Long lead lengths become an external inductance and can affect the results of measurement at the higher frequencies.

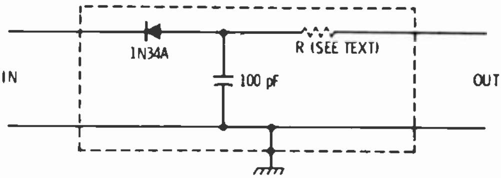


Fig. 2-30. An r-f probe for converting high radio frequencies to dc. The resistor (R) may be omitted if the probe is used with a VTVM.

If you do not have a VTVM with an r-f probe you can make your own. Fig. 2-30 is the schematic. To prevent loading the circuit under test, a series isolating resistor should follow the rectifier. This would be the 1-megohm resistor already part of a VTVM in its d-c probe. If the r-f probe is to be used with a VOM, a resistor should be added. If you are checking filters designed for 52- or 300-ohm loads, a 5000-ohm series resistor will not affect accuracy too much. It will reduce the sensitivity of a 1000 ohms/volt VOM on the 1.5-volt scale by a ratio of 5000/1500. Higher sensitivity VOM's fare better. The output of the probe is direct-current, so the d-c function on the instrument must be used.

While the foregoing measurements were based on the use of a GDO, it should be obvious that a signal generator will do the same thing, provided it has enough output—some don't. In addition, some inexpensive signal generators are somewhat limited in highest frequency fundamental output. Often, the highest band shown on the dial is a calibration of the harmonic of the last fundamental band. Measurements given here, based on a harmonic band, would be completely inaccurate. If you decide to invest in an r-f signal generator be sure to check the highest fundamental frequency it will produce.

CHAPTER 3

Tube and Transistor Testing

As general service instruments, there are two basic forms of vacuum-tube testers. The least-expensive unit tests for emission of electrons from the cathode of the vacuum tube, and is called an *emission tester*. The other applies prescribed voltages to each element of the tubes and is able to measure the G_m—it is called a *transconductance tester*. And among transconductance testers there are two types, one performing a static test and the other a dynamic test, the second including an a-c signal source for the amplification test.

All tube testers also perform other tests on tubes, such as testing for shorts between elements, tests for leakage between heater and cathode, and most also include a test for grid current. Grid current will sometimes result from the contamination of the grid by the bombardment of electrons on it from the cathode, or from a tube becoming gassy.

Tube testers have a number of sockets, each to take a different tube-pin arrangement. The common elements of each socket terminal are wired together. A multi-tap filament transformer connects to a multiple tap switch for selecting the right heater voltage. A bank of switches (one switch for each tube element) sets up the element connections for each tube. A common plate supply becomes self-rectifying through the tube under test. A milliammeter is in series with the plate supply, the cathode, and the other elements of the tube. A variable resistor across the meter is controlled from the front panel. The meter reads "Good-?-Bad" and, in the case of transconductance testers, includes G_m in micromhos. A chart is supplied by the manufacturers with information on setting the element switches, filament voltage, and "plate load" (so-called, but really only a current-adjusting resistor across the meter in the case of emission testers). The chart settings are made by the manufacturer from "bogey," or selected ideal tubes the instrument manufacturer gets from the tube manufacturer.

In an emission checker all the elements from the grid to the plate, inclusive, are tied together. When the variable resistor across the plate milliammeter is set to the number indicated by the chart, pressing the TEST button applies voltage, and the meter will indicate whether the current capabilities or electron flow from the cathode is as much as it should be, is questionable, or is too low for that tube. Assuming no loss of vacuum or contamination of the grid, the result of aging on a tube is loss of electron emission from the cathode. An emission test, therefore, is an indication of age, one of the biggest reasons for tube replacement.

TRANSCONDUCTANCE TUBE CHECKERS

Transconductance or mutual-conductance testers are more complicated and more expensive. They connect the elements of the tube and apply voltages as they would normally be applied to an amplifier, and include a bias voltage on the grid. When the levers and dials are properly set and the TEST button is pressed, the plate meter reads the tubes mutual conductance (G_m) which is a measure of tube gain.

Mutual conductance (also called transconductance and plate transconductance) is a combination of the amplification factor and the *dynamic* plate resistance of a tube. The formula is:

$$G_m = \frac{\mu}{r_p}$$

Amplification factor (the symbol is μ) is the ratio of a change in plate voltage to a change in grid voltage. Dynamic plate resistance equals the plate-voltage change divided by the resultant plate-current change:

$$r_p = \frac{\Delta E_b}{\Delta I_b}$$

where,

r_p equals dynamic plate resistance,

ΔE_b equals change in plate voltage,

ΔI_b equals change in plate current.

G_m may also be defined as a *small change* in plate current divided by a *small change* in grid voltage with plate voltage held constant. This more correctly defines dynamic mutual conductance. This is measured only by the best of tube checkers; most mutual-conductance testers check with fixed voltages and perform what is called a *static* test.

The figure for mutual conductance is in *mhos* (the inverse of ohms, and is *ohms* spelled backward). Actual figures for *mho* is in small fractions, so the scale is marked in *micromhos*, or millionths of a *mho*.

A better test of a vacuum tube, therefore, is on a mutual-conductance tube checker, with the dynamic mutual-conductance test preferred. Even this is not a positive assurance the tube will work in a particular circuit. It is still possible that the tube will not work as an oscillator, or as an amplifier at very high frequencies. The final evaluation is how it performs in the circuit for which it is intended. A check on the tubes in a tube checker does reduce the possibility of circuit malfunction being due to the tube or tubes.

IN-CIRCUIT TUBE TESTING

Tubes are easily checked while in-circuit by knowing the operating characteristics. Tube manuals give the information. Chapter 1 described voltage checks that can be made at the grid and plate of a tube. By these checks a tube can be evaluated for current, which proves emission as well as other operating parameters.

SEMICONDUCTOR TESTING

There are a number of instruments for testing diodes, transistors, and other semiconductors. The less-expensive instruments test for leakage, and relative transistor *alpha* and *beta*. The more expensive ones include a test for high-frequency cutoff.

Semiconductors do not emit electrons from a cathode or any element, so the most important function of a vacuum-tube tester does not apply to testing semiconductors. Transistors have indefinite life if not abused. Operating them beyond their rated limits will usually destroy them. From a practical standpoint, therefore, the average ham wants to know if a diode or transistor is good or bad. Simple checks take simple instruments.

CHECKING DIODES AND TRANSISTORS OUT OF CIRCUIT

An ohmmeter is a handy instrument for checking out a semiconductor as to whether it is good or bad. It will also check leakage. It will *not* give you gain measurements, but only tell you if the device is open or shorted.

Semiconductor devices have rectifier action. That is the secret of diodes particularly. Current will pass in one direction more easily than the other. Therefore, comparing the current in one direction with the current in the other indicates, to a degree, the quality of a diode or transistor.

The first thing you must do is determine the voltage polarity of the leads out of your VOM or VTVM in the ohmmeter function. As explained in Chapter 1, the ohmmeter function is a battery in series with resistors. In VOM's the ground lead is positive (+) and the red lead negative (−). In VTVM's it may be reversed. Check the schematic of your instrument to ascertain which polarity of the battery goes to ground.

Start with a diode. With the positive lead connected to the anode and the negative to the cathode, your VOM or VTVM should read a low value. The amount of resistance measured will vary with the diode and with the range selected on the ohmmeter. Now reverse the connections to the diode. You should read infinity or nearly so on silicon diodes, and several hundred thousand ohms on germanium diodes. This is a measurement of reverse leakage, and it will be higher for silicon diodes. A low reading means high leakage, and the diode should be rejected.

The forward reading (+ to anode, − to cathode) will vary depending on the range being used. At the beginning of conduction, diodes are not linear. A typical curve looks something like Fig. 3-1. There is

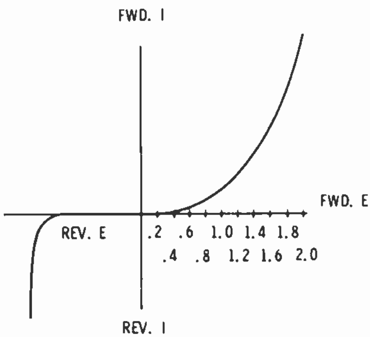


Fig. 3-1. Typical characteristic curve for a semiconductor diode. Note nonlinearity of curve at low current values.

very little forward conduction below the knee of the curve. (There is even less in the reverse direction until you get to the point of avalanche where current increases quite suddenly. This is the phenomenon used for zener diodes.) You can calculate the current drawn in the forward direction by remembering the theory of the ohmmeter.

The internal resistor in series with the battery is the center-scale reading times the range-multiplier figure. If the center scale is 10 and you are on the $\times 1$ range, the internal resistor is 10 ohms. If you are on the $\times$ 100K ohms range, the internal resistor is 1 megohm ($10 \times 100{,}000 = 1{,}000{,}000$).

Let's take an example. A good "top-hat" power-type silicon diode, the 1N1653, will measure infinity in the reverse direction. It measures 200,000 ohms on the $\times$ 100K range and only 9.5 ohms on the $\times 1$ range, in the forward direction. How much current is it passing? On the 100K range, the total resistance is 200K (the amount measured) plus 1,000,000 (the internal resistor), or 1,200,000 ohms. Dividing 1.5 (the battery voltage) by this value is 1.25 microamperes. On the $\times 1$ range the total resistance is the 9.5 ohms measured plus 10 ohms for the internal resistance, or 19.5 ohms. Dividing into 1.5 volts, the current is 77 mA. At 1.25 μA there is hardly any conduction and the internal forward resistance is high. At 77 mA the current is in the range of normal operation for this diode. This condition is true for all diodes although the ratio between high and low current may not be as great, and will not be as great for germanium diodes, especially the small-signal types. On small-signal diodes it is possible to exceed the current ratings when on the $\times 1$ range and destroy the diode. You are pretty safe in using the $\times 10$ range, and all that is important is that there be a reasonably high difference between forward and reverse measurements.

TRANSISTORS

The same method is used to check whether a transistor is passable or bad. You can tell if it should work or not work, but not *how well it will work*. In transistors you have 6 combinations of connections—3 combinations each in two directions of current. On silicon transistors the resistance between collector and emitter will be high (ohmmeter reading nearly infinity) in either direction. On germanium transistors the readings will be fairly high, but higher in one direction than the other. The more important test is between base and collector and base and emitter. In the forward-current direction the resistance readings should be fairly low as compared to the reverse-current readings. Again the actual reading will depend on the ohmmeter range being used. A zero reading means the element is shorted, and an infinite reading that the element is open. The actual gain of the transistor cannot be determined by an ohmmeter test. Again, it is necessary to use caution against applying too much current to the base in the forward direction. In operation, the base bias current is in the range of microamperes (μA); they are just not built to take too much current.

Keep in mind that whether you are applying forward or reverse current depends on whether the transistor is an NPN or a PNP type. An easy way to remember forward-biasing polarity is that the center letter is the required battery polarity for forward biasing when applied to the base. An NPN transistor requires a positive (+) voltage applied to the base and a PNP transistor a negative (−) polarity to the base. In schematic diagrams the emitter arrow points to the outer circle on an NPN transistor, and to the base vertical line on a PNP transistor. The arrow points in the direction of *current* flow but opposite the direction of *electron* flow. This adds a bit to the confusion

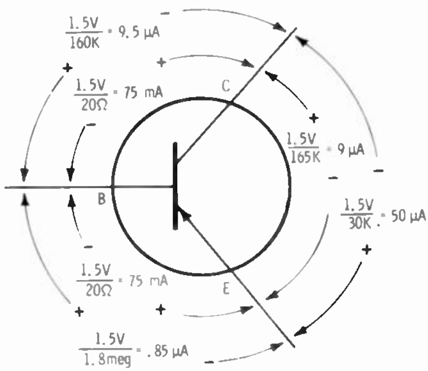


Fig. 3-2. The results of testing an unmarked transistor with a VTVM having an ohmmeter battery of 1.5 volts.

about in what direction electricity flows. The arrow points in the direction of the movement of "holes" in semiconductors, and hole drift is called current in this case.

Fig. 3-2 shows VTVM readings on an unmarked transistor. Note the very low resistance (and dangerously high current) when forward-biasing voltage is applied between either base and emitter or base and collector.

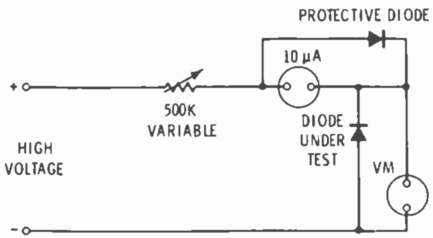


Fig. 3-3. The hookup for a PIV test on a silicon power-rectifier diode.

DIODE MEASUREMENTS

Small-signal diodes used as detectors need only to show good forward-to-reverse current ratios. The check for this is with an ohmmeter as described previously.

Power-supply rectifiers are another thing. The important parameters are the current-carrying capacity, and the peak-inverse-voltage limit. The maximum current rating is the current the diode can carry without overheating the junction internally, with consequent destruction of the diode. The peak inverse voltage (called PIV for short) is the highest reverse voltage that can be applied without breakdown. Approximate measurement of these parameters can be made by the amateur with a minimum of equipment.

For a PIV measurement you will need a source of high voltage (the amount depending on the rating of the diode), a microammeter, a variable resistor for adjusting current, and another diode for protecting the meter against excessive current flow. The circuit is shown in Fig. 3-3. Say you have a number of "top-hat" silicon rectifiers. Most of them have a PIV rating of 400 volts, and a forward-current rating of 750 mA. Your power supply should be able to produce about 500 volts output. Connect as shown in the sketch. The diode under test must be connected in reverse polarity. Start with the full resistance of the variable resistor. The microammeter may read around 10 μA, with only a slight rise as you reduce the value of the series resistor. Watch the meter carefully. As soon as the meter begins to show a rapid rise in current, note the voltage across the diode and immediately shut off the power. The voltage across the diode *before* a fast current rise is the PIV rating.

The forward-current test requires a low-voltage, high-current supply. (Fig. 3-4). A battery eliminator is a good one. The series resistor must have a high current rating. A 200-ohm, 200-watt variable resistor is required for this diode example. As you turn down the re-

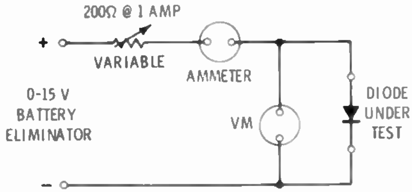


Fig. 3-4. Hookup for checking power rating of diodes. Voltage across the diode times current through it is the power dissipation. If the figure is higher than the published rating, discard the diode.

sistance value, watch both the ammeter and voltmeter, until the product of the two equals the wattage rating of the diode. The rectifier mentioned for this example may have a 1-watt rating. On this basis, the voltage across the diode would be about 1.5 for a current of 750 mA. If the voltmeter reads higher than 1.5, discard the diode. Higher-rated diodes will have a lower internal resistance, and voltage across the diode will be something less than 1.5. The higher the current rating is the lower is the voltage across it. The manuals on rectifier-type silicon diodes show the normal voltage to be expected across them.

ZENER- AND SIGNAL-DIODE MEASUREMENTS

You can run a curve on zener diodes and on small-signal diodes using the simple setup of Fig. 3-5. The finished curve will look something like that in Fig. 3-1. That part of the curve to the right of the vertical curent line is made with the battery polarity applying positive voltage to the anode of the diode under test. To complete the curve to the left of the vertical line, reverse the battery polarity. The schematic shows a 1000-ohm series limiting resistor to protect the diode against excessive current which might destroy it. You can either use a volt-

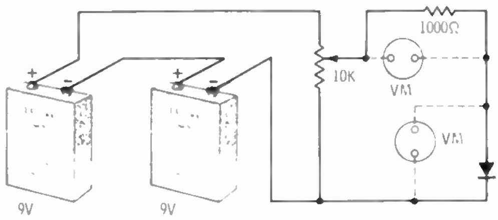


Fig. 3-5. The hookup for drawing a characteristic curve for a diode.

meter to read the voltage drop across this resistor and convert to current ($I = E/R$), or connect a milliammeter in series. Make a chart of current versus voltage as you turn the potentiometer up, then transfer the points to a graph. Reverse the battery polarity and plot a similar curve. Reverse current will be extremely low as you go up in voltage in the negative direction, until you reach a point where the current will suddenly take a jump. This is the *avalanche* point. It establishes the maximum inverse voltage you should safely apply to a diode (except zeners).

Zener diodes are specially designed to run continuously at the avalanche point. The zener voltage-regulating rating is the voltage reached at the point of avalanche. As you increase voltage beyond this point, only the current will increase; the voltage across the diode will remain constant. This is what makes zener diodes good regulators. Fig. 3-6 shows a simple low-voltage power supply with a zener diode

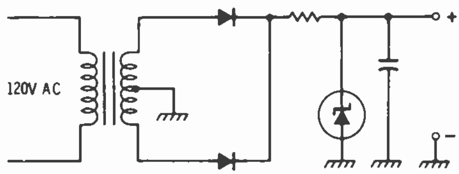


Fig. 3-6. A constant output voltage at the zener diode rating is the feature of this power-supply circuit.

rated at 8 volts across the output. The voltage to the diode may vary due to fluctuations in the a-c line voltage, but the voltage across the diode will always be 8 volts. The current through the diode will increase or decrease with voltage fluctuations preceding it.

For measuring zener diodes with a higher than 8-volt rating, you will need a higher-voltage supply, of course.

TRANSISTOR MEASUREMENTS

Instruments for measuring the standard bipolar transistor vary from a simple go/no-go type (a simple one you can build yourself is described at the end of this chapter) to laboratory types with high-frequency oscillators built in. The most popular instruments make leakage tests and measure d-c alpha or beta, which is considered sufficient to check out transistor quality.

To determine what it is that is important in transistor specs, it is necessary to review the meaning of the specs, or at least those in which hams would be interested, and on which they might be able to make some measurements. The data sheets and transistor manuals are full of spec symbols each of which has a specific meaning. There are over a hundred symbols, but only a few are significant to the amateur. Most symbols are almost self-explanatory. Here are a few of the important ones:

V_{CBO} Volts between collector and base with the remaining element (emitter) open. The first two letters of the subscript may be any two of the three elements of a transistor, and if followed by an *O* means the third element is open. If the last letter is an *R*, it means a resistor of specified value is connected between the remaining element and the second letter subscript. If the last letter is *S* it means shorted, etc. for other letters.

I_C Collector current, usually in milliamperes.

h_{FE} Forward-current or static transfer ratio for a common emitter configuration. It is also known as d-c beta.

$$\text{\textit{d-c beta}, or } h_{FE} = \frac{I_C}{I_B}$$

The ratio will vary with different values of collector current, and with temperature. Spec sheets usually show a minimum figure.

h_{fe} Small signal forward-current transfer ratio, with a-c output short-circuited. It is the ratio of a small change in collector current to a small change in base current. The purpose of the small change is to keep it on the linear portion of the curve. At low audio frequencies the h_{fe} is often the same as the h_{FE}, but is lower at higher frequencies. Spec sheets frequently give the figures for two frequencies; for example, at 1 kHz and at 20 MHz.

f_t Gain-bandwidth product frequency at which h_{fe} is unity. It is the highest frequency at which there is no further amplification.

Alpha (α) is the ratio of collector current to emitter current for a common-base circuit configuration, and beta (β) is the ratio of collector current to base current, usually in a common-emitter configuration. The two ratios are related, as follows:

$$\alpha = \frac{\beta}{1 + \beta}$$

and

$$\beta = \frac{a}{1 - a}$$

COLLECTOR CHARACTERISTICS CURVE

Probably the most important graphical presentation of the operating characteristics of a transistor is in the collector-characteristics graph, which plots collector volts against collector current for various base currents. The graph for a typical amplifier transistor looks like Fig. 3-7. From it, both d-c and a-c beta (h_{FE} and h_{fe}) can be determined.

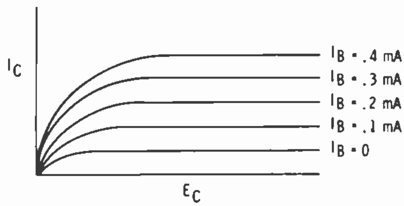


Fig. 3-7. With the setup shown in Fig. 3-8, you can run transistor characteristic curves, which look like this.

With simple equipment you can run your own set of collector-characteristics curves.

Fig. 3-8 is a schematic of a circuit using inexpensive batteries for the base and collector power sources—10,000-ohm potentiometers are used for adjusting battery voltages. While separate meters are shown in each circuit (for measuring base current, collector current, and collector voltage), a single sensitive VOM could be used by switching it around from place to place. A voltmeter could be used to measure base current by measuring the voltage across the 1000-ohm base-limiting resistor and using Ohm's law ($I = E/R$). Adjust the base voltage for 0.5 mA and plot collector current for each 5-volt change in collector voltage. Move the base voltage up to 1 mA base current and plot another curve of collector current versus collector voltage. Draw additional curves for increased values of base current. The final set of curves may look something like Fig. 3-7. Small-signal transistors may require making the base-current increments in microamperes, not exceeding 1 mA. This set of curves will tell you about all you need to know about the amplifier characteristics of an unknown transistor, except its high-frequency response.

Fig. 3-9 is the hookup for observing the collector characteristics on an oscilloscope. The low-voltage half-wave power supply supplies

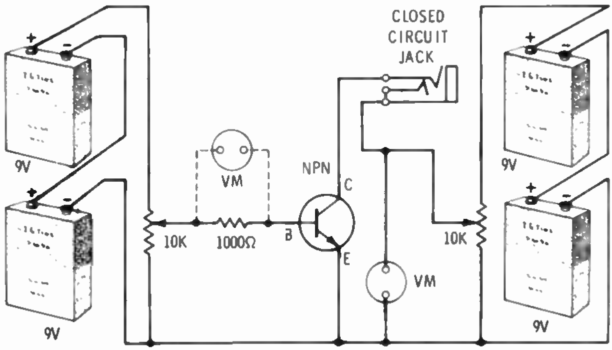


Fig. 3-8. A simple setup for running transistor characteristic curves.

an a-c voltage superimposed on a d-c voltage. The horizontal sweep is carried across the CRT during the forward conduction of the rectifier

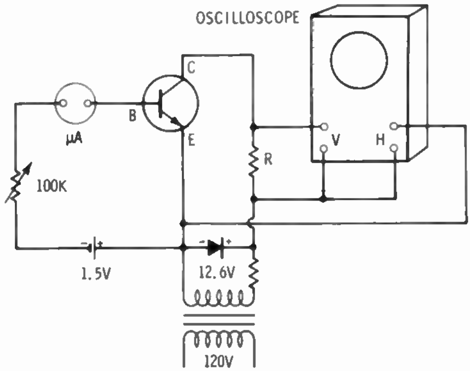


Fig. 3-9. With a hookup like this, a continuous curve of transistor collector characteristics can be observed on an oscilloscope. The curve will be different for each value of base current.

diode. The trace returns to the beginning during the second half of the wave when there is no voltage. At the same time the pulsating d-c voltage is applied to the collector through a load resistor. The vertical deflection on the scope is from the a-c voltage across the load resistor. The value of *R* should be as low as possible and still provide full-scale vertical deflection. The more sensitive the vertical amplifier of the scope is, the lower the value of *R* can be.

Observe the trace on the scope with each change in current in the base circuit. The meter in series with the base circuit can be a microammeter or a milliameter, depending on the transistor under test. By taping a piece of translucent paper over the face of the CRT, you can trace the curves for each of the settings of base current. It will look something like Fig. 3-10.

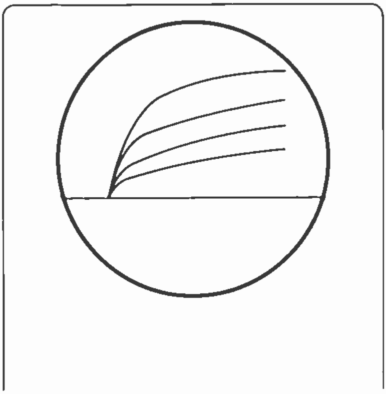


Fig. 3-10. Place a piece of translucent paper over the face of the CRT of Fig. 3-9. Trace each curve for each change of base current. The final tracing will look something like this.

FIELD-EFFECT TRANSISTORS

The transistors we have been talking about are called *bipolar* transistors because their operation depends on the movement of both electrons and holes in the material. A more recent development, and one that is more important to the amateur, is the *unipolar* transistor, called FET (field effect transistor). Some FET's operate on the movement of electrons, some on the movement of holes, but not both.

The FET has characteristics very much like the vacuum tube. Its input (the terminal is called gate instead of base) has very high impedance. It does not load down an input resonant circuit as does the conventional bipolar transistor, and it, therefore, makes an excellent r-f amplifier.

Tests and measurements on the FET depend on the type. Most FET's are of the *depletion* type. In them a slab of N-type silicon is connected between the *drain* and *source* (so-called instead of collector and emitter) on each side of which is a glob of P-type material connected to the *gate*. With no voltage on the gate, current will flow in the drain-source circuit if a voltage is applied. As a negative bias is applied to the gate, electron depletion occurs around the P-type material resulting in a constriction in the flow of current from drain to source. As bias is increased, a point is reached at which no current flows in the drain-source circuit. This is the *cutoff point* similar to vacuum tubes (Fig. 3-11).

Some FET'S are *enhancement* type, with exactly the opposite effect. Current does not flow in the drain-source circuit until a bias is applied. Some FET's are called enhancement/depletion type in which some current flows in the drain-source circuit with no bias, and current is increased or decreased by the application of bias, either negative or positive. Most FET's are the earlier mentioned, depletion type.

The all important drain-source current characteristic chart or graph will show drain-source current versus voltage plotted against gate-difference voltages (instead of current) and will look like Fig. 3-12. The method of measurement is similar to that of the bipolar transistor,

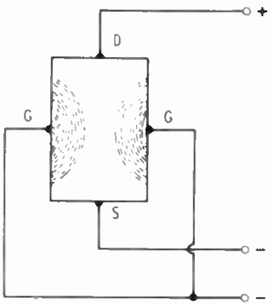


Fig. 3-11. As bias voltage to the gate of a FET is increased, the area around the gate becomes more and more depleted of electrons. The resulting constriction is the control of current through the transistor.

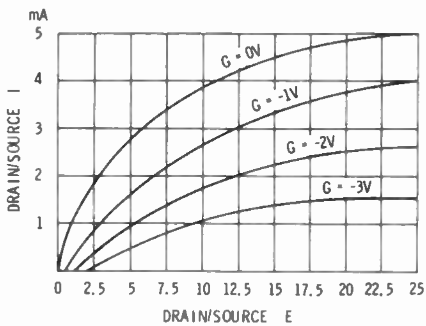


Fig. 3-12. The drain/source characteristic curves for a FET transistor.

except for a voltmeter across the input voltage source instead of a milliammeter. The circuit of Fig 3-18 is easy to put together for the measurement.

Begin measurements by setting all voltages to 0. Plot the figures of drain voltage versus drain current for small increments of drain voltages from 0 to maximum with the gate voltage at 0. Depending on the transistor, saturation may flatten out the curve as you go up in voltage. If it does not, be sure you do not exceed the maximum rated current and voltage listed in the transistor manual. Turn the gate voltage up to −1 and plot another curve under conditions of −1-volt bias. It will fall somewhere below the curve for 0 bias. Draw additional curves with increased bias voltage. At one bias-voltage point there will be no, or very little, drain current, regardless of drain voltage. This is the cutoff bias.

HANDLING FETS

One of the most important points to remember in testing a FET is the handling. You may have noticed that when you buy a FET it comes with the leads twisted together or wrapped in a metallic foil. The leads are shorted together to prevent a static charge from accumulating on the gate. The gate P-material is insulated from the N-type slab of the drain-source circuit by an oxide coating which acts as an insulator. This is one of the important factors which gives the gate such high-impedance characteristics. The metallic oxide coating is very thin, and static charge accumulation on the gate could puncture the thin coating and make the FET inoperative. Even the static charge of the human body, if you pick up a FET by the gate lead and touch the drain or source lead to ground, may be enough to puncture the coating.

A TRANSISTOR TESTER

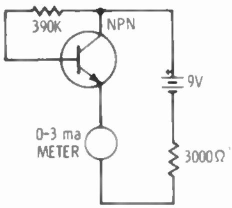
The instrument shown in Fig. 3-13 is home constructed. Anyone can build one. It is a tester of small-signal type transistors to show, to a limited degree, whether they are good or bad.

This tester will indicate the presence of excessive leakage between the collector and emitter, and it will indicate the forward-current transfer quality. While it does not provide the actual figure for d-c beta (h_{FE}) you can determine whether one transistor has more d-c forward gain than another. It will determine whether the transistor is NPN or PNP if this is not already known. The universal-type sockets will accommodate the TO-5 type transistors with leads in a

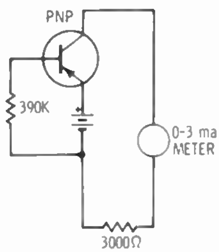
Fig. 3-13. A home-made small-signal transistor tester. It shows relative leakage and static direct-current gain.

semicircle, or the flat-type transistors with leads in line. More important than anything else, the cost to build it is small.

In Fig. 3-14 you will recognize the simplest of amplifier circuits, with the base-bias resistor returned to the collector. This provides some positive (+) bias for the NPN transistor. The 390 K resistor limits the base current to a few microamperes. While a lower value resistor here will give a better deflection of the meter, any transistor under test must be protected against damage by the instrument. The deflection is adequate, and all transistors are fully protected.



(A) Circuit for NPN transistors.



(B) Circuit for PNP transistors.

Fig. 3-14. The fundamental circuit of the tester is shown here in two parts. In the actual unit, polarities are switched.

Fig. 14B is adapted for use with PNP transistors. The 390K resistor is returned to the negative (—) side of the battery for biasing, and the collector-to-emitter polarity is reversed.

Any 3-ma meter movement, even the cheapest you can buy, will do for an indicator in the collector-to-emitter circuit. A 9-volt battery, of the type found in so many transistor radio sets, provides more than enough current for about any of the small transistors. The 3000-ohm resistor is subject to change, depending on the meter movement you use. The resistor value to choose is that which, when a short is applied between the collector and emitter terminals of the socket, the meter will just read full scale. You can determine this by Ohm's law or by trial. The Ohm's law formula is:

$$R = \frac{E}{I}$$

If E is 9 volts and I is 3 ma (or .003 amp), substituting,

$$R = \frac{9}{.003} = \frac{9000}{3} = 3000$$

The resistance of the meter movement must be deducted from the figure of 3000 ohms. Inexpensive meter movements may have several hundred ohms of resistance. If, for example, the resistance of the meter alone is 700 ohms, the series resistor will be 3000 — 700 or 2300 ohms. The meter movement used in the instrument shown in these photos had a resistance of 50 ohms (this one just happened to be handy). The 50-ohm resistance was negligible compared to the total resistance so was not figured in the calculations. A total resistance of slightly more than the 3000 ohms, merely means slightly less deflection in gain tests. Even a 10-percent error is not important to the function of this instrument.

The trial method of selecting the collector limiting resistor is to place a 3000-ohm variable resistor in the circuit and, with a short between collector and emitter, adjust the variable resistor for full-scale deflection. Then measure the value with a VOM and replace the variable resistor with one of fixed value, or the nearest standard value. You can, however, leave the variable resistor in the circuit: in fact, the variable resistor allows you to make adjustments later to bring the indication up to full scale, as the battery voltage drops from use or age.

The complete circuit is shown in Fig. 3-15. To reduce switching complexity, separate sockets are used for NPN and PNP transistors. The only switch in the circuit is used to provide bias of the correct

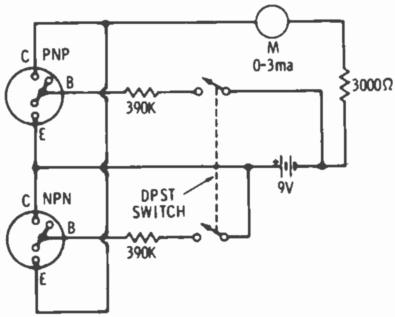


Fig. 3-15. The complete circuit diagram. The switches shown are parts of one double-pole single-throw (dpst) switch.

polarity to the base of the transistor for gain measured. With the switch down, the meter reads forward-current gain.

Basic construction is on the panel of a small inexpensive meter case. You may build your tester any way you wish, but complete enclosure in a case, as shown, makes a neat compact instrument. Parts placement and wiring are not critical as to layout, and any arrangement of sockets, meter, and switch will do. The method shown here is simple and inexpensive (Fig. 3-16).

Before you do any cutting you will need to have the parts on hand. Different meters require different sized holes for mounting them on the front panel.

Besides the meter case, panel, and the hardware, you will need:

- 1 0-3 ma d-c meter, any quality and any size.
- 1 9-volt battery (Neda 1604).
- 1 dpst (double-pole, single-throw), spring-return switch.
- 2 universal transistor sockets.
- 1 2-terminal wiring tie-point.
- 1 390,000-ohm, ½-watt resistor.
- 1 3000-ohm, ½-watt resistor.
- 1 snap-on battery connector.

The small panel for the two sockets is 6" × 3½". The sockets are mounted on the front panel.

Follow the layout shown in the back view if you wish. A bracket for holding down the battery is bent from a thin sheet of metal fastened to one of the meter-mounting screws, but any method you devise will do.

The meter scale is home made, as can be seen from the photo (Fig. 3-17). The original scale showing 0-1-2-3 in milliamperes may be used as well. To make your own like this requires removal of the case from the meter, and careful handling to prevent damage to the needle and internal mechanism. With the case off, measure the diameter of the scale on the meter, and draw a circle of the same size on a blank sheet of paper. Estimate the arc made by the needle and draw an arc on the new scale. Put in the markings shown. Cut out the new scale, and cut off the lower ⅓ of the circle. Apply rubber cement to the back and *carefully* slide the new scale under the needle and press it down against the present scale on the meter move-

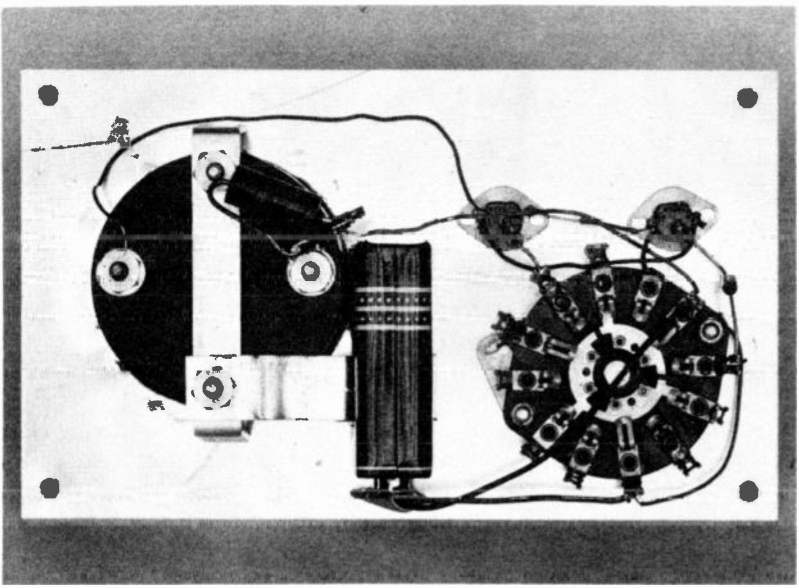


Fig. 3-16. Rear view of the transistor tester showing simplicity.

ment. You may have to cut a small semicircle out of the lower part of the scale to keep from interfering with the movement of the needle. Determine this before you apply the rubber cement, by trying it in place. The PNP and NPN markings on the panel were made from transfer type covered by a coat of clear-plastic spray.

The transistor sockets are available with mounting flanges or for ring mounting. This model used ring mounting. A 11/32-inch hole was a good friction fit for the sockets without the need for the rings. The four holes provide for the insertion of the two common types of transistors. The three holes in a semicircle are for TO-5 type transistors. The in-line holes are for in-line transistor leads, with the two holes at the ends common to collector and emitter for both

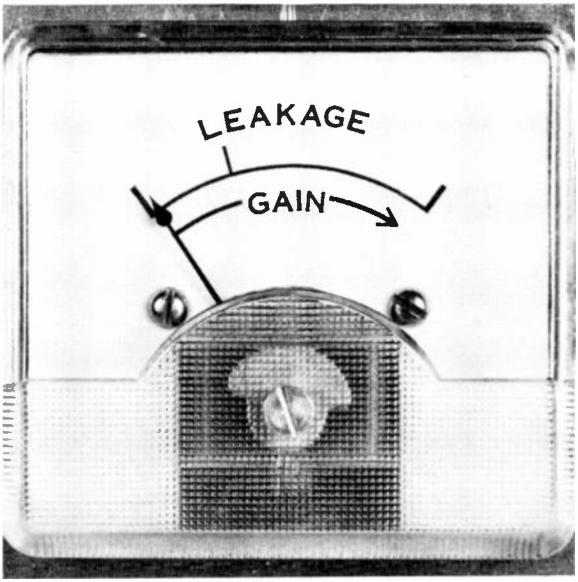


Fig. 3-17. Closeup of the home-made meter scale.

lead styles. The one close to the emitter hole is base, and is wired to the other base terminal.

All connections, except to the meter posts, are by soldering. Use a resin-core solder only, and a well-tinned soldering iron. One of the small pencil-type irons rated from 23.5 to 40 watts is ideal. Any size up to 100 watts will do, however. Hold the hot iron to the terminal to be soldered for the count of three, then apply the solder. Just a small amount of solder should flow into the crevices of the connection, then remove the iron, and wait until the connection has cooled before disturbing it.

You can purchase snap-on connectors for the battery, but since the battery will last for a long time, solder connections to the terminals can be used if you so desire. It will be a long time before replacement becomes necessary.

USING THE TESTER

If the transistor is a PNP type in a TO-5 case or a similar one, with the three leads in a semicircle, insert the leads into the three outermost holes of the PNP socket. Be sure the leads are gripped by the metal terminals inside. Try pulling gently on the leads. If the transistor is a silicon transistor you may get no or very little deflection before you press the switch. Any movement at this point up scale indicates leakage between collector and emitter. Germanium transistors may show about ¼-scale movement upward, and this may be normal. Pressing the switch puts forward bias on the base of the transistor and the needle should move upward, about ⅓ scale. Any deflection higher than the leakage deflection means the transistor is in working condition. If there is no movement upward, try the transistor in the other socket, as it may be incorrectly identified. An NPN transistor in the PNP socket will show no deflection, and vice versa.

If the needle goes to full scale when the transistor is first inserted, the transistor is shorted and should be discarded.

This instrument is for checking small-signal NPN and PNP transistors only; it is not for checking FET (field-effect transistor) or UJT (uninjunction transistor) types. Both of these types will show what appears to be high leakage when first inserted into their sockets, and this is normal. The FET will show a slight drop in deflection when the switch is depressed. The UJT will show no change when the switch is pressed.

TRANSISTOR MANUALS

In addition to individual spec sheets on transistors, the principal manufacturers also produce complete manuals on their line of transistors. Complete specifications are shown, and general data and typical circuits are frequently included. These are available at your electronic parts dealer. Perhaps the most complete listing of all transistors is the Howard W. Sams "Transistor Specifications Manual."

CHAPTER 4

Receiver Measurements

Whether it is a complete receiver alone, or the receiver portion of a transceiver, measurement of receiver performance is desirable either in part or complete. Important performance characteristics of a receiver are *sensitivity, selectivity, image response, calibration accuracy,* and sometimes *input impedance* and *noise factor.*

Sensitivity is the receiver's ability to amplify weak signals. The limit of sensitivity is external and internal noise, which masks weak signals no matter how much gain the receiver may have. This is referred to as the S/N Ratio, or *signal-to-noise* ratio. A more accurate designation is S + S/N or *signal plus signal-to-noise* ratio. Overall gain of a receiver is stated in microvolts (millionths of a volt) signal input that produces a given audio output—usually one-half watt. But a receiver with high gain can lose its value in on-the-air performance if the internal noise is high and masks weak signals. Both the S + S/N ratio and gain are important in evaluating a receiver's ability to perform on weak signals. Communications receiver sensitivity is expressed in microvolts of r-f input to produce a signal 10 dB above the receiver internal noise. Selectivity is the ability of the receiver to separate signals so the desired signal stands out from all others. Selectivity is expressed in a *bandwidth in kHz,* which represents a band of frequencies, the end frequencies of which are 6 dB down from the center frequency. When a signal is 6 dB down it is received at half the power it would be as compared to when tuned on-the-nose. High selectivity (about 0.5 kHz) is desirable for c-w signals; wider selectivitv is necessary (about 3 to 5 kHz) for phone signals. Too high selectivity on phone clips the sidebands and affects intelligibility.

Image response is the ratio of sensitivity to the signal being tuned to the sensitivity to the image of that signal. An image of the signal can be heard at twice the intermediate frequency above or below the signal frequency, depending on whether the high-frequency oscillator of the receiver is above or below the signal frequency. Image response is affected by the selectivity of the "front end," or those stages which tune the signal frequency and which precede the i-f stages. Poor image response can result in interference from another signal twice the intermediate frequency away, if it is a strong signal.

Calibration is the ability to read the received frequency accurately on the calibrated tuning dial. It is expressed in percent of the dial reading and is a function of the accuracy with which the high-frequency oscillator tracks with dial calibrations.

Input impedance is the load in ohms the receiver input reflects to the antenna feeder as a source of signal. When the load impedance is equal to the source impedance, best transfer of power is accomplished in any electrical system. No receiver has a constant and perfect load impedance clear across each band. Input-impedance matching on receivers is a matter of good design, but a reasonable mismatch should not be taken seriously. It has little effect on overall performance. Published figures represent sort of an average across the band, on each band.

Noise figure has to do with noise generated internally in a receiver. It is expressed by the letters "NF" and is given as so many dB above a theoretically perfect noise-free receiver. Internal noise is greatest in the mixer stage and is primarily due to "shot effect," which is the random irregularity of electron flow. Good preamplification overrides mixer noise. Noise can originate in preamplifier stages and in tuned circuits, and dielectric resistance leakage in components and transit time in tubes produce noise. The higher the frequency of the tuned circuits is, the higher is the noise. In addition, the motion of molecules in tuned circuits produces thermal agitation noise and is related to temperature.

MEASURING INSTRUMENTS

Instruments are important to measurement, of course. Some are elaborate, and some are simple enough to build yourself.

The most important instrument for work on a receiver is the *signal generator*. Basic signal generators consist of a high-frequency oscillator, an a-m modulator, sometimes a buffer stage, and, of course, the power supply to power this instrument. All include an attenuator of some sort, and some have an r-f voltmeter connected to the output, and preceding the attenuator.

Courtesy Electronic Instrument Co., Inc.

Fig. 4-1. The Eico Model 324-K signal generator has output from 150 kHz to 145 MHz on fundamentals. The built-in 400-Hz modulator is variable up to 50 percent.

Quality varies considerably in signal generators. Better instruments will have more stable oscillators with more accurate calibration of the tuning dial. Wide frequency coverage is achieved in the same way as in receivers, by bandswitching (Fig. 4-1). An audio oscillator modulates the r-f oscillator. Inexpensive units have only one audio frequency (usually about 400 Hz) and the percent of modulation is about 30 percent. Better instruments may have a variable-frequency audio oscillator and provide up to 100-percent modulation by modulating a later stage. For amateur measurements, a single audio frequency with 30-percent modulation is usually sufficient. When the built-in r-f voltmeter is employed, the instrument is better known as a *Microvolter*. Any good external VTVM with high r-f response can can be used to set the initial reference output voltage, however. Other factors affect quality and therefore usability. Good r-f shielding is essential to prevent r-f leakage from the internal circuitry to the receiver in order to be able to read low signal levels more accurately. A well-designed step attenuator will have a minimum of stray capacitance in order to maintain accuracy of attenuation at the higher frequencies, and be itself well shielded. The power supply should be regulated to prevent line-voltage fluctuations from affecting the frequency of the oscillator.

A sweep generator (Fig. 4-2) is something like a signal generator, but with some important differences. The amplitude modulator is eliminated. The high-frequency oscillator is made to *sweep* across a band of frequencies at a regular rate. The sweep generator must be

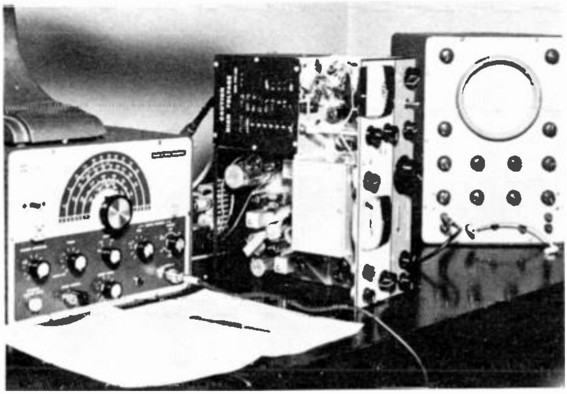


Fig. 4-2. A sweep generator and scope connected to an amateur tranceiver. The trace of Fig 4-3 is an actual photograph of the selectivity curve obtained from this hookup.

used with an oscilloscope. When the generator sweep rate is synchronized with the horizontal sweep on the oscilloscope, the scope pattern becomes a visual picture of the selectivity response of the receiver to the frequencies being swept. Sweeping the frequency means that the oscillator frequency is changing constantly over a range of frequencies. For example the oscillator can be set for a center frequency of 7.4 MHz and swept over a range of from 7.35 MHz to 7.45 MHz. If the receiver is tuned to 7.4 MHz, the scope will show the highest vertical deflection to be at 7.4 MHz, and the lower amplitudes on each side, indicating the attenuation of frequencies below and above 7.4 MHz. This is seen on the scope as a curve like the one shown in Fig. 4-3, which is called the *selectivity curve* of the receiver.

A number of methods are used to sweep the high-frequency oscillator. The oldest uses a variable capacitor, the rotor of which is attached to a motor. The motor spins the rotor of the capacitor. The capacitor is in the high-frequency oscillator circuit. With its constantly changing capacity there is a constantly changing frequency. At least one generator uses a powdered-iron slug attached to the voice coil of a speaker. The speaker is made to vibrate at a 60-Hz rate from the power-line frequency, thus causing the slug to move in and out of an inductor, which is part of the high-frequency oscillator circuit. A currently common method used in inexpensive sweep generators uses a *saturable reactor*. The band coils of the oscillator, either as separate coils or as taps on a common coil, are wound on a powdered-iron core. Also on the core is a winding of heavy wire. A varying rate of direct current is passed through this extra winding. At high current the core

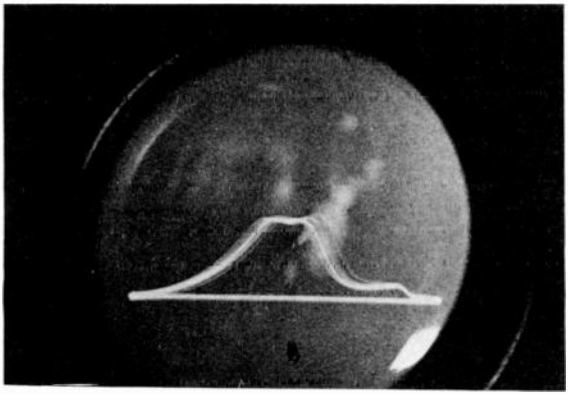
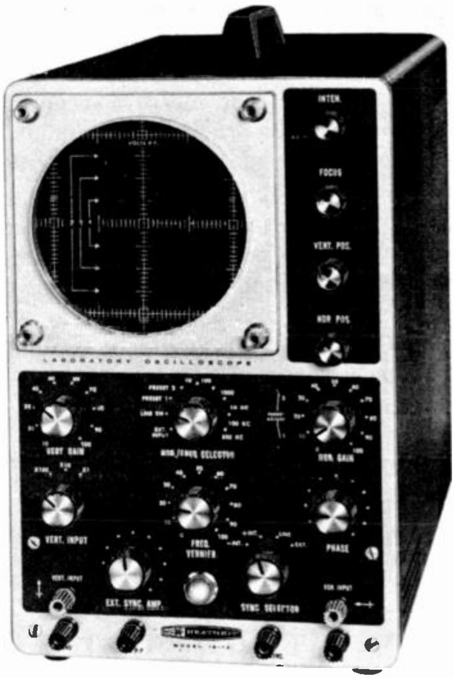


Fig. 4-3. Selectivity curve as shown on the scope face.

of the inductor saturates, and the inductance is lowered. At lower current, the inductance increases. As the inductance is made to change, the oscillator frequency changes in step with the scope horizontal sweep. A fourth method uses either a variable-reactance tube circuit, or a *varicap* which is a solid-state diode whose internal capacitance changes with varying direct current through it. These act as variable capacitors in the oscillator circuit. These last two methods are electronic. In these methods the variable voltage is obtained from a sawtooth oscillator which provides a sweep linearly from one frequency to another, then a quick return to the first frequency.

An oscilloscope (Fig. 4-4) must be used with a sweep generator. An oscilloscope is also important in observing audio distortion and audio-frequency response in speech amplifiers and modulators. Since transmitter modulators are intentionally limited in frequency response, and a small amount of distortion is tolerable, measuring the response and distortion only satisfies one's curiosity. A scope may be considered an inertialess VTVM. The beam of the cathode-ray tube is deflected without inertia and is therefore almost instantaneous. The deflection plates of the CRT in the scope take several volts to deflect the electron beam, therefore they are preceded by amplifiers, one for vertical deflection and one for horizontal deflection. Quality and price of a scope are based on size of the CRT, gain and bandwidth of the amplifiers, and variety of functions. Even the most inexpensive scope is adequate for most amateur needs. A modification of the oscilloscope is used to monitor the modulation of transmitters. This will be described in greater detail in a later chapter.

A *crystal calibrator* (see Chapter 7) is a crystal-controlled oscillator that produces accurate signals for pickup in the receiver. The



Courtesy Heath Co.

Fig. 4-4. A wide-band oscilloscope which can be built from a kit. The vertical amplifier sensitivity is .025V/inch deflection, and has a response of ±1 dB from 8 Hz to 2.5 MHz.

simplest consists of a 100-kHz crystal in a single tube or transistor oscillator, designed to be rich in harmonics. When coupled to a receiver, a signal will be heard every 100 kHz as far up in frequency as 30 MHz. A small variable capacitor in the crystal circuit allows adjusting the frequency to beat with WWV as a standard, thus providing extremely accurate frequency spots. A simple 100-kHz crystal calibrator should be permanently connected to your receiver for a quick check of the dial frequency every time you use it. Better crystal calibrators include multivibrator stages, one set to oscillate at 10 kHz, and often one oscillating at 1 kHz. These lock in with the 100-kHz oscillator for accurate 10- and 1-kHz spots. Some calibrators start with a 1-MHz crystal and include multivibrator circuits to give outputs every 1 MHz, 100 kHz, 10 kHz and 1 kHz. Each can be heard separately by switching.

Noise generators are just what the name implies. The shot effect of a solid-state or vacuum-tube diode produces a random-frequency output which covers the entire useful amateur range. They are excellent for peak-aligning the front end of a receiver. Tracking of r-f stages in a receiver is easy with a noise generator. It is only necessary to peak trimmers and coil slugs for maximum noise in the output of the receiver. An easy-to-build noise generator uses a solid-state diode. A more elaborate one, and one that can be used to make actual S+S/N measurements, uses a 5722 vacuum-tube diode which is designed to produce a measured amount of noise. Circuits for both are given later in this chapter.

The grid-dip oscillator (GDO) should be the number-two instrument in the ham shack, next to a VOM or VTVM. Among its many uses is that of a signal generator. It is too unstable and inaccurate for precision frequency measurements, but it does produce tunable signals for rough calibration and for alignment. It is an excellent instrument for receiver design use, in that it may be used to set all tuned circuits very near the proper frequency while cold, that is, with no power applied to the receiver.

MEASURING SENSITIVITY

Overall sensitivity is expressed as the signal level required to produce a certain amount of audio power in the output (usually one-half watt) or the signal level required to overcome the internal noise. The latter is used more frequently for ham-type receivers, and is the signal input needed to produce an output of 10 dB above the internal noise. A common specification may read: Sensitivity—1 μV for 10 dB S/N. This means 1-microvolt, 30-percent amplitude-modulated signal will be heard in the output, 10 dB above the internal noise.

It must be kept in mind that sensitivity figures published in the literature or in catalogs attempt to show the best figures, and so apply to measurements made on the lowest band. Sensitivity is hardly ever the same across any one entire band or on each different band of a receiver. Since receiver noise increases at higher frequencies, and circuit *Q*'s are lower, it would be a mighty difficult feat to engineer the same high sensitivity into the higher bands as the lower.

It takes a good signal generator or microvolter to make meaningful sensitivity measurements on a high-gain ham receiver. The signal generator must be well shielded, and it must have a good step attenuator. For signal outputs down around 0.5 to 1 microvolt, all of the signal must come from the output jack and be properly stepped down from the attenuator. With poor shielding, your attenuator may read 1 microvolt, but the actual signal may be much more, either from direct radiation from the instrument or around the attenuator network to the output jack. The higher the frequency is the more critical r-f leakage becomes.

In addition to the signal generator you will need a VOM or VTVM for making output readings. It is connected to the output across a resistor which replaces the speaker. The VOM or VTVM must be sensitive enough to read the internal noise of the receiver on its a-c or output function.

A speaker is an inductive load to the receiver output and is, therefore, frequency sensitive. That is, voltage readings across the speaker may be different for different audio frequencies of the same level. The voltage reading of random noise may be different from that of a 400-Hz audio tone, for example. Therefore, output measurements are made across a resistance of the same impedance value as the speaker substituted for the speaker.

SENSITIVITY HOOKUP

Temporarily disconnect one leg of the speaker from the amplifier output, and connect a 3.5- or 4-ohm resistor across the output. If your receiver has a 500-ohm line output, so much the better. Disconnect the speaker and connect a 500-ohm resistor across the 500-ohm output. Set the VOM or VTVM to the a-c or output function, and connect it across the resistor. If the resistor is a 3.5- to 4-ohm resistor, you will need to use the most sensitive range; if the output is 500 ohms, a higher range setting may do. Unless the VOM or VTVM has a very low, low-range scale, the speaker output may not have enough voltage to give a good reading. In this case, connect the VOM or VTVM across the primary of the output transformer.

Fig. 4-5 is the hookup for measuring sensitivity. Connect the signal generator to the input of your receiver with shielded cable. Most signal generators have a 50-ohm termination at the end of the output cable. If you operate your receiver from a 50-ohm coaxial antenna feeder, connect the generator output cable directly to the antenna input of the receiver. If you normally operate your receiver from a random length of wire for an antenna, a "dummy-antenna" network must be connected between the output of the generator and the receiver. A dummy-antenna is easy to make; the schematic is shown in Fig. 4-6. Keep leads as short as possible and enclose the network in a metal box—it must be well shielded to prevent stray pickup.

Turn on the receiver with the r-f sensitivity control at maximum and bandswitch on the lowest band. If the measurement is to be made

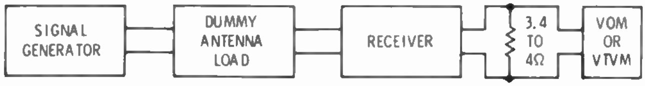


Fig. 4-5. Hookup for sensitivity measurement. If you operate your receiver from a 50-ohm feeder, omit the dummy-antenna network.

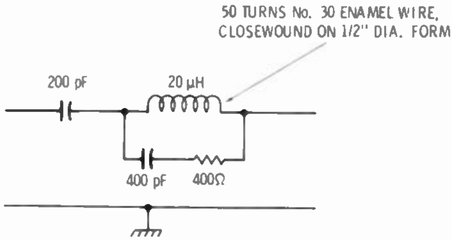


Fig. 4-6. This dummy-antenna network acts like a random-length antenna. Use it between the signal generator and receiver input.

on the basis of input microvolts versus audio power out, the audio volume control must also be at maximum. If the measurement is to be made for dB above the noise, the setting of the audio volume control is not important, since very little noise is created in the audio stages. Be sure the receiver is not tuned to a strong signal by plugging in the headphones and listening. You should hear only noise as a rushing sound. Between the audio volume control and range switch on the VOM or VTVM, make adjustments so the noise reads 0 (zero) dB on the meter scale. You are now set up for a beginning reference point.

It is best to stabilize the signal generator by turning it on about one-half hour before measurements are made. Set it up for 30-percent internal modulation. Adjust the tuning dial and band to the frequency of the receiver. If the signal generator has a built-in reference r-f voltmeter, adjust the variable-output control for a 0.1-volt reading on the meter. If it does not have a built-in meter, disconnect the generator from the receiver, transfer your VTVM (a VOM won't do here) adjusted to read 0.1 r-f volt to the output of the generator. With the step attenuator for minimum attenuation or maximum output, adjust the variable control for 0.1-volt indication on the VTVM. Reconnect the VTVM to the output of the receiver, and the signal generator to the receiver input. Do not touch the variable control on the signal generator; adjustments of output level must be made with the step attenuator, until the end of the test when the variable control may need to be used.

The step attenuator may be calibrated in dB. It must be translated to output in microvolts. Set at zero-dB attenuation, the output is 0.1 r-f volt. At 20 dB, the output is 1/10 of 0.1 volt, or .01 volt. At 40 dB, divide by 10 again (which equals .001 volt) and so on down the line. For each 20-dB position of the switch, move the decimal point to the left one digit. At 100-dB down, the output will be 1 microvolt (μV). For readings between 0.1 μV and 1 μV, and between 1 μV and 10 μV, adjustments will be made with the variable control for multiplying factor.

With the attenuator set at 60 or 80 dB, rock the tuning control on the signal generator for maximum reading on the output VOM or VTVM. If the reading is off scale, move the attenuator down another 20 dB. Set the attenuator for the nearest reading on the VOM or VTVM to 10 dB above the zero-noise reference point. Should the nearest setting be, for example, 8 dB, then turn the variable control on the generator up to where the output meter on the receiver reads 10 dB above 0. The voltage reading on the signal-generator output meter will be the multiplying factor for voltage output above 1 μV; in this case 1.5 μV, which means that the sensitivity of the receiver is 1.5 μV for 10 dB S/N. If the receiver output meter reads, for example, 13 dB, reduce the variable control on the generator to the 10-dB point on the receiver output meter, and multiply the 1 μV reference by the generator meter reading. In this case a reading of 0.7 μV on the generator would bring the output meter down to 10 dB, and the sensitivity of the receiver would be 0.7 μV for 10 dB S/N.

A rating in dB is an expression of *power* ratio and is based on the formula:

$$\text{dB} = 10 \log \frac{P_{out}}{P_{in}}$$

Every 10 dB represents a 10-times change in power ratio. Since, in Ohm's law power varies as the *square of voltage*, voltage ratios expressed in dB are based on the formula:

$$\text{dB} = 20 \log \frac{E_{out}}{E_{in}}$$

For every 20-dB change in voltage there is a 10-times change in voltage ratio.

Many less expensive signal generators have their step attenuators marked in ratio of voltage output, such as × 0.1, × .01, etc. These provide a direct ratio of generator reference voltage to its output.

In a number of medium-priced generators the output step attenuator will not go down as low as 1 μV. The output of these generators at minimum is much too high to measure a good communications receiver, but they can be used with an external fixed attenuator. Fig.

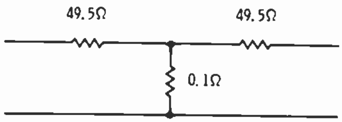


Fig. 4-7. A 60-dB attenuator made of three resistors.

4-7 is the schematic of a 60-dB attenuator which may be connected between the output of the signal generator and the input to the receiver, and will divide the generator output by another 1000.

Sensitivity measurement based on power output is more like an overall gain measurement, disregarding internal noise. The setup is like the foregoing dB measurement but without the use of zero-dB noise-reference reading on the VOM or VTVM output meter. Instead it is based on the modulated signal required at the input to produce a certain power level (usually 0.5 watt) at the receiver output.

The hookup is the same as before. Adjust the signal-generator output to produce 0.5 watt of audio power in the output. In this measurement, the audio volume control must be at maximum. Since the reading is in volts, the selected voltage must be related to the power using Ohm's law:

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

or,

$$E = \sqrt{RP}$$

Assuming a 4-ohm resistor and solving for E at 0.5-watt output:

$$\begin{aligned} E &= \sqrt{4 \times .5} \\ &= \sqrt{2.0} \\ &= 1.4 \text{ volts} \end{aligned}$$

where,

P equals output power in watts,
E equals voltage across the load resistor,
R equals the resistance of the load.

Adjust the signal generator to produce 1.4 volts output. The r-f voltage into the receiver is the sensitivity of the receiver using this method.

STAGE-BY-STAGE GAIN

If you have checked your receiver for sensitivity and found it wanting, you will be curious to know where the trouble is. A stage-by-stage gain measurement may be made which may reveal one stage in trouble. If the zero-dB noise reference can be established, but the sensitivity figure is way below what it ought to be, the trouble is probably in the front end, or r-f stages. If no noise is measurable, or is barely heard with gain wide open in a normally sensitive receiver, gain has been lost in the i-f or audio stages.

Most signal generators have a jack for using the 400-Hz modulator externally. Connect this to the high side of the audio volume control or across the second-detector load resistor. An audio signal of 0.5 to 1 volt in should produce the rated power output. The output should be measured across a dummy-load resistor replacing the speaker as described before. If, for example, the output is rated at 1 watt, a voltage of 2 volts across the resistor should be obtained with the 0.5 to 1-volt input to the audio stages. Again, Ohm's law is used:

$$E = \sqrt{RP}, \text{ or } E = \sqrt{4 \times 1}$$

or 2 volts, if a 4-ohm dummy-load resistor is used.

A modulated r-f signal is used to measure the gain of i-f and r-f stages. Connect your VTVM across the second-detector load resistor and set the VTVM function to read dc on the 1-volt range. Connect the signal generator to the primary of the last i-f transformer and adjust the variable control on the generator for a 0.1-volt reading on the VTVM. The signal generator should be set at the intermediate frequency of the receiver (or approximately so) to begin with. Single-sideband (ssb) receivers do not have a second-detector load resistor. In that case, make connection to the secondary of the last i-f transformer, or to the grid of the tube or base of the transistor which follows.

On ssb receivers you must kill the bfo (or carrier-injection signal) by pulling out the tube if the receiver is a vacuum-tube receiver, or by shorting the oscillator coil if the receiver is a transistor receiver.

A demodulator probe must be used on the VTVM or VOM to convert the modulated-rf to audio. Fig. 2-30 in Chapter 2 is the schematic of a simple demodulator probe you can make if necessary.

Move the signal generator output cable to the grid of the last i-f tube, or the base of the last i-f transistor. Turn the signal-generator attenuator down until you get the same reading (0.1 volt) on the

VOM or VTVM. The difference in reading is the gain of the stage. If the attenuator is marked in decimal dividers, the gain is a direct reading with in-between attenuator range readings taken by adjusting the variable control. If the attenuator is marked in dB, the terms must be converted to gain as a direct ratio, although some gain measurements are made in dB. Since the gain is in voltage rather than power, interpreting dB to ratio means there is a voltage gain of 10 for each 20 dB.

Move the signal generator output cable to the next preceding i-f stage and make a gain measurement. The added attenuation required to reduce the output is the gain of that next preceding stage in dB. For voltage gain ratio, divide the overall gain of both stages by the gain of the last stage, to get the gain ratio of the stage alone.

When moving the signal generator from stage to stage it is important that the output leads are dc-isolated to prevent shorting the stage inputs. Check between hot terminal and ground with a VOM on ohms. If there is continuity, use a .01-μF capacitor in series with the hot lead of the signal generator. An isolating capacitor is not necessary at the receiver antenna input.

Continue on down the line with each stage from right to left (as viewed on a schematic) until the antenna terminals are reached. Intermediate-frequency stages have very high gain because of the very high Q of the i-f coils—gains of several hundred are possible. The converter stage usually has little gain, perhaps only about 10 or less, because they are designed for best conversion efficiency rather than gain. R-f stages, too, have much lower gain than i-f stages, because of the relatively poor Q of the tuned circuits. A receiver fault is found by noticing a substantial difference between two similar stages. If you find this is not so, then suspect rather bad misalignment, which we will discuss later.

Stage gain is the voltage ratio between output and input:

$$\text{Gain } (G) = \frac{E_{out}}{E_{in}}$$

MEASURING SELECTIVITY

The selectivity of a receiver may be measured, or a curve plotted, by the same signal generator described before. In fact it need not be as good a signal generator. The most important quality the signal generator must have is good bandspread on the dial, and a tuning mechanism with no backlash. It is important to be able to read small frequency changes each side of a center frequency.

A sweep generator and oscilloscope are ideally suited to measuring selectivity, while at the same time observing the shape of the selectivity curve. This setup is a great time saver when making adjustments to the tuned circuits— you see what happens as the adjustments are made. An auxiliary variable-frequency signal generator or crystal calibrator is also necessary to properly use a sweep generator. It adds small pips to the curve on the scope screen to establish reference points, and is called a marker generator.

While some selectivity is contributed by a well-designed front end in a receiver, nearly all meaningful selectivity is in the i-f stages. Signal frequencies are converted to a single intermediate frequency for the very reason that at one lower frequency, circuit Q is much higher. While overall selectivity can be measured by connecting the signal generator to the antenna terminals, it is easier to make the measurement at the intermediate frequency, and the results are practically the same.

THE SELECTIVITY HOOKUP

To make proper selectivity measurements it is necessary to make a couple of minor operations on the receiver. Because avc action affects measurement results, it will be necessary to kill the avc. With the bottom plate of the receiver off, and the receiver turned on but with no signal tuned in, measure the fixed bias in the avc line with a VTVM. Parallel this fixed bias with an external voltage source of the same value. The easiest method is to connect some flashlight cells in series, and connect the string across the avc line, making sure the polarity is the same as the original. If, for example, you have measured −3 volts of fixed bias on the avc line, two flashlight cells in series will do. The low internal resistance of the cells will prevent an increase in bias when a signal is applied, and a fixed-bias value of −3 volts will be retained. The second operation is to kill the high-frequency oscillator, since it may beat with the signal generator and show false signals. In a vacuum-tube receiver pull out the h-f oscillator tube. If the tube is multi-purpose, or if the receiver is transistorized, you can kill oscillation by inserting a piece of wire or metal foil between the plates of the oscillator section of the tuning-capacitor gang. Be careful not to be bend the plates—insert metal just enough to short the rotor and stator plates of the capacitor.

Connect a VTVM across the diode load resistor in the second detector with the function switch to read d-c voltage, or connect a VOM or VTVM across the speaker terminals with the function switch set to read a-c volts. Ssb receivers with a product detector only do not have a second-detector diode necessary to convert the r-f signal to audio for measurement by the VTVM. In this case it is necessary to provide a diode detector by using a demodulator probe on the VTVM.

A connection must be made between the signal generator and the input of the converter stage. If the receiver uses vacuum tubes one of the easiest methods is to lift the shield from the converter tube from above the grounding tabs and clip the hot lead of the signal generator to the shield and the ground lead to the chassis. This provides capacitive coupling to the elements of the converter tube. On a transistorized receiver clip the hot lead to the base of the converter transistor. Tune the signal generator to the intermediate frequency, and set the function for r-f output only if the VTVM is across the diode load resistor, or set to 400-Hz modulated-rf if a VOM or VTVM is connected to the output. The hookup is shown in Fig. 4-8.

In a dual-conversion receiver the generator should be connected to the first converter which follows the preamplifier stage or stages. While the higher intermediate frequency of a dual-conversion receiver is used to reduce image response, it does add to selectivity as well, and should be included in an overall selectivity measurement.

With the receiver on and the r-f sensitivity control full on, adjust the frequency of the signal generator for maximum deflection on the

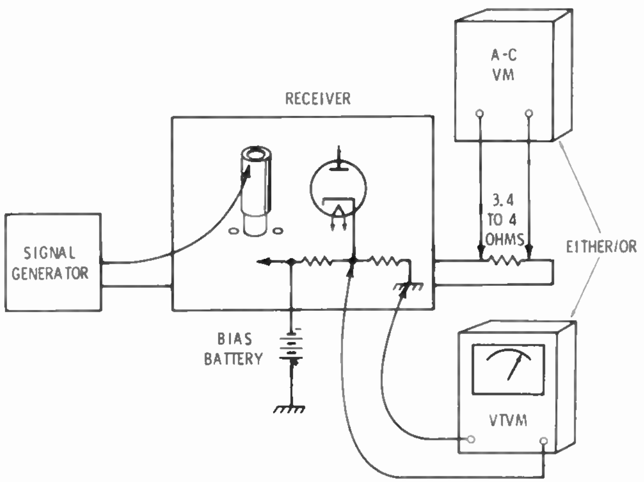


Fig. 4-8. Hookup for checking selectivity. The lifted converter-tube shield provides capacitive coupling to the plate of the tube.

meter. Keep the level of the signal generator as low as possible, and yet get a good deflection on the meter. An overloaded receiver will give erroneous results. Adjust the level of the signal generator to give zero-dB reading on the dB scale of the meter. Carefully note the exact frequency at which peak output is obtained. Slowly move the signal generator tuning dial above the peak frequency until the meter reads −6dB, and note the frequency. Do the same below peak frequency. Subtract the lower frequency from the higher frequency, and the difference is the bandwidth of the receiver at the 6-dB points. For example, the receiver has a 455 kHz i-f. Peak output on the VOM or VTVM occurs when the signal generator is tuned to 455 kHz. As we move the signal generator dial down and watch the meter, the −6dB point is reached when the signal generator dial reads 453.5 kHz. Moving the tuning dial up in frequency, the meter needle rises again to the zero-dB point and drops again above 455 kHz. It reads -6dB at 456.5 kHz. The 6-dB points are, therefore, 453.5 kHz and 456.5 kHz, and the difference between the two is 3 kHz, which is the bandwidth of this receiver at the 6-dB points.

Selectivity figures are sometimes also given for points at 50-dB down. The method of measuring is the same as for the 6-dB points except the frequency extremes are observed for a meter drop to 50 dB below the zero point both above and below the center frequency. If, in the foregoing example, the bandwidth were 30 kHz at the 50-dB points, it would indicate a good receiver for phone reception.

A selectivity curve can be plotted using a signal generator. The method is as described above, except that dB figures are plotted against frequency over a wider range below and above the center frequency. In the previous example, begin by setting for zero dB at the center frequency, then move the signal generator down to about 440 kHz, and chart dB against frequency for each kHz up to about 470 kHz. The chart may look something like this:

CENTER FREQUENCY IS 455 kHz			
Dev. From Center (kHz)	dB	Dev. From Center (kHz)	dB
−5	−50	+1	−1
−4	−48	+1.5	−6
−3	−45	+2	−20
−2	−20	+3	−45
−1.5	−6	+4	−48
−1	−1	+5	−50
0 (455 kHz)			

The next step is to convert this to ruled paper to see the selectivity curve. Fig. 4-9 is a selectivity curve based on the figures above. Note that this is in dB which is a logarithmic method of expressing a ratio. Since our output meter measures in volts, basically, the dB scale on it follows the ratio dB $= 20 \log \frac{E_1}{E_2}$. If your meter does not have a dB scale you will need to convert volts to dB using this formula. The right hand Y axis of Fig. 4-9 shows the voltage ratio and how it relates to the logarithmic ratio.

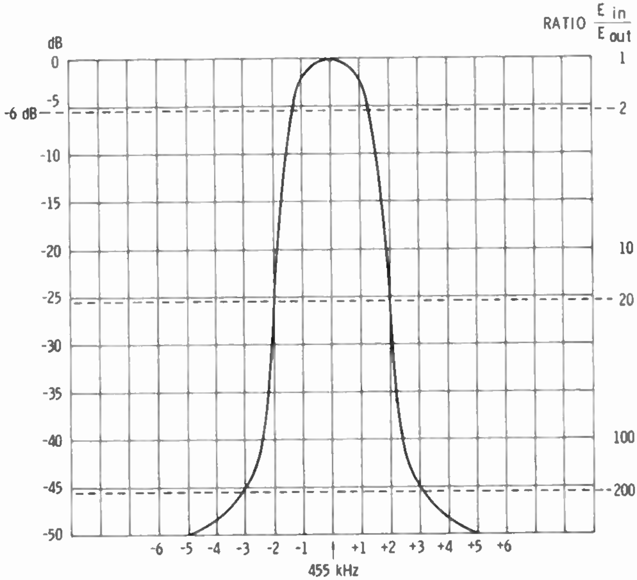


Fig. 4-9. A receiver selectivity curve plotted from the figures tabulated in the text.

On a selective receiver it is often not possible to reach the lower dB outputs by detuning the signal generator. At a bandwidth of 10 kHz, the output meter may appear not to read anything. In such cases a reverse method may be used. As you tune this signal off center frequency, increase the output of the signal generator to maintain a zero-dB reading on the output reading. Plot the amount of increased output in dB or voltage ratio required of the signal generator to maintain the zero reference, using the step attenuator. An advantage of this method is that it gives you a feel of amount of power an interfering station must have off frequency to interfere with the station you have tuned-in on frequency. In the curve of our example a station 5 kHz off frequency would have to have a voltage at the antenna terminals about 630 times the center-frequency station to compete on an equal level. The 630-times voltage ratio is about 50 dB.

THE SWEEP GENERATOR FOR SELECTIVITY

There is a great advantage to being able to see an instantaneous response curve of selectivity, especially if adjustments are to be made to the i-f tuned circuits. The sweep-generator frequency modulates the r-f signal at a 60-Hz rate taken from the a-c line frequency. The signal of varying frequency passes through the receiver i-f system, whose response or amplification varies with the frequency. The varying response is shown on a scope and the pattern or trace is a direct product of the selectivity of the receiver. The horizontal, or X, axis of the trace is frequency, and the vertical, or Y, axis is voltage. Precise frequency identification is obtained by moving a marker pip along the trace line.

Some sweep generators have built-in marker generators, a second variable-signal generator, or a crystal-controlled oscillator. These are superimposed on the sweep frequency and show up as "pips" on the scope trace and identify the frequency at any spot along the trace. A variable marker can be made to move along the trace, while a crystal oscillator makes several pips along the trace. Also used instead of a variable oscillator is a built-in absorption circuit. The absorption type tends to "suck-out" some of the output frequency, and so shows up as a dip on the scope trace. The absorption type is actually easier to work with since it does not distort the trace, as oscillators sometimes tend to do. The best method for showing frequency marks on the trace is with a marker adder. This is usually a separate instru-

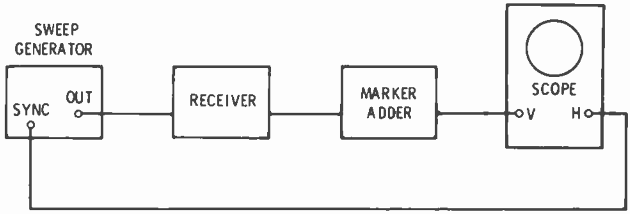


Fig. 4-10. A marker adder is a combination mixer and variable or crystal oscillator, which adds a marker signal between the receiver and scope. It does not distort the trace.

ment, into which the output of the receiver is fed, and which in turn feeds the scope. It is a mixing and amplifying instrument. Since the marker-generator output does not go through the receiver, it cannot affect the gain, and result in trace distortion (Fig. 4-10).

READING THE SCOPE

The Y-axis response being voltage means that the 6-dB down point is at half scale, or a two-to-one voltage ratio between peak response and the 6-dB point (dB $= 20 \log \frac{E_2}{E_1}$, or 20 log 2, or 6).

Thus, with markers you can measure the bandwidth at 6 dB down. Since the vertical amplifier responds linearily to voltage, it is next to impossible to measure the bandwidth at higher attenuation such as 50 dB down.

THE SWEEP HOOKUP

Connect the sweep generator and scope in the same way as described for a signal generator and VTVM, with one exception. A VTVM has an isolation resistor in its probe to prevent the capacitance of the cable from affecting response at the detector-diode load resistor; a scope has not. Therefore, it is necessary to add a resistor in series with the hot lead of the scope input cable. A 50K resistor is about right. In the case of a product detector, again, a demodulator must be connected to the i-f output, followed by the 50K resistor, and the scope lead (Fig. 4-11).

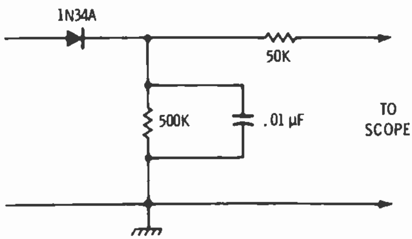


Fig. 4-11. Demodulator and isolating resistor circuit, for use between the ssb receiver i-f output and a scope.

After the hookup described, the sweep generator should be turned on for about 20 or 30 minutes to stabilize its frequency, or the pattern will drift across the scope screen. After warmup, turn on the receiver and scope. Set the scope horizontal-amplifier sweep-selector control to LINE, or 60-Hz sweep. This will match the usual 60-Hz sweep rate of the sweep generator. Adjust the focus and intensity for a horizontal line on the CRT. Turn the sweep generator sweep-width control to zero. With no sweep the generator acts like a standard signal generator, but is not amplitude modulated. Set the band-switch and tuning dial on the sweep generator for the input i-f of the receiver. Rock the tuning control until a deflection is seen on the scope trace. Since the sweep generator is not modulated, there should be no pattern on the scope when on frequency, but rocking the tuning dial develops a low-frequency audio voltage in the output which causes the horizontal line on the scope to move up and down. Now, turn the sweep control up a little at a time. A hump should appear on the horizontal line of the scope. This is the selectivity curve. The curve may appear right side up or upside down, depending on the phase, which in turn is dependent on the number of amplifying stages involved. It is not important which way the curve appears. An inverted curve can be righted by reversing the connections to the vertical-deflection plates on the CRT, but it usually is not worth the trouble. The sweepwidth adjustment should be just enough to show some horizontal line at each end of the curve. The object is to achieve a fully visible trace of the selectivity curve, not too crowded and not overly wide, but just enough to see all parts of the line, its shape and width, as well as height.

Center the trace curve by making slight adjustments to the frequency dial on the sweep generator. If it tends to distort the pattern, make the adjustment simultaneously with adjustment of the phase control on either the sweep generator or the scope. The object is to obtain a centered pattern with straight lines out from each end of the curve.

The output of the sweep generator should be the minimum amount that will provide a clean pattern on the scope without a lot of "fuzz" (noise). Remember that the receiver is to be operated with its sensitivity control wide open. Overloading the receiver with too great a signal will distort the pattern.

Some sweep generators do not go low enough in frequency to work directly into the low i-f stages of a communications receiver. Most low-priced sweep generators are designed for use with TV sets, with i-f's in the megahertz range. In that case, feed the sweep generator into the antenna input with the receiver tuned to one of the lowest amateur bands. In this method, the high-frequency mixer oscillator must not be killed, of course. Connection between the output cable and the antenna terminals must be as close as possible using a minimum of unshielded wire on the hot lead to avoid external signal pickup.

Some sweep generators have a blanking circuit which eliminates the return trace on the scope, and only one pattern appears on the CRT. On sweep generators without the blanking circuit, there will be a curve when the trace moves from left to right and another when the trace moves from right to left. Tuning the sweep-generator frequency-control dial makes the two patterns move in opposite directions. In this case, try to get the patterns together so that they overlay each other. Since sweep generators are primarily designed for TV-set adjustment their maximum sweep width is quite wide. This will require a very careful adjustment of the width control, with the control barely cracked open for the very narrow bandwidth of a ham receiver.

Adjust the scope centering controls (height and width) to fill about ¾ of the CRT face.

ADDING MARKERS

If the sweep generator has a built-in variable marker oscillator, set its frequency to that of the sweep-generator frequency, and turn up its output control a little at a time, until a pip is seen at or near the top of the curve (bottom if the curve is inverted). The pip size should be just large enough to be seen. If the pip is too large, the curve will be distorted. Move the marker tuning dial back and forth and see the pip move off its spot, down one side of the selectivity slope or down the other side, depending on the position of the tuning dial. The pip size will vary depending on where it is on the curve, since it is affected by lack of gain in the receiver the farther off center frequency you go. To maintain a size that can be seen, you will need to readjust the marker output control as you change frequency. This is not necessary with marker adders.

A built-in crystal oscillator marker will produce harmonics of the crystal and show several markers along the curve. A 100-kHz crystal will produce marker pips every 100 kHz along the curve. This may be fine for TV i-f curves but is too far apart for communication receivers. Crystal oscillators are good for finding the center frequency with crystal accuracy. By plugging in a 455-kHz crystal a pip will appear at the 455-kHz point on the selectivity curve. Five-MHz crystals are common and may be used for 5-MHz i-f's in many ssb transceivers.

Any external signal generator or source of r-f energy may be coupled into the i-f strip of the receiver to show up as pips. Coupling is easy—merely lay the hot lead of the output cable of the external signal generator near the i-f section, or connect it in parallel with the output of the sweep generator to the input of the i-f section of the receiver. A 100-kHz crystal oscillator with 10-kHz and 1-kHz outputs is excellent for putting pips on the trace close enough to measure by. A grid-dip oscillator can also be used, but its accuracy is far from good enough for anything but rough approximations as to frequency.

As mentioned before, the marker adder instrument is ideal for showing marker pips, since the marker-generator output does not go through the receiver.

On a variable marker generator, tune the dial to slide the pip down one slope of the curve until the pip appears halfway between the peak and the base line. Note the frequency. Slide the pip up the slope and over to the other side and half-way down. Note the frequency on the marker-generator dial. The difference between the two frequencies is the bandwidth of the receiver at the 6-dB points.

MEASURING IMAGE RESPONSE

An r-f signal beats with the high-frequency oscillator to change the signal frequency to the intermediate frequency for further amplification. The local high-frequency oscillator may be below the signal in frequency, or above the signal frequency by the amount of the intermediate frequency. If the oscillator frequency is above the signal frequency, as is usually the case at the lower amateur-band frequencies, it would oscillate, for example, at 3.955 MHz to convert a 3.5-MHz signal to 455 kHz. Another signal at 4.410 MHz would also beat with the 3.955-MHz oscillator and produce an i-f signal of 455 kHz. If front end (r-f) selectivity were poor and the 4.410-MHz signal were strong, it would also be heard when the receiver is tuned to 3.5 MHz and interfere with a 3.5-MHz signal. The image is twice the intermediate frequency away from the desired frequency—twice the intermediate frequency above the signal if the oscillator frequency is above the desired signal frequency, or twice the intermediate frequency below the signal frequency if the oscillator is below the desired signal frequency.

With a 455-kHz intermediate frequency, a strong image from one end of the 10-meter band can interfere with the reception of a signal at the other end. Most interference is from other than amateur signals, and they can be strong enough to be quite bothersome. The answer to reduced image interference is improved front-end selectivity or a higher intermediate frequency. Most good ham receivers or transceivers use an intermediate frequency much higher than 455 kHz, especially if they use a dual-conversion circuit. Better amateur equipment i-f frequencies are dictated by the mechanical or crystal i-f filters used in ssb equipment. A common mechanical-filter frequency is 2.1 MHz. Crystal filters also are often in the vicinity of 5.0 MHz. They may center on 5174.5 or 5501.5 MHz, for example. An image-frequency signal at twice one of these higher frequencies hardly has a chance to get through almost any front end.

The setup for making an image-ratio test is similar to the setup for a sensitivity test described earlier. Connect a signal generator with a calibrated step attenuator to the input of the receiver, and turn on the internal modulator. Tune the signal generator and the receiver to the same frequency, around 3.5 MHz. Adjust the signal generator for a low output that will not overload the receiver—something like 50 μV should be good. Note the reading of the VOM or VTVM connected to the output, and note the signal-generator attenuator setting. With the receiver untouched, tune the signal generator to a frequency twice the intermediate frequency above 3.5 MHz, or about 4.410 Mhz if the intermediate frequency is 455 kHz. Turn up the attenuator for more output until a signal is heard. Rock the tuning on the signal generator for on-the-head tuning of the image. Adjust the attenuator until the output meter shows the same output as originally obtained on the first signal frequency. The difference in the two readings from the attenuator, in dB, is the *image-response ratio.* If the attenuator is marked in voltage ratio, you will need to convert to dB with the formula:

$$\text{dB} = 20 \log \frac{E_{\text{higher}}}{E_{\text{lower}}}$$

If, for example, the image voltage is 600 times greater than the original signal voltage to produce the same output, the dB attenuation is: 20 log 600 = about 55.5 dB. In other words the image is 55 dB below the signal. It is sometimes called *image-rejection ratio, image response, image ratio* and *signal-to-image ratio.*

NOISE FIGURE

The limit of the ability to copy a signal is noise. Noise may be external, atmospheric, or manmade; or it may be internal, the result of thermal agitation in components, shot effect in tubes, etc. We noted the sensitivity of a receiver is the signal required to overcome internal noise by 10 dB. The next logical question is, just how much noise does my receiver develop? The answer is the dB of noise developed in the receiver over a perfect noiseless receiver. Receiver noise is designated as N or NF, either of which stands for noise factor or noise figure.

In a good ham receiver, noise developed in the converter stage is overridden by signal amplification in the preamplifier stages ahead of it. In fact the signal affecting receiver noise comes down to the first r-f stage of the receiver. The amount of noise developed is related to the temperature of resistive components in the input circuit, input impedance reflected from the antenna circuit, and impedance of the first tuned circuit. The higher the frequency is, the greater is the noise developed. From a practical standpoint, internal noise becomes important in the r-f stages only above about 25 MHz. Most noise is developed in the mixer stages.

To measure NF of a receiver one needs a source of noise, the output of which is known, or is measurable. The circuit of Fig. 4-12 is a noise generator using a 5722 vacuum-tube diode. It is one you can build yourself and should give you a little better than 20-percent accuracy in making receiver noise measurements. Any accuracy better than that requires a calibrated laboratory-type noise generator.

The 50-ohm resistor in the diode load circuit represents the equivalent of the antenna feeder. For minimum receiver-developed noise, the input impedance of the receiver should match the antenna feeder. Should it not, by measurement described later, you can get a lower

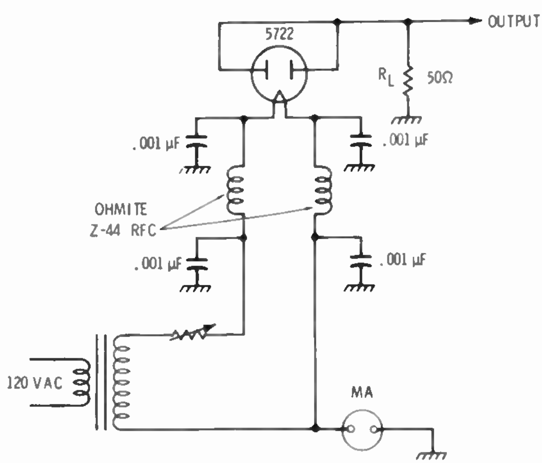


Fig. 4-12. Schematic of a noise generator using a 5722 vacuum-tube diode. Adjust the heater current to provide a 3-dB increase in noise in the receiver. The noise figure is 20 IR_L, where I is the current through the diode, and R_L is the load resistor—in this case 50 ohms.

NF by changing this resistor (R_L) to a value equal to the receiver input impedance. However, you would enjoy an actual decrease in noise level only if the antenna impedance were equal to the receiver input impedance.

The noise generator should be built into a metal case with coaxial cable output connectors to provide a thoroughly shielded connection to the receiver.

Connect a VOM or VTVM to the output of the receiver. Set its function to read dB or a-c output. The receiver must be operated wide open and with the avc disabled by jumping it with a fixed-voltage source equal to the no-signal bias developed. An excellent and temporary bias voltage source is a series of flashlight cells. The bfo must be off, although the carrier-reinsertion oscillator must be on for a ssb receiver using a product detector. Connect the noise generator to the antenna input, with close coupling. It is important that no external noise, manmade or atmospheric, enter the receiver.

With the noise generator off and the receiver wide open, read the receiver noise output on the output meter. Turn on the noise generator and turn the filament current up until the output meter reads an increase of 6dB or twice the receiver noise alone. When the output is doubled it means the input noise from the generator is equal to the internal noise. Generator noise for this circuit is:

$$20\, IR$$

where,

I is the current through the diode circuit,
R is the diode load resistor.

When a 50-ohm resistor is used, as shown in the diagram, NF becomes a direct reading of current in mA. The lower the current is, the lower the external noise is, and, of course, the lower is the internal noise that equals the external noise.

This is a noise ratio that now must be converted to dB by the formula dB $= 20 \times \log$ current ratio. If, for example, the current is 5 mA, the dB figure for NF is $20 \times \log 5$, or $20 \times 0.7 = 14$ dB.

A good communications receiver runs about 5 to 50 dB for NF at frequencies below about 25 MHz. From a practical standpoint these figures are good for reception of signals through the 15-meter band. At higher frequencies lower noise figures become increasingly important, but the actual receiver noise figures must be expected to increase at the higher frequencies.

Current through a silicon diode will also produce random noise. A noise generator using a silicon diode in a simple circuit is described later in this chapter. It is not useful in measuring noise figure of a receiver because the output is not measurable, but it makes a good noise source for aligning receiver front ends.

INPUT IMPEDANCE

In receiving, the antenna is the signal source, with a feed line connecting the antenna to the receiver. Consider the antenna and feeder, then, as a generator and the receiver input as the load. When the impedance of the load is equal to the generator impedance, maximum energy transfer takes place, with the least amount of input circuit noise developed.

Receiver input impedance is a design factor based on the ratio of primary winding to secondary tuned circuit in the antenna coil. Designers do the best they can to make the match perfect, but must compromise for different bands and different frequencies within a band. Sometimes a compromise is made to accommodate the use of random-length antennas instead of a 50-ohm coaxial fed transmitting antenna, or for use with transmission lines of other impedance values.

Chapter 7 describes an impedance bridge of simple design and easy construction. It requires a source of r-f energy, and a null indicator. The r-f source can be a grid-dip oscillator, a regular signal generator, or the reduced output of a transmitter driver. The indicator is a high ohms-per-volt VOM, or VTVM. The device acts as an a-c bridge with the GDO or other source of r-f as input, and the input of the receiver as the unknown arm of the bridge. When the resistance of the potentiometer equals the unknown impedance, the bridge is in balance and an indicator (VOM or VTVM) will show zero voltage when the unknown is a pure resistance. When the receiver is tuned to the same frequency as the signal source, the receiver input is nearly a pure resistance, but not quite. There will nearly always be some reactance, and this reactance will prevent an absolute null and cause some inaccuracy in the impedance reading. But we are checking for approximate input impedance, so for purposes of seeing how far off the input impedance is from the source impedance (such as a 50-ohm coaxial feed line) the error is unimportant.

Solder a single-turn piece of wire between the inner terminal and the shell of a coaxial connector. Couple the loop to a coil of the GDO for signal pickup. Connect the output of the bridge to the receiver input with coaxial cable. Connect a VTVM to the indicator terminals, with the function switch set to read dc and the range switch on the lowest range.

Adjust the panel control on the impedance bridge for a minimum reading on the meter. Tune the receiver for other frequencies and other bands, and readjust the bridge for minimum each time. The resistance of the control will be the approximate input impedance of the receiver. The value may vary considerably from band to band, and even from one end of a band to the other.

RECEIVER ALIGNMENT

Aligning a receiver means setting the calibration of the tuning dial to read correct frequency as near as possible, adjusting for peak sensitivity, setting i-f tuned circuits for optimum (not necessarily sharpest) selectivity, and setting the bfo or carrier-insertion oscillator for proper sideband operation in an ssb receiver.

Alignment varies from a touchup to peak the performance of a receiver to a first-time alignment of a receiver just built. The amount of work involved varies between the two extremes, and also depends on the type of receiver, whether a single-conversion, dual-conversion, a-m, or ssb. In addition there are the filters used in many receivers. There are the sharp crystal filters, and the square-top steep-skirt filters, whether crystal or mechanical.

Three things are needed to align a receiver: a source of steady rf (preferably amplitude modulated), a means of reading the output, and the proper-fitting alignment tools. The best input signal device is a signal generator with a good step attenuator in the output. It should have amplitude modulation built-in, as most do. Other acceptable signal sources include the 100-KHz crystal oscillator, either external or built into the receiver. The vfo of your transmitter, if you keep the signal output down (even if it requires a dead short circuit across the output), makes an excellent r-f source. The grid-dip oscillator is an acceptable r-f source for a rough alignment. A noise generator is good if the receiver is not far out of alignment to start with.

Output indicating devices are simple and can take many forms. Even the ear can be used by listening to the increase of sound level in the speaker, or the hiss from an unmodulated r-f source, but the ear is not as reliable as a meter indicator. The S-meter on the receiver is a good indicator for an unmodulated r-f signal source. A VTVM connected across the avc or agc bias bus serves in the same way if the receiver has no S-meter. The most frequently used output indicator is a VOM or VTVM across the audio output of the receiver. If the VOM or VTVM has a very low range a-c scale, it can be connected across the speaker voice-coil leads, or across a resistor whose value is about equal to the speaker impedance and which is wired in temporarily in place of the speaker. If more voltage is needed, connect the VOM or VTVM across the primary of the output transformer.

A VTVM connected across the second-detector diode load resistor is the best position and type of output indicator, but with avc action replaced with fixed bias. For product detectors a demodulator probe must precede the VTVM. A suitable circuit was described earlier. With a demodulator preceding it, the VTVM is used on a low-range d-c function.

Alignment tools to fit the i-f and r-f trimmers are needed. Where screws are used, as in compression trimmer capacitors, an insulated screwdriver is required. One of the best consists of a tiny metal blade set in a plastic shaft. An all-plastic screwdriver is fine but the end can break off on a hard-to-turn screw. Slug-tuned cores in r-f coils and i-f cans may have a hex-shaped hole or a slot, each requiring a special insulated tool to fit. If a metal screwdriver is used on them, the metal will affect the inductance, and the hardness of the metal could chip out some of the powdered-iron core, even ruining it for use with the proper tool. A hex-ended tool with a narrower neck is handy for adjusting both end cores in an i-f transformer from one end, the hex part fitting all the way through the top core into the bottom for separate adjustment. The tool looks something like Fig. 4-13,

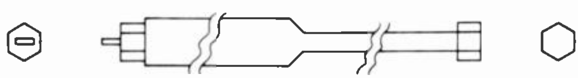


Fig. 4-13. A combination insulated alignment tool for i-f transformers. The narrow neck at the right allows the hex end to protrude through the top slug into the bottom. Both primary and secondary core slugs can be adjusted from one end of the i-f can.

All tuned circuits on a home-brew receiver should be brought into close frequency range with a grid-dip oscillator. With the receiver off, couple the GDO to each of the front end r-f and oscillator coils and adjust them to frequency by setting parallel trimmer capacitors for the high end of each band and adjusting the core slugs (or series padder capacitors if used) for the low end of each band. Intermediate frequency transformers are supplied close enough to frequency so that a rough setting with a GDO is usually not necessary. A check can be made on whether or not they are far off frequency by coupling the coil of the GDO to the coil in the i-f can through the open end of the can. Be sure to use the proper GDO coil, and tune it to the intermediate frequency.

For peaking a receiver for highest sensitivity, the signal source need not be accurately calibrated, but it must be a stable source and capable of fine frequency adjustment, particularly where crystal i-f filters are used in the receiver. For accurately calibrating the receiver tuning dial, the signal source must be close to frequency; a 100-kHz crystal calibrator is excellent for use after rough frequency adjustments are made from some other source.

The receiver should be operated at full r-f gain control setting. Best results are obtained if the avc or agc bias is replaced with fixed bias, in the manner described earlier in this chapter for sensitivity measurements. The r-f signal input should be the minimum amount possible and yet show a good indication in the output metering method. It is important not to overload the receiver.

If the receiver is only slightly out of alignment, the signal source may be connected to the antenna input; if it is badly out of alignment, it may be necessary to make a stage-by-stage alignment in the same manner as stage-by-stage gain measurements were made earlier. In most cases the r-f source is coupled to the converter stage for i-f alignment, and later transferred to the antenna input for r-f adjustments. For purposes of an example, this method will be described in detail.

Lift the shield from the converter tube off the grounding tabs at the socket. The shield will serve as capacitive coupling between the signal generator and the converter tube plate. If the receiver is transistorized, coupling is made to the base of the converter transistor using a .01-μF capacitor between the hot lead of the generator and the base terminal of the converter transistor. Frequently the base input circuit is of such low impedance that the output of the signal generator is insufficient to drive a signal through the converter. In such cases the base connection must be lifted and a 1000-ohm resistor connected from the base to the low end of the coil to which the base was connected. If there already is a resistor from base to ground or to the avc or agc bus, and the base is capacitance coupled to the preceding tuned circuits, disconnect the coupling capacitor, and connect the signal generator to the base terminal of the transistor. In most signal generators a terminating resistor (usually about 50 ohms) is in the lead, right at the output end. When connecting directly to the base of a transistor (or the grid or plate of a tube) the d-c return path of the terminating resistor can upset bias or plate voltages. If a capacitor is not already part of the circuit under test, add a .01-μF capacitor in series.

If the receiver has a built-in high-selectivity crystal in the r-f circuit, connect a VTVM across the second-detector plate load or diode resistor, and set the VTVM function switch for dc. The signal generator should be used unmodulated. The receiver selectivity with crystal in is too great to use a modulated signal because the modulation sidebands would be cut off. If the receiver does not have the crystal, either the connections mentioned or a VOM or VTVM in the output circuit may be used. If connected to the output, set the meter function switch to read OUTPUT or AC on a low range. The signal generator must then be internally modulated. For either method set the meter for a low range, one that reads the internal receiver noise at about one-third scale.

Set the signal generator to minimum output, and rock the tuning dial around the input intermediate frequency of the receiver. Set the dial at the point of maximum reading on the output meter. The output reading should be about twice that of the receiver internal noise. Set the signal generator dial carefully for a peak reading on the meter. Now, with the proper alignment tool, adjust each i-f coil for a maximum reading on the output meter. The i-f coils may have trimmer capacitors across each winding. Use an insulated-blade screwdriver, and adjust both trimmers, usually located on the top of the i-f can. If the coils are slug-tuned with powdered-iron cores, use the proper tool and adjust both cores. Slotted cores will require adjustment from top on one core, and from the bottom on the other. Hex cores are usually hollow, and the tool illustrated in Fig. 4-13 will tune both cores from the top. The narrow neck allows you to pass the tool head through the top core down into the bottom core. Adjust each i-f trimmer or core for highest reading on the output meter. If the signal generator indicates the intermediate frequency is off according to the manual specifications, move the signal generator toward the correct frequency a little at a time, each time readjusting the i-f transformers to peak with the signal generator output.

Receivers with crystal-selectivity positions must be adjusted with the crystal in. The crystal establishes the correct intermediate frequency. The procedure is the same as just described except that the signal generator must be much more carefully set since the peaking frequency is very sharp. It is important that the generator has been thoroughly warmed up before making any adjustments, otherwise the slightest drift off frequency will affect your alignment. After alignment with the crystal in its sharpest selectivity position, set the receiver selectivity on a lower position, and, moving the signal generator dial slightly to one side and the other of center frequency, note the output meter readings. The peak reading should be less than with the crystal in, but the dropoff on each side of maximum should be slower (Fig. 4-14).

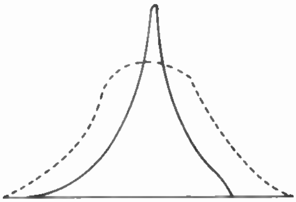


Fig. 4-14. Solid line is selectivity curve with crystal in. Dotted line is curve in broad selectivity position.

SSB RECEIVERS

There are two basic differences between aligning an ssb receiver and the usual a-m type. The ssb receiver does not use a diode second detector (some have both types of detectors and are switchable), and the i-f circuit usually includes a broadband mechanical or crystal lattice filter. To develop dc for the VTVM as an output indicator a demodulator probe or circuit must be connected in front of the VTVM. The input of the demodulator circuit connects to the output of the product detector but with the carrier oscillator killed. The product detector then acts like a simple amplifier. The easiest way to kill the carrier oscillator is to pull the crystal or crystals out of their sockets. If two are used (for upper and lower sidebands), be sure to mark them for return to the proper sockets. The demodulator probe must include an input capacitor to prevent shorting the dc in the plate circuit of the tube or collector circuit of the transistor.

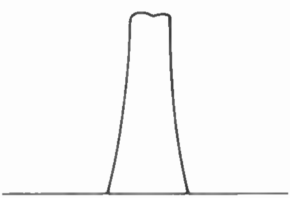


Fig. 4-15. Typical curve of i-f selectivity where a crystal-lattice or mechanical filter is used.

Mechanical i-f filters are fixed tuned, and are not adjustable. Crystal lattice filters are also not themselves adjustable, although some have an adjustment to a bifilar winding to make the selectivity top flatter. These filters establish the intermediate-frequency center. They provide the i-f section with a broad-topped selectivity curve with sharp sides. The broad top is usually about 3 kHz wide. The curve looks something like Fig. 4-15. This calls for rather careful adjustment of the i-f transformers to maintain the proper curve. The real selectivity is provided by the filters rather than the i-f transformers. In many cases the i-f stages do not use dual-winding transformers, but single coils, the purpose of which is to provide the high plate impedance for the i-f tubes for maximum gain.

As an example, the old Swan SW-240 has a four-crystal, crystal-lattice filter. Two crystals are cut for a frequency of 5175.5 kHz and two for 5173.5 kHz. The 6-dB down points are 5176 kHz and 5173 kHz, for a bandwidth of 3 kHz. The top of the curve has two small humps, reflecting the frequency of the crystals, affected by the bifilar winding. The curve looks like Fig. 4-16. The Swan 500 has a 2700-Hz wide passband in the crystal-lattice filter, the extremes of which are 5500.3 kHz and 5503 kHz.

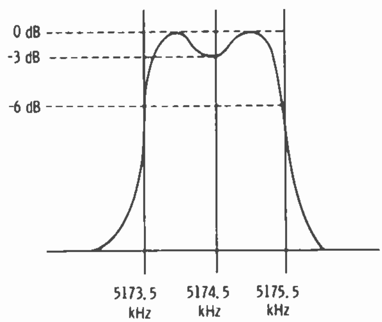


Fig. 4-16. Selectivity curve for the Swan SW-240 transceiver using a crystal-lattice intermediate-frequency filter.

As you move the signal generator through and to each side of i-f resonance, note carefully the output readings. Look for symmetry in the output on each side of center resonance. The output should be fairly flat for a width of about 3 kHz (check your receiver manual for exact width). There will be slight variations in output between the frequency extremes; but if the variations are less than 3 dB and are symmetrical, it is considered acceptable. If one side of the flat top is higher than the other, adjustment of the i-f transformers is prescribed. Adjust for equal output of the various humps of the curve. If there is a bifilar-winding adjustment on the crystal lattice filter, and the humps are more than 3 dB above the dip, adjust the bifilar winding, and retrim the i-f transformers for equal flat top.

SWEEP GENERATOR

Alignment of the i-f stages is much easier when a sweep generator and oscilloscope are used. You can "see" what you are doing much easier. The scope will display either the peak reading of a selective a-m receiver or the flat-topped response of an ssb receiver. The great value in using the sweep method is the ease with which symmetry is achieved. Adjustments are made while observing the scope pattern, and every movement made with the alignment tool will be seen as a change in the response on the scope trace.

BFO ADJUSTMENT

On an a-m/c-w receiver, where the bfo does not have a front-panel frequency control, bfo adjustment is made from any steady r-f signal source, even a station carrier. Tune in the signal from the signal generator or other source. The output indicator will be your own ear. Adjust the bfo trimmer for a pleasant tone either side of zero beat. *Do not* adjust the bfo for zero beat. The higher the audio frequency tone you adjust for, the greater will be the attenuation of the signal on the other side of zero beat when you tune your receiver on c-w signals. But the tone must be pleasant and should not tire you while copying code for long periods.

SSB CARRIER-INSERTION ADJUSTMENT

The reinserted-carrier oscillator for the product detector of an ssb receiver is somewhat similar to the bfo in a cw receiver, but with greater stability a requirement. For that reason most good ssb receivers or transceivers use crystal-controlled carrier oscillators, often with two crystals for switchable sideband selection. The crystals are ground to provide a carrier-reinsertion frequency about 300 Hz below the lower limit of the mechanical or crystal-lattice filter for upper-sideband reception, or 300 Hz above the upper limit of the filters for lower-sideband reception. The carrier oscillator crystals have trimmer capacitors across them to adjust them exactly right.

The carrier-reinsertion oscillator beats with the ssb signal in the product detector and produces the audio frequencies of the voice. The frequency of the carrier oscillator must be such as to result in an audio passband between 300 and 3000 Hz, approximately. If the receiver i-f bandpass is 3 kHz wide at the 6-dB down points, that would put 300 Hz down 6 dB at one end, and 3.3 kHz down 6 dB at the other end of the passband. The carrier frequency would then fall farther down on the selectivity curve.

Where the frequency is adjustable it may be done aurally, or by use of a signal generator with built-in variable audio-frequency modulator, or with an audio generator in the transmit mode if it is a tranceiver. This last method will be mentioned in the chapter on transmitter measurements.

Aurally, adjustment is made by first tuning in a good ssb station. Remove the carrier-reinsertion crystal, and tune the station for highest S-meter reading. Reinsert the crystal and adjust the trimmer across it for best sound quality with proper balance between lows and highs in the speech frequency. If the carrier frequency is set too low on the curve the lower voice frequencies will be lost and the sound will be raspy. If set too high, some of the higher frequencies of the voice will be lost with some loss of intelligibility. In addition when the carrier is set high on the filter response curve, attenuation of the other sideband will be less.

Using a signal generator with built-in variable audio-frequency modulator, connect the generator as described before for i-f alignment. Connect a VOM or VTVM to the output with the meter function switch on OUTPUT or A-C VOLTS. Modulate the signal generator at 1000 Hz, and tune the receiver for maximum output-meter reading. Turn the internal modulator of the signal generator to 300 Hz; then carefully adjust the trimmer across the carrier-oscillator crystal until the 300-Hz signal is 6 dB below the 1000-Hz signal on the output meter.

RECEIVER CALIBRATION

Tuning-dial frequency accuracy depends on proper adjustment of the high-frequency oscillator. Even when correctly adjusted, only the finest receiver will be accurately calibrated clear across each amateur band. The object, then, is to obtain the best compromise of dial markings versus frequency across the band. Making accurate frequency adjustments about ¼ or ⅓ of the way in from each end of the band will make the ends and the center of the band fall quite close to exact calibration.

Calibrator adjustments are only as good as the accuracy of the signal source used. If a signal generator is used, it must be a very good one, or one with a built-in crystal-controlled secondary standard which is used to beat against the variable oscillator to set the dial calibration on the head. An excellent source for calibration purposes is the 100-kHz crystal oscillator. Of course, a 100-kHz crystal calibrator limits the dial settings to even 100-kHz points, but this is better than using a less-than-accurate signal generator.

The spread, or tuning coverage, of each band is controlled by proper adjustment of both the capacitance and inductance of each high-frequency oscillator tuned circuit. The *high* end of the band is set by the *capacitance* across the coil, and the *low* end by the *inductance* of the coil. Low-capacitance trimmer capacitors are across the coil, and are used to fix the high-end limit of the frequency range of the parallel main-tuning capacitor. The high-frequency trimmers are usually screwdriver-adjusted compression types, or screwdriver-adjusted cylindrical chassis-mounted types. For the low end, high-capacitance compression capacitors are connected in series with the coil and main tuning capacitor, or adjustable coil cores inside the coils are used to adjust inductance. Both types of capacitors are screwdriver adjusted. The series capacitors are called series *padders*, and they set the lower limit of frequency effect of the main tuning capacitor. The adjustable powdered-iron cores determine the inductance of the coil.

Connect the signal source to the antenna input. The connection may be made directly to the antenna terminals, or the output cable may be laid near the antenna input. It is only necessary to get a signal into the receiver. Almost any type of output indicator may be used. The S-meter on the receiver is fine. A VTVM across the second-detector diode load resistor, or through a demodulator circuit if the receiver has only a product detector, is OK for unmodulated signal sources. If a modulated signal generator is used, a VOM or VTVM at the audio output is OK. Allow the receiver and signal generator to warm up for about half an hour.

Start with the 80-meter band and set the receiver tuning dial exactly on 3.9 MHz. Set the signal generator exactly on 3.9 MHz. Adjust the parallel trimmer capacitor for maximum reading on the output meter. Set the receiver tuning dial and signal generator on exactly 3.6 MHz and adjust the series padder or coil core for peak meter reading. This will upset the adjustment made at the high end of the band, so go back to 3.9 MHz and adjust the trimmer again. Again, this adjustment will change the low-end adjustment, so return to 3.6 MHz and retune. You will find each adjustment affects the other and requires that you repeat each until there is no effect on accuracy. Each time you go from one end to the other the adjustment change becomes smaller and smaller until there is accurate dial readings at both ends. The middle and far ends of the bands should now read accurately also. If they do not, it may require a mechanical shift of the dial or bending the plates of the main tuning capacitor. These changes affect all bands, however, and are not recommended unless analysis indicates a common error on all bands.

A handy device for knowing which way in frequency an r-f circuit is off is a "tuning wand," which is a long plastic tool with a short piece of copper at one end, and a short piece of powdered-iron slug at the other. Inserting the iron slug into a coil increases its inductance and decreases frequency. Inserting the copper slug into a coil decreases inductance and increases frequency. When doing front-end alignment you can tell which way to turn a trimmer or slug by using the tuning wand to determine if the resonant circuit is above or below the desired frequency.

In those few cases where the front-end resonant circuits have neither an adjustable slug in the coil nor a series padder capacitor, it may be possible to squeeze or expand the coils if they are air-wound and self-supported.

Move on to the next-higher band, 40 meters. Use 7.1 and 7.2 MHz for alignment points if your signal source is a 100-kHz crystal calibrator. With a signal generator, use 7.075 and 7.225 MHz for calibration points. Then continue on to the higher bands, picking the nearest 100-kHz point about ¼ or ⅓ in from the band ends.

Some ssb tranceivers cover only the phone portions of the bands. For these, select the nearest 100-kHz points in from the ends, or at the band ends marked.

If a 100-kHz crystal calibrator is used as a signal source it should itself be calibrated against WWV. All crystal calibrators have a trimmer capacitor across the crystal which allows you to zero beat the crystal against the WWV signal standards. If your receiver is a ham-band only receiver, you will need to enlist another general-coverage receiver somewhere. WWV broadcasts from Ft. Collins, Colorado are on 2.5, 5, 10, 15, 20, and 25 MHz. Also, WWVH in Hawaii is on 5, 10, and 15 MHz. The signals provide pure cw, 400- and 600-Hz tones, time, and propagation information. Zero-beat the 100-kHz crystal during the pure unmodulated signal periods.

Using the 100-kHz crystal calibrator on the higher amateur bands requires care to select the right 100-kHz harmonic. Use a signal generator or other reasonably accurate signal source to find any even 1-MHz point, and pick up the 100-kHz crystal calibrator signal near it. From there you can find the proper 100-kHz points inside the edges of the bands by counting the beat notes up or down from the 1-MHz point. In the vhf bands, finding the right 100-kHz point is very difficult, and may be impossible. The 4-MHz bandwidth of the 6-and 2-meter bands contain 41 100-kHz points from one end to the other. For the vhf bands the 1-MHz points from a decent signal generator usually give sufficiently accurate calibration points, except for the very particular ham.

If we seem to have overlooked the popular grid-dip oscillator for alignment, it is only because its accuracy is not nearly good enough for calibration, and too often its signal is not stable enough for front-end peaking. Battery-operated GDO's have better stability since they are independent of the a-c line, and could be used for peaking, lacking a better signal source.

FRONT-END ALIGNMENT

It is the alignment of the preamplifier stages that provides the best usable sensitivity. Peaking the front end will not provide a great deal of overall again improvement, but peaking does provide the improved signal-to-noise factor. The more you can bring the signals up over the internal receiver noise the better the receiver performance will be.

The setup with a signal source is similar to the one for calibration but without the need for calibration accuracy. The same kind of adjustments are made to the antenna and r-f coils, and in the same manner, as for the oscillator circuits. Keep going back and forth between the two ends of the bands and adjust parallel trimmers and series padders or coil cores for highest reading on the output meter.

One of the best signal sources for improving signal-to-noise ratio is a noise generator. Where we just want an improvement, but not measurement, a simple solid-state diode circuit for a noise generator is good enough, instead of the vacuum-tube noise generator described

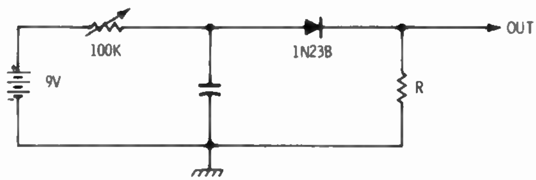


Fig. 4-17. An uncalibrated but adequate noise generator for peaking the receiver front end.

earlier in this chapter. Fig. 4-17 is the circuit diagram of a simple noise generator anyone can build. The variable resistor varies the current through the diode and the output of the noise generator.

USING THE NOISE GENERATOR

The noise generator must be well shielded, and should be connected to the receiver antenna terminals as close as possible. It is important that a minimum of outside noise pickup get through. Turn the receiver gain controls wide open, and get a noise reading on the output meter with the noise generator off. You will be reading the receiver noise. Turn on the noise generator and adjust the variable control until you read about twice the noise that you did from the receiver alone. Adjust the antenna and r-f stages at about ¼ or ⅓ in

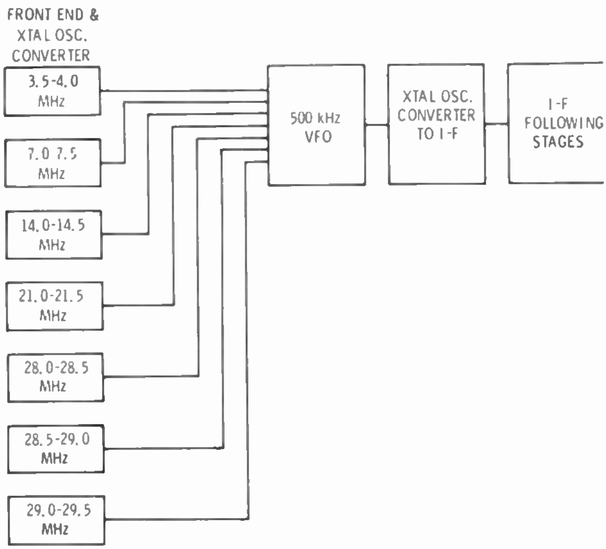


Fig. 4-18. Very fine receivers have a single vfo with only 500-kHz range (some 200-kHz) for a high stability and accuracy. Front-end sections are each 500-kHz wide, and have separate crystal-controlled converters to make use of the single vfo.

from each end of each band for an increase in output of noise. Reduce the noise-generator output with each increase in output-meter reading as you make the peaking adjustments. Go back and forth between the two ends of the bands, on each band, until there is no increase in noise signal with adjustments.

HIGH-PRICED RECEIVERS

The very best receivers have multiple-conversion circuits, with segmented or multiple-band front ends. A common 500-kHz wide vfo is used on all bands, and provides high stability and accuracy. It may feed directly into the IF stages, or be followed by a higher-frequency, crystal-controlled conversion stage before the i-f stages. The vfo is preceded by separate 500-kHz wide front-end sections (Fig.4-18), each with its own crystal-controlled conversion stage to convert for use with the common 500-kHz vfo. There are variations of the one sketched here. Some receivers locate the vfo in another part of the circuit. At least one receiver has 200-kHz segments.

Calibration is simplified, in that there is only one variable-frequency oscillator to calibrate. The other conversion mixers being crystal-controlled need no calibration adjustments. Front-end peaking is the same as described before, except there will be more bands to peak (there are three segments to the 10-meter band) with 500-kHz segments, and still more with 200-kHz segments.

Because of the very high sensitivity and selectivity of these fine receivers, it takes a laboratory type signal generator to do a good job of alignment.

CHAPTER 5

Transmitter Measurements

There is hardly a QSO on the air that does not include a report on the transmitter being used, and power rating of that transmitter. Power is the standard of comparison between transmitters. A power of 1 kW is the legal limit of input for amateurs in the United States.

Power is work, it is energy, and when used right, it is what pushes effective radio waves off the antenna. Power is the result of voltage (E) forcing current (I) through a load (R). It is expressed in various ways by Ohm's law:

$$P = IE$$

$$P = I^2R$$

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

These expressions will be used in considering power measurements.

DEFINITIONS OF POWER

The most common understanding of power by amateurs is the input power to a continuous-wave generator which is the output stage of a transmitter. With key down on a c-w rig, or unmodulated carrier on for an a-m phone rig, the input power is simply the plate voltage times the plate current ($P = EI$). When an a-m rig is modulated, power is being added to the carrier in the form of sidebands. Each sideband, for 100-percent modulation, is equal in power to 25-percent of the carrier. Thus, for 100-percent modulation, a modulator must have a power capability equal to 50-percent of the r-f output-stage power capability. A kilowatt rig is putting out the equivalent input of a 1500-watt rig, when it is 100-percent modulated. The 1-kW limit is exceeded technically but not legally. Rated input power is still *E* times *I* to the r-f amplifier, whether modulated or not.

Measurement, therefore, is simple—multiply the plate current reading by the plate voltage on c-w and a-m phone transmitters. Do not hold the key down too long on a c-w rig, or you may be damaging the final tube. A c-w transmitter is designed for intermittent service if it is designed to give the most for the money, and closing the key could cause the final tube to overheat. In a well-designed a-m phone rig, the power-input reading should not be affected by whether the rig is modulated or not—the plate current should be the same.

The input power to an ssb transmiter is a little different. There is no carrier until the transmitter is modulated. But FCC regulations for setting the legal 1-kW limit are liberal. They will accept the reading of plate current and voltage during modulation, provided the current meter does not have a time constant longer than .25 second. Under these conditions it is still plate voltage times plate current on the final. So if you are running a 1-kW ssb transmitter, use a plate-current meter with a .25-second time constant and keep your eye on it while modulating. Most commercial milliammeters have about the right time constant.

A-C POWER OUT

Up to now we have been talking about the power supplied to the final by the power supply; it has been all d-c power. But d-c power does not indicate the amount of power coming out of the final stage into the antenna for a signal. R-f power can be measured with an r-f ammeter or an r-f voltmeter, and can be "seen" with an oscilloscope. R-f power output will be something less than the d-c power input—no r-f amplifier is 100-percent efficient. As a starter let's "look" at the rf with a scope, and see what goes on.

CONNECTING AN OSCILLOSCOPE

The frequency of the r-f output of a ham rig is much too high for the input amplifier of the average scope. Many moderately priced scopes have a vertical-amplifier response that is flat up to around 5 MHz, and these could be used at 80 meters. But to see r-f at higher frequencies, you must apply the r-f directly to the vertical-deflection plates of the cathode-ray tube. Many scopes have direct access on the back of the case where the terminals of the deflection plates are brought out to jumper bars. By removing the jumper bar of the vertical plates, direct connection can be made. If this feature is not on your scope, install a connection just for the vertical-deflection plates. Remove the scope from the case. Solder a .01-μF disc capacitor to each of the two socket terminals of the CRT vertical-deflection plates. Cut a hole in the back of the case (if there is not already an opening), and bring the two capacitor leads out to a two-terminal tie point, or to binding posts. This type of connection does not interfere with the normal centering controls on the scope, and it leaves the scope free to be used for other oscilloscope duties. Fig. 5-1

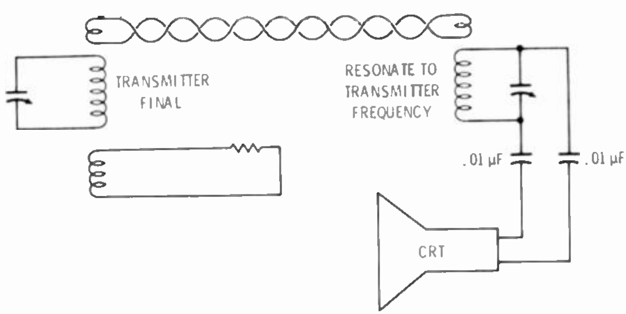


Fig. 5-1. The most effective way of coupling a transmitter to the vertical-deflection plates of an oscilloscope.

shows the r-f pickup circuit. A two-turn loop permits you to adjust the voltage by varying the coupling to the final tank coil, but observe caution in handling, since you are in the presence of high voltage. The resonant circuit tunes to the signal frequency for maximum signal to the scope. Often there will be enough r-f pickup by connecting to the center lead of your output coaxial cable. Use a Tee coaxial connector for this. It may be possible to eliminate the tuned circuit, depending on the power of the transmitter and the deflection sensitivity of the CRT in the scope. Simple capacitive coupling may be enough.

The transmitter must be fully loaded with a dummy load. Don't clutter up the air with unnecessary signals while you are making transmitter observations or measurements, except when it is required that the antenna be connected. Adjust the pattern size by the loop coupler and/or the tuning of the resonant circuit. Fill about ⅔ of the CRT with vertical deflection with the internal sweep on the scope set at a slow rate of about 30 Hz.

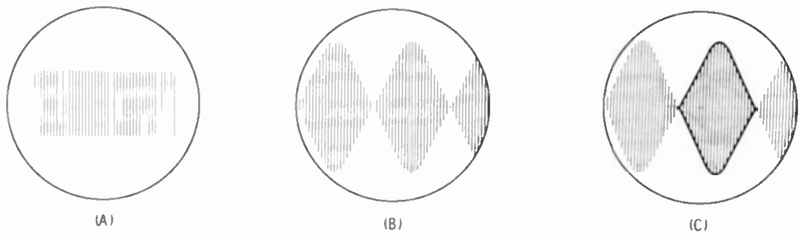


Fig. 5-2. Pattern on scope of a c-w signal is shown at A. This is also the pattern obtained by modulating an ssb rig by a single audio tone. B is the pattern of an a-m transmitter modulated by a 1000-Hz tone, and an ssb transmitter with a combination of two tones 1000 Hz apart. C outlines the power portion of an ssb signal.

WHAT YOU SEE

The trace for a c-w signal will look like a bar (A in Fig. 5-2). The height from top to bottom is the peak-to-peak relative value of the r-f voltage. The rms value, if the signal is a pure sine wave, is the peak-to-peak voltage value divided by 2.828. At the moment the actual value has no meaning. Modulate an a-m rig with a pure 1000-Hz tone from an audio generator, and the pattern will look like B in Fig. 5-2 at 100-percent modulation. The total height of the pattern doubles. At double the voltage, the instantaneous peak power is *four times* the unmodulated power, since power varies as the square of the voltage ($P=E^2/R$). It is the doubling of the voltage during modulation that accounts for the need for high-voltage r-f output capacitors in phone rigs as compared to straight cw. Modulate an ssb transmitter with a single 1000-Hz tone and it will look like a c-w carrier (Fig. 5-2A). What is observed in Fig. 5-2A is the sideband only, or an r-f signal whose frequency is the basic suppressed carrier plus 1000 Hz. For example, 3.9 MHz plus 1000 Hz equals 3.901 MHz, and that is what you will see. But, modulate an ssb transmitter with two tones simultaneously, and it will look like B of Fig. 5-2. Here you are seeing the effect of two r-f signals and the beat between the two. In our example a 3-kHz tone and a 4-kHz tone beat with 3.9 MHz will produce signals of 3.903 MHz and 3.904 MHz, and the beat between 3.903 and 3.904 MHz is 1 kHz. The two-tone test is the most popular test for adjusting an ssb transmitter for proper linearity. The ssb pattern shown helps to explain what is meant by PEP for an ssb transmitter. PEP means *peak envelope power,* and is the power shown under one curve on the pattern outlined in bold in Fig. 5-2C. It is not the instantaneous peak mentioned before, but the power at modulation peaks. At the time it occurs it is about twice the equivalent d-c input-power average. This accounts for the 2000-watt PEP ratings of ssb rigs in the 1-kW class.

POWER EFFICIENCY

Outside of the need to know that the legal limit of 1-kW d-c input to the final is not being exceeded, the real concern with power is to know how well the transmitter is doing in getting a signal into the antenna and out on the air.

The efficiency of the output stage is a ratio of the r-f power output to the d-c power input to the last stage. It may be expressed as:

$$Eff = \frac{P_{out}}{P_{in}}$$

A class-C amplifier runs about 75-percent efficiency. A class-B amplifier as used in ssb transmitters runs about 65-percent efficient. These are typical maximums. The difference between power in and power out is heat, heat that must be dissipated by the final tube or tubes. It is the ability to dissipate this lost heat that determines the rating of the tubes used as final amplifiers. Coupling to the output also enters into efficiency, and optimum coupling is obtained when the load impedance of the tube is matched.

The maximum power which can be fed into a tube or set of tubes is the highest voltage the tube will take without arcing over, times the current at saturation, above which there just can be no increase in the electron flow from the cathode or filament. Operating a tube under saturation conditions will result in calling on the tube to dissipate much more heat than it is designed for. But another factor enters the picture, and that is time on. If the tube has a short on-time cycle it can handle more heat dissipation for that moment than if it is required to throw off the heat continuously. With this in mind tubes may be operated above their continuous-duty ratings when keyed for cw, and when used in ssb transmitters. In a-m phone the carrier is on all the time the transmitter is in transmit mode, so the final tubes must be used within their normal rating if a long life is expected.

MEASURING POWER OUT

Output power may be measured by an r-f ammeter, or r-f voltmeter. Most r-f ammeters read the voltage developed in a *thermocouple* through which the rf is flowing. An r-f voltmeter is a VTVM with an r-f probe, designed to read the r-f voltage in a rectifier circuit in terms of the dc it develops. Some VTVM's have a vacuum-tube diode in a special probe; others use a solid-state diode in the probe. You can build your own probe from the circuit shown in Fig. 5-3. The components must be shielded from stray pickup. An excel-

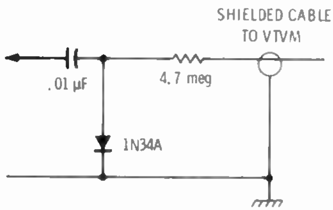


Fig. 5-3. Circuit of an r-f or demodulator probe. With these constants a VTVM will read the peak value of rf on the d-c scale.

lent enclosure is the tube shield of a miniature tube. After the components are wired in a tight bundle and a shielded cable is connected to the output, place the circuit into a tube shield with a test probe sticking out the top and pour an epoxy filler into the tube shield. The epoxy is available at many of the leading radio- and TV-parts supply companies. The VTVM is set on its d-c function. The d-c reading on the scale will be approximately correct for the rms value of the rf.

Diodes in r-f voltmeters are limited as to the voltage they can handle, and this limit must be observed. On solid-state diodes the limit on most is about 20 volts. A resistor voltage divider must be used externally for reading higher voltages. VTVM's with vacuum-tube diodes are usually good for reading voltages up to about 500 volts without damage to the diode. In the voltage divider circuit of Fig. 5-4, the voltage out is related to the voltage in by the value of R_2 to the sum of R_1 and R_2. This ratio can be simplified as follows:

$$\frac{E_1}{E_2} = \frac{R_1 + R_2}{R_2}$$

$$= \frac{R_1}{R_2} + 1$$

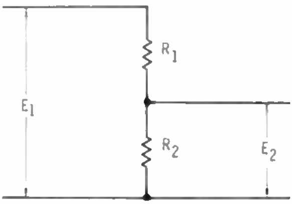


Fig. 5-4. A carbon-resistor voltage divider for reducing rf to the low maximum limits necessary for instrument demodulators.

or

$$\frac{E_2}{E_1} - 1 = \frac{R_1}{R_2}$$

Thus if the ratio of E_1 to E_2 is to be 10 to 1, subtract 1 from the ratio, which makes it 9 to 1 as the ratio of R_1 to R_2. For example a voltage divider using 6200 ohms for R_1 and 670 ohms for R_2 will provide about a 9-to-1 resistance ratio, or a 10-to-1 voltage division. These values must be selected from 5-percent carbon resistors. This example is a good one for use on 52-ohm coaxial lines, for output wattages to 800 watts. To maintain accuracy and prevent phase shift, keep the leads short and the input and output well separated from each other.

A handy replacement for an r-f ammeter is a resistor in series with the output, across which the voltage drop is measured and then converted to current by $I = E/R$. Having the current, calculate power by $P = I^2R$.

A 1-ohm, 2-watt carbon resistor in series with the ground side of the r-f output (the ground side because the metal case of the VTVM could be accidentally grounded) deducts very little from the output and gives you the equivalent of the circuit shown in Fig. 5-3. The r-f probe converts the high-frequency rf to dc, which is then read on a low d-c scale of the VTVM. A 2-watt resistor will handle up to 2 amperes of current. Into a 50-ohm coaxial load, powers up to 200 watts may be read ($P = I^2R = 2^2 \times 50 = 200$). For higher power, a higher-power resistor must be used. Four 1-ohm 2-watt resistors in series-parallel will handle 4 amperes, or 800 watts output.

THE DUMMY LOAD

According to Ohm's law you will need another value in addition to current or volts, and that is resistance (R). The resistance is the output load impedance when the antenna is at resonance and the feeder is perfectly matched to the antenna, or a dummy load that is nonreactive and of known value. If these conditions are not met the load impedance will be more or less reactive and the reading will not give you true power output. It is obvious from this that initial measurements and adjustments should be made into a nonreactive dummy load. A dummy load can be made up of a combination of carbon resistors in parallel or series-parallel. Power handling will depend on the number and rating of the resistors. Fig. 5-5 shows the assembly of 2-watt carbon resistors to make a 100-watt dummy load. Purchase twenty-five 680-ohm 10-percent resistors and twenty-five 620-

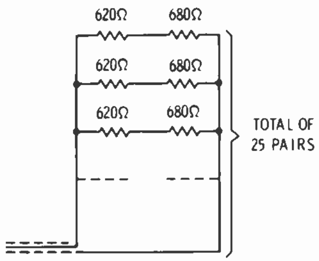


Fig. 5-5. A 52-ohm, 100-watt dummy load can be made up of paralleled 2-watt carbon resistors.

ohm 5-percent resistors (620-ohm resistors come only in 5-percent types), and connect a 680- and a 620-ohm resistor in series. Connect the 25 series-connected pairs in parallel. Keep pigtails short and stray capacity low. Connect a coaxial cable to the ends as close as possible. From about 14 MHz up, this dummy load will begin to show some reactance, because the total length of the bundle will act like a wire loop. For amateur use at the usual lower-frequency bands this makes a fairly nonreactive dummy load, and the net load resistance will be 52 ohms. Higher-powered dummy loads are constructed similarly, but with higher wattage resistors or more of them. Dummy loads are available commercially (Fig. 5-6).

With an r-f ammeter the power output will be the current squared times the dummy-load resistance ($P = I^2R$). With an r-f voltmeter it will be the voltage squared divided by the load resistance ($P = E^2/R$). It can be seen that the power goes up as the square of the current or voltage. Any small increase in current or voltage that can be obtained by making adjustments represents quite an increase in power.

The antenna may be used instead of a dummy load only when it is a purely resistive load to the output. This becomes true only when the antenna is resonant at the operating frequency, and the feeder is perfectly matched to the antenna; *or* when the antenna is resonant, and the feeder is an electrical half-wavelength long. In this latter case it will be necessary to measure the reflected impedance by the use of an impedance bridge, which is described in the next chapter. For the first case an SWR bridge designed for 52-ohm loads, showing exactly a 1:1 ratio will indicate a resonant antenna and a perfect feeder match of 52-ohm coaxial cable.

If you have both an r-f ammeter and an r-f voltmeter, the power output is the product of the two readings regardless of the value of the load impedance, provided the load is a pure resistance.

The foregoing applies to a key-down c-w transmitter and an unmodulated carrier from an a-m phone transmitter. Measuring the output of an ssb transmitter is a bit more complicated. To measure the output of an ssb transmitter, it must be modulated by a two-tone audio signal, or by a single-tone audio signal with some carrier insertion. If a single tone and carrier insertion is used, adjust both until you get all patterns of equal height, as shown on the scope. On an ssb transmitter the final must be operating in a linear manner with correct load, bias, and drive (more on this appears later in this chapter). The scope pattern should look like Fig. 5-7.

Courtesy Waters Manufacturing, Inc

Fig. 5-6. A commercial dummy load, with 52-ohm impedance. This one is good to 230 MHz.

Turn the audio gain up on the ssb transmitter until the highest waveform like that shown in Fig. 5-6 is seen on a scope without distortion or flattopping. An r-f voltmeter, if peak reading (as it will be if it uses a single diode), will read the voltage as indicated between the extremes of E on the pattern of Fig. 5-7. This peak amplitude is converted to rms by multiplying by .707, so the formula for reading PEP on an ssb transmitter is as follows:

$$\text{PEP} = \frac{(E_{\text{peak}} \times .707)^2}{R}$$

where,

R is the impedance of the dummy load or antenna feeder.

For PEP input to the final, the plate-current meter can be used with a two-tone signal by reason of the following. With a two-tone signal the plate current will read about 64-percent of a single-tone signal on class-B amplifiers only. Multiply the plate-current meter indication by the reciprocal of .64 (or 1.57) to get the equivalent plate current, which, when multiplied by the plate-supply voltage, is the

PEP power input. This method will not meet FCC requirements when you are operating at or near the 1-kW mark, nor, for that matter, does the FCC recognize a PEP figure. The FCC requirements reads something like this:

> "The maximum d-c plate-power input to the radio-frequency tube or tubes supplying power to the antenna system of a single-sideband suppressed-carrier transmitter, as indicated by the usual plate voltmeter and plate milliammeter, shall be considered as the "input power" insofar as Sections 12.131 and 12.126(d) of the Commission's rules are concerned, provided the plate meters utilized have a time constant not in excess of approximately 0.25 second, and the linearity of the transmitter has been adjusted to prevent the generation of excessive sidebands. The "input power" shall not exceed one kilowatt on peaks as indicated by the plate meter readings."

The peak meter readings based on the above will be about one-half the meter reading of the plate current using the two-tone method described.

LOADING THE TRANSMITTER

When the antenna system includes an antenna tuner, all reactance can be tuned out and the load to the transmitter is a pure resistance. When the load is a pure resistance, final plate resonance occurs at the current-dip point as indicated on the plate-current meter. In the case of an a-m rig, the plate current and voltage must not exceed the rating of the tube or tubes in the final, and coupling to the antenna should be for achieving the desired current at the plate-current dip. Because of intermittent on-time for c-w and ssb transmitters, the final can be loaded heavier without exceeding the average dissipation rating. Higher current to the final is permissible, up to plate-current saturation. Some form of output indicator should be used for fully loading a c-w or ssb transmitter. It may be an r-f ammeter, r-f voltmeter, or the forward position of an SWR meter.

For a c-w rig with antenna tuner, adjust the loading for a higher and higher plate-current at the dip, until the output indicator shows no further increase in output. Do not keep the key down for more

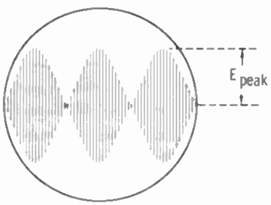


Fig. 5-7. A demodulator probe on a VTVM using a single diode will read the peak value of an ssb signal.

than 30 seconds at a time while making the adjustment. If your setup does not include an antenna tuner, but you are feeding a matched (hopefully) line to the antenna, the plate-current dip is resonance only if you are perfectly matched, and the feeder-antenna load is a pure resonance. Since this seldom occurs, disregard the plate-current dip, and tune for maximum output only. If maximum output is found to be off the plate-current dip, it means the load has reactance, and the final plate capacitor is, in effect, tuning out the reactance.

The foregoing is also true for an ssb transmitter. Insert a carrier to read about half the normal plate current expected. Make adjustments for loading to achieve maximum output on the output indicator. Turn the carrier up until there is no more output. The plate current times the voltage is the peak input power, and the transmitter loaded for maximum output. However, maximum ssb output under modulation is another thing. It is now necessary to establish maximum *linear* modulation limits, and only a scope will do this for you. The proper method is with a two-tone input to the microphone to form a bow-tie pattern on the scope to look like Fig. 5-8, with carrier

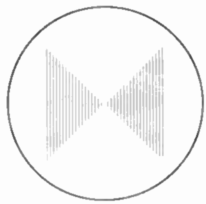


Fig. 5-8 Bow-tie pattern on scope when the ssb transmitter is modulated with two tones (or one tone and some carrier) and audio from the speech amplifier is applied to the horizontal amplifier of the scope.

suppressed or with one tone and carrier. Mark the limits of maximum deflection on the scope face under linear-modulation conditions. To obtain this kind of pattern it will be necessary to pick off audio from the speech amplifier of the transmitter and couple it to the horizontal input of the scope.

Leave the scope permanently connected for voice monitoring. While talking, never exceed the limit of the marks, and your output will be as high as it can be and still maintain linearity without flattopping, which is important. During voice modulation, plate current will kick to about half the two-tone maximum plate current. Switch the scope to the internal horizontal sweep generator for voice monitoring. See Chapter 7 for hookup details.

MODULATION MEASUREMENTS

Important to the proper operation of a phone transmitter are the amount and quality of modulation. It is desirable to know the percent of modulation, the modulator capability, the linearity, and the amount of distortion.

Modulation Percent

The laws governing amateur transmitter operation decree that the percentage of modulation should never exceed 100, nor exceed the capability of the modulator. When an a-m carrier is modulated to where the output peaks are twice the peaks of the unmodulated carrier, it is said to be modulated 100 percent.

The best way to examine the effects of modulation is to observe the shape of modulation patterns on an oscilloscope. Any inexpensive standard oscilloscope or just a cathode-ray tube with power supply is sufficient. Since both vertical and horizontal deflection may be made directly to the deflection plates of the CRT, the tube and power to run it are all that is really needed. However, since there are a number of inexpensive scopes on the market, especially in kit form, the treatment here will assume a regular oscilloscope (Fig. 5-9). Radio-frequency output from the transmitter is coupled to the vertical-deflection plates directly. Audio from the modulator may be connected to the horizontal plates, or it may be fed into the horizontal amplifier of a standard scope.

As mentioned earlier for output measurements, a pickup loop is used to obtain some rf from the final tank. A tuned circuit will often help to supply more r-f voltage to the deflection plates. R-f can also be picked off the coaxial feeder by a Tee connection, and even a small pickup rod acting as an antenna may be used. For these latter two methods, a tuned circuit is almost certainly needed. Fig. 5-10 shows three methods for getting rf to the vertical-deflection plates of a scope CRT. In each case a tuned circuit is included. The tuned circuit can be any convenient inductor and variable-capacitor combination, which will resonate in the band being checked. The variable capacitor also can act as a level control for setting the height of the pattern on the CRT face. Fig. 5-10A shows a pickup loop consisting of a couple of turns of insulated wire, with the lead from it to the test gear twisted. Care must be exercised in coupling the loop to the final tank in the transmitter to avoid shock. In transmitters that are fully enclosed, try the Tee connnection into the coaxial feeder. This is shown in Fig. 5-10B. For noncoaxial feeders, try the pickup rod shown in Fig. 5-10C. The r-f connection is all you will need to observe wave shapes on the scope. Horizontal deflection will be obtained from the sweep circuit in the scope itself. This method is excellent if an audio generator is available to modulate the rig with a single sine-wave tone. Speech waveforms, if used, are so random

Courtesy RCA

Fig. 5-9. A medium-priced oscilloscope available in kit form or wired. The Model WO-33A is switchable from a medium-gain, wideband (to 5.5 MHz) amplifier to a high-gain, medium-bandwidth (to 150 kHz) amplifier.

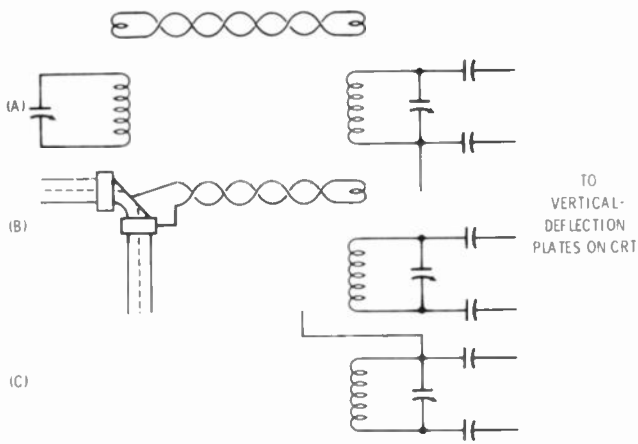


Fig. 5-10. Three methods of coupling the vertical-deflection tuned circuit to transmitter. At A is a 2-turn pickup loop. Use the loosest coupling that will give convenient height to the pattern. At A a slice is cut at the bend of a right-angle connector and a solder connection is made to the inner conductor of the coax feeder. A tee connector could also be used. A 3-foot piece of stiff wire acts as an antenna at C.

that about all you can see is clipping in overmodulation in the negative direction, or flat-topping in the positive direction. It takes a pure signal tone to observe the quality of amplitude modulation.

By picking off audio from the modulator in addition to the r-f, you can produce a trapezoid pattern which may be used to observe voice modulation, thus eliminating the need for a sine-wave audio tone. A tap should be made to the high side of the secondary of the modulation transformer. Connect a .01-μF capacitor with a voltage rating about twice the value of the final supply. This should be followed by a string of 1-watt carbon resistors, and end with a volume control to ground. The values of the resistors should be such that the voltage from the modulator does not exceed the dissipation rating of the resistors. The hookup is shown in Fig. 5-11, and the values are approximately right for use on a transmitter with a 1000-volt plate supply. The higher-value control is for going directly to the horizontal-deflection plates of the CRT. For feeding the horizontal amplifier of a standard scope, use about 1/100 that value. It goes without saying that good insulation is necessary between the top of the series string of resistors and any metal of the chassis or cabinet or other parts. The shunt fixed resistor across the control prevents

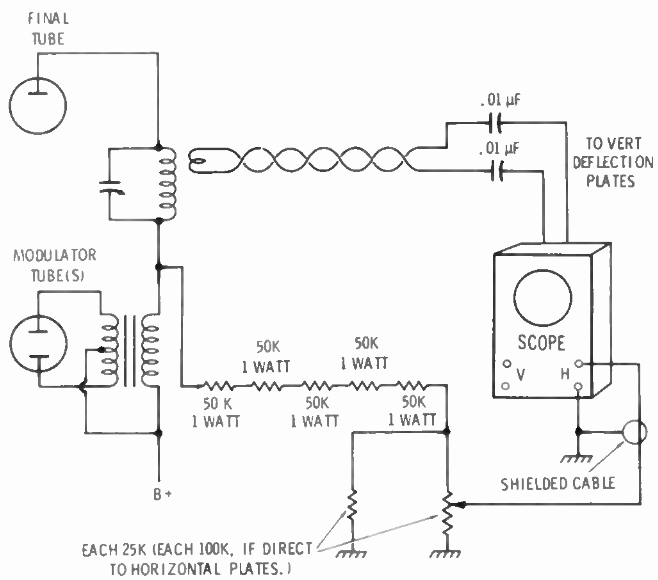


Fig. 5-11. Hookup for a trapezoid pattern from an a-m transmitter.

high-voltage arcing across the wiping contacts of the control should it be dirty or loose.

The modulator connection just described produces a trapezoidal pattern regardless of the type or source of the audio signal to the speech section of the modulator. Voice modulation may be used. As the modulated rf deflects the pattern up and down, the audio voltage from the modulator moves the pattern trace laterally. Modulation positive peaks correspond to highest modulated rf, and negative peaks correspond to minimum instantaneous rf, therefore the trapezoid.

To illustrate modulation percentage, both the wave envelope and trapezoid are shown together. The modulation ratio is the ratio of the difference between modulation peaks and valleys to the sum of the peaks and valleys. Fig. 5-12A shows the carrier only of an unmodulated a-m transmitter; at B is a 50-percent modulated pattern; at C is 100-percent modulation. The formula based on these patterns and the values shown in Fig. 5-12B is:

$$\text{Mod }\% = \frac{PP^1 - VV^1}{PP^1 + VV^1} \times 100$$

where,

PP^1 is the peak-to-peak value of the crest of the modulated carrier,

VV^1 is the peak-to-peak value of the trough of the modulated carrier.

Another way of expressing it is the ratio of the increase in carrier modulated peaks to the carrier itself. If the peaks during modulation are twice the carrier only, it is a 100-percent increase in peak value, or 100-percent modulation.

Frequently there is a difference between the value of the positive excursion of the audio wave and the negative excursion. This may be noted by observing the deflection each side of the center horizontal reference line and the distance to the crest of the wave both above and below the center. The formula given above may be applied to half the wave height, measuring from the center line, instead of peak-to-peak. If the figures are different for upward deflection and downward deflection, modulation is not uniform, although this is not necessarily serious. It is quite common, for example, for the human voice to deflect higher in one direction than the other. Advantage can be taken of this to increase output slightly if the modulator has sufficient capability. Using voice modulation and the trapezoid-pattern system, carefully observe the shape of the top and bottom peaks

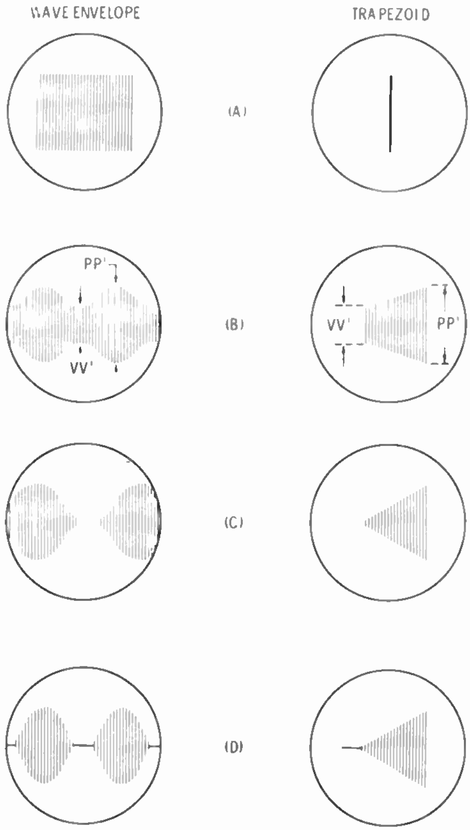


Fig. 5-12. Wave envelope (using the scope internal sweep) and trapezoid (audio from modulator) for increasing percentages of a-m modulation. A is carrier only. B is 50-percent modulation. C is 100-percent modulation. D is overmodulation with negative clipping.

and their distance from the center. Reverse the connections to one winding only of any transformer in the audio system, and make a second inspection of the shape and height of the pattern. The higher of the two, provided the peaks are clean and sharp, will give more push to your modulation without negative clipping.

Negative Clipping

The most serious consequence of overmodulation is negative clipping. When the instantaneous negative excursion of the modulating audio signal exceeds the plate supply voltage to the final r-f amplifier, there is a short duration of voltage shutoff to the final amplifier, and no carrier for that duration. This sharp break in output results in setting up very high harmonics and an interfering signal at other than the operating frequency is created. This is shown in Fig. 5-12D. Note the gaps created at the center line.

A handy circuit for monitoring an a-m transmitter for overmodulation is shown in Fig. 5-13. During the momentary bursts of overmodulation resulting in negative clipping, the plate supply goes negative. If a diode is connected to the top of the modulation transformer secondary with its cathode facing the connection, current will flow during negative clipping, and no current will flow at the time the voltage is positive. Any indicator in the circuit will show the current flow while negative, thus indicating overmodulation. With the values shown in the circuit a VOM on its 10-volt d-c range will kick up noticeably. The speech level should be kept down to where the VOM never kicks up. A similar circuit can be built into the transmitter

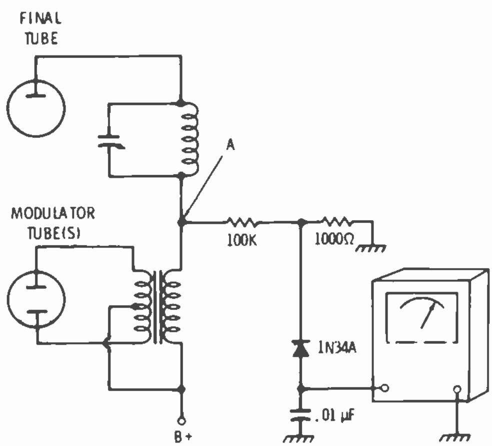


Fig. 5-13. Circuit for monitoring an a-m transmitter for negative overmodulation. No current flows through the meter until point A becomes negative.

using a neon bulb. A forward bias of about 65 volts must be supplied to the neon bulb to overcome the ignition voltage.

Modulation Capability

A plate modulator must be capable of supplying an undistorted (or nearly so) signal with power output equal to one-half the input power to the final r-f amplifier, in order to provide 100-percent modulation. A 250-watt input to a final, for example, requires a 125-watt modulator. Without this capability the positive modulation peaks will flatten out. Not only will you not be producing the 25 percent of extra power for each sideband, but the audio will distort badly by pushing it with higher audio control settings. Any kind of distortion results in harmonics. Audio distortion produces higher-order audio tones that are not part of the intelligence requirement, but produce additional sideband frequencies higher than necessary, and so occupy a wider-than-necessary bandwidth.

Here is how the power-ouput capability of a modulator can be measured: Determine the load impedance the final r-f amplifier presents to the modulator by dividing the plate voltage by the plate current. This should be done in order to properly match the modulator to the final anyway. Disconnect the modulator output from the final, and connect a fixed resistor of the value of plate impedance to the output of the modulator. It will take a husky-rated resistor, of course. The wattage rating would be equal to the final voltage times the current. Tap off a small portion of the output load resistor for a feed to the vertical input amplifier to your scope. Another, smaller, resistor in series at the bottom is equivalent to a tap. Connect a VOM or VTVM across the entire load resistor and set the function switch for a-c volts on a high range. If the expected maximum voltage exceeds that of the highest range on the voltmeter, tap down on the resistor load, but establish a multiplying factor for voltage by noting the ratio of the tapped portion of the resistor to the total resistance. Temporarily connect the r-f amplifier voltage feed point to the bottom of the modulation transformer secondary.

Connect an audio generator to the microphone input of the speech amplifier, and turn the attenuators down to a signal of about 10 millivolts. Set the tone for about 1000 Hz. The level of audio must not overload the speech amplifier, and should be about the right amount for the usual volume-control setting for normal transmitting modulation.

Turn everything on. Bring up the audio signal while observing the quality of the pattern on the scope. The vertical gain control will need to be readjusted on the scope as you turn up the audio in the modulator to maintain a constant pattern size for comparison. As soon as the pattern shape changes and is no longer a sine wave, back the audio down just a little. Adjust for maximum voltage reading on the output without distortion in the sine-wave. This is the point of maximum capability. Measure the voltage output on the a-c meter. The square of the voltage divided by the resistance of the load (P =

E^2/R) is the power output of the modulator. If this is 50 percent or more of the power input to the r-f final, you have sufficient modulation capability.

If the power output of the modulator is below design value, check the tubes, and check all voltages. Low emission on the power tubes will limit the power output.

Other factors affecting modulation power are: low bias or excitation to the r-f final, and the final too lightly loaded to the antenna. When the load is lightened, the input impedance to the final is higher, and the modulator will not be matched. With the audio generator connected to the input of the speech amplifier and a scope at the output, tests can also be made for clipping in the speech stages. If the output shows adequate power for modulation purposes, but the waveform is not a true sine wave, suspect clipping and attendant distortion somewhere in the modulator stages. Clipping produces harmonics, which add extra sideband frequencies and increase the bandwidth transmitted. A search of the offending stage can be made by moving the scope input to each stage output. Use a .01-μF isolating capacitor in series with the scope lead. Move the capacitor connection from the output to the plate of the modulator driver stage, then to the next preceding stage, and so on down the line until there is no evidence of clipping. Be sure, however, that there is no overloading by having the audio generator set at too high an output. The output that just supplies 100-percent modulation is about right. Clipping shows up as a flattening of one crest of the sine wave. It is the result usually of incorrect bias on a tube, or current saturation due to a tube with low emission.

The same setup as above is useful in observing the action of speech clippers. This should be done with speech input instead of as audio generator. Observe the clipping action while talking in a normal tone of voice. Set the audio gain and clipping control so that speech at moderately low level just fills the extremes of the pattern; then high clipping will take place at loud speech levels. Clipping from a speech clipper also causes harmonics, but it is assumed that your clipper circuit also includes the necessary filters to cut off harmonics developed by the clipping action.

C-W KEYING

The oscilloscope is an excellent device for studying the keying waveform of a c-w transmitter. With output rf applied to the vertical-deflection plates by one of the systems shown in Fig. 5-10, key the transmitter with dots by use of a "bug." Either an electronic or a

Fig. 5-14. Observing uniformly spaced c-w dots on a scope. With no keying filter they will be perfectly rectangular as in A. Proper filtering results in rounding the corners, which reduces key clicks, as in B.

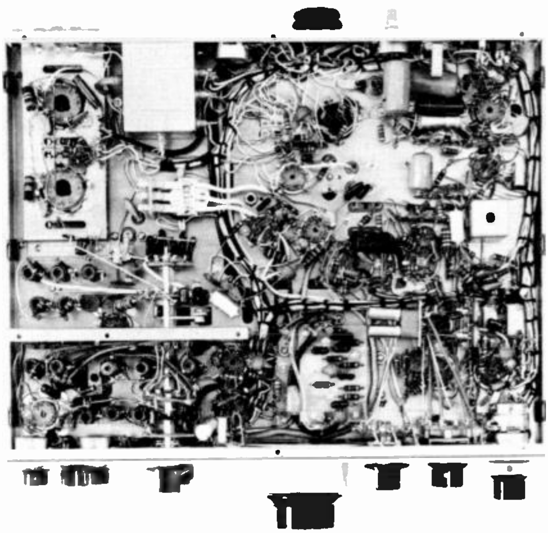
mechanical semiautomatic key will do, but a straight hand key will not. Adjust the horizontal-sweep frequency on the scope to a slow rate to make the keying pattern stand still. It will look like one of the illustrations of Fig. 5-14. With no keying filter you should see square waveforms like Fig. 5-14A. A perfectly square wave produces an infinite number of harmonics every time the key is closed and opened. This produces off-frequency interference and is heard as key clicks. A properly designed keying filter will produce waveforms with rounded corners at the lead edge and the trailing edge. The ideal pattern is shown in Fig. 5-14B. Interfering harmonics will be reduced, and the keying will have a good sharp sound at the receiving end. Too much filtering will round the corners more, but will not have a nice, sharp make and break sound at the receiving end. At high speed, dots will tend to run together and be almost indistinguishable.

SSB MODULATION PERCENT

Modulation percent for an ssb transmitter has no real meaning in the same sense as for an a-m transmitter. The limit of modulation on an ssb transmitter is set by the point at which flattening begins at maximum modulation. Any "percent" would be any ratio of actual modulation level to the maximum clean-signal modulation. Again, the oscilloscope pattern is the method used for seeing the modulation and measuring the height of peaks. Because voice frequencies are complex, a trapezoid pattern like that obtained from an a-m transmitter cannot be obtained from an ssb transmitter in a clean form worth evaluating. Horizontal deflection must be from the horizontal-sweep circuit in the scope. Set it for a slow sweep, as you voice-modulate your ssb rig. Having established the maximum possible height by placing markers on the CRT face from output measurements, voice peaks, for maximum modulation, should just barely touch these marks on the scope. Any attempt to "hit it harder" will result in flattopping, and splatter.

TRANSMITTER ALIGNMENT

Unlike receivers, alignment of transmitters is generally rather simple, becoming a little more involved in ssb transceivers. An a-m or c-w transmitter is rather straightforward when no heterodyning stages are used. From the vfo output up to but not including the driver stage, tuned circuits are usually broad-tuned for each band. These



Courtesy Swan Electronics

Fig. 5-15. Transmitter mixer and driver plate coils are shown in the lower left corner of this photo of a Swan 500C. Adjustment is to the cores of the coils with an alignment tool.

tuned circuits are the vfo plate circuits, the multipliers, and the buffers, (unless the last buffer is also the driver to the final grid circuit). These coils are small, either chassis- or switch-mounted, and have adjustable ceramic or powdered-iron cores. Vfo oscillator coils require calibration, and driver coils are slug-tuned and are set for proper coverage with the driver variable-capacitor coverage of each band. Output adjustment was covered earlier in this chapter.

Begin by unsoldering the B+ from the screen or screens of the output stage if tetrodes are used. If triodes are used, disconnect the plate supply to that stage only. Connect a VTVM across the self-bias grid resistor of the output stage, or across the resistor in series with the fixed bias if that is used. Set the function and range for reading −15 volts d-c. In this position the VTVM reads excitation to the final.

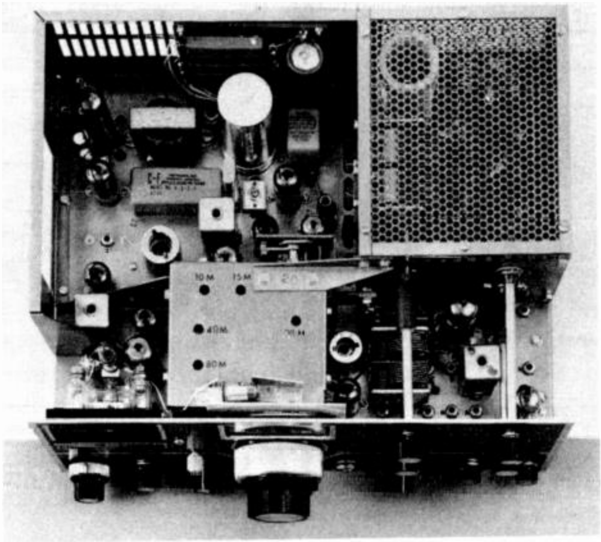
Place the vfo at midband on the 80-meter band, and the driver front-panel-tuned variable capacitor at about midscale. Using the appropriate adjusting tool (usually a plastic tool with a hex head) adjust the vfo plate coils, and each coil from there to the output stage for a peak meter reading. Switch to the 40-meter band and repeat this for each 40-meter coil. Do the same for each band (Fig. 5-15). For ssb transmitters put the function switch on transmit and insert some carrier.

Some ssb transceivers have a trap from grid to ground on the first r-f stage of the receiver circuit. Tuning it requires an r-f signal generator. Connect the signal generator to the antenna terminals, and set the frequency of the generator to the carrier-insertion frequency (5500 kHz in the case of a Swan 500C, for example). You should hear a beat note like that of a signal carrier. Adjust the carrier trap for minimum beat note as heard in the speaker or headphones.

VFO CALIBRATION

Calibration of the vfo is similar to that of the high-frequency oscillator of a receiver. On transceivers using a common receive-transmit tuning dial, the adjustment of the receiver calibration automatically calibrates the dial for transmit (Fig. 5-16).

Heterodyne frequency instruments are ideal for calibration purposes but are expensive. Heterodyne frequency meters include a built-in closely calibrated oscillator and a mixer for mixing some of the output of the transmitter with the local calibrated oscillator. Beating the two permits adjusting the transmitter to match the oscillator standard. A much less expensive method is to use a 100-kHz crystal calibrator and a receiver. It is to be assumed, of course, that the vfo calibration is only a few kHz off frequency in order that the right 100-kHz harmonic is used. The crystal calibrator must have been previously set precisely by zero-beating with WWV on a general-coverage receiver.



Courtesy Swan Electronics

Fig. 5-16. The vfo components are enclosed in a shield in the Swan 500C. Marked access holes are provided to slug-tuned coils.

The military BC 221, which was once available on the surplus market for a low price, makes an excellent moderately priced heterodyne frequency meter.

For either method above, output must be reduced by the removal of the screen voltage from the output stage or working into a dummy load, with receiver antenna shorted, if a receiver is used for the heterodyne tone.

Set the vfo for the nearest 100-kHz calibration if a 100-kHz crystal calibrator is used (or about ¼ of the way in from the high end of the band), for each band to be calibrated. With the dial set for the frequency, adjust the parallel trimmer located on or near the coil until a zero beat is heard in the receiver. Move the dial to the nearest 100-kHz point (or about ¼ of the way in from the low end of the band) and adjust the iron core inside the appropriate coil for zero beat. Go back to the high end and repeat the parallel-trimmer adjustment. Again adjust inductance at the low end. It may be necessary to move back and forth between the two ends several times before correct calibration is obtained. Each adjustment at one end of the dial affects the other end, but each time you go from one to the other, the calibration gets closer.

Make the adjustment for each band for which the vfo is equipped. On some transceivers, there are only one or two bands because a band will be heterodyned for setting frequency on another. Consult your transceiver manual for a check on this. Return power to the final by reconnecting the screen of the output stage or stages, or restoring power to the plate if triodes.

SSB CARRIER ADJUSTMENT

The mechanical or crystal-lattice i-f filters used currently in ssb transmitters to attenuate one of the two sidebands are designed to pass a band of radio frequencies, the width of which is the audio-frequency response. For example, the filter may have a response of 5500.3 kHz to 5503 kHz. The difference between the extremes is 2700 Hz, the planned audio frequency response. The actual response of the filter is not perfectly flat between 5500.3 kHz and 5503 kHz, but has a rolloff at the ends. The extremes of the frequency response figures are 6-dB down from the top, and the curve looks something like Fig. 5-17.

A crystal-controlled carrier oscillator of appropriate frequency beats with the audio frequencies of the voice to produce sum and difference frequencies resulting in a lower sideband, the carrier, and

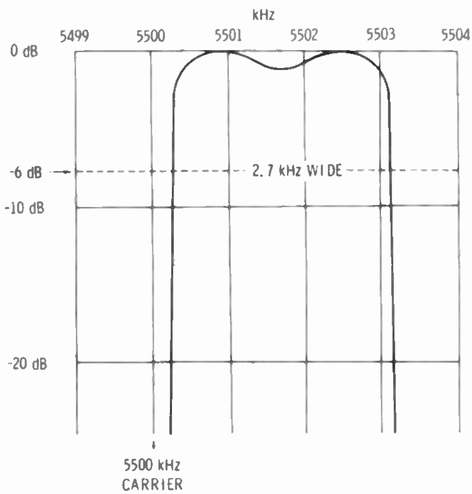


Fig. 5-17. Filter response curve of a typical ssb transceiver. The carrier frequency is adjusted to a 6-dB drop in output with a 300-Hz modulaton tone.

an upper sideband. The carrier is balanced out, and one sideband is attenuated by the filter. The other sideband is heterodyned to the operating frequency of the band and passed on to the r-f driver stage. A small variable capacitor across the crystal in the carrier oscillator adjusts the carrier to precise frequency.

For example, if the carrier is adjusted to 5500 kHz an audio range, of 300 Hz to 3000 Hz will result in an upper sideband of 5500.3 kHz to 5503 kHz. (the bandwidth of the filter), or the sum of the audio and the carrier. If the carrier frequency is 5503.3 kHz, the lower sideband will be 5500.3 kHz to 5503 kHz, or the difference.

Adjustment of the carrier frequency is best done with an audio signal generator. Connect a variable audio signal generator to the mike input, with the generator attenuated to about normal microphone level. Connect a dummy load to the output, and connect a VTVM with an r-f probe across the output. With the transmitter tuned up for normal operation and the audio generator set for 1000 Hz, turn up the mike gain for about normal output, and note the signal level on the VTVM. If the VTVM has a dB scale, set the signal level and VTVM range for a reading on the 0-dB mark. Now turn the signal generator down to 300 Hz. If the output has dropped 6 dB, the carrier frequency is correct; otherwise adjust the trimmer across the carrier crystal for a -6 dB reading on the VTVM. The same adjustment should be made for both the upper and lower carrier frequency if your rig has both. If yours is a transceiver, this adjustment will also be the correct one for receiving sideband signals with the recommended audio-frequency response of 300 to 3000 Hz. This is the frequency response of the example described here; it may be different for other transmitters. Consult your manual for the range of the filter and the oscillator frequency. The method of adjustment is the same.

If the design of your transceiver is such that one carrier is an exact multiple of 100 kHz (the Swan 500C has one carrier on 5500 kHz, for example), you can set it exactly to frequency by connecting a 100-kHz crystal calibrator to the antenna input and adjusting the carrier frequency to zero beat with the 100-kHz crystal calibrator on receive.

If your VTVM does not have a dB scale, select any multiple of 10 as a reference. The -6 dB point will be at 5, or half voltage.

NEUTRALIZING

In spite of excellent physical isolation between input and output of the final r-f stage, and the use of tetrodes and pentodes with good

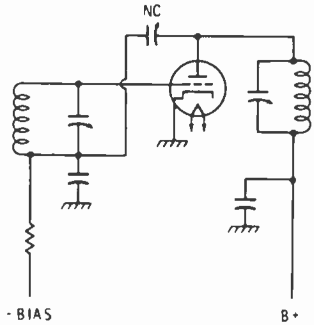


Fig. 5-18. The usual method of neutralizing a final is to capacitively couple some output rf, out of phase, back to the grid circuit.

internal isolation between grid and plate, it seems that some energy from the plate gets back to the grid. When the input and output stages are tuned to the same frequency (as they usually are) any feedback will result in oscillation or some instability because of the tendency to oscillate. This feedback is positive because it is *in phase.*

To overcome positive feedback it is only necessary to introduce negative feedback in the same amount. This is called neutralizing and is usually accomplished by adjusting a small capacitor of about 15 to 20 pF, which is connected between the top of the output tank coil and the bottom of the ungrounded driver coil, or final grid coil (NC in Fig. 5-18).

There are several indications for lack of neutralization. The most serious is a tendency to take off by itself into self-oscillation when excitation is removed from the final r-f stage. Sometimes this occurs when the final tank capacitor is tuned slightly off the operating frequency rather than at resonance. The test is to remove excitation (only when there is protective fixed bias on the final) and tune the final tank capacitor back and forth. The plate meter should remain at zero through the full range of the tuning capacitor if the stage is perfectly neutralized. Another indication is the relation between the plate and grid currents. Normally the plate meter will dip at resonance and the grid current will increase slightly at the same time. If the grid current appears to rise slightly to one side of the plate-current dip, a recheck of neutralization is called for.

A check for neutralization involves determining how much energy gets from the grid to the plate of the output stage without benefit of amplification from the output stage. To do this remove power from the output stage by disconnecting the screen supply from the output stage if tetrodes or pentodes are used, or the plate supply if triodes are used. The screen (or plate in the case of triodes) must be bypassed to ground for both dc and rf. There should be no load attached to the output, which can be accomplished by removing the antenna connection.

Make up a simple indicator for rf like the schematic of Fig. 5-19. This device rectifies any rf in the pickup loop and the resulting dc will show on the VOM. Couple the loop to the output tank. If high voltage is on the output tank, as in the case of tetrodes or pentodes, be very careful in coupling the loop that you do not touch the output coil with the pickup loop, unless the pickup loop is well insulated.

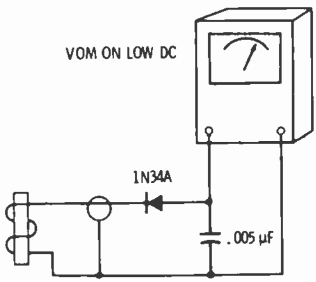


Fig. 5-19. An r-f indicator for finding parasitics. A few turns of wire on an insulated rod will pick up rf when coupled to various points in the circuit of the transmitter, and shows as dc on the meter.

Operate the transmitter in the usual manner except for the dc on the output. If it is an ssb rig, unbalance the carrier-balance control to insert some carrier. Bring excitation up slowly to prevent possible damage to the VOM as an output indicator.

If the output indicator meter deflects, the output stage needs adjustment for neutralization. Using a screwdriver with an insulated shaft, slowly rotate the neutralizing capacitor until there is no output shown on the indicator, as you move the output tank capacitor back and forth each side of resonance.

PARASITICS

Any piece of wire has inductance and distributed capacitance and will resonate at some frequency. Interconnecting wires in the final r-f stage of a transmitter, together with bypass or tuning capacitors, can be shock-excited into oscillation, or they can be resonant with another set of similar circuits and go into oscillation as a tptg (tuned-plate tuned-grid) oscillator. Radio-frequency chokes in both the grid and plate circuits are perhaps the worst offenders. Their inductances, together with bypass capacitors, create their own resonant circuits, which may be the same frequency in the grid and plate sides. The cure lies in changing the values of r-f choke inductance or bypass capacitors, or in swamping them with resistance either in parallel with the chokes or in series with them. Stages preceding the final, because of their lower power and usually better shielding, are not so prone to parasitics.

Unlike harmonics (which can often be easily traced because they are multiples of the operating frequency), parasitic oscillations can occur at any frequency, depending on the resonance of the offending circuit. Thus, parasitic oscillations can cause interference within an amateur band, outside the amateur bands to other services, to TV (which is the worst of all interference), and even to broadcast-band receivers. If the parasitic oscillation is near the operating frequency, it can cause instability in tuning the final. A tube carrying two frequencies can be overloaded unnecessarily, which means less power available for the signal you want to put out.

Here is how to check for parasitics: Place a short across the driver plate coil, and disconnect the antenna. Put a lamp in series with the 120-volt input to the transmitter to prevent overdissipation in case the plate current runs away. If the final has fixed bias replace it with a grid-leak resistor of about 10K to 20K. Turn the transmitter on, and revolve the final tuning capacitor. If there is any kick in the plate meter you have parasitic oscillation.

If there is an increase in plate current, leave the capacitor set at the point where the increase occurs and couple a grid-dip oscillator, switched to be an absorption wavemeter, to the output **(CAREFUL! HIGH VOLTAGE).** Go through the entire set of GDO coils and tune across each coil band and watch for an upward kick of the GDO meter. When the meter kicks up, this is the frequency of parasitic oscillation. If there is no indication from the output tank, try coupling the GDO to other wires in the final, both input and output. When the frequency of oscillation is found, turn the transmitter off and, using the GDO as a grid-dip oscillator, couple to all circuits in the final until you find one resonating at or near the frequency found by the

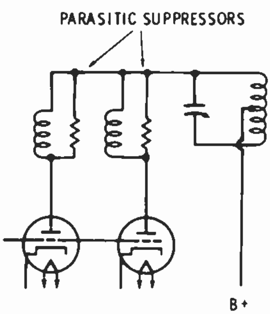


Fig. 5-20. Parallel a 1-mH choke with a 100-ohm resistor to make a parasitic suppressor. A 2-watt 100-ohm carbon resistor with about 10 turns of heavy wire wound over it and connected in parallel is often a good suppressor.

absorption wavemeter method. You may find one in the output and one in the input near each other in frequency.

Be particularly suspicious of r-f chokes. Check their resonant frequency with the GDO. If resonance is in one of the TV bands, replace the choke or chokes with other values. Parasitics are also cured by changing the length of wires, preferably shortening them where possible, or by introducing parasitic suppressors. A parasitic suppressor is an r-f choke with a resistor across it or in series with it. Fig. 5-20 shows the circuit for the most popular type.

Another method for checking for parasitics and determining their frequency is to couple the GDO into the driver coil with the driver plate tuning capacitor disconnected. Pull out the driver tube. (The GDO becomes the excitation to the final.) The rest of the setup is as described previously. Tune the GDO through all of its frequencies. If plate meter kicks up, that is the frequency of the parasitic. With the transmitter off use the GDO set for the same frequency that caused the kick, and go searching for the circuit resonating at the frequency.

CLASS-C OPERATION

The final r-f stage for c-w and a-m phone is operated at twice cutoff for bias and is called *class C*. The object in adjustments is to attempt to get the most out for the d-c input without exceeding the rated dissipation. This is done by loading, and differs whether it is for c-w intermittent service or a-m phone continuous-carrier service. The object is to achieve as near the maximum theoretical efficiency of 75 percent as possible. Linear plate modulation of a class-C amplifier depends on matching the impedance and having a modulator with sufficient power capability.

CLASS-B LINEARITY

When modulation occurs ahead of the final r-f stage, then the final *must* have linear amplification characteristics; that is, the output must be a duplicate of the input, at greater power. In audio amplification this is achieved by properly designed single-ended or push-pull class-A; or by push-pull class-AB_1, class-AB_2, or class-B amplifiers. These classes differ as to the amount of grid bias used, and whether or not the grids are driven positive at times. The latter three types must be used in push-pull to cancel even-order harmonics in audio amplifiers. The most efficient of these is class-B, with the bias at or near the cutoff point. At cutoff the no-signal plate current is nearly zero, and plate-current flow is intermittent. Thus the stage can be more heavily loaded without exceeding its dissipation rating.

Linear r-f power stages are operated in class-B bias conditions. Because of the flywheel effect of the output tank, only one tube need be used, push-pull operation not being necessary. A previous stage may be amplitude-modulated (although this is not usually done, because you are ahead in a-m phone operation with plate modulation of the final). Class-B amplifiers are used, almost exclusively, for ssb transmission.

If a class-B stage is not operated properly it will not be a linear amplifier. Nonlinearity produces distortion, both harmonic and intermodulation. Intermodulation distortion results in sum and difference products of frequencies beating with each other and producing other frequencies. Here is an example to illustrate this: Modulate a 3.9-MHz signal with two audio frequencies, 3000 Hz and 4000 Hz. With the carrier suppressed, the transmitted sidebands will produce radio-frequency signals of 3.897 MHz and 3.896 MHz (lower sideband). Harmonic distortion will double the 3.897-MHz signal to 7.794-MHz. Intermodulation distortion will beat the 7.794-MHz signal with the 3.896-MHz signal producing a new signal at 3.898 MHz. At voice frequencies the harmonics of any frequency will beat with any other frequency to produce new signals. This is transmitted as sideband splatter and gives rise to interference to other stations.

The only way to properly adjust a class-B amplifier for linearity is by the use of an oscilloscope. In addition to a scope a two-tone audio source is most helpful, but a single tone plus some carrier can be used. The human voice cannot be used.

Using a dummy load, couple the vertical plates of the CRT of the scope to the output using one of the methods described before. Couple the output of the speech amplifier to the horizontal input of the scope. With the transmitter tuned for normal voice operation apply sine-wave audio to the microphone input (a tone of about 3000 Hz and one about 4000 Hz). The output of two variable-frequency audio generators can be paralleled, with the signal level from each equal. Turn up the microphone gain on the transceiver and observe the pattern of the scope. It will be a bow-tie pattern like Fig. 5-21. Now turn up the microphone gain to the point just before the top

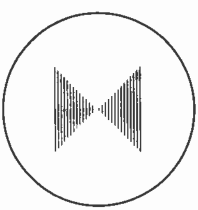


Fig. 5-21. A bow-tie pattern on a scope from the two-tone (or one-tone and carrier) test method. Sharp corners, and a perfect X outline means good linearity, without flattopping.

and bottom peaks on the bow-tie lose their sharpness. At this point you are at maximum output just before flat-topping. The signal must not be held there for more than about 30 seconds at a time. One easy method of observing the maximum signal trace without over-dissipation is to insert a semiautomatic key into the key jack and key a series of fast dots. This provides an intermittent signal for observation. The pattern may flicker but you can see what is happening at maximum output without worrying about overdissipation.

Rounded top and bottom corners indicate flat-topping. Flat-topping can result from three things: undercoupling to the antenna, overexcitation from the r-f driver, or too high a speech level (the most common reason). Increase antenna loading as you observe the pattern. If the flattening does not clear up, you were previously loaded enough, and probably you are just overdriving the input to the amplifier by too much rf or audio. The crossover point between the two parts of the bow-tie should be a perfect X. If there is a tapering at the points, it is usually a sign of too much bias. Adjust the bias for a perfect X at the same time not exceeding the recommended resting current to the plate with no modulation. Observe the sides of the pattern. They should be straight lines. If all voltages to the final are correct and antenna loading is the right amount, the sides will be straight and the stage will be operating in a linear manner.

CHAPTER 6

Antennas and Feeders

With the transmitter tuned up for maximum output from the r-f final amplifier, the object now is to get the energy into the antenna with the least amount of loss, and to tune the antenna for most effective radiation of the signal. Factors which affect the signal are: antenna resonance at the operating frequency, feed-line losses, and matching the feeder impedance to the antenna.

The most important antenna factor is antenna resonance. Any piece of wire, because of the inductance and capacitance distributed along its length, is resonant at some frequency. It is a resonant circuit like a coil and capacitor, and, theoretically, looks something like Fig. 6-1.

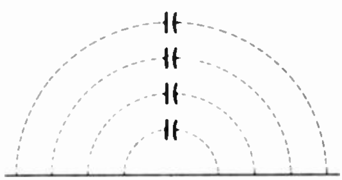


Fig. 6-1. Any piece of wire has inductance and distributed capacitance; therefore, any wire will resonate at some frequency.

Its length is closely related to the resonant frequency. A piece of wire 80 meters, or 262.5 feet, long will resonate at about 80 meters (3.75 MHz). The wire is said to be one wavelength long. It will also resonate at one-half wavelength long, and it then becomes a *half-wave dipole.*

The mechanical length of a half-wave dipole is slightly shorter than its electrical length. A factor (K) is involved in the formula. K is affected by the ratio of wire length to diameter of wire, and is theoretically 1 for an infinitely long wire of infinitesimally thin diameter. The factor K is .95 for most practical antennas for frequencies up to about 30 MHz, considering the usual wire size and length and the fact that it is not truly in free space (infinitely high). Taking into account $K = .95$, and converting to feet, the usual formula for a half-wave dipole is:

$$\text{Length (in feet)} = \frac{468}{f}$$

where,

f is the frequency in MHz.

A piece of information important in tuning an antenna is what it looks like to the end of the feed line. This is the factor of *radiation resistance*. The very center of an ideal half-wave dipole (one with a factor of $K = 1$) at resonance is equivalent to pure resistance of 73 ohms. That is, if it were replaced by a noninductive resistance of 73 ohms, the end of the feeder would not know the difference. At the center of an antenna the reactances (C and L) exactly cancel and the net effect is zero reactance or a pure resistance. The impedance (a combination of resistance and reactance) increases farther out from center, and is theoretically infinite at the ends. Maximum current and minimum impedance occur at the center; zero current and maximum impedance is at the ends. An antenna is affected by many factors: length of wire to diameter; height above ground; capacitance effect of nearby metal objects, reflectors and directors; etc. The center impedance of a 3-element beam is more like 15 ohms, for example, and the center of a half-wave dipole fairly clear above ground is usually around 72 ohms. What is important to tuning an antenna is to know that the center looks like a pure resistance, whatever its value may be.

The point to remember is that the R of the center of the antenna is the *load* on the transmitter, whether it is applied through a feeder or a combination of feeder and some transformer device.

TUNING THE ANTENNA

If you can get to the center of the antenna there are two methods of measuring for resonance. Each uses inexpensive equipment, the GDO (grid-dip oscillator) and the impedance bridge.

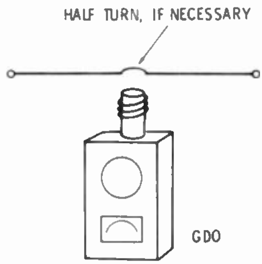


Fig. 6-2. An a-c operated or sensitive battery—powered grid-dip oscillator will measure resonance when coupled to the center of a half-wave dipole.

Couple the coil of a GDO to the closed center of the antenna (Fig. 6-2) and tune the GDO for a dip in the meter. For least effect of the GDO on the resonance of the antenna, the coupling should be as loose as possible and still be able to discern a dip in the meter. The best type of GDO out in the field away from a-c power is a battery-operated GDO using transistors or a tunnel diode. After getting the dip bring the GDO into the shack and check the frequency with your receiver. The accuracy of calibration of the usual GDO is not good enough for exact frequency measurement.

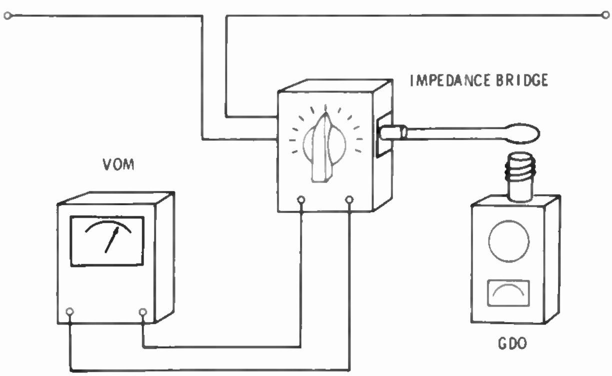


Fig. 6-3. Driving an impedance bridge with a GDO will give you antenna resonance and impedance at the center.

Method two uses a GDO and an impedance bridge (Fig. 6-3). The X (unknown) terminals of the bridge connect to the center of the antenna. The GDO is the signal source of the bridge. The output of a battery-operated GDO is pretty low, so it takes a very sensitive meter in the bridge circuit for an indicator. A 50-μA meter should be used. Adjust both the GDO and the bridge impedance arm for a zero reading on the indicating meter of the bridge. When the meter reads zero, the GDO is at the resonant frequency *and* the bridge will read the actual impedance of the center of the antenna. If no zero can be obtained it means you are not at the exact electrical center of the antenna. You may be physically, but the capacitance effects of nearby metal objects to the end of an antenna will shift the electrical center.

To find the electrical center, take a plier in your hand and touch along the antenna wire on each side of the feed point and watch the meters. Touching the antenna will cause both meters to jump a little. When you have touched a point on the antenna where neither meter is affected, you have found the electrical center. You can move the center closer to the feed point in the manner in which you trim your antenna for resonance to the operating frequency. If the electrical center is found to be to the right of the feed point, add wire to the right end if the measured frequency is too high, or cut wire from the left end if the measured frequency is too low. If the electrical center is to the left of the feed point, add wire to the left if the antenna is too short, or cut wire from the right if too long.

To accomplish a feed-line condition with no standing waves on the feeder, the feeder must connect to the exact electrical center. However, a small amount of reactance by being slightly off center will not have too much effect, and is usually not worth correcting.

The real problem in measuring antenna resonance is a physical one, how to get up there to the center of the antenna. High-frequency antennas and beams mounted on a roof are sometimes within reach. Long-wire dipoles for the lower frequencies are strung between poles or similar structures and just cannot be reached easily. The answer is to do your measuring at the lower end of a feeder—of a *half-wave* feeder, that is.

THE HALF-WAVE FEEDER

A feeder of any type, which is exactly one-half wavelength long, will reflect the same load impedance to the transmitter as it sees at the antenna. When a half-wave feeder is connected to the feed point of an antenna, the radiation resistance of the antenna at resonance will also appear at the input end of the feeder. Therefore, the same measurements can be made to the antenna (via a feeder) in the shack.

The characteristic impedance of the feeder is not important. It is also not necessary that the feeder match the antenna at the feed point. It is only important that the feeder be exactly one-half wavelength long electrically.

Coupling a GDO alone to the end of the feeder will not indicate antenna resonance. You will get a dip in the meter, but the resonance shown in the GDO will include any reactance in the feeder in case it is not an exact half wave in length. The best method of making a measurement is to use the number-two method mentioned earlier for measuring at the antenna, use a bridge and a GDO (or signal generator or even the transmitter vfo). The setup is shown in

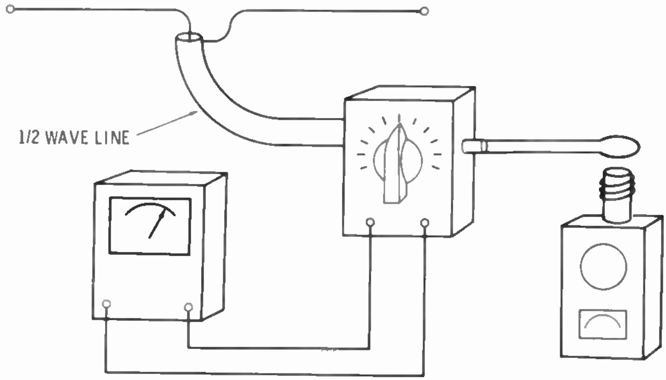


Fig. 6-4. Measurement of antenna resonance and impedance can be made from the shack with a half-wave feed line coupling your test equipment to the antenna.

Fig. 6-4. There will be a slight amount of error resulting from the use of a coupling loop at the input of the bridge because the reactance of the loop affects the frequency of the GDO. It can be eliminated by injecting a signal from a signal generator or your transmitter vfo and listening for the beat in the headphones plugged into the GDO, or by using a signal source such as a signal generator whose output cable is terminated with a resistor. Most signal generators worthy of the name have built-in terminations of 50 or 75 ohms. The direct connection to the bridge eliminates the coupling loop.

Set the GDO to the operating frequency (the frequency for which the feed line has been cut). Turn the bridge control arm for a zero reading on the indicating meter of the bridge, or the VOM connected to it. If the indicator reads zero, the antenna is resonant at that frequency, and the center feed-point impedance is the impedance shown on the bridge. If the bridge-balance indicator cannot be zeroed, there is reactance at the feed point to the antenna, meaning either the antenna is not exactly resonant or the feed point is not at the electrical center. If a perfect balance is not obtained, readjust the GDO for the best dip in the GDO meter. The new frequency reading will indicate the direction in frequency the antenna is off. From then on it is a matter of cut and try on the antenna until operating resonance is reached. If, with the antenna at resonance and a perfect balance in the bridge (indicator at zero) the bridge impedance shown is not the same as the rated characteristic impedance of the feed line, there is a mismatch between feeder and antenna. The antenna is at resonance but the feed point is not a perfect match to the feed line. This may or may not be important, depending on how far off the mismatch is, and how long the feed line is. Power loss in the feeder will be a combination of its construction, effect due to an SWR (standing-wave ratio) other than 1:1, and its length. Table 6-1 lists popular coaxial cables for feeders and important characteristics, including the velocity factor which will be discussed shortly.

The impedance bridge used in this test is the one described in the next chapter. A 50-μA meter movement gives about a one-third scale reading in the unbalanced condition when used with a GDO whose oscillator is a 6C4 tube powered from the 120-volt a-c line. The author's portable GDO did not have sufficient oscillator power to operate the balance indicator.

HOW TO CUT A HALF-WAVE LINE

If you purchase brand-new, good-quality feed line (coax cable, for instance), be sure a recognized manufacturer's name and type number appears imprinted on the cable at intervals. Your choice of cable will depend on the amount of power you expect to handle and how much you can afford to spend to keep losses down. First determine how many feet you will need from Table 6-1. Use the velocity factor shown in the table in the following formula for length needed, and add a few feet for error and trimming:

$$\frac{\text{Length in feet}}{492} = \frac{\text{velocity factor}}{f \text{ in MHz}}$$

This formula is set up as a ratio for easy slide-rule use. A half-wave feed line is open at both ends. Since it is difficult to couple a GDO to an open-ended feed line, short one end and measure it as though it were a quarter-wave long, but at half the frequency. Skin back one end and connect the inner conductor to the outside braid, and form a small loop, about the size of the GDO coils. Measure the exact physical length and calculate what the resonant frequency should be using the following ratio:

$$\frac{\text{Length in feet}}{246} = \frac{V}{f}$$

where,

V equals velocity factor,
f equals frequency in MHz.

This ratio is based on the formula for a quarter-wave stub, and the figure for frequency used is one-half the frequency to which you want to cut the half-wave line. Now couple the GDO (tuned for half the desired frequency) to the loop and adjust the tuning dial for a dip in the indicating meter. It should be close to the frequency calculated based on its length and published V (velocity factor). If the frequency does not agree and you have used a non-branded and unmarked cable, then suspect a difference in the velocity factor. Should you find this to be true, then solve for V in the foregoing ratio, and use the new figure for V to calculate the length of cable you will need for the frequency at which it is to operate. Cut the cable so it is only a few inches longer than desired.

Turn on your receiver and tune it to the middle of the band for which you want the half-wave cable. Couple the GDO to the loop on the cable and listen for its harmonic on the receiver (remember, the GDO is measuring at half the frequency). Cut off an inch at a time from the far, open end of the cable and redip the GDO until you hear the harmonic and see the dip at the same time. The feeder is now a half-wave line in the band you want. The results are the same whether the cable is stretched out or coiled up while making the frequency measurement.

An impedance bridge may also be used to measure a feed line for a half-wave length. Short the far end of the line and connect the open end to the X terminals of the bridge. Place the balance control at zero. Adjust the GDO used to drive the bridge for a zero balance on the bridge indicating meter. The line is a half-wave long at the indicated frequency. The measurement should agree with the formula:

$$L \text{ in feet} = \frac{492\,V}{f}$$

or the ratio for a slide rule is;

$$\frac{L \text{ in feet}}{492} = \frac{V}{f}$$

Since a half-wave line reflects the same impedance at one end of the line as the other, a short at one end will measure zero impedance at the other. The hookup is shown in Fig. 6-5.

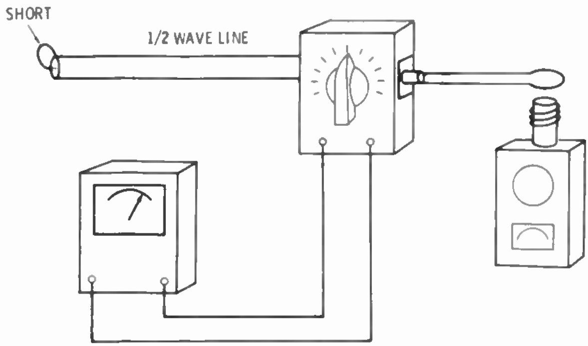


Fig. 6-5. A GDO and impedance bridge may be used to measure for a half-wave line length. With a short at the end of the line and zero setting on the bridge, the GDO will read the frequency when the indicator meter is at null.

While the line is set up for measurement for length and velocity factor you should make another check, and that is for characteristic impedance. This is particularly important on unbranded cable or on open-wire lines that you have made yourself. The hookup for this test is the same as in Fig. 6-5 and employs the impedance bridge and GDO. Instead of shorting the other end of the line, connect a carbon resistor whose value is that of the supposed characteristic impedance of the line between the two conductors. With the impedance-bridge control arm set at the value of the characteristic impedance, the balance meter should read zero. Furthermore, the meter should remain at zero with changes in frequency on the GDO. Should the meter not remain at zero with frequency changes, the characteristic impedance of the line is something other than the resistor value used at the end of the line. Substitute other values of resistors until there is no change. The final resistor value is the characteristic impedance of the line. The line can be any length for this test. With each change in resistor be sure to reset the balance arm on the impedance bridge to the same value as the resistor used.

MEASURING ANTENNA IMPEDANCE

With the half-wave feeder, GDO and impedance bridge, you are in a position to make a check of the impedance of the antenna at the center. This assumes that the antenna has been adjusted for resonance at the operating frequency.

The hookup is as shown in Fig. 6-4. Adjust the GDO to the frequency of the antenna and couple it to the input of the impedance bridge. Adjust the balancing arm of the impedance bridge for a zero reading on the balance indicator. Since a half-wave feed line reflects the same impedance it sees at the far end, the impedance value indicated on the balance arm of the impedance bridge is the impedance of the antenna where the feeder connects to it. If it is the same as the characteristic impedance of the feeder you are matched, and any length of the same feeder, either longer or shorter than the half-wave length, may be used. However, if coaxial feed line is used you should correct the unbalance between the use of an unbalanced line to the balanced feed point on the antenna. This is corrected by using a 1:1 balun transformer at the antenna or by using a "gamma" matching device.

A word of caution on using the impedance bridge. Do not attempt to use the transmitter as the signal source of the impedance bridge. The components of the bridge have limited power-handling capability, and if more than about 4 watts are used to drive the bridge, the components will burn out.

ABOUT SWR

If the above hookup shows a difference in impedance between the antenna and the characteristic impedance of the line, the mismatch will result in standing waves on the line during transmission if it is not corrected. The extent of the standing waves is measured in terms of the ratio between the forward wave in the line and the portion returning, or reflected, back into the line as a result of the mismatch. Knowing the impedance of the feed point on the antenna and the characteristic impedance of the line, the SWR (standing-wave ratio) is:

$$\text{SWR} = \frac{Z_A}{Z_L},$$

where,

Z_A is measured impedance of the antenna,
Z_L is the characteristic impedance of the line.

THE VALUE OF A LOW SWR

While the insulating material used inside the coaxial cables between the inner conductor and shielded braid are of low-loss quality in the cables of recognized brands, there is some loss, nevertheless. The higher the voltage developed in the cable for a given power level is, the higher is the percentage of loss. The higher the frequency is, the higher is the percentage of loss, and, the longer the cable is, the greater the loss. All of this is expressed in dB (decibels) down from a perfect cable. Each 3 dB of loss is a 50-percent loss of power; this is a convenient figure to keep in mind.

Table 6-1 listing popular feeders shows the attenuation in dB for each 100 feet of cable at various frequencies. From this you can see

Table 6-1. Attenuation per 100 Feet of Common Coaxial Cables

RG TYPE	Attenuation in dB at MHz					Velocity Factor	Power Rating in Watts at 50 MHz	Characteristic Impedance in Ohms
	1	10	50	100	200			
8A/U	.15	.53	1.3	1.9	2.7	66	1500	52
11A/U	.21	.66	1.5	2.3	3.3	.66	1000	75
17A/U	.06	.24	.62	.95	1.5	.66	5400	52
58A/U	.44	1.4	3.3	4.9	7.4	.66	425	52
59A/U	.33	1.1	2.4	3.4	4.9	.66	540	73

Courtesy Amphenol Corp.

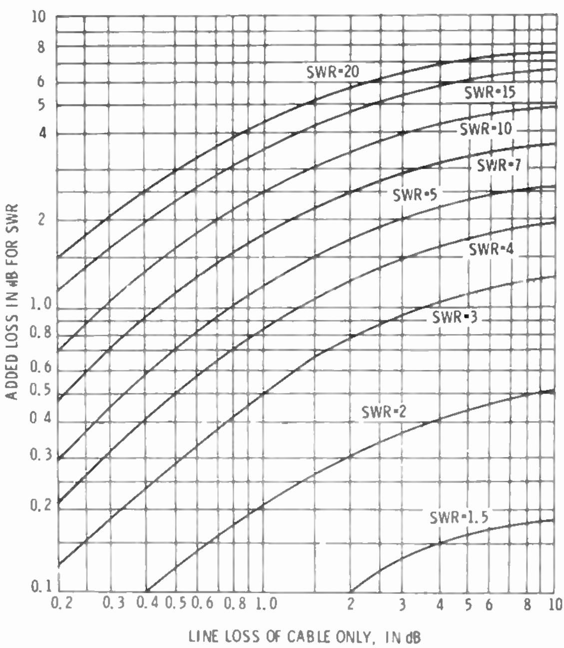


Fig. 6-6. Chart shows additional loss in dB when the SWR is greater than 1:1.

that long cables at very-high frequencies can mean a considerable loss of power; this is power that dissipates as heat in the line and never gets to the antenna. In addition a high SWR can increase the losses because of the higher voltages resulting from the higher SWR. The chart in Fig. 6-6 shows the added dB loss resulting from SWR ratios higher than 1:1. This figure is added to the dB figure based on length and frequency for a figure of total loss. For example:

A 50-ft. length of RG-58/U cable has about a 2.3-dB loss per hundred feet at 30 MHz. The center of the antenna was measured to have an impedance of 73 ohms at resonance (a perfect situation). The nominal characteristic impedance of the cable is 53.5 ohms. The SWR because of mismatch is $\frac{73}{53.5}$ or 1.36:1 by calculation. From the chart of Fig. 6-7, it can be seen that the additional loss because of this mismatch is so small as to be off the chart, therefore in-

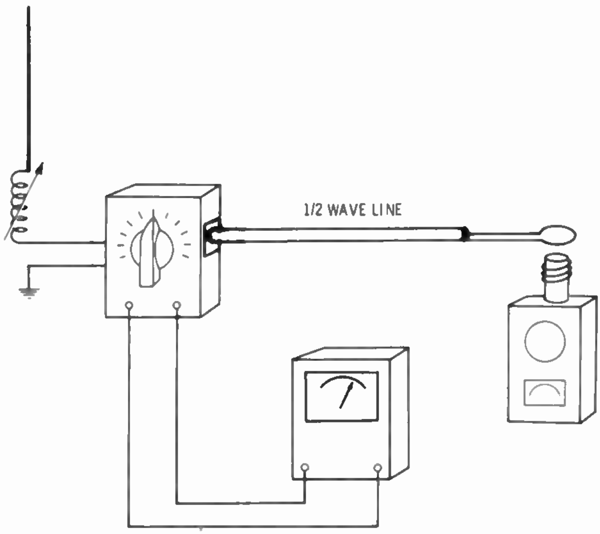


Fig. 6-7. Recommended instrument hookup when adjusting the base loading coil on a vertical antenna.

significant. The loss due to line length is one-half of 2.3 dB (for 50 ft) or 1.65 dB. Converted to power loss, as obtained from any power-versus-dB chart or table or using a slide rule or logarithmic table, mathematically, we solve for:

$$dB = 10 \log \frac{P_O}{P_I}$$

where,

P_O is power into the feeder,

P_I is power into the antenna.

Therefore, one-tenth 1.65 is .165, and the antilog of .165 is 1.46. This means that for every watt into the antenna, it takes 1.46 watts into the feed line based on cable type, length, and frequency only. However, if the SWR is 3:1, the chart will show that an additional 0.6 dB of loss must be added to the 1.65-dB figure, for a total loss of 2.25 dB. The log tables will show it takes 1.67 watts into the feeder to get 1 watt into the antenna, or about 40 percent of the power is lost in the cable compared to a 31.5-percent loss from the cable alone. The RG-58/U is obviously not a good cable for long lines at high frequency. A better choice would be the popular RG-8/U.

TUNED LINES

All that has been said about antenna resonance and its measurement with a half-wave line applies only to "matched" type or untuned lines, whether balanced (parallel-wire type) or unbalanced (coaxial type). The output of most transmitters is designed for direct unbalanced connection of a 50-ohm line, which is the most popular line in use. However, tuned feeders are still popular with many amateurs and have the advantage of multiband operation without the use of antenna traps, and the use of tuned feeders reduces the problems of resonating the antenna precisely. The disadvantage of tuned feeders is the need for careful installation, particularly with regard to the method by which they are brought into the shack from the outside. Also, tuned-feeder systems require an antenna tuner in the shack, which is used to resonate the system and couple it to the transmitter.

When a tuned feeder is connected to the center of a dipole, there is no need to trim the antenna to exactly one-half wave length. It is sufficient to cut approximately based on the standard formula for antenna length.

With the feeders connected to the center, currents and voltages in the two legs of the feeder will always cancel and radiation is only from the antenna. Of course, standing waves will be high on the feeder, but wide-spaced conductors with good insulating spacers keep losses down. Even if the antenna length is off, there is always balance in the line. And since the entire system (antenna and feeder) is resonated at the antenna tuner, resonance for any frequency is easily accomplished.

"Zep"-fed (end-fed) antennas can result in some unbalance in the feeder if the antenna is not at resonant length. While an end-fed antenna with tuned feeders can be still be resonated with the antenna coupler system in the shack, if the antenna length is too far off there will be unbalanced voltages and currents in the feed line and some feed-line radiation. Usually that unbalance is not important enough to go through a lot of trouble to achieve exact resonance of the antenna, and the formula for cutting the antenna mechanically is good enough.

TUNING A VERTICAL

A vertical antenna is considered to be a quarter-wave antenna with the ground acting as the other quarter wave when the antenna is fed at the point where the vertical connects to the ground, usually right at the ground point. Because of the conductivity of the earth where radials are buried in it, and the infinite length of cold-water pipes when they are used as the ground, the *total* length of the antenna is anything but a halfwave, or twice the vertical quarterwave. You might say it is resonant at any frequency, and tuning means shifting the center current loop to the feed point at the frequency you want to operate.

The length of a simple vertical is based on the formula:

$$L\text{ (ft)} = \frac{234}{f\text{ (MHz)}}$$

Vertical antennas can be tuned to precise frequency (current loop at the feed point) with a small reactance (either capacitive or inductive) in series; a variable capacitor in series if the antenna is too long, or an adjustable inductance in series if it is too short. Because of their height, vertical antennas can become pretty clumsy in the lower amateur bands. They are common for mobile operation on 10 meters, but for fixed or mobile operation at lower frequencies, short verticals with base-loading coils or traps are more common. For very short dimensions (as for mobile operation) better performance is obtained if the loading coils are up higher in the antenna. Since the area of maximum r-f current in the antenna is the most effective radiator, it should be left free to become part of the vertical antenna portion. Common low-frequency verticals for fixed-station operation have resonant traps which give them multiband resonances as well as shortening the overall length. These are not subject to tuning when purchased complete.

Among home-made fixed-station verticals the most common is the short vertical rod (usually about 22 feet high and base-loaded with a tapped coil). Trimming the antenna means adjusting the tap on the coil for the purpose of placing the current loop at the feed point. While a perfect vertical without lumped inductances and with a perfect ground of a number of buried radials typically has a center impedance of about 30 ohms, the shortened version with a base-loading coil, especially if the ground system is not perfect, can have a center impedance as high as 200 ohms. With a poor ground it is sometimes impossible to find resonance by coupling a GDO to the antenna at the feed point. The best method is to use the GDO as a signal source and adjust the antenna for the lowest impedance at the feed point with an impedance bridge. Fig. 6-7 is the hookup for tuning a one-band vertical with a base-loading coil. This hookup is for rough adjustment at the antenna. The feed line must be a half wavelength long at the operating frequency in order to show the same impedance at each end.

Set the GDO to the operating frequency at the shack end of the line. Adjust to nearly exact frequency by listening to the beat in the re-

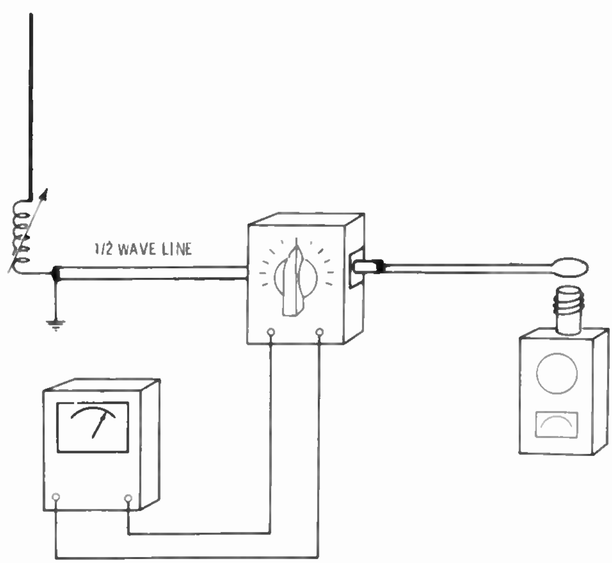


Fig. 6-8. After adjusting the vertical with the setup of Fig. 6-7, bring the impedance bridge into the shack for the final touchup.

ceiver with the bfo on. Connect the impedance bridge and meter between the feeder and the feed point of the antenna. Adjust the taps on the coil, at the same time varying the bridge arm for lowest reading on the balance meter. Disconnect the bridge and reconnect the feed line. Bring the bridge inside and connect it as shown in Fig. 6-9. This time juggle the bridge arm and the frequency setting of the GDO for lowest reading on the bridge meter. You will probably find that a complete zero can be obtained at a frequency slightly off the frequency at which you first set the GDO for the outdoor check on impedance. If the frequency of perfect bridge balance is higher than the desired operating frequency, add a little inductance to the base coil at the antenna. If lower, move the tap to reduce the inductance. You may find that even only a quarter turn on the coil will make a big shift in frequency. Your antenna is resonant to the desired frequency when the meter balance is zero, and the antenna impedance is the impedance read from the bridge arm. That impedance may be anything from about 30 ohms to 200 ohms; but regardless of the impedance, the antenna is resonant, which is the important thing. If the impedance happens to come close to the characteristic impedance of the feed line, the mismatch is unimportant and can be ignored. If the mismatch is 2:1 or greater, check for power loss as described

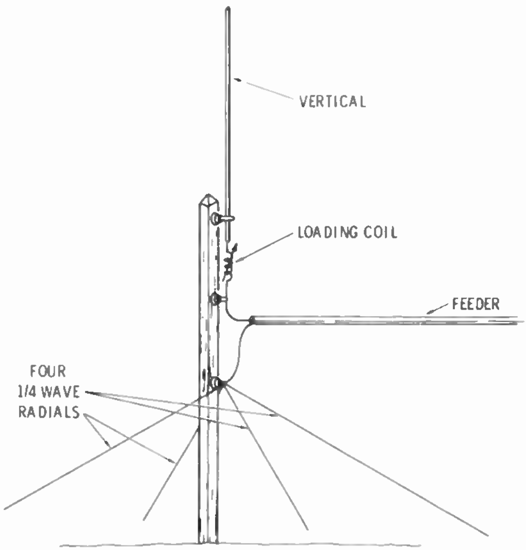


Fig. 6-9. When good grounding is impractical, quarter-wave radials above ground will give superior results.

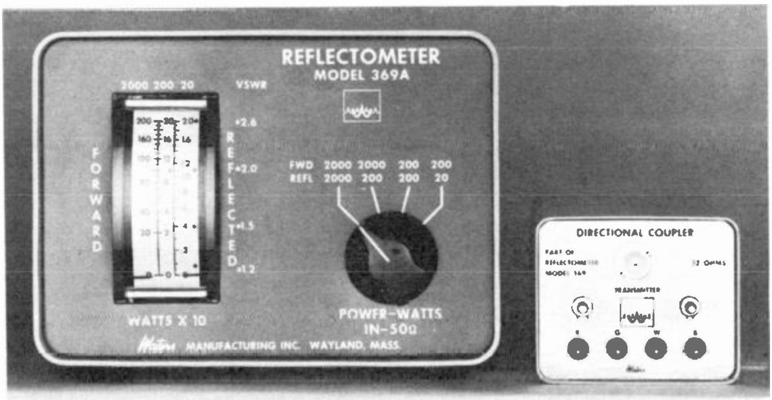
before in this chapter, and correct the mismatch if power loss is excessive. The power loss will depend on the operating frequency, the quality of the feed line, and its length

When the impedance bridge is connected to the base of the vertical as shown in Fig. 6-7 it introduces discontinuity and stray capacitance or inductance that may prevent obtaining a zero balance on the impedance meter. With the bridge at the input end of the cable, the normal operating connection at the antenna gives better balancing results (Fig. 6-8).

If the resulting impedance is far from the 30-ohm average of a vertical antenna, it is because of the use of the lumped inductance at the base of the antenna, a less than perfect ground, or the use of an unbalanced cable (coaxial cable, for example) connected to a probably balanced feed point. Although the antenna is resonant (and this is most important), the mismatch to the cable characteristic impedance may result in a greater-than-desired power loss, and an inability of the transmitter to load properly.

ABOUT VERTICAL ANTENNA GROUND

Probably the most important consideration in the performance of a vertical antenna is a good ground. In many areas of the country, such as the Southwest where the climate is dry and the soil sandy, buried radials of random length for a ground are almost useless. Un-



Courtesy Waters Manufacturing, Inc.

Fig. 6-10. This reflectometer has a meter with two pointers. One reads forward power and the other reflected. When loaded with exactly 50 ohms, the power readings are actual. VSWR is shown in the vertical panel scale to the right of the meter.

Fig. 6-11. The Model W-4 Wattmeter reads actual power into a 50-ohm load. Flipping a switch changes the reading from forward to reverse power. A calculator, supplied with the instrument is used to convert to SWR.

Courtesy R. L. Drake Co.

Courtesy Philco-Ford Corp.

Fig. 6-12. The Sierra Bidirectional Power Monitor reads forward power and SWR directly on the meter scale. Plug-in units, on top, are multipliers for reading power in watts into a 50-ohm load.

less the buried radials are themselves a quarter-wave long it is better to use above-ground radials (Fig. 6-9). With the antenna up in the air, and about four quarter-wave radials coming off at about 45°, you will have less trouble finding resonance and matching the impedance. The center impedance of an antenna built like this is about 72 ohms.

THE SWR METER

The antenna impedance bridge described previously cannot be left in the feed line while operating the transmitter. A device for indicating relative output power and SWR that can be left in the line is the SWR meter (Figs. 6-10, 6-11, and 6-12). The construction of one is described and shown in the next chapter. The circuit diagram is duplicated in Fig. 6-13 for illustrative purposes. The SWR meter reads

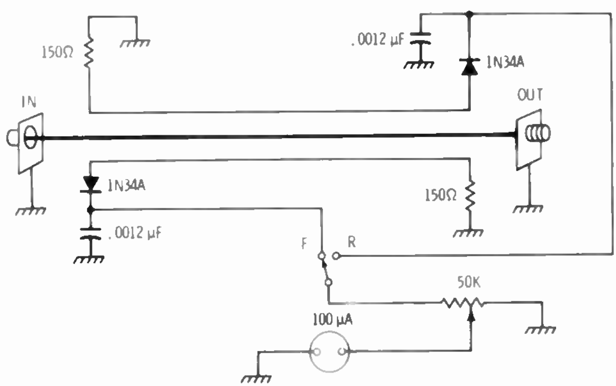


Fig. 6-13. Circuit diagram of the SWR meter described in Chapter 7.

the forward-going wave and the reflected wave which is due to mismatch, and from this information the SWR ratio is obtained. The values of R_1 and R_2 are selected for use on a particular line characteristic impedance, and the readings will be true only when the line used between the SWR bridge and antenna has that impedance.

The SWR or standing-wave ratio is a function of the forward voltage versus the reflected voltage, and it is based on the following formula:

$$\mathrm{SWR} = \frac{V_O + V_R,}{V_O - V_R}$$

where,

V_O is the outgoing voltage,

V_R is the voltage value of the reflected wave.

The voltage need not be absolute, but only relative. For example, the forward voltage is at full scale of the meter, which is 1 on a 100-μA meter. Switching to reflected wave the needle reads 0.5, or midscale. Substituting in the formula:

$$\text{SWR} = \frac{1 + .5}{1 - .5}$$

$$= \frac{1.5}{.5}$$

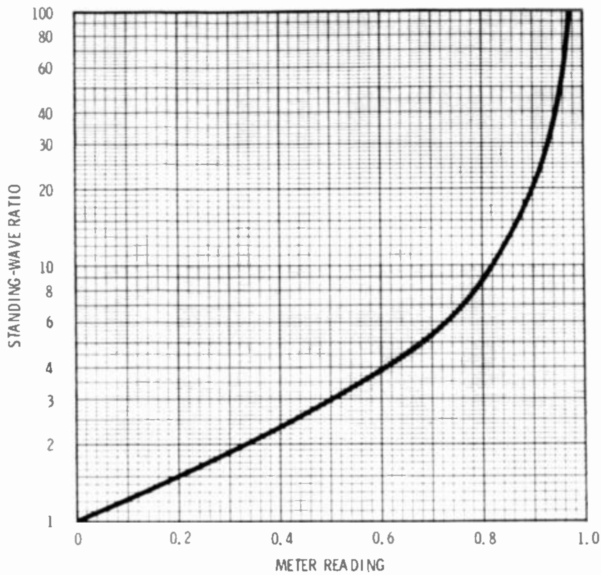
$$= 3$$

A handy device for reading SWR directly is the chart of Fig. 6-14.

USING THE SWEEP GENERATOR FOR ANTENNA RESONANCE

An excellent method for "seeing" the results of antenna off-resonance and adjusting for resonance is through the use of a sweep generator and scope, employed in the same way as using a sweep generator for aligning a receiver. In this case it is the resonance of the antenna that is under investigation.

The sweep generator connects to the same bridge as used previously. It is the signal source, but it sweeps a range of frequencies instead of supplying a single frequency. The balance indicator is a scope instead of a meter. The horizontal sweep of the scope is synchronized with the sweep of the generator. As the generator sweeps past the point of antenna resonance the scope will show a dip in the trace, which is equivalent to the zeroing of an indicating meter. The trace will look something like Fig. 6-15A or B. In Fig. 6-15A the dip is easily recognized. In Fig. 6-15B the dip is inverted and looks like a peak, but it is a dip nevertheless. It is inverted because of the number of stages in the scope and is a matter of phase. Consider the trace with relation to the base line, regardless of the direction. Fig. 6-16 is the hookup for using the sweep. It is similar to the hookup using a straight signal generator and a microammeter balance indicator, except for the type of instruments. If the sweep generator does not have a built-in marker oscillator, it will be necessary to use one externally. The marker generator places a pip on the trace and is the method for determining exact resonance of the antenna. Many sweep generators have facilities for plugging in a crystal for marker purposes. An amateur-frequency crystal for the band for which you are trimming your antenna is an excellent marker.



Courtesy Allied Radio Corp.

Fig. 6-14. A chart like this can be used with any meter scale. Adjust the forward reading to 100-percent of scale, and convert reflected reading to SWR from this chart.

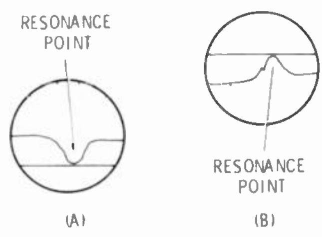


Fig. 6-15. Appearance of the dip on a scope indicating antenna resonance with a sweep generator. The customary dip is at A. At B it appears as a peak, but it is the "dip." The difference is the result of phasing.

For this method the sweep generator must have fairly high output, and the scope must have good sensitivity but not necessarily high-frequency response. There is considerable loss of signal in the bridge and it takes about 1-volt output from the generator, plus a sensitivity of about .025 V/inch on the part of the scope, to get a normal-sized pattern. No demodulator probe is necessary for the scope since the rectifier in the bridge acts as one.

Set up the sweep generator for about 2-MHz sweep width, and the place the tuning dial to the frequency at which the antenna is to be

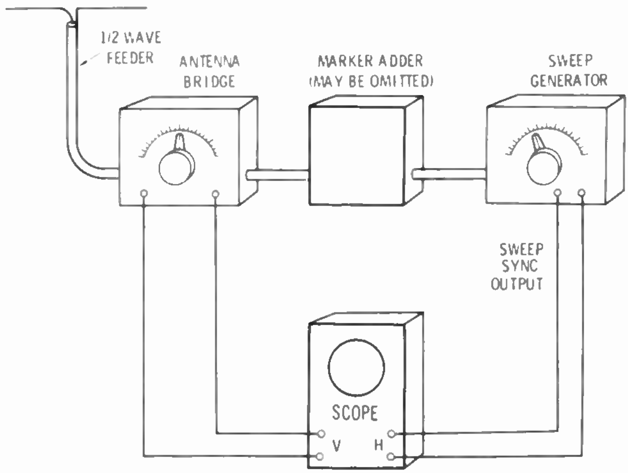


Fig. 6-16. Hookup for using a sweep generator and scope to observe antenna resonance.

measured. Place the bridge control arm at about 50 ohms. The pattern you will see will be nearly like one of those illustrated here, but the dip will probably not touch the base line. Now adjust the balance arm on the bridge until the dip touches the base line. The reading on the bridge arm will be the impedance of the antenna at the center, and the point where the dip touches the base line is the resonant frequency. Turn up the gain on the crystal marker (if one is used) and see where the pip appears. If an external variable signal generator is used as a marker, adjust it until the pip appears at the center of the dip. The frequency of the marker generator is now at, or near, the resonant frequency of the antenna.

If resonance appears at a frequency other than the frequency for which the half-wave feed line was cut, reactance in the feed line will affect the exact resonant frequency. However, an antenna resonance off the desired frequency of operation is indicated. How far off and the direction is the real point of interest. When the antenna is trimmed to the desired operating resonance (equal to the half-wave length of the feeder), the frequency will be correct.

Fig. 6-15B was drawn from an actual test for resonance of a new 40-meter vertical antenna. The off-center marker pip was from a 7.150-MHz crystal. It indicated that antenna resonance was too low, and needed trimming to bring the peak of the dip (inverted here) at the pip marker.

A variable marker can be moved back and forth on the scope screen to determine how broad the antenna resonance is. If you experience intermittent waviness in the trace lines, it is the result of a close-by broadcast station.

FRONT-TO-BACK ANTENNA RATIO

Beam element length and spacing, as given by standard formulas or supplied by the manufacturer (if purchased) usually provide results so nearly optimum it does not pay to make field adjustments. However, it is sometimes desirable to know that everything is working as it should. Resonance and feed-point matching measurements are made as described before. Forward-gain measurements or front-to-back measurements require some means of picking up the forward and rear power and comparing them. This is done with a field-strength meter whose pickup antenna is placed in front of the beam at least three wavelengths away.

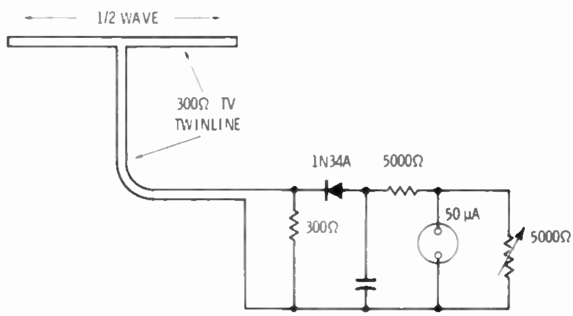


Fig. 6-17. Field-strength meter using 300-ohm TV twinline as a dipole and feeder.

Fig. 6-17 is the sketch of a simple field-strength meter for this purpose. It consists of inexpensive 300-ohm TV twinline, both as a dipole and a feeder. At the end of the feed line is a 300-ohm load resistor, followed by a rectifier, r-f bypasses, and a sensitive meter. A variable shunt across the meter reduces sensitivity for higher-power transmitters. A series resistor is needed for obtaining linearity in the meter reading because of the nonlinear characteristics of the diode. It reduces sensitivity, of course, and may be eliminated if only a relative reading rather than an absolute reading is acceptable. The meter may be a 50-µA to a 1-mA type depending on need for greater sensitivity. If a 50-µA meter is used, the variable shunt resistor should be 5000 ohms, down to 200 ohms for a 1-mA movement.

Where high sensitivity is not necessary the pickup antenna need not be a full half wave. The entire r-f portion may be built on a plastic board with short dipole elements sticking out each end, all the r-f components on the board, and any kind of lamp cord feeding away to the meter located nearer the antenna so it can be seen if adjustments are made. The meter may be mounted to the board also if any F/B (front-to-back) measurements are to be made.

The field-strength dipole must be hung parallel to the antenna under measurement. It is important that polarization be the same. Unless the field-strength dipole is hung several wavelengths away from the transmitter antenna it may be part of the inductive field and affect the antenna dimensions. The field-strength antenna feeder must be brought away from its dipole at right angles; it must not pick up signal from the transmitter. The same is true of the transmitting-antenna feeders; they must not radiate energy. If the antenna is not gamma-matched or does not use a balun (if feeding a balanced antenna with unbalanced coax), the feeder is apt to radiate and give wrong readings.

If the series resistor is used on the meter, the meter readings may be taken as a measure of relative voltage. Rotate the beam to point to the field-strength antenna, and adjust the meter for full-scale reading (E_1). If it cannot be adjusted to full scale, note the reading. Rotate the beam to point in the opposite direction and note the meter reading (E_2). Convert the readings to dB with the following formula:

$$dB = 20 \log \frac{E_1}{E_2}$$

where,

E^1 is field-strength reading off the front,
E^2 is field-strength reading off the back.

If you make adjustments to the elements of the beam, remember a few things: Increasing the recommended length of the reflector often results in increased F/B ratio, but not an improvement in power gain. Best F/B ratio is seldom the same as best power gain. Any adjustment to any element will affect the length of other elements. Recheck particularly the driven element for resonance and feed-point match. If the driven element needs adjustment, go back over the other elements again. You may find it necessary to repeat the adjustments to each element until a nearly perfect ratio (resonance and feeder match) is obtained.

MATCHING TRANSMISSION LINE TO ANTENNA

A feed line of the untuned type has a characteristic impedance based on physical dimensions determined by the diameter of the conductors and the distance between them. When terminated by a pure resistance equal to the characteristic impedance, there are no standing waves and loss is at a minimum. The center of a resonant antenna is a pure resistance, and when its value is the same as the characteristic impedance of the feed line, a direct connection can be made. When the antenna impedance is different from the line impedance, standing waves will be set up in the line, and the losses will go up. If the mismatch is great enough, the losses become important enough to do something about correcting the mismatch. There are three popular methods of correcting the mismatch.

THE GAMMA MATCH

One of the most convenient and effective (and most popular) methods of matching the low impedance of the driven element of a beam is the gamma-match method. In addition to increasing the feed-point impedance, it provides for changing from a balanced antenna center to an unbalanced coaxial cable. The diagram is shown in Fig. 6-18.

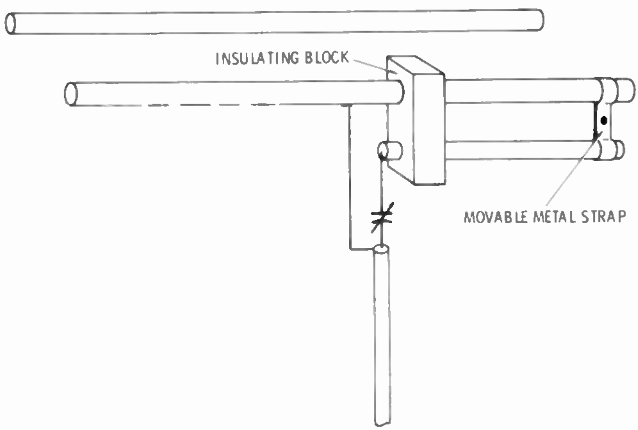


Fig. 6-18. One of the most popular methods of matching a coaxial cable to the center of the driven element of a beam is the gamma match. Adjust the position of the slider and the value of the capacitor for a 1:1 SWR. Voltage is low at the feed point, so the capacitor may be a receiving-type variable.

The half-wave driven element is not broken in the center, but is made continuous. A similar aluminum rod is supported about two inches below the driven element, extending from the center to about 40 percent of the length of one side. It is insulated from the driven element at the center. At the far end is a metal support, which is constructed to permit sliding it up and down the matching element.

The shield side of the coaxial-cable transmission line connects to the center of the driven element. The inner conductor connects to a variable capacitor (about 150 pF for 14 MHz; more for lower frequencies, less for higher frequencies) connects between the inner conductor and the beginning of the matching element. Since the matching element is shorter than a quarter wave, it will have inductive reactance. The capacitor tunes this out.

Adjustment is made by sliding the end strap back and forth, and at the same time adjusting the variable capacitor, until an SWR of 1:1 is read at the input of the transmission line.

THE Q-BAR

A quarter-wave length of feeder open at both ends acts like a transformer. When it is terminated in its characteristic impedance at one end, the other end will look like the same impedance. When the one end is terminated in an impedance higher than its characteristic impedance, the other end will look like an impedance lower than its characteristic impedance—thus the transformer action. For matching feed lines to antennas, this quarter-wave section has been popularly known as a "Q-Bar."

Two unequal impedances can be matched when the impedance of the Q-bar has a characteristic impedance that is the mean value between the two unequal impedances. The formula is:

$$Z_Q = \sqrt{Z_1 \times Z_2}$$

For example, the center of a resonant antenna was found to have an impedance of 200 ohms by measurement with an impedance bridge as previously described. It is desired to match a 50-ohm line to that antenna. Using the above formula the Q-bar impedance must be:

$$\begin{aligned} Z_Q &= \sqrt{200 \times 50} \\ &= \sqrt{10{,}000} \\ &= 100 \end{aligned}$$

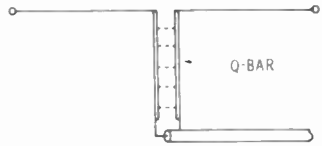


Fig. 6-19. The usual Q-bar quarter-wave matching section consists of parallel conductors, spaced with insulators. The desired impedance of the Q bar can be predetermined by formula, and constructed by spacing and radius of the conductors.

A Q-bar is usually made up of a pair of parallel conductors one-quarter wave long (Fig. 6-19), and its characteristic impedance is reached by observing the following formula:

$$Z_Q = 276 \log \frac{d}{r}$$

where,
- d is the distance between conductors, center-to-center,
- r is the radius of the conductors.

Once the Q-bar is made to the above dimensions the exact quarter-wave length can be measured by shorting one end and coupling a GDO to the short. Start with a length slightly longer than a quarter wave, and cut from the open end while checking the frequency of the dip on the GDO until resonance at the operating frequency is reached.

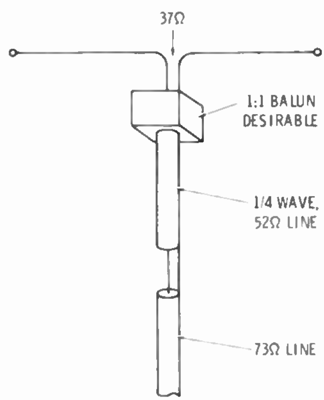


Fig. 6-20. Under certain conditions, coaxial cable may serve as the quarter-wave matching transformer.

Although parallel bars can be constructed to meet a specific impedance need, it is sometimes possible to find manufactured cable to meet the impedance needed for a quarter-wave Q-bar. Fig. 6-20 shows one example. The center of the driven element of a beam was found to have an impedance of about 37 ohms. By using an electrical quarter-wave length of 52-ohm coaxial cable, an impedance transformation was made to 73-ohm coaxial cable for the feed line.

MATCHING STUBS

When the transmission line and antenna feed point are mismatched, there will be standing waves along the line. By attaching a prescribed length of feeder at a prescribed distance down the feed line, the reactance of the unmatched line can be "tuned-out," and at the same time a value of impedance created at the junction that will match the line. This is known as *stub-matching* and it looks like Fig. 6-21 for both open lines and coaxial cable.

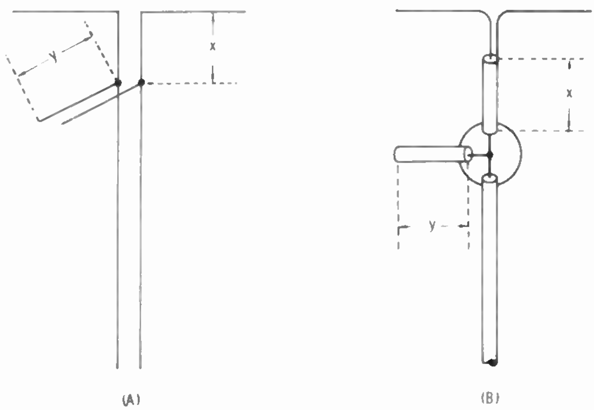


Fig. 6-21. Matching stubs may be added to open transmission lines (A), or to coaxial lines (B). Dimensions for X and Y are obtained from formulas and the SWR before correction.

The length of the stub and the point of attachment is determined by formulas. The only thing we need to measure beforehand is the SWR between the unmatched line and antenna as determined by the ratio indicated on an SWR meter.

Connect the feeder directly to the center of the antenna in the usual way. Measure the SWR at the input of the line (at the transmitter). (Note: when using coaxial cable, which is an unbalanced transmission line, better results will be obtained if a balun is added to the balanced center of the antenna. A balun changes the balanced center to an unbalanced line, and prevents rf from appearing on the outside of the line. The most convenient is a balun transformer wound on a toroidal ferrite core. These may be purchased or you can make your own from information in amateur radio handbooks.)

The following formulas are for use where the stub is made of the same type line (that is, same impedance) as the transmission line. Working out the formulas will give the distance (X) and length of stub (Y) as shown in Fig. 6-22. First, solve for length in degrees, then convert degrees to electrical length at the desired frequency.

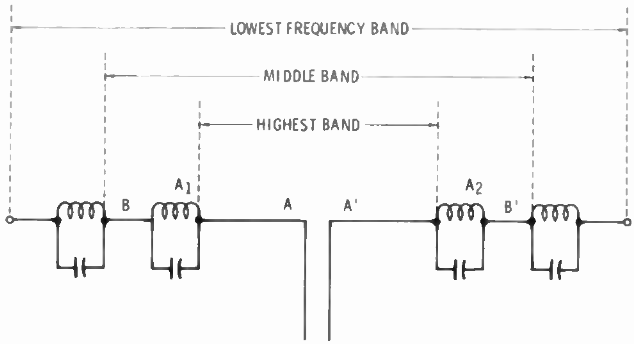


Fig. 6-22. All-band trap antennas use tuned circuits to isolate the higher band from the next lower band. At a lower frequency, the inductance of the higher-band traps become part of the added elements.

Where the antenna impedance is greater than the line impedance, the formula is:

$$\tan X = \sqrt{\text{SWR}} \qquad \text{and} \qquad \cot Y = \frac{\text{SWR} - 1}{\sqrt{\text{SWR}}}$$

Where the antenna impedance is less than the line impedance, the formula is:

$$\cot X = \sqrt{\text{SWR}} \qquad \text{and} \qquad \tan Y = \frac{\text{SWR} - 1}{\sqrt{\text{SWR}}}$$

The degrees obtained from these formulas must be converted to electrical length from:

$$\text{degrees} = 360 \text{ times length in wavelengths}$$

The answer will be a decimal fraction of a wavelength, which is added to the standard formula for length in feet:

$$\text{Length in feet} = \frac{984 \times V}{f\ (\text{MHz})} \times \text{wavelength decimal}$$

where,

V is the velocity factor of the cable,
f is the frequency.

Example: A 52-ohm coaxial cable is to feed the center of the driven element of a beam. The SWR reading before any correction was applied is 3:1. The coaxial cable has a velocity factor of .66, and the frequency of operation is 14.25 MHz.

$$\tan X = \sqrt{\text{SWR}} = \sqrt{3} = 1.7321$$

Checking a trigonometry table, or using the slide rule, the angle for which $\tan X = 1.7321$ is 30°. 30° is .083 times 360°. The length in feet of section X, or the tap on the line will be:

$$X\ (\text{feet}) = \frac{984\ V}{f\ (\text{MHz})} \times .083$$

$$= \frac{984 \times .66}{14.25} \times .083$$

$$= 3.81 \text{ feet}$$

Therefore, distance X will be 3 feet 9¾ inches from the antenna feed points, for operating at 14.25 MHz.

The stub length (Y) for the same case is:

$$\cot Y = \frac{\text{SWR} - 1}{\sqrt{\text{SWR}}} = \frac{3 - 1}{\sqrt{3}} = \frac{2}{1.7321} = 1.155.$$

The angle for which $\cot Y$ is 1.155 is 41° (approx). The angle 41° is .114 times 360°. The length in feet of stub Y will be:

$$Y\ (\text{feet}) = \frac{984\ V}{f\ (\text{MHz})} \times .114$$

$$= \frac{984 \times .66}{14.25} \times .114$$

$$= 5.18 \text{ feet, or 5 feet } 2+ \text{ inches}$$

Because the stub Y is shorter than a quarter wave, its reactance tuned out the reactive component of section X, so that the line is looking at a pure resistive load of 52 ohms at the junction. While standing waves will appear on the length X plus Y, the feed line will be flat and should then show a 1:1 SWR ratio at the transmitter end.

ANTENNA TRAPS

A popular method of constructing a multiband antenna is with resonant traps. Fig. 6-22 shows a dipole for three-band operation using two pairs of traps. Traps also reduce the overall length.

The dimensions of sections A and A′ of the sketch are tuned to the frequency of the highest band as if there were no traps or other element lengths there. Traps A_1 and A_2 are parallel-resonant tuned circuits. Their resonant frequency is the frequency of the highest band, or dipole section A—A′. The high impedance of the parallel-resonant trap isolates the high-band dipole (A—A′) from the rest of the elements when a signal of that frequency is fed to the antenna. For the next band, the traps are no longer resonant but do represent an inductance in series with the elements of A and A′. Elements B and B′ are added and tuned to the next or middle band. The middle-band dipole now consists of B—A_1—A, and A′—A_2—B′ resonant to the middle band. Another pair of resonant traps tuned to the midfrequency of the second band are added, followed by elements to resonate to the lowest band with all the elements of the high band, middle band, and the inductances. Thus we have a three-band antenna, resonant to each of three band frequencies fed to it from the same feed line and the same center feed point.

Traps can be set precisely to frequency by coupling a GDO to the coil and adjusting to resonance, provided the coils are not enclosed in a metal sleeving, as some are. You can make your own traps from air-wound coils and parallel capacitors, setting them to frequency with a GDO. Enclose the combination in a polystyrene sleeve for weather protection. All this can be done in the workshop ahead of time. Except for the highest-band elements, which may be cut to the standard formula, the rest is cut and try. The elements added beyond the highest band elements must be determined experimentally because the inductance added by the nonresonant traps depends on the ratio of L to C; the higher the L is, the shorter will be the added elements.

CHAPTER 7

Instruments Every Amateur Should Have

On the basis of frequency of use and low-cost investment, the instruments described in this chapter should be a part of every hamshack equipment list. Most of these can either be purchased at low cost ready to operate; or they can be built from reliable kits, or from amateur manuals or magazine articles. Information is given in this chapter on how to build the SWR bridge, impedance bridge, and the coupling and phasing network of a monitoring scope. There are any number of magazine articles on how to build a crystal calibrator, but the one shown here is as inexpensive coming from a ready-to-assemble kit as building from scratch. Grid-dip oscillators are easily home-built, also, except for the need to calibrate them for each coil of the set. Kits include the calibration and may be a better bet for the amateur than building from scratch. The transistor checker described in Chapter 3 is a simple go-no-go type that will tell if a transistor is good or bad, but it will not indicate the gain in numbers. It is a good instrument to have around to check out a drawer full of transistors, or for selecting the best of a package of bargain transistors.

VOM

With the large number of high-sensitivity VOM's being imported, there is no excuse for an amateur not having a VOM in his shack. These days it is important to use a high-sensitivity VOM. The old standard 1000 ohms-per-volt VOM should be considered out of date. Transistors operate at a very low voltage, which means that a VOM will be operated on its lowest scale most of the time. A 1000 ohms-per-volt VOM will usually be used on its lowest scale. If this is the 5-volt scale, it will look like a 5000-ohm load at all times when making transistor measurements. This could be much too low a load, especially when testing for base voltage.

VOM's of 10,000 ohms/V and 20,000 ohms/V "look like" resistor loads 10 times and 20 times that of a 1000 ohms/V VOM. A high quality 20,000 ohms/V is a bit expensive, but it should be your first consideration; if you cannot afford one, a lower-cost imported 20,000 ohms/V VOM is better than a high-quality 1000 ohms/V VOM.

Another place where a high-sensitivity VOM is important is when one is used as a bridge indicator, as in the antenna impedance bridge described later. A grid-dip oscillator may be used as the signal source if a high-sensitivity VOM is used for a null indicator.

The VOM has the advantage of being portable. This is especially handy when it is necessary to go outdoors and make impedance or resonance checks on an antenna (Fig. 7-1).

VTVM

Much higher sensitivity is an advantage of the VTVM over the VOM. It usually has a constant input resistance of at least 11 megohms. More and more transistorized VTVM's are coming on the market, and these will provide the portability of a VOM (Fig. 7-2).

Fig. 7-1. This pocket-sized VOM measures only 2-3/4" x 4-1/4" x 1-5/16". It has 20,000 ohms-per-volt sensitivity.

Courtesy Triplett Electrical Instrument Co.

Courtesy RCA

Fig. 7-2. In every respect this instrument is like an a-c operated VTVM, but it is transistorized and battery-operated.

However, these are generally higher priced than the a-c operated VTVM, which has the high sensitivity so important to checking amateur equipment but depends on an a-c source for prime voltage (Fig. 7-3). If cost is a factor, consider building a VTVM from a kit from one of several manufacturers with established reputations.

Since amateur equipment operates at high frequencies, a worthwhile investment is an r-f probe for the VTVM. With it you can make actual measurements of voltage at high radio frequencies.

SWR METER

The SWR meter may be called by several names: SWR bridge, reflected-power meter, VSWR meter, etc. The instrument reads the ratio of forward power to reflected power going from the transmitter into the feed line, and indicates the mismatch of the antenna to the

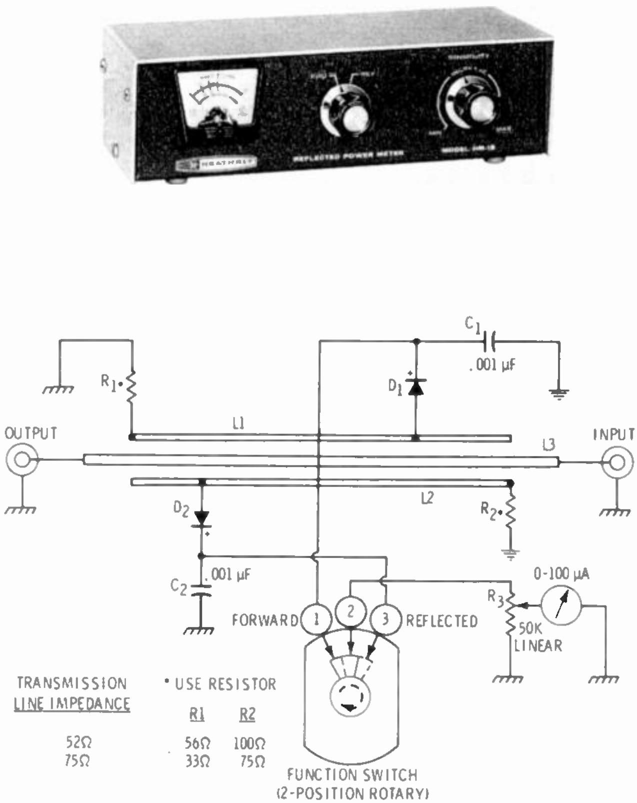
Courtesy Simpson Electric Co.

Fig. 7-3. This VTVM features an unusually large meter scale. The d-c input resistance is 16 megohms, as compared to the usual 11.

feedline. SWR means Standing Wave Ratio. When power is reflected back down the feed line due to a mismatch of the antenna to the feed line, standing waves are developed on the feed line. The greater the mismatch is, the greater are the standing waves. A perfect match leaves no standing waves, and is said to have a 1:1 standing-wave ratio.

A low SWR is important only in coaxial-type cable using a solid dielectric between the inner conductor and the outer shield. The dielectric introduces losses. Tables on cables show this as so many dB of attenuation per hundred feet of cable, and the higher the frequency is, the higher the attenuation is. The losses shown in the tables apply to the forward current only; that is, when there are no reflected waves, or the SWR is 1:1. A mismatch between the feed line and the antenna will result in a higher than 1:1 ratio, and the standing waves will introduce additional losses, depending on the ratio. The higher the ratio is, the higher the losses will be on the feedline. See Chapter 6 for more information on this and appropriate tables on cable attenuation.

The big advantage of an SWR bridge like those shown here is that it can be left in the line at all times. It can handle the full out-



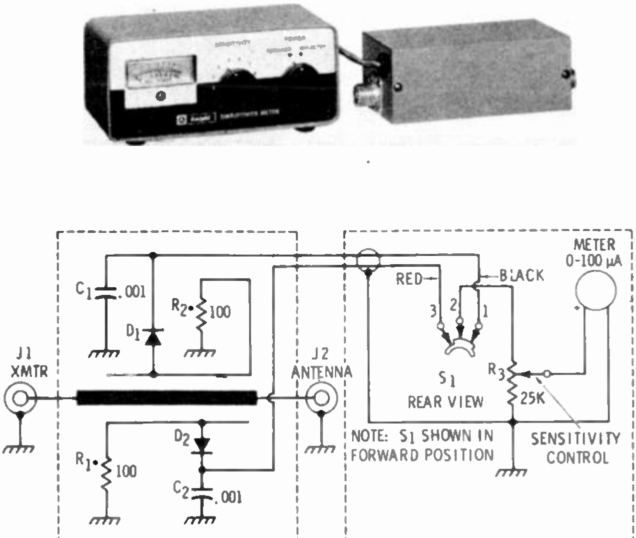
Courtesy Heath Co.

Fig. 7-4. A popular SWR meter in kit form is the *Heathkit* unit. Both the pickup section and indicator are in a single case.

put power of any transmitter and introduces practically no losses. In this way you have a constant check on output conditions. Figs. 7-4 and 7-5 show two SWR meters available in kit form.

BUILDING THE SWR METER

An SWR meter is easy to build. Construction is mostly mechanical. The one described here is in two parts, the part connecting in series with the transmission line is separate, so it may be connected without bringing the transmission line up to the operating table. The indicating meter and switch are a separate unit to be located at the operating position for easy visibility. A 5-inch × 2¼-



NOTES

1. CAPACITORS INDICATED IN MICROFARADS.
2. RESISTORS INDICATED IN OHMS.
3. K = 1000 OHMS
• VALUES SHOWN ARE FOR 72Ω LINE.
• WHEN USING A 52Ω LINE, VALUES OF R_1 AND R_2 ARE 160Ω

Courtesy Allied Radio Corp.

Fig. 7-5. The *Knight-Kit* SWR meter is in two parts. The pickup unit can be located near the transmitter output and the indicator placed anywhere on the operating table.

inch × 2¼-inch *Minibox* contains the transmission-line elements. Fig. 7-6 is a photograph of the front view of the indicator part and closed *Minibox*. The meter, switch, and sensitivity control can be mounted in a metal box also. Shown here is a piece of perforated *Masonite* supported vertically to a wood base. The indicator elements need not be shielded. Fig. 7-7 is the rear view of the indicator section and the open box of the bridge parts. Fig. 7-8 is the schematic diagram.

Mount the female coax chassis connectors one to each end of the U-shaped half of the *Minibox* (like the one shown in Fig. 7-5). Cut two ¾-inch squares of polystyrene sheet 1/16-inch thick. Drill ¼-inch holes in the centers of the polystyrene squares, and slip them

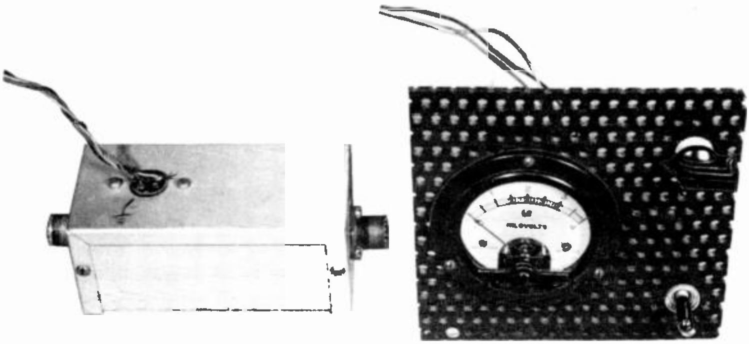


Fig. 7-6. Front view of the home-built SWR meter. The panel meter is mounted on a *Masonite* board, but it can be enclosed in a box if preferred.

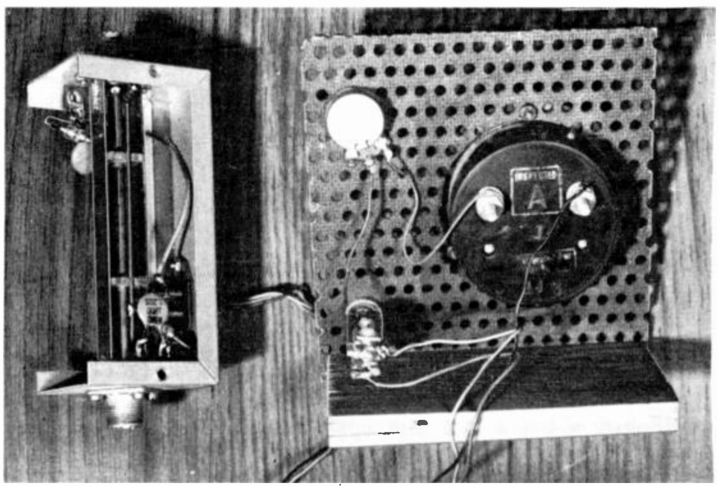


Fig. 7-7. Rear view of the SWR meter with cover removed from pickup section.

over the copper tubing. The polystyrene blocks will act as spacers for the metal strips and will support the pickup wires. Solder the ends of the copper tubing to the center terminals of the connectors. Place solder lugs under each opposite screw used for mounting the coax connectors, and solder the ends of 5-inch × ⅝-inch metal strips to the solder lugs so the strips face the copper tubing on opposite sides and are spaced ¼-inch from it. Any kind of metal will do except aluminum, as long as you are able to solder to it.

Cement 4-inch lengths of copper wire to the polystyrene blocks, one on each opposite side of the tubing, using *Duco* or similar

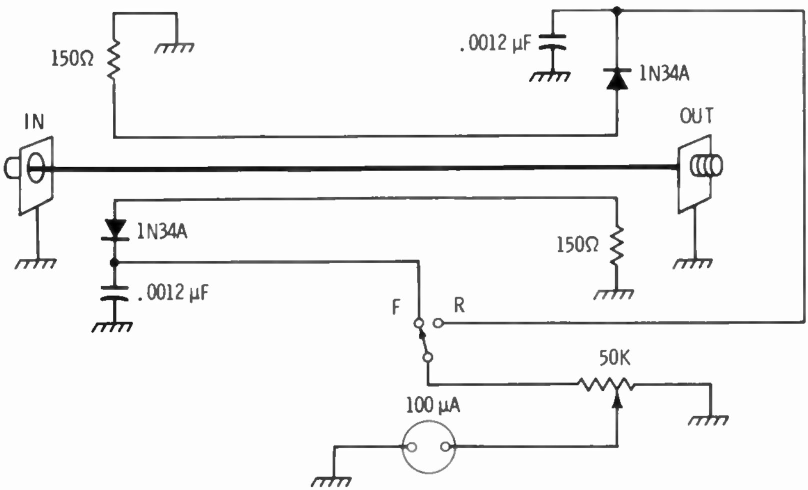


Fig. 7-8. The schematic diagram is fairly standard, and this one is similar to many others. A lower-sensitivity meter may be used, but it will take higher power at the lower frequencies to read full scale.

PARTS LIST			
1	100-μA meter movement	2	150-ohm, ½-watt resistors
1	spdt switch	2	.0012-μF ceramic capacitors
1	50,000-ohm potentiometer	2	4-inch pieces #12 copper wire
1	5″ × 2¼″ × 2¼″ *Minibox*	2	¾-inch square polystyrene blocks
2	5″ × ⅝″ metal strips (solderable)	1	4¹³⁄₁₆-inch piece ¼-inch copper tubing
2	female coax chassis connectors.	2	2-terminal wiring tie points
2	IN34A (or any r-f type) matched diodes		

cement. Wire the rest of the components between the copper pickup wires and ground lugs or terminal tie points, as shown in the schematic diagram. Be sure to keep leads as short as possible. This will require using heat sinks on the diode leads as well as on the resistor leads. Be sure the polarities of the diodes are the same. If the SWR meter is to be used with 52-ohm cable, use 150-ohm resistors for pickup-wire terminating resistors, or if it is for 75-ohm lines, use

100-ohm resistors. For greater accuracy use close-tolerance resistors and matched diodes.

Sensitivity depends on the length of pickup unit, sensitivity of the meter movement, and the frequency involved. The components and construction shown here will operate at 80 meters with power inputs of about 75 watts and up. For greater sensitivity use a 50-μA meter movement, or add a transistor amplifier ahead of the meter.

The meter panel shown in the photograph is part of a "junk-box" meter with markings from 1-4. SWR figures can be calculated from any meter markings on the basis of assuming a figure of 10 for the top of the scale, using this formula:

$$\mathrm{SWR} = \frac{10 + x}{10 - x}$$

where,

x is the reflected-power reading.

for example, if the reflected reading is 2,

$$\mathrm{SWR} = \frac{10 + 2}{10 - 2} = \frac{12}{8} = 1.25 \text{ or a ratio of } 1{:}1.25$$

In use the switch is set *Forward* and the sensitivity control is adjusted for full-scale indication of 10. The switch is then flipped to *Reflected* and the meter is read, and the reading applied to the above formula.

The forward position may be used to determine relative power output for transmitter tuning. Since power output varies as the square of output current or voltage, make up a table of relative power versus the meter figures. With a full scale of 10 the table will look like this:

Meter Reading	Relative Output		
1	×1		× .01
2	×4		
3	×9		
4	×16		
5	×25	or	× 2.5
6	×36		
7	×49		× 5
8	×64		
9	×91		
10	×100		×10

You can make your own overlay scale for the meter with two ranges, one for *Forward* relative power and the other for *Reflected*, but in terms of SWR. The SWR scale will be about like this:

Meter	SWR
0	1:1
1	1:1.2
2	1:1.5
3	1:1.8
4	1:2.5
5	1:3
6	1:4
7	1:5.5
8	1:9
9	1:20
10	1: Infinity

ACCURACY

Several factors affect the accuracy of a home-built instrument such as this. Stray component capacities and inductances of their leads will make some higher-frequency readings different than lower frequencies. Diodes are inherently nonlinear at low current. Low power through the SWR meter will have greater inaccuracy than high power. Harmonics generated by the transmitter will affect the readings. Twenty-meter harmonics from a 40-meter output will make a 40-meter antenna look highly mismatched. Reflected power from the 20-meter harmonics will show up as a higher-than-correct SWR on the meter.

USING THE SWR METER

The SWR meter (or the pickup part if it is a two-unit instrument) is connected in series with the transmission line between the transmitter and the transmission line. The design of the unit (determined by the pickup lead terminating resistors) must match the characteristic impedance of the transmission line. Turn on the transmitter and resonate it for the frequency you are interested in, matching to the resonant frequency of the antenna. Turn power on and adjust the sensitivity of the SWR meter for full-scale reading with the switch in the *Forward* position. Without touching anything else, flip the the switch to *Reflected* and read the SWR, using the foregoing table if a scale has not been made.

Regardless of the accuracy of the SWR meter, a high SWR reading means there is a poor match between the feed point of the antenna and the transmission line and/or the antenna is not resonant at the operating frequency. The object is, of course, to trim the antenna to exact resonance, and to match the impedance of the center of the antenna to the characteristic impedance of the line. Each adjustment can be made using the SWR meter. Trim the antenna a little at a time until the SWR meter indicates the lowest ratio. The antenna is then resonant. Then adjust for a match by means of a T- or gamma-matching system (or a proper matching stub) for a ratio of 1:1 on the SWR meter.

IMPEDANCE BRIDGE

A true bridge has four ratio arms. When each pair of resistances are equal the bridge is in balance. Fig. 7-9 shows the basic circuit of the impedance bridge we are about to describe. Two 51-ohm fixed resistors comprise one pair of arms. It is not important that these resistors be exactly 51 ohms but they must be *exactly equal to each other.* The other pair of arms consists of a variable resistor and the unknown resistance. When the variable resistor is adjusted to be equal

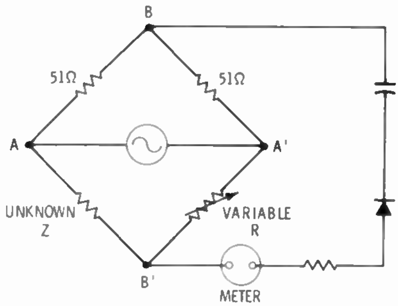


Fig. 7-9. The basic schematic of a bridge. If the variable arm and the unknown impedance are the same, the bridge is in balance, and the meter will read zero.

to the unknown resistance, the bridge is in balance, and the value of the unknown impedance is read on the calibrated variable-resistor dial scale. Note the connection of the source of voltage and the indicator in the circuit. When a source of a-c r-f voltage is applied to terminals

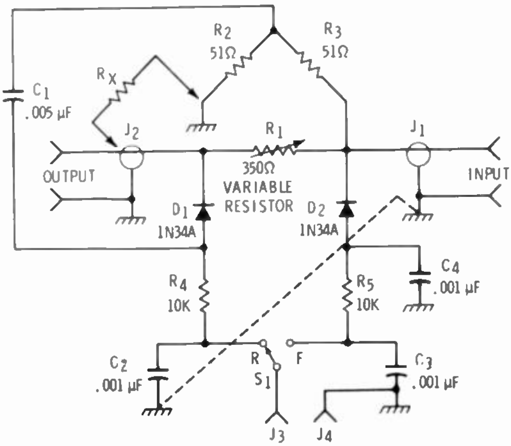


Fig. 7-10. The schematic of the homebuilt impedance bridge. The part of the circuit below and to the right of the dashed line is for reading r-f input—an added feature.

PARTS LIST

J_1, J_2	coax female chassis connectors	D_1, D_2	1N34A diodes
J_3, J_4	test-lead pin jacks	C_1	.005-µF ceramic capacitor
R_1	350-ohm, 2-W carbon potentiometer	C_2, C_3	.001-µF ceramic capacitors
R_2, R_3	51-ohm, 1-W resistors (matched to each other, or 1%)	S_1	spdt switch
R_4, R_5	10,000-ohm, ½-W resistors	2	pointer knobs
		1	4" × 2" × 1½" *Minibox*

A and A′, no voltage will develop between terminals B and B′, and the meter will read zero if all arms are in balance.

The schematic of the impedance bridge we are describing is shown in Fig. 7-10. It is similar to the basic bridge circuit except for the addition of a means of reading relative input voltage, and switching from input reading to bridge reading. The addition is below and to the right of the dashed line. This part may be omitted, and is in many such circuits. The input reading permits readings for SWR. The primary object of this instrument is to read input impedance to a transmission line or the center of an antenna, however. Fig. 7-11 is a front view, and Fig. 7-12 is a rear view with the cover off.

This impedance bridge will do a fairly accurate job of reading impedance in all amateur bands through the 10-meter band. Above that frequency stray capacitances and lead inductances affect accuracy. Commercial bridges are available with precision construction for reading impedance at much higher frequencies (Fig. 7-13).

None of the parts for the bridge are critical except the two 51-ohm resistors. (R_2, R_3) which should both be the same resistance, even if not precisely 51 ohms. All parts fit inside the 4" × 2" × 1½" *Minibox*. You could use a larger box and mount a 50-μA meter movement in it. This instrument is designed for use with an external VOM of

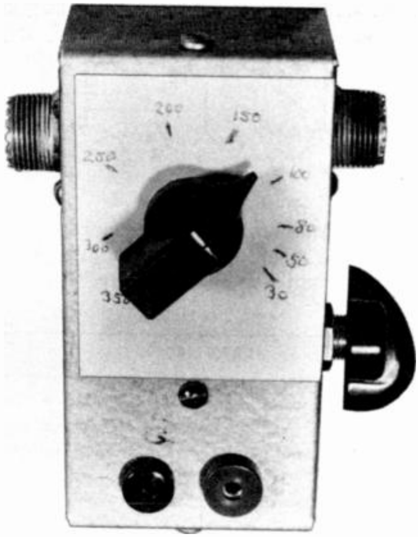


Fig. 7-11. This front view shows the homemade scale for the balance arm.

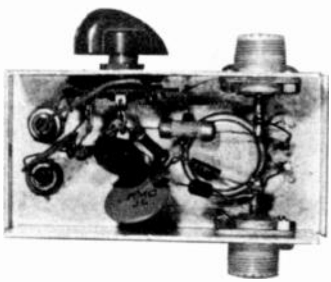


Fig. 7-12. Rear view of bridge showing parts. The case measures only 4" × 2" × 1½".

Courtesy James Millen Mft. Co. Inc.

Fig. 7-13. An impedance bridge for use up to 140 MHz. It has a variable capacitor for a bridge arm instead of a variable resistor. Note the loop on back for coupling to a GDO for signal source.

high sensitivity. This keeps the cost of the bridge down, although one with a built-in meter is handier to use.

CONSTRUCTION

Those parts that are part of the bridge circuit must have short leads and be spaced for a minimum of capacitive coupling between them. This involves the bridge resistors, the two diodes, and the .005-μF capacitor. The rest of the parts are decoupled from rf and may be placed anywhere. If the metal box is painted, be sure to scrape the paint away where grounds are made.

After the instrument is assembled and wired, cut out a white card and place it under the potentiometer mounting nut. Connect an ohmmeter to the terminals of the pot and mark off important resistance points on the card as measured on the ohmmeter. A more precise calibration is made with a signal generator connected as a voltage source and a variety of accurate resistors connected to the output coax connector. Set the generator for the highest frequency you expect to use, and balance the bridge for minimum reading on the VOM with each resistor, marking the value on the scale card.

USING THE IMPEDANCE BRIDGE

Unlike the SWR meter this impedance bridge cannot be left in the transmission line, nor used with high power. Any more than four watts of r-f fed into the bridge is likely to burn out the variable resistor arm of the bridge. Users find this is the one component most frequently replaced when the transmitter is used as the signal source and the output is accidentally allowed to exceed the power handling limit of the bridge. The transmitter may be used as a source of signal input provided its output is held down, by a voltage-divider resistor network, limited carrier injection in the case of a ssb transmitter, or by using the exciter only. A better source of rf is a signal generator or a vacuum-tube type grid-dip oscillator, although these require the use of a sensitive VOM or VTVM as a null indicator. Fig. 7-14 shows in block diagram form how a GDO is coupled to the bridge to measure the impedance reflected from the transmission line. Fit a coaxial connector with a 1- or 2-turn link having a diameter just right for going over the GDO coils, and tightly coupled.

Be sure to mark the side of the case as to which side is for the unknown (X transmission-line connection), and which is for the source

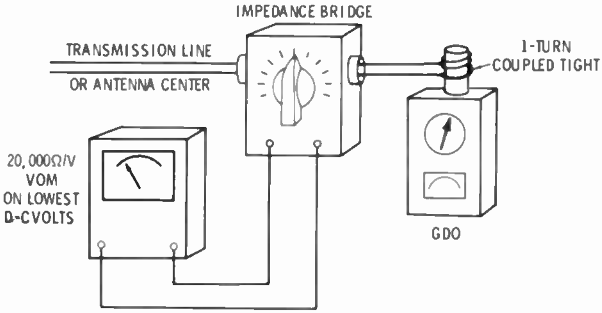


Fig. 7-14. The hookup of an antenna impedance bridge.

of r-f power. Use adhesive tape and write on it, or punch a tape from a label maker. Refer to the schematic diagram for "In" and "Out." "In" is the signal-source side.

Depending on the sensitivity of the VOM used as the balance indicator, the GDO control may have to be turned all the way up for enough signal power to give a good indication on the meter. When the variable resistor is set at a point where a null in the meter reading is indicated, the bridge is balanced and the unknown impedance can be read from the calibrated dial scale. This measurement is made with the switch in the "R" position.

The bridge may also be used to measure SWR. Furthermore, it may be used for unbalanced lines of any impedance. Set the balance arm for the characteristic impedance of the line, and place the switch on the side for reading input rf ("F" position). Adjust the coupling of the GDO and its control for full scale reading on the VOM. If full-scale reading cannot be obtained, use any voltage figure on the VOM scale for a reference, and call it 10. Switch to bridge reading on the side switch (position "R"), read the voltage on the VOM, and convert it to ratio of the reference 10. Refer to the graph of **Fig. 6-14.** of Chapter 6, and read SWR.

100-kHz CRYSTAL CALIBRATOR

Inexpensive to buy and easy to build, a 100-kHz crystal oscillator should be permanently connected to every amateur's receiver, if it is not already built in. The one shown in Fig. 7-15 may be attached to

Fig. 7-15. The Knight-Kit 100—kHz crystal calibrator fastens to the back of a receiver, and takes its power from the receiver.

Courtesy Allied Radio Corp.

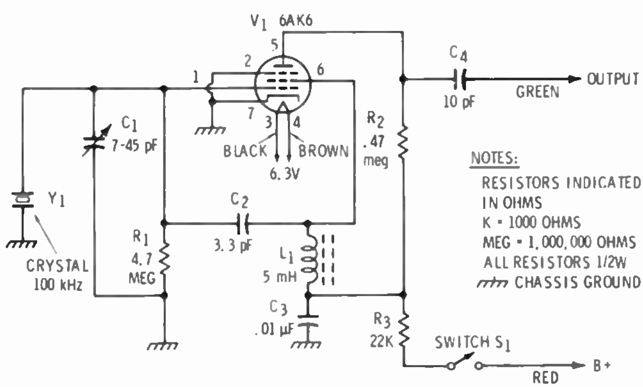


Fig. 7-16. The schematic diagram of Fig. 7-15. Add a 1N34A diode in series with the output for richer harmonics, reaching well over 30 MHz.

the back side of a receiver, or mounted inside if there is room. The calibrator receives its power from the receiver power supply. The schematic (Fig. 7-16) is typical of circuits using a vacuum tube.

Design is such that the oscillator is rich in harmonics, and so produces not only a signal at 100 kHz, but harmonics every 100 kHz up to above 30 MHz. The higher the harmonic number is, the weaker is the signal produced. In this particular circuit adding a 1N34A diode to the output between the 10-pF capacitor (marked C_1) and the output will increase the strength and limit of harmonics.

Almost any transistor is capable of oscillating at 100 kHz and may be used as an oscillator controlled by a 100-kHz crystal. One of the

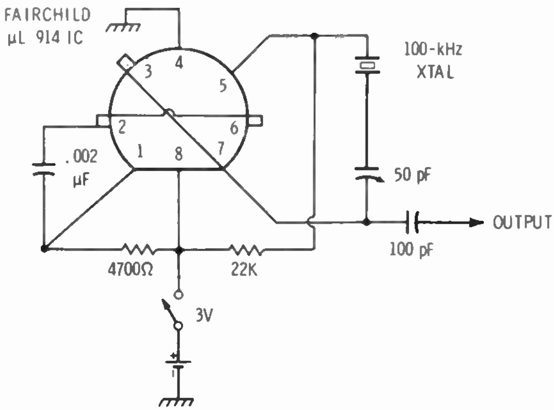


Fig. 7-17. An integrated circuit (IC) 100-kHz crystal calibrator. It operates from a 3-volt battery.

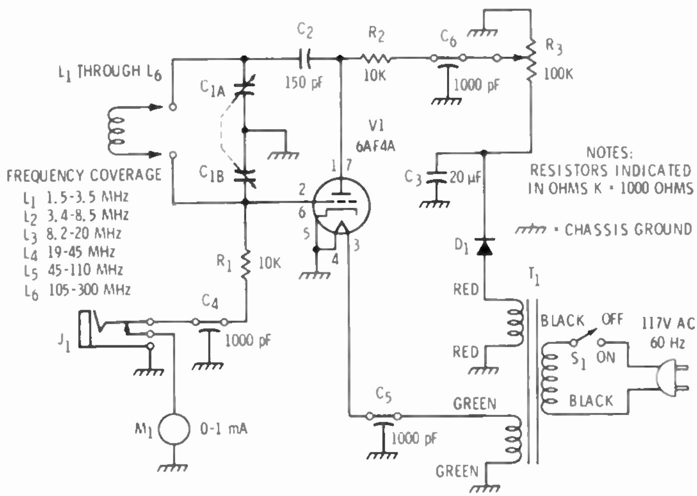
latest uses an inexpensive Fairchild µL914 IC (integrated circuit). The circuit used is shown in Fig. 7-17. The current drain is so small that the calibrator will just about run forever on the 3-volt battery.

The amateur magazines frequently show frequency-divider circuits that may be added to the output of the 100-kHz oscillator to supply markers at 10 kHz and even 1 kHz. These are flip-flop circuits whose constants are such that the 100-kHz oscillator will lock the 10-kHz oscilllating circuit into step to produce exact 10-kHz markers, etc.

The quartz crystals are cut in such a way that a small variable capacitor across or in series with the crystal will adjust the crystal frequency to exactly 100 kHz. By using an all-wave receiver and tuning to one of the WWV signals (carrier only) the crystal may be adjusted for zero beat. Thereafter exact 100-kHz markers can be read on the receiver.

GRID-DIP OSCILLATOR

The name itself describes the grid-dip oscillator operation—an oscillator in which the grid current dips when energy is absorbed from its resonant circuit. As with all vacuum-tube oscillator circuits, feedback from plate to grid sustains oscillation. The r-f voltage across the resistor between grid and ground produces a d-c current through the resistor as a result of the rectifying action between grid and cathode (Fig. 7-18). A meter in the grid circuit will read the direct



Courtesy Allied Radio Corp.

Fig. 7-18. The schematic of the *Knight-Kit* grid-dip oscillator is typical of many using a vacuum tube. The Colpitts oscillator circuit permits using two-terminal plug-in coils.

current. In a free-running oscillator the voltage at the grid is high, and, therefore, the current is high. Power absorbed from the resonant circuit reduces the grid voltage and, thus, the current. When the coil of the resonant circuit is coupled to an external resonant circuit, power is absorbed when the two circuits are mutually resonant. Thus a dip in grid current indicates mutual resonance, and the calibrated dial indicates the frequency of the external circuit.

To increase the utility of this valuable little instrument, the circuit is usually designed to permit other functions. By reducing the plate voltage to zero, oscillation stops and the grid and cathode act as a diode to indicate energy picked up from an external "live" resonant circuit, and so the GDO is used as an absorption wavemeter. By plugging headphones in series with the plate and operating the circuit as an oscillator, you can hear the beat note between its own oscillations and an external signal—thus it can be used as a beat-frequency detector. Of course, as an oscillator, its signal can be picked up in a receiver, and it becomes a signal generator.

Plug-in coils are used to cover a wide range of frequencies, up to as high as 300 MHz. The highest-frequency coil is usually a hairpin or half-turn coil.

SEMICONDUCTOR GDO's

Vacuum-tube GDO's are vigorous oscillators and quite sensitive to the effects of an external resonant circuit. Battery-powered, portable GDO's using semiconductors are handier to use, especially outdoors on an antenna. High-frequency transistors, especially the high-impedance FET type, make excellent oscillators, although they frequently require a following amplifier built in to increase their sensitivity and operate a meter.

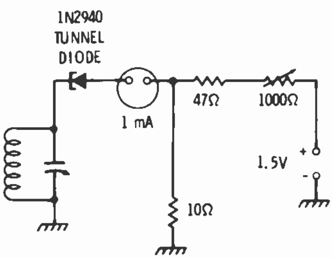
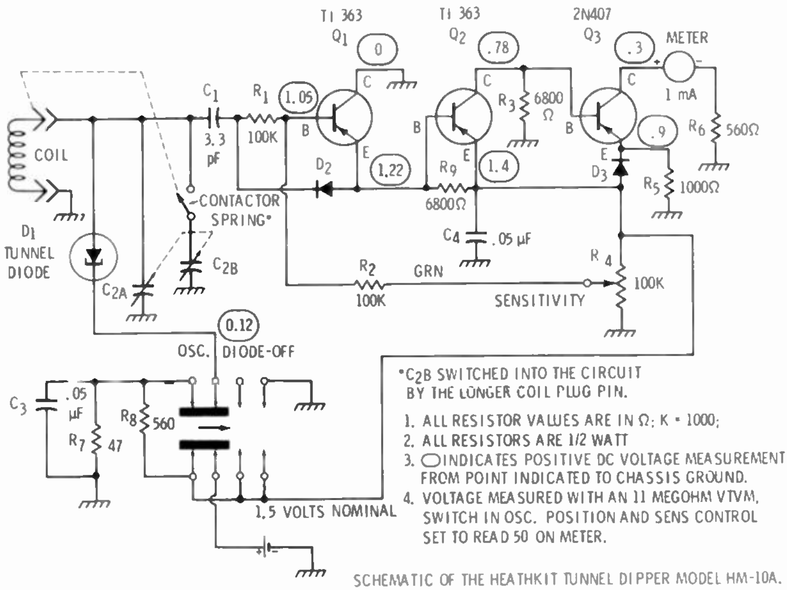


Fig. 7-19. The basic tunnel-diode GDO circuit. Sensitivity is low, and the meter should be preceded by an amplifier, as shown in Fig. 7-20.

The tunnel diode has negative resistance characteristics and makes an excellent oscillator good for very high frequencies. A fundamental circuit is shown in Fig. 7-19. While this circuit is usable as shown, meter deflection is quite small unless it is coupled tightly to an external resonant circuit. The circuit of Fig. 7-20 is a complete tunnel diode type GDO with a three-stage transistor amplifier to increase its sensitivity for more positive meter action. The photograph of Fig. 7-21 shows the completed unit, which is built from a commercial kit.

MONITORING SCOPE

Because the electron beam in a cathode-ray tube has no inertia, it reacts instantaneously to voltages impressed on its deflection plates.



Courtesy Heath Co.

Fig. 7-20. A three-stage transistor amplifier is used to increase sensitivity in the tunnel-diode GDO.

The best way to "see" what is happening in the output of a transmitter is by means of an oscilloscope, not only for alignment and adjustment, but as a continuous monitor while transmitting.

The method used is to apply the r-f voltage of the transmitter output to the vertical-deflection plates, while audio from the transmitter or the internal sweep of a standard scope is applied to the horizontal plates.

Only expensive laboratory type scopes will handle rf at amateur frequencies through their vertical amplifiers. However, since the output voltage of a transmitter is high, it can be connected directly to the vertical-deflection plates without going through the internal amplifier. Then, an inexpensive general-purpose scope may be used (Fig. 7-22). Some instruments have facilities on the back for making the direct connections to the vertical-deflection plates. Some instruments have only an access opening on the back of the case. Where there is no access, it will be necessary to cut an opening or install binding posts on the back.

The CRT mounting socket is always on the back, and generally very close to the back panel of the case. Access to the terminals on the socket is usually fairly easy. Do not disconnect any of the circuits to the plates, since dc must remain connected if the rest of the focusing and centering circuit is not to be disturbed. Capacitance coupling to the deflection plates through a pair of .01-μF ceramic capacitors is all that is necessary.

Courtesy Heath Co.

Fig. 7-21. A tunnel-diode GDO made from a kit. Plug-in coils cover the range from 3 to 260 MHz.

Fig. 7-23 is a photograph of the circuitry added to the back of a general-purpose scope. All components are mounted on a *Masonite* shelf. To the right is a tuned circuit using plug-in coils. The tuned circuit is link-coupled to a pickup loop which is coupled to the transmitter output tank, or is capacitance-coupled to the inner conductor of a coaxial transmission line. On the left is an audio phasing circuit to correct the phase of the audio output from the transmitter before applying it to the horizontal input of the scope. The circuit is shown in Fig. 7-24.

A basic scope without amplifiers and a sweep circuit is easy to build, and some are shown in amateur handbooks. These are good only for monitoring an a-m transmitter because of the high voltages needed for the deflection plates. For example, a 3AP1-A CRT commonly used in basic scopes needs peak-to-peak signals of from 50 to 90 volts per inch of deflection. Such voltages are easily available from the modulator and from the r-f output of a-m transmitters; but ssb transmitters will have only very low audio voltages available,

Courtesy Heath Co.

Fig. 7-22. Any inexpensive oscilloscope will serve as a modulation monitor. This one is available in kit form.

which will require further amplification for deflecting the horizontal plates of the CRT. Therefore, a general purpose scope with a horizontal amplifier is needed, and these are better purchased (even if in kit form) than built from scratch. A sweep circuit is needed for voice monitoring an ssb transmitter.

COUPLING RF

Very often rf may be obtained from the coaxial transmission line and coupled directly to the vertical-deflection plates of the CRT. A 200-watt transmitter will develop 100 volts of rf between the center terminal of a 50-ohm line and ground. The peak-to-peak value is about 280 volts, enough to develop a three-inch high pattern on the face of the CRT. Since the voltage to the deflection plates will vary with power and frequency, it is best to provide some means of adjust-

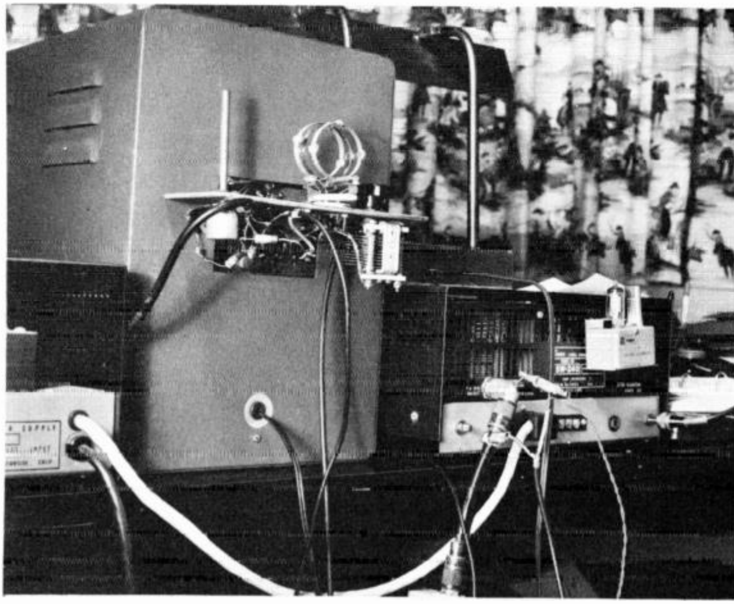


Fig. 7-23. A home-built "add-on" for applying transmitter rf to the vertical plates of a CRT in an oscilloscope. R-f pickoff is from the center conductor of a cut coaxial connector, shown at the lower right of the photo.

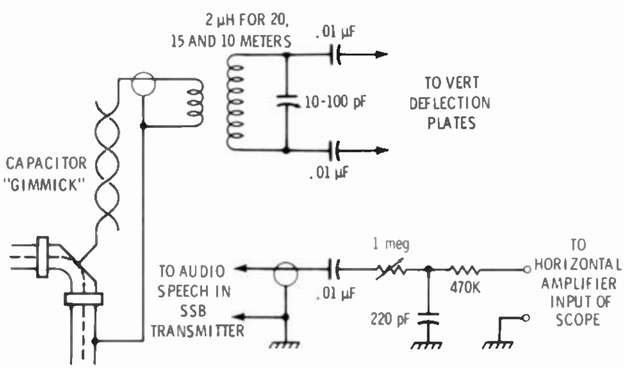


Fig. 7-24. Diagram of the monitoring hookup. A plug-in coil, trimmed to 2 μH covers 20, 15, and 10 meters with a 100-pF variable capacitor. The audio phasing network is shown below.

ment, and of increasing the voltage when necessary. The answer is in a tuned circuit as shown in Fig. 7-24. The coil and capacitor values shown will resonate in the 20-, 15-, and 10-meter bands by adjusting the variable capacitor. The coil used was a cut-down plug-in, coil which was quite popular back in the days of plug-in coils for the bandswitching of transmitters. It includes an adjustable link, although it is not necessary. A coil and capacitor of your own design may be used, with currently available air-wound coils. A 2-inch diameter, 2½-inch long coil with 25 turns will resonate in the 80- and 40-meter bands with a 100-pF variable capacitor. A tap at 5½ turns will have the right inductance to tune the 20-, 15-, and 10-meter bands. Couple a 2-turn link of insulated wire to the bottom of the coil, and use a piece of coaxial cable to the r-f source. When tuned to resonance the voltage available for the deflection plates is quite high. If the pattern goes off screen, merely detune the circuit until the pattern is the desired size.

The lower right portion of the picture of Fig. 7-23 shows a good method of tapping off rf from the transmission line. A slice is cut across the elbow of an elbow coaxial adapter. Make the slice to expose the inner conductor. Solder a heavy wire stub to the inner conductor. Clip or solder a piece of insulated wire about 6 inches long to the stub. Wrap another piece of insulated wire around the first insulated wire, and connect this wire to the center conductor of the shielded cable to the tuned circuit. The twisted wires have capacitance between them. This is known as a "gimmick." The shield of the cable should be grounded to the transmitter ground. With the transmitter tuned to the lowest frequency band you use, resonate the tuned circuit on the back of the scope. If the pattern goes off the screen, clip small amounts off the end of the "gimmick" until the pattern size is right for you. This establishes the maximum pattern size for the lowest frequency. At higher frequencies, detune the capacitor of the resonant circuit for size.

Coupling may be made to the final tank coil of your transmitter by a link of one or two turns of well-insulated wire. Extreme caution must be observed because of the high voltage around the final tank.

At the vhf amateur-band frequencies, the coupling methods mentioned above may introduce serious reactances which will affect output tuning and power output. Connecting capacitors and links may have enough reactance to become part of the final tuned circuit. Links should be small and loosely coupled. Capacitors (the "gimmick" mentioned above) must be of low value.

AUDIO PHASE

The rf deflects the CRT in a vertical direction. To complete the pattern the horizontal trace is developed by the internal sweep of the scope or by straight audio from the transmitter. The audio may be at voice frequencies using the microphone, or a single or double audio frequency from an audio signal generator. When using audio for the horizontal sweep a phase shift often occurs between the modulated rf and the audio itself. The left-hand portion of the scope circuit on the back platform of Fig. 7-23 contains a phase-shifting network to correct the out-of-phase signal. The circuit is shown in the lower portion of Fig. 7-24. To reduce phase shift as much as possible any capacitors in the circuit coupling to the scope must have low reactance compared to any resistance in the circuit. A .05-μF capacitor has a reactance of 10,000 ohms at 300 Hz, the usual lower audio limit of modulators. It is a low enough value if any resistance which follows it is at least ¼ megohm in value.

Fig. 7-25 shows the connection from an a-m transmitter for audio. The .05-μF capacitor must have a high-voltage rating, twice as high as the d-c voltage used in the final of the transmitter. The .05-μF capacitor and .25 megohm resistors should be inside the transmitter. The tap is brought to a .25-megohm pot mounted near the horizontal deflection terminals of the CRT scope. Horizontal pattern width is adjusted with the pot. A trapezoid pattern will be seen on the CRT with voice modulation. Should the pattern not be a perfect triangle,

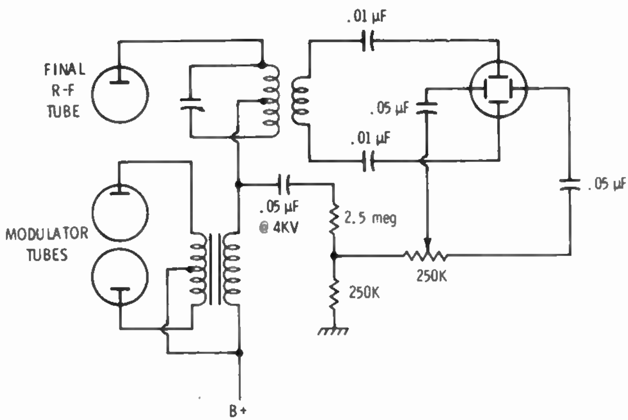


Fig. 7-25. Circuit for picking off audio from a plate modulator, using a resistor voltage divider to reduce voltage.

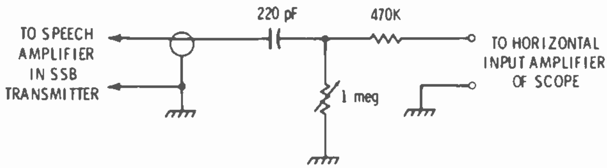


Fig. 7-26. It may be necessary to reverse the audio phase shift by using this circuit for audio instead of the one shown in Fig. 7-24.

but appear to be folded over, there is phase shift. In that case the connection at the scope must be made through a phase-shift network as shown in Fig. 7-24.

Fig. 7-24 shows the connection from an ssb or frequency- or phase-modulated transmitter. The connection in the transmitter is made at the last point of audio. If there is dc at that point (as from the plate of an audio amplifier tube), a blocking capacitor must also be installed.

The phase-shift network described corrects usual phase shifting resulting from lack of wide frequency response in the modulator. However, sometimes the shift is in the other direction. This calls for the use of the phase-shift network shown in Fig. 7-26. If one circuit does not work, change to the other.

Phase-shift networks are frequency discriminating. Adjustment of the variable resistor will correct for a single audio frequency from an audio generator. It sometimes does not correct for the random frequencies developed by the human voice. Perfect correction might involve replacing some coupling capacitors in the transmitter speech-amplifier and modulator circuits. If satisfactory correction is not easily obtained, it is simpler to use the sweep circuit of the scope for your observations.

A commercial monitoring scope with built-in tuned circuits is shown in Fig. 7-27.

OBSERVING PATTERNS

Fig. 7-28 shows a series of patterns obtained from the output of an a-m transmitter. When the carrier is applied to the vertical-deflection plates and no signal or sweep voltage is applied to the horizontal plates, a bright vertical line only is seen as at A. Turn on the internal sweep generator of the scope, using any frequency of sweep, and the pattern will become a solid block as at B. The width of the pattern is arbitrary, and is adjusted to suit the observer. The height is related to the amount of r-f output from the transmitter. Pattern C is

Courtesy James Millen Mfg. Co. Inc.

Fig. 7-27. A commercial monitoring scope with tuned circuits for vertical deflection built in.

what you see when you key the transmitter with a series of dots from an automatic or semiautomatic key. The sweep frequency is set at a low rate in order to synchronize with the speed of the dots to make the pattern stand still. Observe the squareness of the make and break points in the r-f squares. If the pattern is too square it means that there is no keying filter, and spurious frequencies could be created when the circuit is made and broken with the key. There should be a slight taper off the trailing end. When a single sine-wave audio frequency is fed into the microphone jack from an audio generator the pattern will look like D. This is 100-percent modulated. If you drew lines around the edges of the pattern they should make two overlapping sine waves. If the edges are a good sine wave, the modulation quality is excellent. The sweep of pattern D is the internal sweep of the scope and, again, synchronized to the frequency injected. If audio (voice or sine wave) is used to sweep the horizontal trace, the pattern is a trapezoid like pattern E. At 100-percent modulation you will see a perfect triangle with good sharp corners. Overmodulation will have a tip sticking out the left corner, and the top and bottom corners may begin to round out. Less than 100-percent modulation is shown in pattern F. You can measure the height of the pattern vertical sides and find percent of modulation with the following formula:

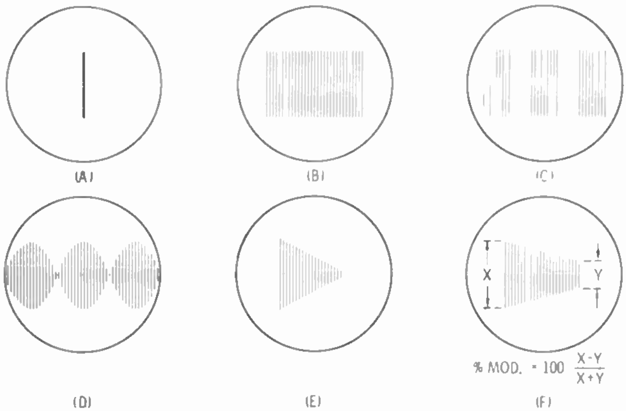


Fig. 7-28. Oscilloscope patterns obtained from a plate -modulated or c-w transmitter. See text for explanations.

$$\% \text{ Mod} = \frac{100\, x - y}{x + y}$$

where,

x is the maximum amplitude,
y is the minimum amplitude.

SSB PATTERNS

With the ssb transmitter but with no modulation, the pattern on the scope trace will look like pattern A of Fig. 7-29—just a dot. A suppressed-carrier ssb transmitter with no modulation has no output, thus no pattern will show on the scope. Turn on the horizontal sweep for a 30-Hz sweep and the pattern will be just a horizontal line from the sweep. Modulate with voice through the microphone and it will look like pattern B. Remove the microphone and connect an audio generator. As you increase the audio input you will see a pattern like C; this is straight cw. The single audio tone has brought the carrier up to a steady state. As you increase the audio a point will be reached where the pattern height does not increase. This is maximum modulation and is the point at which "flattopping" occurs with voice modulation. You have reached plate-current saturation in the final. Mark the top and bottom on the CRT face with a grease pencil for future reference. These will be the limits for monitoring your modulation with voice. (Don't leave this signal on for more than 30 seconds at a time. At this point, with a steady tone, you are exceeding the plate dissipation rating of the final tube or tubes, and damage could occur.)

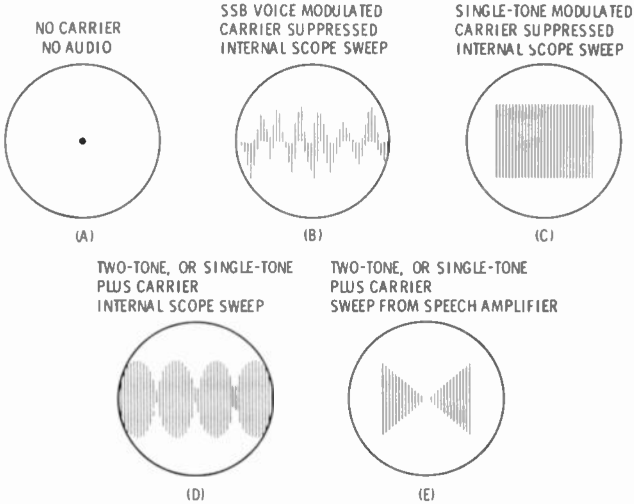


Fig. 7-29. Patterns observed on an ssb transmitter. Each pattern is described in the text.

Parallel two audio generators with signals about 1 kHz apart and of equal level, and the pattern will look like D. This shows 100-percent modulation for an ssb transmitter if the outline of the pattern forms two sine waves overlapping and the top and bottom touches the maximum-modulation reference marks. Lacking two audio generators you can get the same pattern with one tone and carrier insertion.

With the audio off, insert carrier for a bar like pattern C but with half the height marked on the CRT with your grease pencil. Add audio to the height marked on the face of the tube. This is the equivalent to the "two-tone" signal recommended for studying linearity of the final class-B amplifier. With this setup (or the two-tone signal) switch the horizontal amplifier to direct audio from the transmitter. This will show a bow-tie pattern as shown in E. The bow-tie is actually a superior pattern for observing linearity. If the lines cross with a perfect X and the top and bottom corners are sharp, your final has good linearity and is properly loaded.

For continuous monitoring of an ssb transmitter with voice, use the internal sweep of the scope. If your voice peaks just touch the limits you have marked on the scope face, you are modulating within the limits necessary to prevent flattopping. Your average plate current will probably be about ⅓ that of full c-w plate current (as for pattern C).

Index

A

A-c measurements, 23-56
 audio transformers, 35, 36
 capacitors, 36-43
 capacitors and inductors, 32 ,33
 crystal frequency, 49
 filters, 49-56
 frequency, 26, 27, 48, 49
 inductances, 43-45
 oscilloscopes, 25, 26
 power transformers, 33-35
 Q, 45, 46
 r-f and i-f transformers, 46, 47
A-c power out, transmitter, 118
Alignment
 receiver, 104-116
 transmitter, 137, 138
A-m and c-w receivers, 107, 108
Antennas and feeders, 148-177
 attenuation, 157
 front-to-back ratio, 169, 170
 half-wave feeder, 151-155
 impedance, 155, 156
 matching transmission line to antenna, 171-177
 resonance, 148, 149
 sweep generator for resonance, 166-169
 SWR, 156-159
 SWR meter, 165, 166
 tuned lines, 159, 160
 tuning, 149-151
  vertical, 159-163
 velocity factor, 153-155
 vertical-antenna ground, 163-165
Attenuation, 157
Audio generator, 31, 32
Audio phase, 202, 203
Audio transformers, 35, 36

B

Batteries, 20-22
Bfo adjustment, 110, 111

C

Calibration
 receiver, 111-114
 vfo, 138-140
Capacitors and inductors, 32, 33
Carrier adjustment, ssb, 140, 141
Characteristic curve
 collector, 67-69
 diode, 60
 transistor, 68
Class-B linearity, 145-147
Class-C operation, 145
Crystal
 calibrator, 193, 194
 frequency, 49
C-w keying, 135, 136

D

D-c measurements, 7-22
 audio stages, 13-16
D-c measurements contd.
 buffer stages, 16
 oscillators, 16
 power supplies, 22
 r-f stages, 16, 17
 transistor circuits, 18-20
 vacuum-tube amplifiers, 13-17
Definition of power, 117-118
Diode measurements, 63-65
Distributed capacitance, 44, 45
Dummy load, 123-126

E

Efficiency, power, 121
Electrolytic capacitors, 40

F

FET measurements, 69-71
Field-effect transistors, 69-71
Filters, 49-56
 audio, 53, 54
 constant-*k*, 50-52
 formulas, 51
 m-derived, 50-51
 measurements, 54-56
 pi-section, 50
 TVI, 51, 52
Frequency measurements, 48, 49
Front-end alignment, receiver, 114
Front-to-back ratio, 169, 170

G

Gamma match, 171, 172
Grid-dip oscillator, 28, 29, 195, 196
 semiconductor, 196
Ground, vertical-antenna, 163-165

H

Half-wave feeder, 151-155
Handling FET's, 71
Headphone and speaker impedance, 45
Heterodyne frequency meter, 27
High-pass filter, 55

I

I-f transformers, 46, 47
Image response, 99, 100
Impedance, 155, 156
Impedance bridge, 188-193
 construction, 191
 using, 191-193
In-circuit testing, 59
Inductances, 43-45
Input impedance, receiver, 103, 104
Instruments, 10, 178-206
 accuracy, 10
 crystal calibrator, 193-194
 grid-dip oscillator, 195-196
 impedance bridge, 188-193
 monitoring scope, 196-206
 SWR meter, 180-188
 VOM, 178, 179
 VTVM, 179, 180

L

Loading, transmitter, 126, 127
Loading effect, 11, 12

M

Matching stubs, 174-177
Matching transmission line to antenna, 171-177
  d-c, 7-22
  diode, 63-65
  field-effect transistor, 69-71
  modulation, 127-135
  receiver, 78-116
  transistor, 65-89
  Zener and signal diodes, 64, 65
Meter rectifiers, 23, 24
Modulation measurements, 127-135
Modulation percent, ssb, 136, 137
Monitoring scope, 196-206

N

Neutralizing, 141-143
Noise figure, 100-103

O

Ohm's law, 29
Oscilloscope, reading, 96
Out-of-circuit testing, diodes and transistors, 59-65

P

Parasitics, 143-145
Power efficiency, 121
Power out, measuring, 121-123
Power transformers, 33-35

Q

Q, 45, 46
Q bar, 172-174

R

Receiver measurements
  alignment, 104-116
  a-m and c-w receivers, 107, 108
  bfo adjustment, 110, 111
  calibration, 111-114
  front-end alignment, 114
  image response, 99, 100
  input impedance, 103, 104
  instruments, 79, 84
  noise figure, 100-103
  selectivity, 90-99
  sensitivity, 84-88
  ssb receivers, 108-110
  stage gain, 89-90
  using noise generators, 114, 115
Resonance, 148, 149
R-f probe, 56
R-f transformers, 46, 47

S

Selectivity, 90-99
Semiconductor testing, 59
Sensitivity, 84-88
Signal generator, 30-31
Solid state VTVM, 9
Speaker impedance, 45
Ssb
  carrier adjustment, 140-141
  modulation percent, 136, 137
  patterns, 205, 206
  receivers, 108-110
Stage gain, 89, 90
Sweep generator
  for resonance, 166, 169
  for selectivity, 95
SWR, 156-159
SWR meter, 165, 166, 180-188
  accuracy, 187
  building, 183-186
  using, 187, 188

T

Transconductance tube testers, 58, 59
Transistor
  bias, 18
  collector current, 18, 19
  manuals, 77
  measurements, 65-69
  oscillators, 19, 20
  tester, 71-77
  thermal runaway, 18
Transmitter measurements
  a-c power out, 118
  alignment, 137, 138
  class-B linearity, 145-147
  class-C operation, 145
  connecting an oscilloscope, 118, 119
  c-w keying, 135, 136
  definitions of power, 117, 118
  dummy load, 123-126
  loading, 126, 127
  measuring power out, 121-123
  modulation measurements, 127-135
  neutralizing, 141-143
  parasitics, 143-145
  power efficiency, 121
  ssb carrier adjustment, 140, 141
  ssb modulation percent, 136, 137
  vfo calibration, 138-140
Traps, antenna, 177
Tube and transistor testing
  in-circuit, 59
  out-of-circuit, 59-63
  transconductance testers, 58, 59
Tuned lines, 159, 160
Tuning, antenna and feeders, 149-151
Tuning a vertical, 159-163

V

Velocity factor, 153-155
Vertical antenna
  ground, 163-165
  tuning, 159-163
Vfo calibration, 138-140
VOM, 7-9, 178, 179
  practical d-c measurements, 12-23
  sensitivity, 7, 8
VTVM, 8, 9, 179, 180
  sensitivity, 8, 9
  solid-state, 9

Z

Zener diode measurements, 64, 65

AMATEUR TESTS and MEASUREMENTS

by
Louis M.
Dezettel,
W5REZ

The amateur radio operator, by the very nature of his hobby, should be familiar with a few basic test instruments. The amateur **needs** to know whether or not he is within the law as to power output and transmitting frequency. It is also extremely helpful to be able to troubleshoot and repair his own equipment when "Murphy's Law" strikes.

Amateur Tests and Measurements has been written to encourage the amateur to use his test equipment to full advantage. Each chapter takes a related series of tests or measurements and develops methods and techniques available to the average ham for testing and measuring the performance of his rig. Test setups, procedures, and results obtainable are described and illustrated in detatil.

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Every amateur should have **Amateur Tests and Measurements** as a basic part of his reference library to answer questions regarding the uses of test equipment, and to supply the best procedures for achieving the desired results from his transmitter, receiver, and antenna system.

ABOUT THE AUTHOR

Lou Dezettel is well known to readers of electronics literature. The author of several books, he also has over 60 magazine articles to his credit. He worked with a large electronic parts distributor for nearly 25 years, has been a member of IEEE for over 25 years, and an active licensed amateur radio operator (W5REZ, ex W9SFW) for over 30 years. He now devotes his full time to free-lance technical writing in electronics.

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