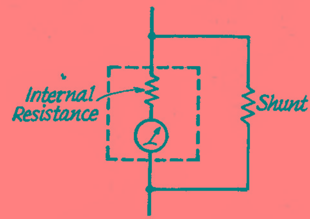


Fig. 17-1 – Use of a shunt to extend the calibration range of a currant-reading instrument.

external resistance in parallel with the internal resistance, as in Fig. 17-1, the current will divide between the two, with the meter responding only to that part of the current which flows through the internal resistance of its movement. Thus it reads only part of the total current; the effect is to make more total current necessary for a full-scale meter reading. The added resistance is called a shunt.

It is necessary to know the meter's internal resistance before the required value for a shunt can be calculated. It may vary from a few ohms to a few hundred, with the higher resistance values associated with higher sensitivity. When known, it can be used in the formula below to determine the required shunt for a given current multiplication:

$$R = \frac{Rm}{n - 1}$$

where R is the shunt, R_m is the internal resistance of the meter, and n is the factor by which the original meter scale is to be multiplied.

Making Shunts

Homemade shunts can be constructed from any of various special kinds of resistance wire, or from ordinary copper wire if no resistance wire is available. The Copper Wire Table in this *Handbook* gives the resistance per 1000 feet for various sizes of copper wire. After computing the resistance required, determine the smallest wire size that will carry the full-scale current (250 circular mils per ampere is a satisfactory figure for this purpose). Measure off enough wire to provide the required resistance.

THE VOLTMETER

If a large resistance is connected in *series* with a current-reading meter, as in Fig. 17-2, the current

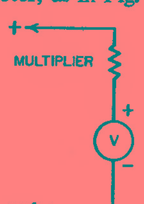


Fig. 17-2 – A voltmeter is a current-indicating instrument in series with a high resistenca, the "multiplier."

multiplied by the resistance will be the voltage drop across the resistance, which is known as a **multiplier**. An instrument used in this way is calibrated in terms of the voltage drop across the multiplier resistor, and is called a voltmeter.

Sensitivity

Voltmeter sensitivity is usually expressed in ohms per volt, meaning that the meter's *full-scale* reading multiplied by the sensitivity will give the total resistance of the voltmeter. For example, the resistance of a 1000-ohms-per-volt voltmeter is 1000 times the full-scale calibration voltage, and by Ohm's Law the current required for full-scale deflection is 1 milliampere. A sensitivity of 20,000 ohms per volt, a commonly used value, means that the instrument is a 50-microampere meter.

The higher the resistance of the voltmeter the more accurate the measurements in high-resistance circuits. This is because in such a circuit the current flowing through the voltmeter will cause a change in the voltage between the points across which the meter is connected, compared with the voltage with the meter absent, as shown in Fig. 17-3.

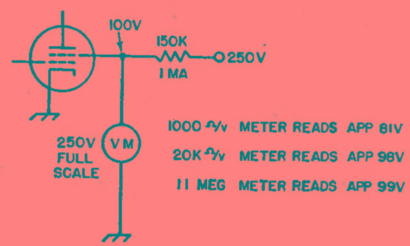


Fig. 17-3 – Effect of voltmeter resistence on accuracy of raadings. It is assumed thet the dc resistanca of the screen circuit is constant at 100 kilohms. The actual current and voltage without the voltmeter connected ara 1 mA and 100 volts. The voltmeter raadings will differ because the differant types of meters draw different amounts of current through the 150-kilohm resistor.

Multipliers

The required multiplier resistance is found by dividing the desired full-scale voltage by the current, in amperes, required for full-scale deflection of the meter alone. Strictly, the internal resistance of the meter should be subtracted from the value so found, but this is seldom necessary (except perhaps for very low ranges) because the meter resistance will be negligibly small compared with the multiplier resistance. An exception is when the instrument is already a voltmeter and is provided with an internal multiplier, in which case the multiplier resistance required to extend the range is

$$R = R_m(n - 1)$$

where R is the multiplier resistance, R_m is the total resistance of the instrument itself, and n is the factor by which the scale is to be multiplied. For example, if a 1000-ohms-per-volt voltmeter having a calibrated range of 0-10 volts is to be extended to 1000 volts, R_m is 1000 X 10 = 10,000 ohms, n is 1000/10 = 100, and R = 10,000 (100 − 1) = 990,000 ohms.

When extending the range of a voltmeter or converting a low-range meter into a voltmeter, the rated accuracy of the instrument is retained only when the multiplier resistance is precise. Precision wire-wound resistors are used in the multipliers of high-quality instruments. These are relatively expensive, but the home constructor can do quite well with 1-percent-tolerance composition resistors. They should be "derated" when used for this purpose – that is, the actual power dissipated in the resistor should not be more than 1/4 to 1/2 the rated dissipation – and care should be used to avoid overheating the body of the resistor when soldering to the leads. These precautions will help prevent permanent change in the resistance of the unit.

Ordinary composition resistors are generally furnished in 10- or 5-percent tolerance ratings. If possible errors of this order can be accepted, resistors of this type may be used as multipliers. They should be operated below the rated power dissipation figure, in the interests of long-time stability.

DC MEASUREMENT CIRCUITS

Current Measurement with a Voltmeter

A current-measuring instrument should have very low resistance compared with the resistance of the circuit being measured; otherwise, inserting the instrument will cause the current to differ from its value with the instrument out of the circuit. (This may not matter if the instrument is left permanently in the circuit.) However, the resistance of many circuits in radio equipment is quite high and the circuit operation is affected little, if at all, by adding as much as a few hundred ohms in series. In such cases the voltmeter method of measuring current, shown in Fig. 17-4, is frequently convenient. A voltmeter (or low-range milliammeter provided with a multiplier and operating as a voltmeter) having a full-scale voltage range of a few volts is used to measure the voltage drop across a suitable value of resistance acting as a shunt.

The value of shunt resistance must be calculated from the known or estimated maximum current expected in the circuit (allowing a safe margin) and the voltage required for full-scale deflection of the meter with its multiplier.

Power

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

The setup for measuring power is shown in Fig. 17-5, where R is any dc "load," not necessarily an actual resistor.

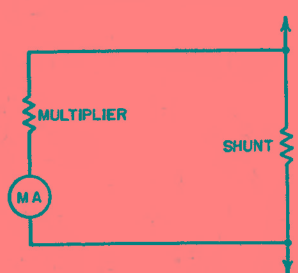


Fig. 17-4 – Voltmeter method of measuring current. This method permits using relatively large values of resistance in the shunt, standard values of fixed resistors frequently being usable. If the multiplier resistance is 20 (or more) times the shunt resistance, the error in assuming that all the current flows through the shunt will not be of consequence in most practical applications.

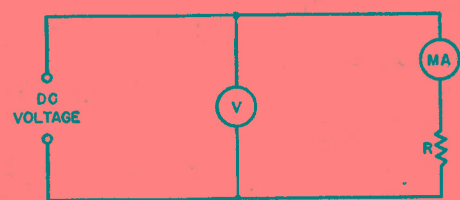


Fig. 17-5 – Measurement of power requires both current and voltage measurements; once these values are known the power is equal to the product – P = EI. The same circuit can be used for measurement of an unknown resistance.

Resistance

Obviously, if both voltage and current are measured in a circuit such as that in Fig. 17-5 the value of resistance R (in case it is unknown) can be calculated from Ohm's Law. For accurate results, the internal resistance of the ammeter or milliammeter, MA, should be very low compared with the resistance, R, being measured, since the voltage read by the voltmeter, V, is the voltage across MA and R in series. The instruments and the dc voltage should be chosen so that the readings are in the upper half of the scale, if possible, since the percentage error is less in this region.

THE OHMMETER

Although Fig. 17-5 suffices for occasional resistance measurements, it is inconvenient when frequent measurements over a wide range of resistance are to be made. The device generally used for this purpose is the **ohmmeter**. This consists fundamentally of a voltmeter (or milliammeter, depending on the circuit used) and a small dry battery, the meter being calibrated so the value of an unknown resistance can be read

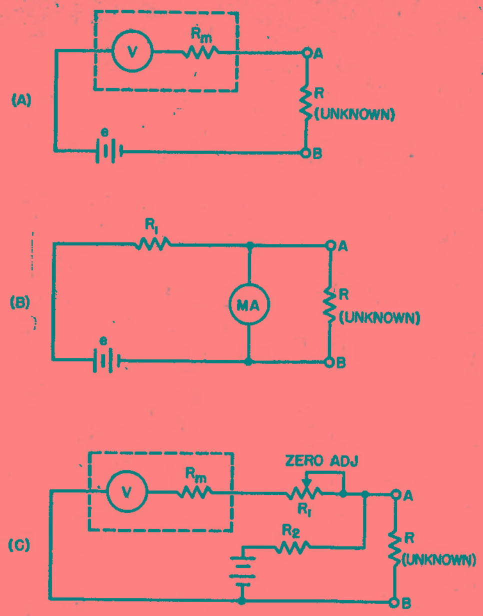


Fig. 17-6 — Ohmmeter circuits. Values are discussed in the text.

directly from the scale. Typical ohmmeter circuits are shown in Fig. 17-6. In the simplest type, shown in Fig. 17-6A, the meter and battery are connected in series with the unknown resistance. If a given deflection is obtained with terminals *A-B* shorted, inserting the resistance to be measured will cause the meter reading to decrease. When the resistance of the voltmeter is known, the following formula can be applied:

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

where R is the resistance to be found,
e is the voltage applied (A-B shorted),
E is the voltmeter reading with R connected, and
R_m is the resistance of the voltmeter.

The circuit of Fig. 17-6A is not suited to measuring low values of resistance (below a hundred ohms or so) with a high-resistance voltmeter. For such measurements the circuit of Fig. 17-6B can be used. The unknown resistance is

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

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

The formula is based on the assumption that the current in the complete circuit will be essentially constant whether or not the "unknown" terminals are short-circuited. This requires that R1 be very large compared with R_m – e.g., 3000 ohms for a 1-mA meter having an internal resistance of perhaps 50 ohms. A 3-volt battery would be necessary in this case in order to obtain a full-scale deflection with the "unknown" terminals open. R1 can be an adjustable resistor, to permit setting the open-terminals current to exact full scale.

A third circuit for measuring resistance is shown in Fig. 17-6C. In this case a high-resistance voltmeter is used to measure the voltage drop across a reference resistor, R2, when the unknown resistor is connected so that current flows through it, R2 and the battery in series. By suitable choice of R2 (low values for low-resistance, high values for high-resistance unknowns) this circuit will give equally good results on all resistance values in the range from one ohm to several megohms, provided that the voltmeter resistance, R_m, is always very high (50 times or more) compared with the resistance of R2. A 20,000-ohm-per-volt instrument (50-μA movement) is generally used. Assuming that the current through the voltmeter is negligible compared with the current through R2, the formula for the unknown is

$$R = \frac{eR2}{E} - R2$$

where R and $R2$ are as shown in Fig. 17-6C,
e is the voltmeter reading with A-B shorted, and
E is the voltmeter reading with R connected.

The "zero adjuster," R_1, is used to set the voltmeter reading exactly to full scale when the meter is calibrated in ohms. A 10,000-ohm variable resistor is suitable with a 20,000-ohms-per-volt meter. The battery voltage is usually 3 volts for ranges up to 100,000 ohms or so and 6 volts for higher ranges.

BRIDGE CIRCUITS

An important class of measurement circuits is the **bridge**, in which, essentially, a desired result is obtained by balancing the voltages at two different points in the circuit against each other so that there is zero potential difference between them. A voltmeter bridged between the two points will read zero (null) when this balance exists, but will indicate some definite value of voltage when the bridge is not balanced.

Bridge circuits are useful both on direct current and on ac of all frequencies. The majority of amateur applications is at radio frequencies, as shown later in this chapter. However, the principles of bridge operation are most easily introduced in terms of dc, where the bridge takes its simplest form.

The Wheatstone Bridge

The simple resistance bridge, known as the **Wheatstone bridge**, is shown in Fig. 17-7. All other bridge circuits – some of which are rather elaborate, especially those designed for ac – derive from this. The four resistors, R1, R2, R3, and R4 shown in A, are known as the **bridge arms**. For the

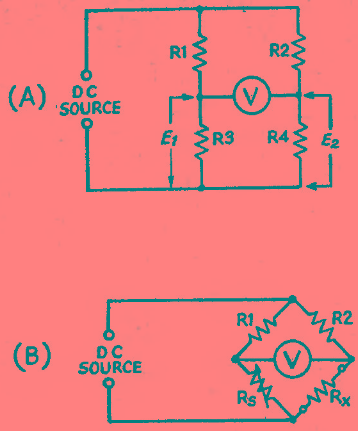


Fig. 17-7 – The Wheatstone bridge circuit. It is frequently drawn as at (B) for emphasizing its special function.

voltmeter reading to be zero, the voltages across R3 and R4 in series must add algebraically to zero; that is E1 must equal E2. R1R3 and R2R4 form voltage dividers across the dc source, so that if

$$\frac{R3}{R1 + R3} = \frac{R4}{R2 + R4}$$

E1 will equal E2.

The circuit is customarily drawn as shown at 17-7B when used for resistance measurement. The equation above can be rewritten

$$R_x = R_s \frac{R2}{R1}$$

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

Note that the resistance *ratios*, rather than the actual resistance values, determine the voltage balance. However, the values do have important practical effects on the sensitivity and power consumption. The bridge sensitivity is the readiness with which the meter responds to small amounts of unbalance about the null point; the "sharper" the null the more accurate the setting of R_s at balance.

The Wheatstone bridge is rarely used by amateurs for resistance measurement, the ohmmeter being the favorite instrument for that purpose. However, it is worthwhile to understand its operation because it is the prototype of more complex bridges.

ELECTRONIC VOLTMETERS

It has been pointed out (Fig. 17-3) that for many purposes the resistance of a voltmeter must be extremely high in order to avoid "loading" errors caused by the current that necessarily flows through the meter. This tends to cause difficulty in measuring relatively low voltages (under perhaps 1000 volts) because a meter movement of given sensitivity takes a progressively smaller multiplier resistance as the voltage range is lowered.

The voltmeter resistance can be made independent of the voltage range by using vacuum tubes or field-effect transistors as electronic dc amplifiers between the circuit being measured and the actual indicator, which is usually a conventional meter movement. As the input resistance of the

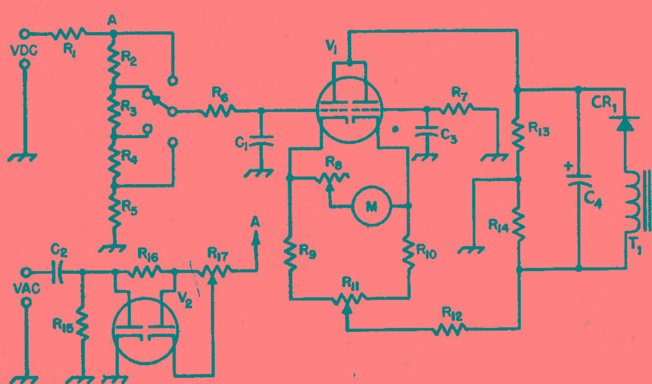


Fig. 17-8 – Vacuum-tube voltmeter circuit.

C1, C3 – .002- to .005-μF mica.
C2 – .01 μF, 1000 to 2000 volts, paper or mica.
C4 – 16 μF electrolytic, 150 volts.
CR1 – 400 PRV rectifier.
M – 0-200 microammeter.
R1 – 1 megohm, 1/2 watt.
R2-R5, incl. – To give desired voltage ranges, totaling 10 megohms.
R6, R7 – 2 to 3 megohms.
R8 – 10,000-ohm variable (calibrate).
R9, R10 – 2000 to 3000 ohms.
R11 – 5000- to 10,000-ohm control (zero set).
R12 – 10,000 to 50,000 ohms.
R13, R14 – App. 25,000 ohms. A 50,000-ohm slider-typa wire-wound can be used.
R15 – 10 megohms.
R16 – 3 megohms.
R17 – 10-megohm variable.
T1 – 130-volt 15-mA transformer (only secondary shown).
V1 – Dual triode, 12AU7A.
V2 – Dual diode, 6AL5.

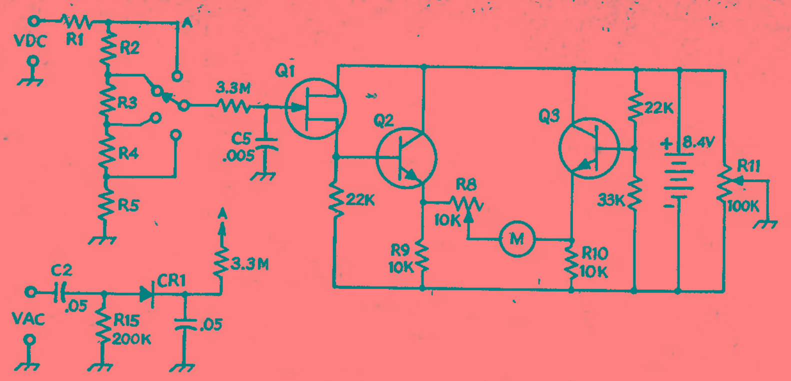


Fig. 17-9 — Electronic voltmeter using field-effect transistor for high input resistance. Components having the same functions as in the VTVM circuit of Fig. 17-8 carry the same designations. (Circuit is basic voltmeter circuit of the Heathkit IM-17.)
CR1 — Silicon diode.
Q1 — Field-effect transistor.
Q2, Q3 — Small-signal audio type.

electronic devices is extremely high – hundreds of megohms – they have essentially no loading effect on the circuit to which they are connected. They do, however, require a closed dc path in their input circuits (although this path can have very high resistance) and are limited in the amplitude of voltage that their input circuits can handle. Because of this, the device actually measures a small voltage across a portion of a high-resistance voltage divider connected to the circuit being measured. Various voltage ranges are obtained by appropriate taps on the voltage divider.

In the design of electronic voltmeters it has become practically standard to use a voltage divider having a resistance of 10 megohms, tapped as required, in series with a 1-megohm resistor incorporated in a probe that makes the actual contact with the "hot" side of the circuit under measurement. The total voltmeter resistance, including probe, is therefore 11 megohms. The 1-megohm probe resistor serves to isolate the voltmeter circuit from the "active" circuit.

The Vacuum-Tube Voltmeter

A typical vacuum-tube voltmeter (VTVM) circuit is given in Fig. 17-8. A dual triode, V1, is arranged so that, with no voltage applied to the left-hand grid, equal currents flow through both sections. Under this condition the two cathodes are at the same potential and no current flows through M. The currents can be adjusted to balance by potentiometer, R11, which takes care of variations in the tube sections and in the values of cathode resistors R9 and R10. When a positive dc voltage is applied to the left-hand grid the current through that tube section increases, so the current balance is upset and the meter indicates. The sensitivity of the meter is regulated by R8, which serves to adjust the calibration. R12, common to the cathodes of both tube sections, is a feedback resistor that stabilizes the system and makes the readings linear. R6 and C1 form a filter for any ac component that may be present, and R6 is balanced by R7 connected to the grid of the second tube section.

Values to be used in the circuit depend considerably on the supply voltage and the sensitivity of the meter, M. R12, and R13-R14, should be adjusted by trial so that the voltmeter circuit can be brought to balance, and to give full-scale deflection on M with about 3 volts applied to the left-hand grid (the voltage chosen for this determines the lowest voltage range of the instrument). The meter connections can be reversed to read voltages that are negative with respect to ground.

The small circuit associated with V2 is for ac measurements, as described in a later section.

As compared with conventional dc instruments, the VTVM has the disadvantages of requiring a source of power for its operation, and generally must have its "cold" terminal grounded in order to operate reliably. It is also somewhat susceptible to erratic readings from rf pickup when used in the vicinity of a transmitter, and in such cases may require shielding. However, its advantages outweigh these disadvantages in many applications.

The FET Voltmeter

The circuit of an electronic voltmeter using a field-effect transistor as an input device is shown in Fig. 17-9. Allowing for the differences between vacuum tubes and semiconductors, the operation of this circuit is analogous to that of Fig. 17-8. Transistors Q2 and Q3 correspond to the dual triode in the VTVM circuit, but since the input resistance of Q2 is fairly low, it is preceded by an FET, Q1, with source-coupled output. Note that in this circuit the "zero" or current-balance control, R11, varies the gate bias on Q1 by introducing an adjustable positive voltage in series with the source. This arrangement permits applying the adjustable bias to the gate through the voltmeter range divider, with no other provision needed for completing the dc gate-source path.

The small circuit associated with CR1 is for ac voltage measurement, to be discussed later.

As the power supply for the FET voltmeter is a self-contained battery, the grounding restrictions associated with a VTVM do not apply. The instrument can, however, be susceptible to rf fields if not shielded and grounded.

Electronic Ohmmeters

Most commercial electronic voltmeters include provision for measuring resistance and ac voltage, in addition to dc voltage. The basic ohmmeter circuit generally used is that of Fig. 17-6C. Since for practical purposes the input resistance of the vacuum tube or FET can be assumed to approach infinity, electronic ohmmeters are capable of measuring resistances in the hundreds of megohms – a much higher range than can be reached with an ordinary microammeter.

AC INSTRUMENTS AND CIRCUITS

Although purely electromagnetic instruments that operate directly from alternating current are available, they are seen infrequently in present-day amateur equipment. For one thing, their use is not feasible above power-line frequencies.

Practical instruments for audio and radio frequencies generally use a dc meter movement in conjunction with a rectifier. Voltage measurements suffice for nearly all test purposes. Current, as such, is seldom measured in the af range. When rf current is measured the instrument used is a thermocouple milliammeter or ammeter.

The Thermocouple Meter

In a thermocouple meter the alternating current flows through a low-resistance heating element. The power lost in the resistance generates heat which warms a "thermocouple," a junction of certain dissimilar metals which has the property of developing a small dc voltage when heated. This voltage is applied to a dc milliammeter calibrated in suitable ac units. The heater-thermocouple-dc meter combination is usually housed in a regular meter case.

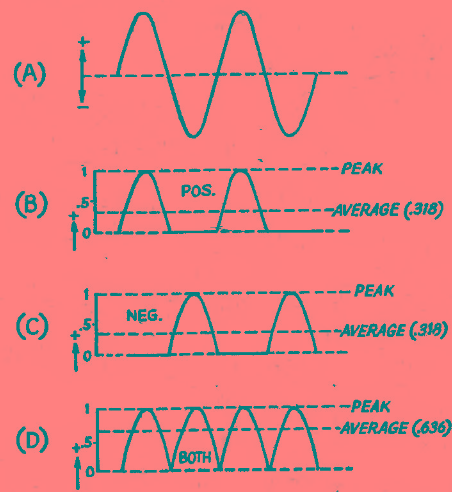


Fig. 17-11 – Sine-wave alternating current or voltage (A), with half-wave rectification of the positive half cycle (B) and negative half cycle (C). D – full-wave rectification. Average values are shown with relation to a peak value of 1.

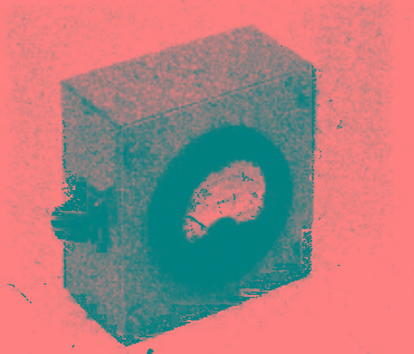


Fig. 17-10 – Rf ammetar mounted in a Minibox, with connectors for placing the meter in series with a coaxial lina. A bakelite-case meter should be used to minimize shunt capacitance (which introduces error) although a metal-cese meter can be used if mounted on bakelite sheet with a large cut-out in the case around the rim. The meter can be used for rf power measurements ($P = I^2R$) when connected between a transmitter and a nonreactive load of known resistance.

Thermocouple meters can be obtained in ranges from about 100 mA to many amperes. Their useful upper frequency limit is in the neighborhood of 100 MHz. Their principal value in amateur work is in measuring current into a known load resistance for calculating the rf power delivered to the load. A suitable mounting for this is shown in Fig. 17-10, for use in coaxial lines.

RECTIFIER INSTRUMENTS

The response of a rectifier-type meter is proportional (depending on the design) to either the peak amplitude or average amplitude of the rectified ac wave, and never directly responsive to the rms value. The meter therefore cannot be calibrated in rms without preknowledge of the relationship that happens to exist between the "real" reading and the rms value. This relationship, in general, is not known, except in the case of single-frequency ac (a sine wave). Very many practical measurements involve nonsinusoidal wave forms, so it is necessary to know what kind of instrument you have, and what it is actually

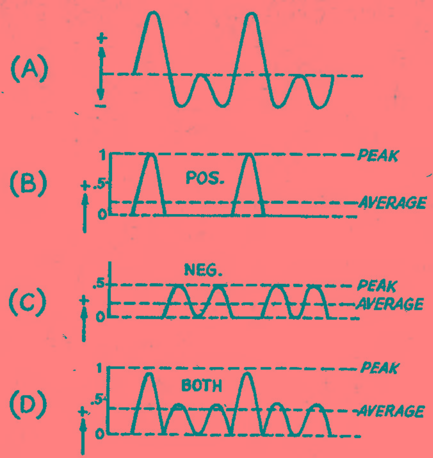


Fig. 17-12 — Same as Fig. 17-11 for an unsymmetrical waveform. The peak values are different with positive and negative half-cycle rectification.

reading, in order to make measurements intelligently.

Peak and Average with Sine-Wave Rectification

Fig. 17-11 shows the relative peak and average values in the outputs of half- and full-wave rectifiers (see power-supply chapter for further details). As the positive and negative half cycles of the sine wave have the same shape (A), half-wave rectification of either the positive half (B) or the negative half (C) gives exactly the same result. With full-wave rectification (D) the peak is still the same, but the average is doubled, since there are twice as many half cycles per unit of time.

Unsymmetrical Wave Forms

A nonsinusoidal waveform is shown in Fig. 17-12A. When the positive half cycles of this wave are rectified the peak and average values are as shown at B. If the polarity is reversed and the negative half cycles are rectified the peak value is different but the average value is unchanged. The fact that the average of the positive side is equal to the average of the negative side is true of *all* ac waveforms, but different waveforms have different averages. Full-wave rectification of such a "lop-sided" wave doubles the average value, but the peak reading is always the same as it is with the half cycle, that produces the *highest* peak in half-wave rectification.

Effective-Value Calibration

The actual scale calibration of commercially-made rectifier-type voltmeters is very often (almost always, in fact) in terms of rms values. For sine waves this is satisfactory, and useful since rms is the standard measure at power-line frequency. It is also useful for many rf applications where the waveform is often closely sinusoidal. But in other cases, particularly in the af range, the error may be considerable when the waveform is not pure.

Turn-Over

From Fig. 17-12 it is apparent that the calibration of an average-reading meter will be the same whether the positive or negative sides are rectified. A half-wave *peak*-reading instrument, however, will indicate different values when its connections to the circuit are reversed (turn-over effect). Very often readings are taken both ways, in which case the sum of the two is the peak-to-peak value, a useful figure in much audio and video work.

Average- and Peak-Reading Circuits

The basic difference between average- and peak-reading rectifier circuits is that in the former the output is not filtered while in the latter a filter capacitor is charged up to the peak value of the output voltage. Fig. 17-13A shows typical average-reading circuits, one half-wave and the other full-wave. In the absence of dc filtering the meter responds to wave forms such as are shown at B, C and D in Figs. 17-11 and 17-12, and since the inertia of the pointer system makes it unable to follow the rapid variations in current, it averages them out mechanically.

In Fig. 17-13A CR1 actuates the meter; CR2 provides a low-resistance dc return in the meter circuit on the negative half cycles. R1 is the voltmeter multiplier resistance. R2 forms a voltage

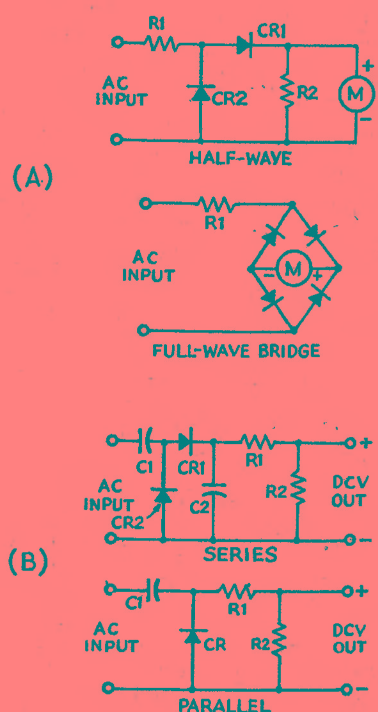


Fig. 17-13 — A — Half-wave and full-wave rectification for an instrument intended to operate on average values. B — half-wave circuits for a peak-reading meter.

divider with R1 (through CR1) which prevents more than a few ac volts from appearing across the rectifier-meter combination. A corresponding resistor can be used across the full-wave bridge circuit.

In these two circuits no provision is made for isolating the meter from any dc voltage that may be on the circuit under measurement. The error caused by this can be avoided by connecting a large capacitance in series with the "hot" lead. The reactance must be low compared with the meter impedance (see next section) in order for the full ac voltage to be applied to the meter circuit. As much as 1 μF may be required at line frequencies with some meters. The capacitor is not usually included in a VOM.

Series and shunt peak-reading circuits are shown in Fig. 17-13B. Capacitor C1 isolates the rectifier from dc voltage on the circuit under measurement. In the series circuit (which is seldom used) the time constant of the C2R1R2 combination must be very large compared with the period of the lowest ac frequency to be measured; similarly with C1R1R2 in the shunt circuit. The reason is that the capacitor is charged to the peak value of voltage when the ac wave reaches its maximum, and then must hold the charge (so it can register on a dc meter) until the next maximum of the same polarity. If the time constant is 20 times the ac period the charge will have decreased by about 5 percent by the time the next charge occurs. The *average* drop will be smaller, so the error is appreciably less. The error will decrease rapidly with increasing frequency, assuming no change in the circuit values, but will increase at lower frequencies.

In Fig. 17-13B R1 and R2 form a voltage divider which reduces the peak dc voltage to 71 percent of its actual value. This converts the peak reading to rms on sine-wave ac. Since the peak-reading circuits are incapable of delivering appreciable current without considerable error, R2 is usually the 11-megohm input resistance of an electronic voltmeter. R1 is therefore approximately 4.7 megohms, making the total resistance approach 16 megohms. A capacitance of .05 μF is sufficient for low audio frequencies under these conditions. Much smaller values of capacitance suffice for radio frequencies, obviously.

Voltmeter Impedance

The impedance of the voltmeter at the frequency being measured may have an effect on the accuracy similar to the error caused by the resistance of a dc voltmeter, as discussed earlier. The ac meter acts like a resistance in prallel with a capacitance, and since the capacitive reactance decreases with increasing frequency, the impedance also decreases with frequency. The resistance is subject to some variation with voltage level, particularly at very low voltages (of the order of 10 volts or less) depending upon the sensitivity of the meter movement and the kind of rectifier used.

The ac load resistance represented by a diode rectifier is approximately equal to one-half its dc load resistance. In Fig. 17-13A the dc load is essentially the meter resistance, which is generally quite low compared with the multiplier resistance R1, so the total resistance will be about the same as the multiplier resistance. The capacitance depends on the components and construction, test lead length and disposition, and such factors. In general, it has little or no effect at power-line and low audio frequencies, but the ordinary VOM loses accuracy at the higher audio frequencies and is of little use at rf. For radio frequencies it is necessary to use a rectifier having very low inherent capacitance.

Similar limitations apply to the peak-reading circuits. In the parallel circuit the resistive component of the impedance is smaller than in the series circuit, since the dc load resistance, R1R2, is directly across the circuit being measured, and is therefore in parallel with the diode ac load resistance. In both peak-reading circuits the effective capacitance may range from 1 or 2 to a few hundred pF. Values of the order of 100 pF are to be expected in electronic voltmeters of customary design and construction.

Linearity

Fig. 17-14, a typical current/voltage characteristic of a small semiconductor rectifier, indicates that the forward dynamic resistance of the diode is not constant, but rapidly decreases as the forward voltage is increased from zero. The transition from high to low resistance occurs at considerably less than 1 volt, but is in the range of voltage required by the associated dc meter. With an average-reading circuit the current tends to be proportional to the *square* of the applied voltage. This crowds the calibration points at the low end of the meter scale. For most measurement purposes, however, it is far more desirable for the output to be "linear;" that is, for the reading to be *directly* proportional to the applied voltage.

To achieve linearity it is necessary to use a relatively large load resistance for the diode – large enough so that this resistance, rather than the diode's own resistance, will govern the current flow. A linear or equally spaced scale is thus gained at the expense of sensitivity. The amount of resistance needed depends on the type of diode;

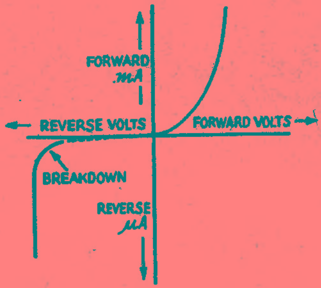


Fig. 17-14 — Typical semiconductor diode characteristic. Actual current and voltage values vary with the type of diode, but the forward-current curve would be in its steep part with only a volt or so applied. Note change in current scale for reverse current. Breakdown voltage, again depending on diode type, may range from 15 or 20 volts to several hundred.

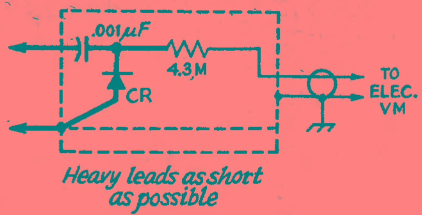


Fig. 17-15 — Rf probe circuit. CR is a small semiconductor rectifier, usually point-contact germanium. The resistor value, for exact voltage division to rms, should be 4.14 megohms, but standard values are generally used, including 4.7 megohms.

5000 to 50,000 ohms usually suffices for a germanium rectifier, depending on the dc meter sensitivity, but several times as much may be needed for silicon. The higher the resistance, the greater the meter sensitivity required; i.e., the basic meter must be a microammeter rather than a low-range milliammeter.

Reverse Current

When voltage is applied in the reverse direction there is a small leakage current in semiconductor diodes. This is equivalent to a resistance connected across the rectifier, allowing current to flow during the half cycle which should be completely nonconducting, and causing an error in the dc meter reading. This "back resistance" is so high as to be practically unimportant with silicon, but may be less than 100 kΩ with germanium.

The practical effect of back resistance is to limit the amount of resistance that can be used in the dc load resistance. This in turn affects the linearity of the meter scale.

The back resistance of vacuum-tube diodes is infinite, for practical purposes.

RF VOLTAGE

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

For general rf measurements an rf probe is used in conjunction with an electronic voltmeter, substituted for the dc probe mentioned earlier. The circuit of Fig. 17-15, essentially the peak-reading shunt circuit of Fig. 17-13B, is generally used. The series resistor, installed in the probe close to the rectifier, prevents rf from being fed through the probe cable to the electronic voltmeter, being helped in this by the cable capacitance. This resistor, in conjunction with the 10-megohm divider resistance of the electronic voltmeter, also reduces the peak rectified voltage to a dc value equivalent to the rms of the rf signal, to make the rf readings consistent with the regular ac calibration.

Of the diodes readily available to amateurs, the germanium point-contact type is preferred for rf applications. It has low capacitance (of the order of 1 pF) and in the high-back-resistance types the reverse current is not serious. The principal limitation is that its safe reverse voltage is only about 50-75 volts, which limits the rms applied voltage to 15 or 20 volts, approximately. Diodes can be connected in series to raise the overall rating.

Linearity at Radio Frequencies

The bypass or filter capacitance normally used in rf rectifier circuits is large enough, together with the resistance in the system, to have a time constant sufficient for peak readings. However, if the resistance is low (the load sometimes is just the microammeter or milliammeter alone) the *linearity* of the voltmeter will be affected as previously described, even if the time constant is fairly large. It is not safe to assume that the voltmeter is even approximately linear unless the load resistance is of the order of 10,000 ohms or greater.

Nonlinear voltmeters are useful as indicators, as where null indicators are called for, but should not be depended upon for actual measurement of voltage.

RF Power

Power at radio frequencies can be measured by means of an accurately-calibrated rf voltmeter connected across the load in which the power is being dissipated. If the load is a known pure resistance the power, by Ohm's Law, is equal to E^2/R, where E is the rms value of the voltage.

The method only indicates *apparent* power if the load is not a pure resistance. The load can be a terminated transmission line tuned, with the aid of bridge circuits such as are described in the next section, to act as a known resistance. An alternative load is a "dummy" antenna, a known pure resistance capable of dissipating the rf power safely.

AC BRIDGES

In its simplest form, the ac bridge is exactly the same as the Wheatstone bridge discussed earlier. However, complex impedances can be substituted for resistances, as suggested by Fig. 17-16A. The same bridge equation holds if Z is substituted for R in each arm. For the equation to be true, however, *the phase angles as well as the numerical values of the impedances must balance*; otherwise, a true null voltage is impossible to obtain. This means that a bridge with all "pure" arms (pure resistance or reactance) cannot measure complex impedances; a combination of R and X must be present in at least one arm besides the unknown.

The actual circuits of ac bridges take many forms, depending on the type of measurement intended and on the frequency range to be covered. As the frequency is raised stray effects (unwanted capacitances and inductances, principally) become more pronounced. At radio frequencies special attention must be paid to minimizing them.

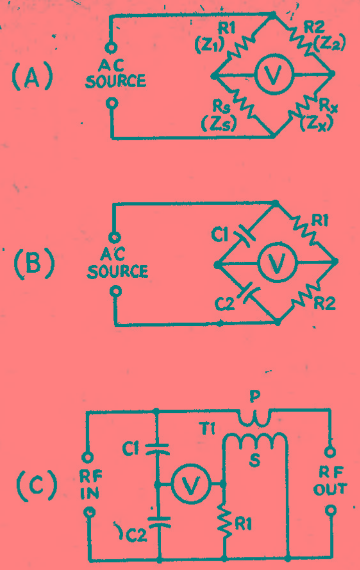


Fig. 17-16 — A — Generalized form of bridge circuit for either ac or dc. B — One form of ac bridge frequently used for rf measurements. C — SWR bridge for use in transmission lines. This circuit is often calibrated in power rather then voltage.

Most amateur-built bridges are used for rf measurements, especially SWR measurements on transmission lines. The circuits at B and C, Fig. 17-16, are favorites for this purpose. These basic forms are often modified considerably, as will be seen by the constructional examples later in the chapter.

Fig. 17-16B is useful for measuring both transmission lines and "lumped-constant" components. Combinations of resistance and capacitance are often used in one or more arms; this may be required for eliminating the effects of stray capacitance.

Fig. 17-16C is used only on transmission lines, and only on those lines having the characteristic impedance for which the bridge is designed.

SWR Measurement – The Reflectometer

In measuring standing-wave ratio advantage is taken of the fact that the voltage on a transmission line consists of two components traveling in opposite directions. The power going from the transmitter to the load is represented by one voltage (designated "incident" or "forward") and the power reflected from the load is represented by the other. Because the relative amplitudes and phase relationships are definitely established by the line's characteristic impedance, its length and the load impedance in which it is terminated, a bridge circuit can separate the incident and reflected voltages for measurement. This is sufficient for determining the SWR. Bridges designed for this purpose are frequently called reflectometers.

Referring to Fig. 17-16A, if R1 and R2 are made equal, the bridge will be balanced when $R_X = R_S$. This is true whether R_X is an actual resistor or the input resistance of a perfectly matched transmission line, provided R_S is chosen to equal the characteristic impedance of the line. Even if the line is not properly matched, the bridge will still be balanced for power traveling *outward* on the line, since outward-going power sees only the Z_0 of the line until it reaches the load. However, power reflected back from the load does not "see" a bridge circuit, and the reflected voltage registers on the voltmeter. From the known relationship between the incident and reflected voltages the SWR is easily calculated:

$$SWR = \frac{V_0 + V_r}{V_0 - V_r}$$

where V_0 is the forward voltage and V_r is the reflected voltage. The forward voltage may be measured either by disconnecting R_X or shorting it.

The "Reflected Power Meter"

Fig. 17-16C makes use of mutual inductance between the primary and secondary of T1 to establish a balancing circuit. C1 and C2 form a voltage divider in which the voltage across C2 is in the same phase as the voltage at that point on the transmission line. The relative phase of the voltage across R1 is determined by the phase of the *current* in the line. If a pure resistance equal to the design impedance of the bridge is connected to the "RF Out" terminals, the voltages across R1 and C2 will be out of phase and the voltmeter reading will be minimum; if the *amplitudes* of the two voltages are also equal (they are made so by bridge adjustment) the voltmeter will read zero. Any other value of resistance or impedance connected to the "RF Out" terminals will result in a finite voltmeter reading. When used in a transmission line this reading is proportional to the reflected voltage. To measure the incident voltage the secondary terminals of T1 can be reversed. To function as described, the secondary leakage reactance of T1 must be very large compared to the resistance of R1.

Instruments of this type are usually designed for convenient switching between forward and reflected, and are often calibrated to read power in the specified characterisitic impedance. The net power transmission is equal to the incident power minus the reflected power.

Sensitivity vs. Frequency

In all of the circuits in Fig. 17-16 the sensitivity is independent of the applied frequency, within practical limits. Stray capacitances and couplings generally limit the performance of all three at the high-frequency end of the useful range. Fig. 17-16A will work right down to dc, but the low-frequency performance of Fig. 17-16B is degraded when the capacitive reactances become so large that voltmeter impedance becomes low in comparison (in all these bridge circuits, it is assumed that the voltmeter impedance is high compared with the impedance of the bridge arms). In Fig. 17-16C the performance is limited at low frequencies by the fact that the transformer reactance decreases with frequency, so that eventually the reactance is not very high in comparison with the resistance of R1.

The "Monimatch"

A type of bridge which is quite simple to make, but in which the sensitivity rises directly with frequency, is the Monimatch and its various offspring. The circuit cannot be described in terms of lumped constants, as it makes use of the distrbuted mutual inductance and capacitance between the center conductor of a transmission line and a wire placed parallel to it. The wire is terminated in a resistance approximating the characteristic impedance of the transmission line at one end and feeds a diode rectifier at the other. A practical example is shown later in this chapter.

FREQUENCY MEASUREMENT

The regulations governing amateur operation require that the transmitted signal be maintained inside the limits of certain bands of frequencies.* The exact frequency need not be known, so long as it is not outside the limits. On this last point there are no tolerances: It is up to the individual amateur to see that he stays safely "inside."

This is not difficult to do, but requires some simple apparatus and the exercise of some care. The apparatus commonly used is the frequency-marker generator, and the method involves use of the station receiver, as in Fig. 17-17.

THE FREQUENCY MARKER

The marker generator in its simplest form is a high-stability oscillator generating a series of signals which, when detected in the receiver, mark the exact edges of the amateur assignments. It does this by oscillating at a low frequency that has harmonics falling on the desired frequencies.

All U.S. amateur band limits are exact multiples of 25 kHz, whether at the extremes of a band or at points marking the subdivisions between types of emission, license privileges, and so on. A 25-kHz fundamental frequency therefore will produce the desired marker signals if its harmonics at the higher frequencies are strong enough. But since harmonics appear at 25-kHz intervals throughout the spectrum, along with the desired markers, the problem of identifying a *particular* marker arises. This is easily solved if the receiver has a reasonably good calibration. If not, most marker circuits provide for a choice of fundamental outputs of 100 and 50 kHz as well as 25 kHz, so the question can be narrowed down to initial identification of 100-kHz intervals. From these, the desired 25-kHz (or 50-kHz) points can easily be spotted. Coarser frequency intervals are rarely required; there are usually signals available from stations of known frequency, and the 100-kHz points can be counted off from them.

Transmitter Checking

In checking one's own transmitter frequency the signal from the transmitter is first tuned in on the receiver and the dial setting at which it is heard is noted. Then the *nearest* marker frequencies above and below the transmitter signal are turned in and identified. The transmitter frequency is obviously between these two known frequencies.

If the marker frequencies are accurate, this is all that needs to be known – except that the transmitter frequency must not be so close to a band (or subband) edge that sideband frequencies, especially in phone transmission, will extend over the edge.

If the transmitter signal is "inside" a marker at the edge of an assignment, to the extent that there is an audible beat note with the receiver's BFO turned off, normal cw sidebands are safely inside the edge. (This statement does not take into account *abnormal* sidebands such as are caused by clicks and chirps.) For phone the "safety" allowance is usually taken to be about 3 kHz, the nominal width of one sideband. A frequency difference of this order can be estimated by noting the receiver dial settings for the two 25-kHz markers which bracket the signal and dividing 25 by the number of dial divisions between them. This will give the number of kHz per dial division.

Transceivers

The method described above is applicable when the receiver and transmitter are separate pieces of equipment. When a transceiver is used and the

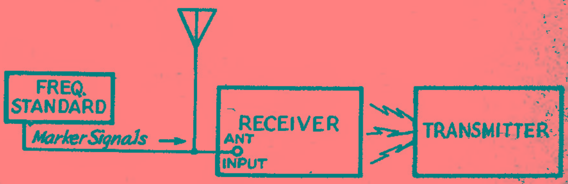


Fig. 17-17 — Setup for using a frequency standard. It is necessary that the transmitter signal be weak in the receiver — of the same order of strength as the marker signal from the standard. This requirement can usually be met by turning on just the transmitter oscillator, leaving all power off any succeeding stages. In some cases it may also be necessary to disconnect the antenna from the receiver.

* These limits depend on the type of emission and class of license held, as well as on international agreements. See the latest edition of *The Radio Amateur's License Manual* for current status.

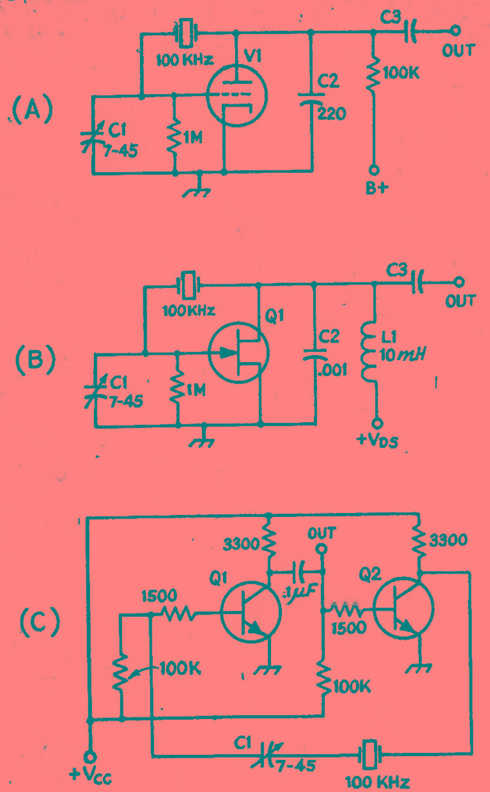


Fig. 17-18 — Three simple 100-kHz oscillator circuits. C is the most suitable of available transistor circuits (for marker generators) and is recommended where solid-state is to be used. In all three circuits C1 is for fine frequency edjustment. The output coupling capacitor, C3, is generally small — 20 to 50 pF — a compromise to avoid loading the oscillator by the receiver antenna input while maintaining adequata coupling for good harmonic strength.

transmitting frequency is automatically the same as that to which the receiver is tuned, setting the tuning dial to a spot between two known marker frequencies is all that is required.

The proper dial settings for the markers are those at which, with the BFO on, the signal is tuned to *zero* beat — the spot where the beat disappears as the tuning makes the beat tone progressively lower. Exact zero beat can be determined by a very slow rise and fall of background noise, caused by a beat of a cycle or less per second.

FREQUENCY-MARKER CIRCUITS

The basic frequency-determining element in most amateur frequency markers is a 100-kHz crystal. Although the marker generator should produce harmonics at 25-kHz and 50-kHz intervals, crystals (or other high-stability devices) for frequencies lower than 100 kHz are expensive and difficult to obtain. However, there is really no need for them, since it is easy to divide the basic frequency down to any figure one desires; 50 and 25 kHz require only two successive divisions, each by 2. In the division process, the harmonic output of the generator is greatly enhanced, making the generator useful at frequencies well into the vhf range.

Simple Crystal Oscillators

Fig. 17-18 illustrates a few of the simpler circuits. Fig. 17-18A is a long-time favorite where vacuum tubes are used and is often incorporated in receivers. C1 in this and the other circuits is used for exact adjustment of the oscillating frequency to 100 kHz, which is done by using the receiver for comparing one of the oscillator's harmonics with a standard frequency transmitted by WWV, WWVH, or a similar station.

Fig. 17-18B is a field-effect transistor analog of the vacuum-tube circuit. However, it requires a 10-mH coil to operate well, and since the harmonic output is not strong at the higher frequencies the circuit is given principally as an example of a simple transistor arrangement. A much better oscillator is shown at C. This is a cross-connected pair of transistors forming a multivibrator of the "free-running" or "astable" type, locked at 100 kHz by using the crystal as one of the coupling elements. While it can use two separate bipolar transistors as shown, it is much simpler to use an integrated-circuit dual gate, which will contain all the necessary parts except the crystal and capacitors and is considerably less expensive, as well as more compact, than the separate components. An example is shown later in the chapter.

Frequency Dividers

Electronic division is accomplished by a "bistable" flip-flop or cross-coupled circuit which produces one output change for every two impulses applied to its input circuit, thus dividing the applied frequency by 2. All division therefore must be in terms of some power of 2. In practice this is no handicap since with modern integrated-circuit flip-flops, circuit arrangements can be worked out for division by any desired number.

As flip-flops and gates in integrated circuits come in compatible series — meaning that they work at the same supply voltage and can be directly connected together — a combination of a dual-gate version of Fig. 17-18C and a dual flip-flop make an attractively simple combination for the marker generator.

There are several different basic types of flip-flops, the variations having to do with methods of driving (dc or pulse operation) and control of the counting function. Information on the operating principles and ratings of a specific type usually can be obtained from the manufacturer. The counting-control functions are not needed in using the flip-flop in a simple marker generator, although they come into play when dividing by some number other than a power of 2.

Frequency Standards

The difference between a marker generator and a frequency standard is that in the latter special pains are taken to make the oscillator frequency as stable as possible in the face of variations in temperature, humidity, line voltage, and other factors which could cause a small change in frequency.

While there are no definite criteria that distinguish the two in this respect, a circuit designated as a "standard" for amateur purposes should be capable of maintaining frequency within at least a few parts per million under normal variations in ambient conditions, without adjustment. A simple marker generator using a 100-kHz crystal can be expected to have frequency variations 10 times (or more) greater under similar conditions. It can of course be adjusted to exact frequency at any time the WWV (or equivalent) signal is available.

The design considerations of high-precision frequency standards are outside the scope of this chapter, but information is available from time to time in periodicals.

OTHER METHODS OF FREQUENCY CHECKING

The simplest possible frequency-measuring device is a parallel *LC* circuit, tunable over a desired frequency range and having its tuning dial calibrated in terms of frequency. It can be used only for checking circuits in which at least a small amount of rf power is present, because the energy required to give a detectable indication is not available in the *LC* circuit itself; it has to be extracted from the circuit being measured; hence the name **absorption frequency meter**. It will be observed that what is actually measured is the frequency of the rf *energy*, *not* the frequency to which the circuit in which the energy is present may be tuned.

The measurement accuracy of such an instrument is low, compared with the accuracy of a marker generator, because the *Q* of a practicable *LC* circuit is not high enough to make precise reading of the dial possible. Also, any two circuits coupled together react on each others' tuning. (This can be minimized by using the loosest coupling that will give an adequate indication.)

The absorption frequency meter has one useful advantage over the marker generator – it will respond *only* to the frequency to which it is tuned, or to a band of frequencies very close to it. Thus there is no harmonic ambiguity, as there sometimes is when using a marker generator.

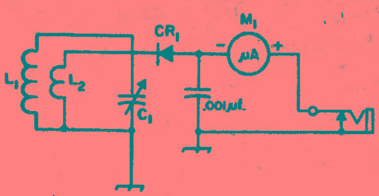


Fig. 17-19A — Absorption frequency-meter circuit. The closed-circuit phone jack may be omitted if listening is not wanted; in that case the positive terminal of M1 goes to common ground.

Absorption Circuit

A typical absorption frequency-meter circuit is shown in Fig. 17-19. In addition to the adjustable tuned circuit, L1C1, it includes a pickup coil, L2, wound over L1, a high-frequency semiconductor diode, CR1, and a microammeter or low-range (usually not more than 0-1 mA) milliammeter. A phone jack is included so the device can be used for listening to the signal.

The sensitivity of the frequency meter depends on the sensitivity of the dc meter movement and the size of L2 in relation to L1. There is an optimum size for this coil which has to be found by experiment. An alternative is to make the rectifier connection to an adjustable tap on L1, in which case there is an optimum tap point. In general, the rectifier coupling should be a little *below* (that is, less tight) the point that gives maximum response, since this will make the indications sharper.

Calibration

The absorption frequency meter must be calibrated by taking a series of readings on various frequencies from circuits carrying rf power, the frequency of the rf energy first being determined by some other means such as a marker generator and receiver. The setting of the dial that gives the highest meter indication is the calibration point for that frequency. This point should be determined by tuning through it with loose coupling to the circuit being measured.

OTHER INSTRUMENTS AND MEASUREMENTS

Many measurements require a source of ac power of adjustable frequency (and sometimes adjustable amplitude as well) in addition to what is already available from the transmitter or receiver. Rf and af test oscillators, for example, provide signals for purposes such as receiver alignment, testing of phone transmitters, and so on. Another valuable adjunct to the station is the oscilloscope, especially useful for checking phone modulation.

Rf Oscillators for Circuit Alignment

Receiver testing and alignment, covered in an earlier chapter, uses equipment common to ordinary radio service work. Inexpensive rf signal generators are available, both complete and in kit form. However, any source of signal that is weak enough to avoid overloading the receiver usually will serve for alignment work. The frequency marker generator is a satisfactory signal source. In addition, its frequencies, although not continuously adjustable, are known far more precisely, since the usual signal-generator calibration is not highly accurate. For rough work the dip meter described in the next section will serve.

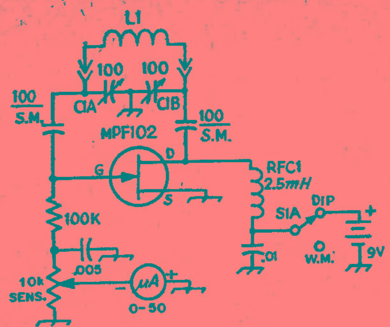


Fig. 17-19B — An FET source-dipper circuit suitable for use from 1.5 to 50 MHz. For operation at vhf and uhf the value of C1 should be made smaller, RFC1 would be a vhf type, end the bypass capacitors would be smaller in value. For uhf use Q1 would be changed to a uhf-type FET, a 2N4416 or similar.

THE DIP METER

The dip meter reverses the absorption-wavemeter procedure in that *it* supplies the rf power by incorporating a tunable oscillator from which the circuit being checked absorbs energy when this circuit and the oscillator are tuned to the same frequency and coupled together. In the vacuum-tube version the energy absorption causes a decrease or "dip" in the oscillator's rectified grid current, measured by a dc microammeter.

The same principle can be applied to solid-state oscillators. In some transistor versions the oscillator rf power is rectified by a diode to provide a meter indication. This technique can result in "dead spots" in the tuning range if the oscillator power is too low to enable the diode to conduct at all times. The circuit of Fig. 17-19B avoids the problem by measuring the changes in source current. In the W.M. (wavemeter) position of S1 the gate-source junction of Q1 serves as the detector diode.

Each tuning range of the dipper should overlap to provide sufficient coverage to check circuits of unknown resonant frequency. Plug-in coils are normally used to allow continuous coverage from 1.5 to at least 250 MHz.

Calibration

A dipper should have reasonably accurate calibration. Calibration of the dipper dial can be effected by monitoring the dipper output signal with a calibrated receiver. Make sure the *fundamental* frequency of the dipper is being used during calibration.

Operating the Dip Meter

The dip meter will check only resonant circuits, since nonresonant circuits or components will not absorb energy at a specific frequency. The circuit may be either lumped or linear (a transmission-line type circuit) provided only that it has enough Q to give sufficient coupling to the dip-meter coil for detectable absorption of rf energy. Generally the coupling is principally inductive, although at times there may be sufficient capacitive coupling between the meter and a circuit point that is at relatively high potential with respect to ground to permit a reading. For inductive coupling, maximum energy absorption will occur when the meter

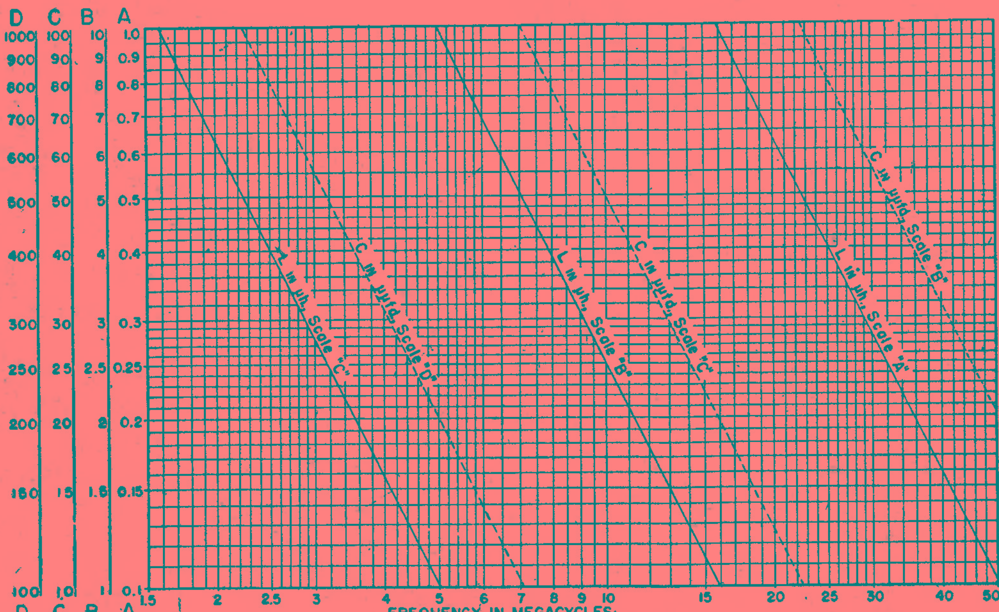


Fig. 17-20 — Chart for determining unknown values of L and C in the range of 0.1 to 100 μH and 2 to 1000 pF, using standards of 100 pF and 5 μH.

Fig. 17-21 — A convenient mounting, using binding-post plates, for L and C standards made from commercially available parts. The capacitor is a 100-pF silver mica unit, mounted so the lead length is as nearly zero as possible. The inductance standard, 5 μH, is 17 turns of coil stock, 1-inch diameter, 16 turns per inch.

is coupled to a coil (the same coupling rules that apply to any two coils are operative here) in the tuned circuit being checked, or to a high-current point in a linear circuit.

Because of distributed capacitance (and sometimes inductance) most circuits resonant at the lower amateur frequencies will show quasi-linear-type resonances at or close to the vhf region. A vhf dip meter will uncover these, often with beneficial results since such "parasitic" resonances can cause unwanted responses at harmonics of the intended frequency, or be responsible for parasitic oscillations in amplifiers. Caution must be used in checking transmission lines or antennas – and, especially, combinations of antenna and line – on this account, because these linear circuits have well-defined series of harmonic responses, based on the lowest resonant frequency, which may lead to false conclusions respecting the behavior of the system.

Measurements with the dip meter are essentially frequency measurements, and for best accuracy the coupling between the meter and circuit under checking must be as loose as will allow a perceptible dip. In this respect the dip meter is similar to the absorption wavemeter.

Measuring Inductance and Capacitance with the Dip Meter

With a carefully calibrated dip meter, properly operated, inductance and capacitance in the values ordinarily used for the 1.5-50 MHz range can be measured with ample accuracy for practical work. The method requires two accessories: an inductance "standard" of known value, and a capacitance standard also known with reasonable accuracy. Values of 100 pF for the capacitance and 5 μH for the inductance are convenient. The chart of Fig. 17-20 is based on these values.

The L and C standards can be quite ordinary components. A small silver-mica capacitor is satisfactory for the capacitance, since the customary tolerance is ±5 percent. The inductance standard can be cut from commercial machine-wound coil stock; if none is available, a homemade equivalent in diameter, turn spacing, and number of turns can be substituted. The inductance will be 5 μH within amply close tolerances if the specifications in Fig. 17-21 are followed closely. In any case, the inductance can easily be adjusted to the proper value; it should resonate with the 100-pF capacitor at 7100 kHz.

The setup for measuring an unknown is shown in Fig. 17-22. Inductance is measured with the unknown connected to the standard capacitance. Couple the dip meter to the coil and adjust the meter for the dip, using the loosest possible coupling that will give a usable indication. Similar procedure is followed for capacitance measurement, except that the unknown is connected to the standard inductance. Values are read off the chart for the frequency indicated by the dip meter.

Coefficient of Coupling

The same equipment can be used for measurement of the coefficient of coupling between two coils. This simply requires two measurements of inductance (of *one* of the coils) with the coupled coil first open-circuited and then short-circuited. Connect the 100-pF standard capacitor to one coil and measure the inductance with the terminals of the second coil open. Then short the terminals of the second coil and again measure the inductance of the first. The coefficient of coupling is given by

$$k = \sqrt{1 - \frac{L_2}{L_1}}$$

where k = coefficient of coupling
$L1$ = inductance of first coil with terminals of second coil open
$L2$ = inductance of first coil with terminals of second coil shorted.

AUDIO-FREQUENCY OSCILLATORS

Tests requiring an audio-frequency signal generally call for one that is a reasonably good sine wave, and the best oscillator circuits for this are *RC*-coupled, operating as nearly as possible as Class A amplifiers. Variable frequency covering the entire audio range is needed for determining frequency response of audio amplifiers, but this is

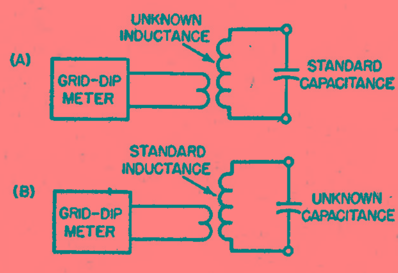


Fig. 17-22 — Setups for measuring inductance and capacitance with the dip meter.

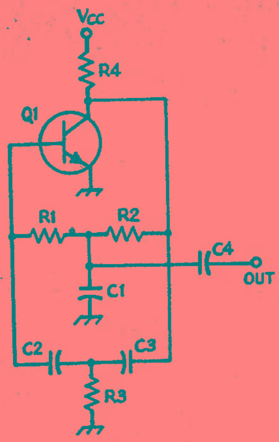


Fig. 17-23 — Twin-T audio oscillator circuit. Representative values for R1-R2 and C1 range from 18kΩ and .05 μF for 750 Hz to 15kΩ and .02 μF for 1800 Hz. For the same frequency range, R3 and C2-C3 very from 1800 ohms and .02 μF to 1500 ohms and .01 μF. R4 should be approximately 3300 ohms. C4, the output coupling capacitor, can be .05 μF for high-impedance loads.

a relatively unimportant type of test in amateur equipment. The variable-frequency af signal generator is best purchased complete; kits are readily available at prices that compare very favorably with the cost of parts.

For most phone-transmitter testing, and for simple trouble shooting in af amplifiers, an oscillator generating one or two frequencies with good wave form is adequate. A "two-tone" (dual) oscillator is particularly useful for testing sideband transmitters, and a constructional example is found later in the chapter.

The circuit of a simple *RC* oscillator useful for general test purposes is given in Fig. 17-23. This "Twin-T" arrangement gives a wave form that is satisfactory for most purposes, and by choice of circuit constants the oscillator can be operated at any frequency in the usual audio range. R1, R2 and C1 form a low-pass type network, while C2C3R3 is high-pass. As the phase shifts are opposite, there is only one frequency at which the total phase shift from collector to base is 180 degrees, and oscillation will occur at this frequency. Optimum operation results when C1 is approximately twice the capacitance of C2 or C3, and R3 has a resistance about 0.1 that of R1 or R2 (C2 = C3 and R1 = R2). Output is taken across C1, where the harmonic distortion is least. A relatively high-impedance load should be used — 0.1 megohm or more.

A small-signal af transistor is suitable for Q1. Either npn or pnp types can be used, with due regard for supply polarity. R4, the collector load resistor, must be large enough for normal amplification, and may be varied somewhat to adjust the operating conditions for best waveform.

RESISTORS AT RADIO FREQUENCIES

Measuring equipment, in some part of its circuit, often requires essentially pure resistance — that is, resistance exhibiting only negligible reactive effects on the frequencies at which measurement is intended. Of the resistors available to amateurs, this requirement is met only by small composition (carbon) resistors. The inductance of wire-wound resistors makes them useless for amateur frequencies.

The reactances to be considered arise from the inherent inductance of the resistor itself and its leads, and from small stray capacitances from one part of the resistor to another and to surrounding conductors. Although both the inductance and capacitance are small, their reactances become increasingly important as the frequency is raised. Small composition resistors, properly mounted, show negligible capacitive reactance up to 100 MHz or so in resistance values up to a few hundred ohms; similarly, the inductive reactance is negligible in values *higher* than a few hundred ohms. The optimum resistance region in this respect is in the 50 to 200-ohm range, approximately.

Proper mounting includes reducing lead length as much as possible, and keeping the resistor separated from other resistors and conductors. Care must also be taken in some applications to ensure that the resistor, with its associated components, does not form a closed loop into which a voltage could be induced magnetically.

So installed, the resistance is essentially pure. In composition resistors the skin effect is very small, and the rf resistance up to vhf is very closely the same as the dc resistance.

Dummy Antennas

A dummy antenna is simply a resistor that, in impedance characteristics, can be substituted for an antenna or transmission line for test purposes. It permits leisurely transmitter testing without radiating a signal. (The amateur regulations strictly limit the amount of "on-the-air" testing that may be done.) It is also useful in testing receivers, in that electrically it resembles an antenna, but does not pick up external noise and signals, a desirable feature in some tests.

For transmitter tests the dummy antenna must be capable of dissipating safely the entire power output of the transmitter. Since for most testing it is desirable that the dummy simulate a perfectly-matched transmission line, it should be a pure resistance, usually of approximately 52 or 73

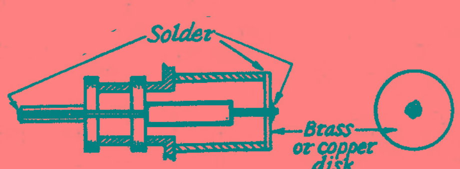


Fig. 17-24 — Dummy antenna made by mounting e composition resistor in e PL-259 coaxial plug. Only the inner portion of the plug is shown; the cap screws on after the assembly is completed.

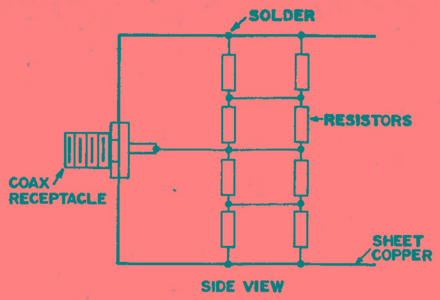


Fig. 17-25 – Using resistors in series-parallel to increase the power rating of e small dummy antenna. Mounted in this way on pieces of flat copper, inductance is reduced to a minimum. Eight 100-ohm 2-watt composition resistors in two groups, each four resistors in parallel, can be connected in series to form a 50-ohm dummy. The open construction shown permits free air circulation. Resistors drawn heavy are in one "deck"; light ones are in the other.

ohms. This is a severe limitation in home construction, because nonreactive resistors of more than a few watts rated safe dissipation are very difficult to obtain. (There are, however, dummy antenna kits available that can handle up to a kilowatt.)

For receiver and minipower transmitter testing an excellent dummy antenna can be made by installing a 51- or 75-ohm composition resistor in a PL-259 fitting as shown in Fig. 17-24. Sizes from one-half to two watts are satisfactory. The disk at the end helps reduce lead inductance and completes the shielding. Dummy antennas made in this way have good characteristics through the vhf bands as well as at all lower frequencies.

Increasing Power Ratings

More power can be handled by using a number of 2-watt resistors in parallel, or series-parallel, but at the expense of introducing some reactance. Nevertheless, if some departure from the ideal impedance characteristics can be tolerated this is a practical method for getting increased dissipations. The principal problem is stray inductance which can be minimized by mounting the resistors on flat copper strips or sheets, as suggested in Fig. 17-25.

The power rating on resistors is a *continuous* rating in free air. In practice, the maximum power dissipated can be increased in proportion to the reduction in duty cycle. Thus with keying, which has a duty cycle of about 1/2, the rating can be doubled. With sideband the duty cycle is usually not over about 1/3. The best way of judging is to feel the resistors occasionally; if too hot to touch, they may be dissipating more power than they are rated for.

THE OSCILLOSCOPE

The electrostatically deflected cathode-ray tube, with appropriate associated equipment, is capable of displaying both low- and radio-frequency signals on its fluorescent screen, in a form which lends itself to ready interpretation. (In contrast, the magnetically deflected television picture tube is not at all suitable for measurement purposes.) In the usual display presentation, the fluorescent spot moves across the screen horizontally at some known rate (horizontal deflection or horizontal sweep) and simultaneously is moved vertically by the signal voltage being examined (vertical deflection). Because of the retentivity of the screen and the eye, a rapidly deflected spot appears as a continuous line. Thus a varying signal voltage causes a pattern to appear on the screen.

Conventionally, oscilloscope circuits are designed so that in vertical deflection the spot moves upward as the signal voltage becomes more positive with respect to ground, and vice versa (there are exceptions, however). Also, the horizontal deflection is such that with an ac sweep voltage – the simplest form – positive is to the right; with a linear sweep – one which moves the spot at a uniform rate across the screen and then at the end of its travel snaps it back very quickly to the starting point – *time* progresses to the right.

Most cathode-ray tubes for oscilloscope work require a deflection amplitude of about 50 volts per inch. For displaying small signals, therefore, considerable amplification is needed. Also, special circuits have to be used for linear deflection. The design of amplifiers and linear deflection circuits is complicated, and extensive texts are available. For checking modulation of transmitters, a principal amateur use of the scope, quite simple circuits suffice. A 60-Hz voltage from the power line makes a satisfactory horizontal sweep, and the voltage required for vertical deflection can easily be obtained from transmitter rf circuits without amplification.

For general measurement purposes amplifiers and linear deflection circuits are needed. The most economical and satisfactory way to obtain a scope having these features is to assemble one of the many kits available.

Simple Oscilloscope Circuit

Fig. 17-26 is an oscilloscope circuit that has all the essentials for modulation monitoring: controls for centering, focusing, and adjusting the brightness of the fluorescent spot; voltage dividers to supply proper electrode potentials to the cathode-ray tube; and means for coupling the vertical and horizontal signals to the deflection plates.

The circuit can be used with electrostatic-deflection tubes from two to five inches in face diameter, with voltages up to 2500. Either set of deflecting electrodes (D1D2, or D3D4) may be used for either horizontal or vertical deflection, depending on how the tube is mounted.

In Fig. 17-26 the centering controls are not too high above electrical ground, so they do not need special insulation. However, the focusing and intensity controls are at a high voltage above ground and therefore should be carefully insulated. Insulated couplings or extension shafts should be used.

The tube should be protected from stray magnetic fields, either by enclosing it in an iron or steel box or by using one of the special CR tube

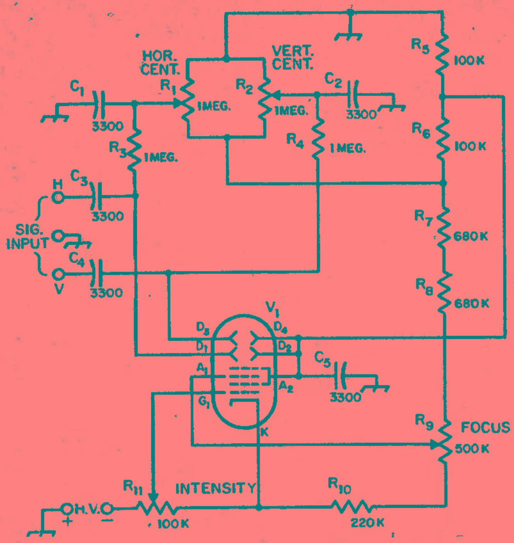


Fig. 17-26 — Oscilloscope circuit for modulation monitoring. Constants are for 1500- to 2500-volt high-voltage supply. For 1000 to 1500 volts, omit R8 and connect the bottom end of R7 to the top end of R9.

C1-C5, incl. — 1000-volt disk ceramic.

R1, R2, R9, R11 — Volume-control type, linear taper. R9 and R11 must be wall insulated from chassis.

R3, R4, R5, R6, R10 — 1/2 watt.

R7, R8 — 1 watt.

V1 — Electrostatic-deflection cathode-ray tube, 2- to 5-inch. 8ase connections and heater ratings vary with type chosen.

shields available. If the heater transformer (or other transformer) is mounted in the same cabinet, care must be used to place it so the stray field around it does not deflect the spot. The spot cannot be focused to a fine point when influenced by a transformer field. The heater transformer must be well insulated, and one side of the heater should be connected to the cathode. The high-voltage dc can be taken from the transmitter plate supply; the current required is negligible.

Methods for connecting the oscilloscope to a transmitter for checking or monitoring modulation are given in earlier chapters.

Quasi-Linear Sweep

For wave-envelope patterns that require a fairly linear horizontal sweep, Fig. 17-27 shows a method of using the substantially linear portion of the 60-Hz sine wave – the "center" portion where the wave goes through zero and reverses polarity. A 60-Hz transformer with a center-tapped secondary winding is required. The voltage should be sufficient to deflect the spot well off the screen on both sides – 250 to 350 volts, usually. With such "over-deflection" the sweep is fairly linear, but it is as bright on retrace as on left-to-right. To blank it in one direction, it is necessary to couple the ac to the No. 1 grid of the CR tube as shown.

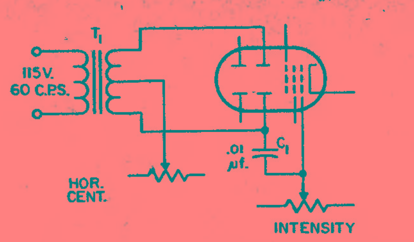


Fig. 17-27 — A quasi-linear time base for an oscilloscope can be obtained from the "center" portion of a sine wave. Coupling the ac to the grid gives intensity modulation that blanks the retrace.

C1 — Ceramic capacitor of adequate voltage rating.

T1 — 250- to 350-volt center-tapped secondary. If voltage is too high, use dropping resistor in primary side.

Lissajous Figures

When sinusoidal ac voltages are applied to both sets of deflecting plates in the oscilloscope the resultant pattern depends on the relative amplitudes, frequencies and phases of the two voltages. If the ratio between the two frequencies is constant and can be expressed in integers a stationary pattern will be produced.

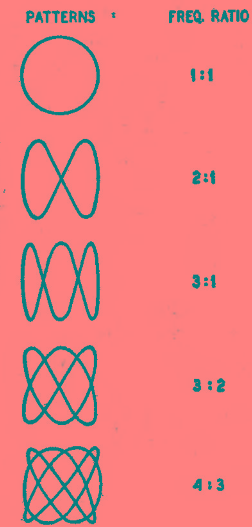


Fig. 17-28 — Lissajous figuras and corresponding frequency ratios for e 90-degrea phase relationship between the voltages applied to the two sets of deflecting plates.

The stationary patterns obtained in this way are called **Lissajous figures.** Examples of some of the simpler Lissajous figures are given in Fig. 17-28. The frequency ratio is found by counting the number of loops along two adjacent edges. Thus in the third figure from the top there are three loops along a horizontal edge and only one along the vertical, so the ratio of the vertical frequency to the horizontal frequency is 3 to 1. Similarly, in the fifth figure from the top there are four loops along the horizontal edge and three along the vertical edge, giving a ratio of 4 to 3. Assuming that the known frequency is applied to the horizontal plates, the unknown frequency is

$$f2 = \frac{n2}{n1} \; f1$$

where $f1$ = known frequency applied to horizontal plates,
$f2$ = unknown frequency applied to vertical plates,
$n1$ = number of loops along a vertical edge and,
$n2$ = number of loops along a horizontal edge

An important application of Lissajous figures is in the calibration of audio-frequency signal generators. For very low frequencies the 60-Hz power-line frequency is held accurately enough to be used as a standard in most localities. The medium audio-frequency range can be covered by comparison with the 440- and 600-Hz modulation on the WWV transmissions. It is possible to calibrate over a 10-to-1 range, both upwards and downwards, from each of the latter frequencies and thus cover the audio range useful for voice communication.

An oscilloscope having both horizontal and vertical amplifiers is desirable, since it is convenient to have a means for adjusting the voltages applied to the deflection plates to secure a suitable pattern size.

MARKER GENERATOR FOR 100, 50 AND 25 KHZ

The frequency generator in the accompanying illustrations will deliver marker signals of usable strength well into the vhf region when its output is connected to the antenna input terminals of a communications receiver. It uses a 100-kHz crystal in an integrated-circuit version of the solid-state multivibrator oscillator shown earlier. The oscillator is followed by a two-stage IC divider which produces 50- and 25-kHz marker intervals. Two inexpensive ICs are used, an MC-724P quad gate and an MC790P dual JK flip-flop. Two of the gates in the MC724P are used for the oscillator and a third serves as a following buffer amplifier and "squarer" for driving the first divide-by-2 circuit in the MC790P. This divider then drives the second divide-by-2 flip-flop. Outputs at the three frequencies are taken through a 3-position switch from taps as shown in the circuit diagram, Fig. 17-30.

Two of the three poles of the 4-position switch are used for controlling the collector voltage for the ICs. Voltage is on the MC724P in all active positions of the switch, but is applied to the MC790P only when 50- and 25-kHz markers are required. This saves battery power, since the MC790P takes considerably more current than the MC724P.

The outputs on all three frequencies are good square waves. To assure reasonably constant harmonic strength through the hf spectrum the output is coupled to the receiver through a small capacitance which tends to attenuate the lower-frequency harmonics. This capacitance, C3, is not critical as to value and may be varied to suit individual preferences. The value shown, 22 pF, is satisfactory for working into a receiver having an input impedance of 50 ohms.

At 3 volts dc input the current taken in the 100-kHz position of S1 is 8 mA. In the 50- and 25-kHz positions the total current (both ICs) is 35 mA. The generator continues to work satisfactorily when the voltage drops as low as 1.5 volts. The oscillator frequency is subject to change as the voltage is lowered, the frequency shift amounting to approximately 30 Hz at 15 MHz on going from 3 to 2 volts. There is a slight frequency shift between the 100-kHz and 50/25-kHz positions, but this amounts to only 6 or 7 Hz at 15 MHz.

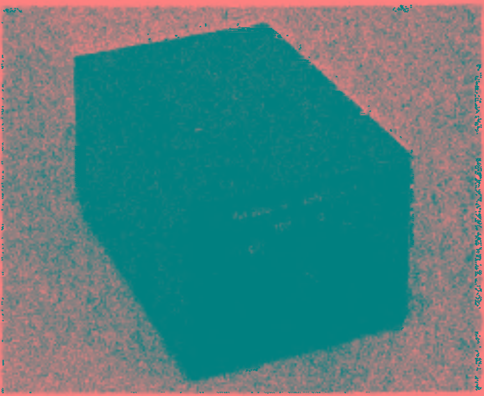
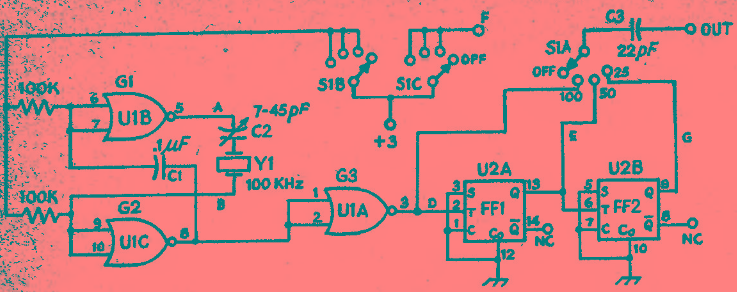


Fig. 17-29 — Frequency marker generating 100-, 50-, or 25-kHz intervals. Battery power supply (two "D" cells) is inside the cabinet, a 3 X 4 X 6-inch aluminum chassis with bottom plate. The trimmer capacitor for fine adjustment of frequency is available through the hole in the top near the left front.

Frequency changes resulting from temperature variations are larger; they may be as much as a few hundred Hz at 15 MHz in normal room-temperature variations. All such frequency changes can be compensated for by adjusting C2, and it is good practice to check the frequency occasionally against one of the WWV transmissions, readjusting C2 if necessary.

Layout and Construction

The physical layout of the circuit can be varied to suit the builder's tastes. The size of the box containing the generator shown in the photographs makes the batteries easily accessible for replacement. The method of mounting the crystal and C2 allows the latter to be reached through the top of the box for screwdriver adjustment, and makes possible the easy removal of the crystal since it plugs into a standard crystal socket. There is ample room for soldering the various wires that lead to the switch from the etched board on which the ICs, resistors, and C1 are mounted. The output



C1 – 0.1µF paper, low voltage.
C2 – 7-45-pF ceramic trimmer.
C3 – 22-pF dipped mica (ceramic also satisfactory).
S1 – 3-pola, 4-position rotary (Mallory 3134J).
U1 – Quad 2-input NOR gata, 1 section unused (Motorola MC724P).
U2 – Dual J-K flip-flop (Motorola MC790P).

Fig. 17-30 – Marker genarator circuit. Pin 4 of both ICs is grounded. Connect pin 11 of U1 to point C, and pin 11 of U2 to point F.

jack is placed at the rear where it is convenient when the unit is alongside a receiver.

An etched board does not have to be used for wiring the ICs and associated parts, although it makes for neatness in construction. The wiring plan used in this one is shown in Fig. 17-32. Fig. 17-32 is not a conventional template, but is a scale drawing showing how the etched connections can run with a minimum number of cross-over points where jumpers are required (only one is needed in this layout). In following the wiring plan the resist can be put on as desired, so long as the separation between conductors is great enough to prevent short-circuits.

Fig. 17-32 shows the front or component side of the board. To get the reversed drawing that would be followed on the copper side, place a piece of paper under the figure, with a face-up piece of carbon paper under it. Then trace the wiring with a sharp pencil and the layout will be transferred to the back of the paper. The points where holes are to be drilled are shown by small dots and circles, the latter indicating the points at which external connections are to be made.

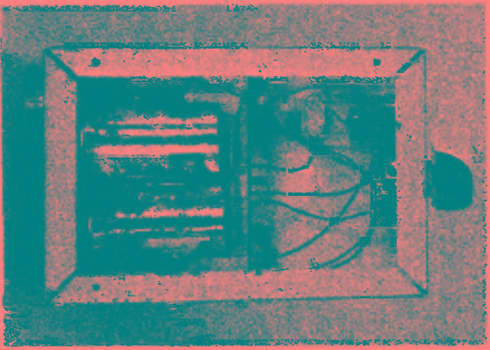


Fig. 17-31 – Integrated circuits and associated fixed capacitors and resistors are mounted on an etched broad measuring 3 3/4 X 2 1/2 inches, supported from one wall by an aluminum bracket. The 100-kHz crystal and trimmer capacitor ara on a 1 X 2-inch plestic strip supported below the top on 1/2-inch spacers, with the capacitor facing upward so it can be adjusted from outsida. The two dry cells ara in a dual holder (available from alectronics supply stores). The output connector is a phono jack, mounted on the rear wall (upper left in this view) with C3.

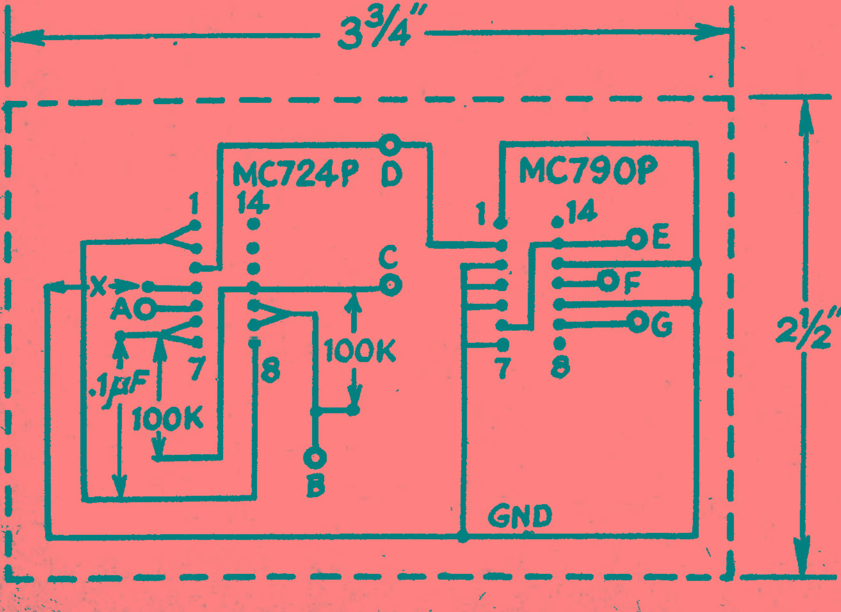


Fig. 17-32 – Wiring plan for the circuit board, componant side. Dimensions for placement of parts are exact. X – jumper. Other letters indicate external connection points, corresponding to similarly lettared connections in Fig. 17-30.

AN ELECTRONIC CLOCK

An Electronic Clock

The device described here will provide accurate time indication in either 12 hour or 24 hour service. Special effort was given to construct this model as compact as possible and accordingly the dimensions given below need not be followed. The chassis with the top cover in place measures (HWD) 2-3/4 × 4-1/4 × 5-1/2 inches. The top cover was fabricated using perforated aluminum stock. Since not much heat is generated by the internal components, a solid cover would be satisfactory. Screws are used to attach the cover to each angle bracket mounted on the chassis. The digital display is the only front-panel mounted component. Since the sheet metal used for the cabinet is a thin gauge, it was necessary to provide a shim under the bezel. Five switches (which are used for setting the time and number of digits displayed) are located along the top of the rear panel. The builder has the choice of 12- or 24-hour time, and the seconds

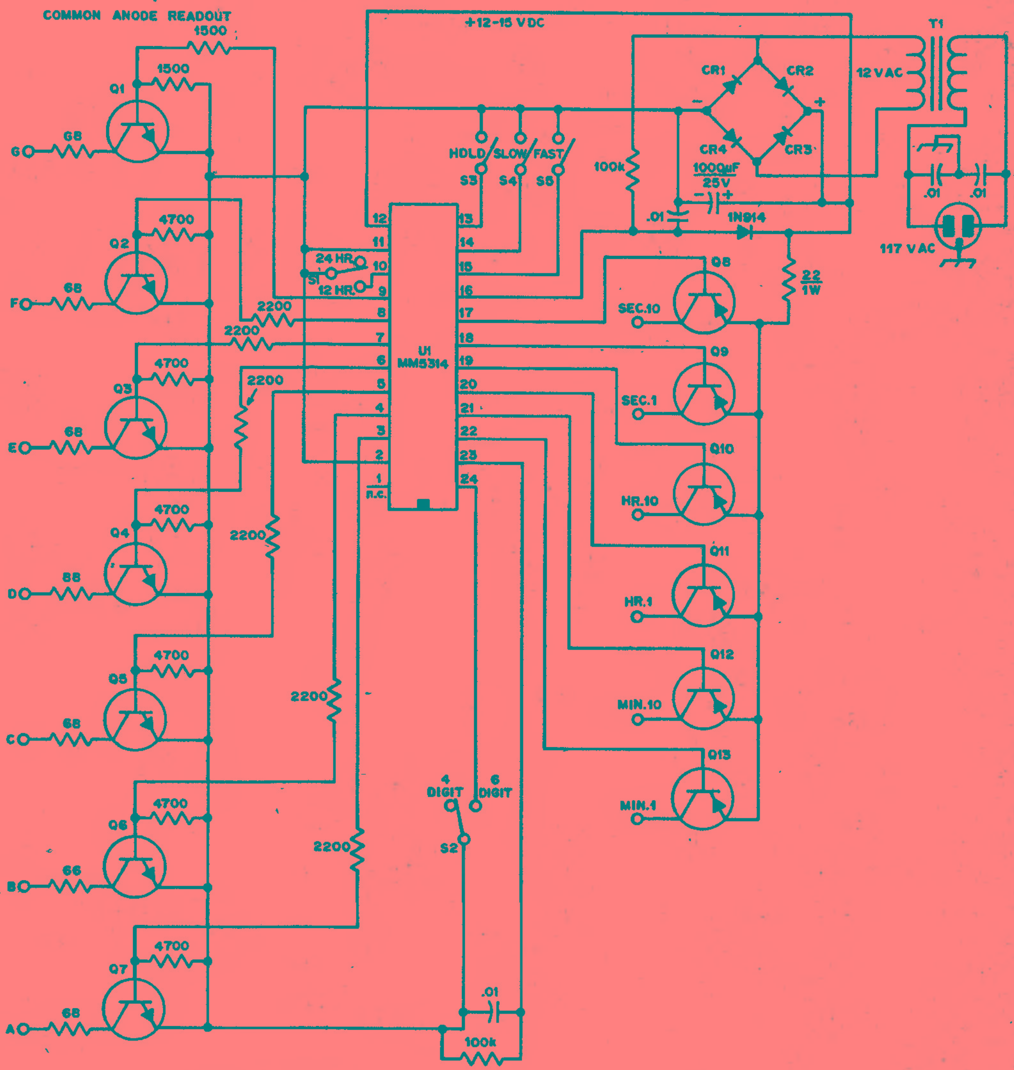


Fig. 1 — Circuit diagram for the common-anode configuration digital clock. Resistors are 1/4 or 1/2 watt composition; capacitors are disc ceramic. Readouts are Opcoa SLA-1. A-G letters correspond to segments on the readouts. S10, M10, and H10 correspond to tens of seconds, minutes, and hours respectively. S1, M1, and H1 correspond to single seconds, minutes, and hours respectively.
CR1-CR4, inc. — 50 PRV 1 ampere rectifiers.
Q1-Q7, inc. — 2N2222, or 2N3904, or HEP736.
Q8-Q13, inc. — 2N4403, or 2N2907A, or 2N3628, or HEP52.
T1 — 12 Vac, 500 mA transformer.
U1 — National Semiconductor MM5314.

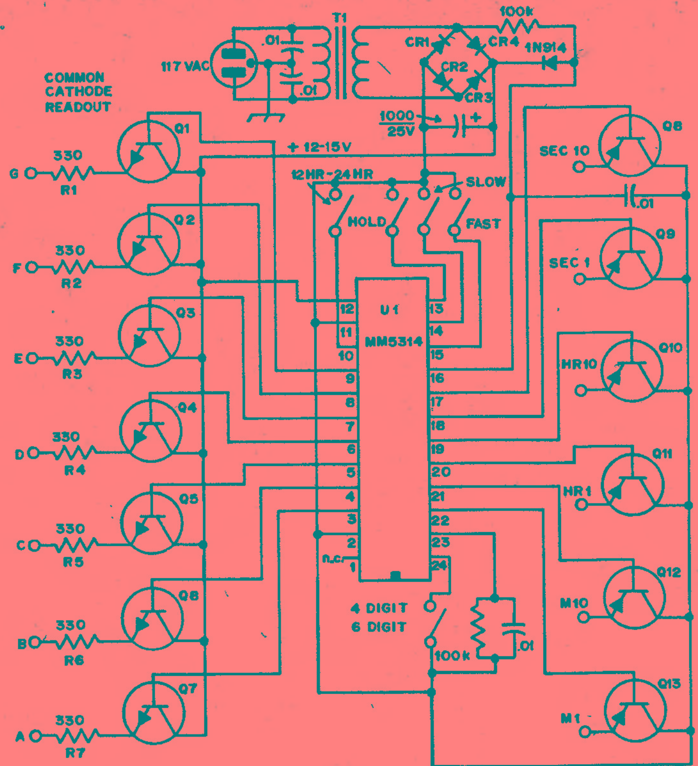


Fig. 2 — Circuit diagram for common-cathode configuration digital clock. Unless otherwise specified, resistors ere 1/4 or 1/2 watt composition and capacitors ara disc ceramic. Readouts are Radio Shack 276-055, or Litronix 704. A-G letters correspond to segments on the readouts. S10, M10 and H10 correspond to tans of seconds, minutes, and hours respectively. S1, M1, and H1 correspond to single seconds, minutes, and hours respectively.

CR1-CR4, inc. — 50 PRV 1 ampere rectifiers.
Q1-Q7, inc. — 2N2222, or 2N3904, or HEP736.
Q8-Q13, inc. — 2N4403, or 2N2907A, or 2N3638, or HEP52.
R1-R7, inc. — 330 ohms for Radio Shack 276-055 readouts, 100 ohms for Litronix 704 readouts.
T1 — 12 Vac, 500 mA transformer.
U1 — National Semiconductor MM5314.

digits may be suppressed (turned off) to eliminate display flicker and operator distraction. In the common-anode circuit the segments didn't always turn off properly. A 22-ohm resistor was added to the collector line of Q8-Q13 to correct the problem.

The display of readouts is mounted approximately 1/16 inch behind the lens of the bezel to maximize visibility. The mounting circuit board is supported by an L bracket secured to the chassis with a 3/4 inch spacer. All of the other components are mounted on a single circuit board which is positioned a half inch above the bottom floor of the cabinet. The power transformer is positioned directly behind the circuit board. As with all station equipment which is ac operated, a three-wire cord should be used. This is especially necessary when the input ac line is bypassed with a large value of capacitance.

Several different versions of this circuit can be used to provide features like battery operation, display shut down, and mobile operation. Modifications of this nature are described by Kelley in *QST* for November, 1974.

DIP METERS FOR THE HF-VHF-UHF RANGE

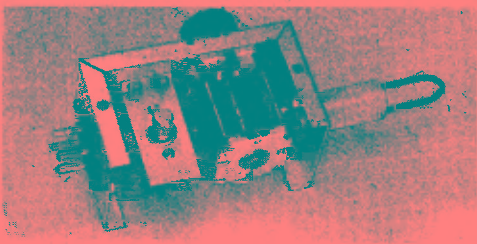


Fig. 17-37 — Dip meter covering the range 1.7 to 275 MHz, with the 90- to 165-MHz coil in place.

Figs. 17-37 through 17-41 show representative construction of vacuum-tube dip meters for the frequencies of interest to amateurs. Two separate designs are required to cover the lower frequencies along with the vhf-uhf range. The same power supply/meter unit serves for both. The 6CW4 Nuvistor triode is used in both meters.

Referring to the circuit in Fig. 17-38, a resistor, R2, is plugged in with each coil (the resistor is mounted in the coil form). It forms a voltage divider with the normal grid leak, R1, and brings the metering circuit into the best range for the transistor dc amplifier.

A small aluminum bracket supports the Nuvistor socket within the 2 1/4 X 2 1/4 X 4-inch Minibox that is used as a housing. A 5-pin socket (Amphenol 78-S5S) is mounted at one end of the Minibox, and the variable capacitor stator leads are soldered directly to two of the pins. Coils in the low-frequency ranges are wound with enameled wire on 3/4-inch diameter forms. In the intermediate ranges coil stock (B&W Miniductor) is mounted inside the coil forms, with one end of the coil close to the open end of the form, for ease in coupling. The two highest-range coils are hairpin loops of No. 14 wire, covered with insulation as a safety precaution. In every case the associated R2 is mounted in the coil form. The highest range requires that only the base of the coil form be used, since the loop is shorter than the form.

The power supply may be included with the oscillator, but since this increases the bulk and weight a separate supply is often desirable. The power supply shown in Fig. 17-40 uses a miniature power transformer with a silicon rectifier and a simple filter to give approximately 120 volts for the oscillator plate. It is also built in a 2 1/4 X 2 1/4 X 4-inch Minibox. The two Miniboxes are connected by a length of 4-conductor cable.

Either meter may be used as an indicating-type absorption wavemeter by removing the plate voltage and using the grid and cathode of the tube as a diode.

UHF Grid-Dip Oscillator

The range of the grid-dip meter shown in Fig. 17-39 is from 275 to 725 MHz, a higher range than most of the inexpensive meters now available. It is able to cover these high frequencies by virtue of the 6CW4 tube and the series-tuned circuit.

The uhf grid-dip meter is built in a 2 1/4 X 2 1/4 X 4-inch Minibox. The "heart" of the meter is the oscillator section, which is built on a 1 3/4 X 1 7/8-inch piece of 1/8-inch thick polystyrene. The Nuvistor socket is mounted in one corner and the tuning capacitor is mounted a little above center. The coil socket, a National CS-6, is mounted on the end of the Minibox. The polystyrene sheet is supported by four 1-inch 6-32 screws, and the sockets and variable capacitor are positioned so that direct connections can be made between plate pin and coil socket, capacitor rotor and coil socket, and capacitor stator and grid pin. The various resistors and rf chokes are supported at one end by a multiple-terminal tie strip mounted on the polystyrene sheet and at the other end by the socket pins and other terminals.

The coils are made from No. 10 tinned copper wire; as a safety precaution they are covered except at the tips by clear plastic insulation. Details are given in Fig. 17-41.

Frequency calibration of the meter can be started by reference to uhf TV stations in the area, if any, or by reference to 420-MHz amateur gear.

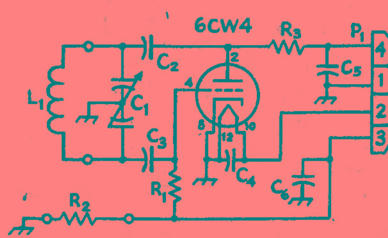


Fig. 17-38 — Circuit diagram of the hf-vhf dip meter.
C1 — 50 pF per section (Johnson 167-11 with stator bars sawed between 6th and 7th plates).
C2, C3 — 100-pF ceramic.
C4, C5, C6 — .001-μF disk ceramic.
P1 — 4-pin chassis plug (Amphenol 86-CP4).
R1 — 47,000 ohms, 1/2 watt.
R2 — See table below.
R3 — 10,000 ohms.

Range	*L1*	*R2*
1.7-3.2 MHz	195 turns No. 34 enam.*	680
2.7-5.0	110 turns No. 30 enam.*	470
4.4-7.8	51 1/2 turns No. 30 enam.*	470
7.5-13.2	24 1/2 turns No. 30 enam.*	470
12-22	31 t. No. 24 (B&W 3004)**	1000
20-36	14 t. No. 24 (B&W 3004)**	680
33-60	8 1/2 t. No. 20 (B&W 3003)***	680
54-99	3 3/4 t. No. 20 (B&W 3003)***	1000
90-165	3 3/8-inch loop No. 14 1/2-inch separation	1500
150-275	1 1/4-inch loop No. 14, 1/4-inch separation	3300

* Wound on 3/4-inch diameter polystyrene form.
**32 tpi ***16 tpi

Fig. 17-39 – Dip meter for the 300- to 700-MHz range. The oscilletor section is at the left in its own case, and the power supply plus trensistorized indicator is at the center and right. In the oscillator section, the 6CW4 (Nuvistor) socket is to the left of the tuning capacitor.

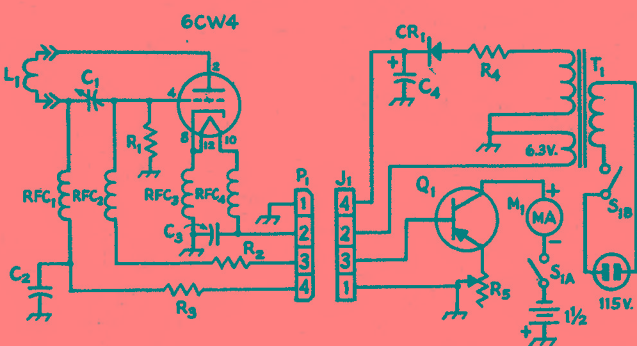


Fig. 17-40 – Circuit diagram of the uhf dip meter.

C1 – 8-pF midget variable (Hemmarlund MAC-10 with one rotor plete removed).
C2 – 150-pF ceremic.
C3 – .001-μF ceramic.
C4 – 20μF at 450-V electrolytic.
CR1– 400 PRV rectifier (Sarkes Tarzian 2F4).
J1 – 4-pin tube socket.
M1 – 0-500 microammeter.
P1 – 4-pin plug (Amphenol 86-CP4).
Q1 – Motorole HEP253 transistor.
R1 – 330 ohms, 1 watt.
R2 – 47,000 ohms, 1/2 watt.
R3 – 10,000 ohms.
R4 – 22 ohms, 1/2 watt.
R5 – 10,000-ohm potentiometer.
RFC1, RFC2 – 22-μH rf choke (Millen 34300-22).
RFC3, RFC4 – 0.82-μH rf choke (Millen 34300-.82).
S1A, S1B – Dpst, part of R5. Switches should be open when R5 is at maximum resistance.
T1 – 6.3- end 125-V transformer (Knight 61 G 410).

Range	*Dimension "L"*	*"M"*
271-324 MHz	2 3/4	11/16
312-378	3 1/8	–
372-463	2	–
413-519	1 5/8	–
446-565	1 1/4	–
544-730	1/2*	–

* Shape closed end to be nearly square.

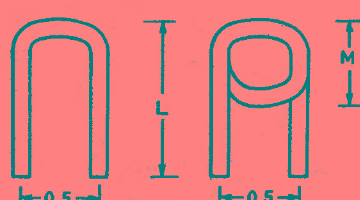


Fig. 17-41 – Details of the coils used in the uhf grid-dip meter. The material is No. 10 tinned-copper wire. One turn in end of low-frequency coil.

A CALIBRATED FIELD STRENGTH METER

There are many occasions when it is desirable to determine the relative performance of an antenna. While near-field pattern measurements are generally not accurate, they do show trends in terms of front-to-back ratio and may be used to determine what adjustments, if any, should be made to an existing system. The field-strength meter described here will detect large as well as small changes in radiated power from an antenna. For instance, the pattern of an hf-band Yagi may be checked by placing the meter and an associated sampling antenna several hundred feet from the beam. A watt or two of power is needed to make tests above 21 MHz, but for frequencies below this point, a grid-dip oscillator may serve as a "transmitter."

Fig. 1 gives the circuit diagram of the calibrated field-strength meter. L1 and L2 are resonated to the desired frequency with C1 to tune the hf bands. Adjustment is made to produce maximum meter deflection of the signal being sampled. Should the signal cause the meter to deflect off scale, the attenuator, R4, may be reset to reduce the level of the incoming energy.

Two operational amplifiers comprise a logarithmic circuit which produces a voltage output at pin 10 of U1B that is proportional to the logarithm (thus dB) of the input voltage. Forward bias is applied to CR1 via a 1-megohm resistor to improve conductivity at low signal input values. The output voltage from U1B is displayed by M1, a conventional milliammeter. Two scale ranges are available, 20 dB and 40 dB. With no signal applied, a small amount of quiescent current will appear on

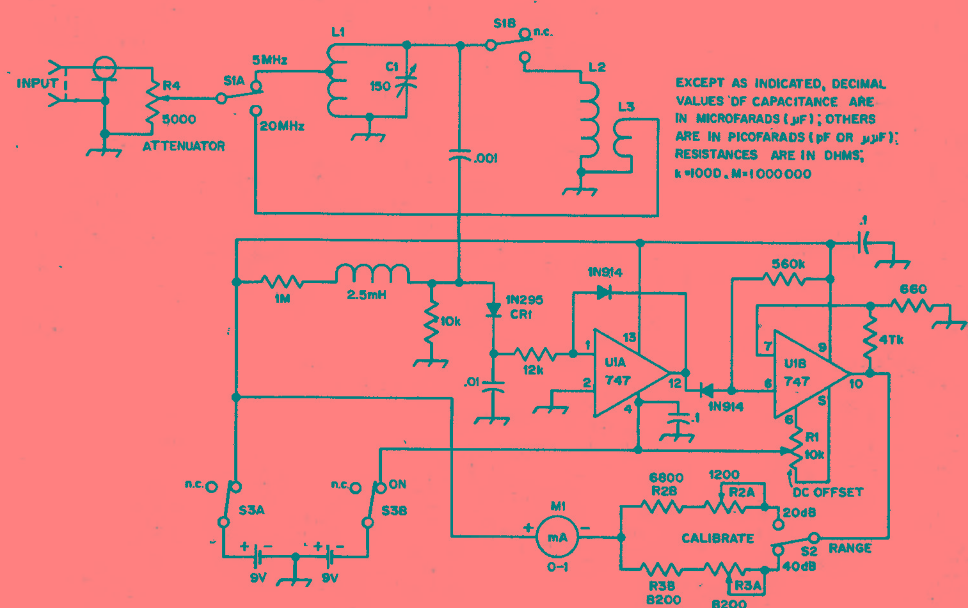


Fig. 1 — Circuit diagram for the calibrated field strength meter. Component designations not listed below are for taxt reference.

C1 — Variable capacitor, 140 pF maximum.
L1 — 44 turns of No. 24 enam. on a T-68-2 core tapped four turns from the ground end.
L2 — 15 turns of No. 24 enam on e T-68-2 core.
L3 — Two turns of No. 24 enam. wound over L2.
U1 — Dual 747 operational amplifier.
S1 — Dpdt rotary.
S2, S3 — Miniature toggle.

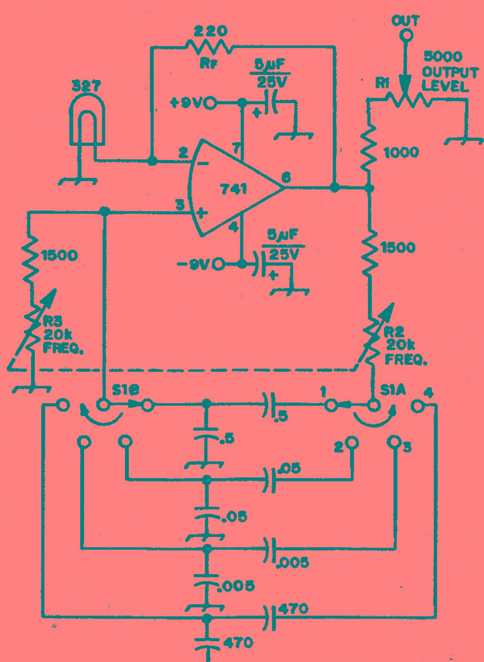
Inside view of the field strength meter. Most of the components are mounted on a circuit board.

M1. Readings made near this level will not be quite as accurate as those made in the upper portion of the scale. Accuracy is within one dB. About 1000 microvolts of signal is necessary to provide a meaningful movement of M1. R1 is the dc offset control and is mounted on the rear panel. It permits some variation of the absolute readings by shifting the dc levels at the output of U1B and may be used to set the meter to some convenient reference mark. The combined values of R2A and R2B should be 8000 ohms. R2A is a trim pot to allow proper adjustment to exactly that value. R3A and R3B serve a similar purpose and should be set for a total resistance of 16 kΩ.

AN AUDIO OSCILLATOR

A wide-range audio oscillator that will provide a moderate output level can be built from a single 741 operational amplifier (Fig. 1). Power is supplied by two nine-volt batteries, from which the circuit draws 4 mA. The frequency range is selectable from 15 Hz to 150 kHz, although a 1.5- to 15-Hz range can be included with the addition of two 5-μF nonpolarized capacitors and an extra switch position. Distortion is approximately one percent. The output level under a light load (10 kΩ) is 4 to 5 volts. This can be increased by using higher battery voltages, up to a maximum of plus and minus 18 volts, with a corresponding adjustment of R_f.

Pin connections shown are for the TO-5 case. If another package configuration is used, the pin connections may be different. R_f (220 Ω) is trimmed for an output level about five percent below clipping. This should be done for the temperature at which the oscillator will normally operate, as the lamp is sensitive to ambient temperature. Note that the output of this oscillator is direct coupled. If you are connecting this unit into circuits where dc voltage is present, use a coupling capacitor. As with any solid-state equipment, be cautious around plate circuits of tube-type equipment, as the voltage spike caused by charging a coupling capacitor may destroy the IC. This unit was originally described by Schultz in *QST* for November, 1974.



SWITCH 1	
POS.	FREQ.
1	15–150Hz
2	150–1500Hz
3	1500Hz–15kHz
4	15-150kHz

EXCEPT AS INDICATED, DECIMAL VALUES OF CAPACITANCE ARE IN MICROFARADS (μF); OTHERS ARE IN PICOFARADS (pF OR μμF); RESISTANCES ARE IN OHMS; k=1000, M=1000 000

Fig. 1 — A simple audio oscillator that provides a selectable frequency range. R2 and R3 control the frequency and R1 varies the output level.

A TESTER FOR FET AND BIPOLAR TRANSISTORS

The circuit shown is intended solely as a tester for npn and pnp transistors, junction FETs, and dual-gate MOSETs. This equipment is not for use in checking audio or high-power rf transistors.

The circuit of Fig. 1 is an oscillator which is wired so that it will test various small-signal transistors by switching the battery polarity and bias voltage. A crystal for the upper range of the hf spectrum is wired into the circuit permanently, but could be installed in a crystal socket if the builder so desires. A 20-MHz crystal was chosen for this model. Any hf crystal cut for fundamental mode operation can be used.

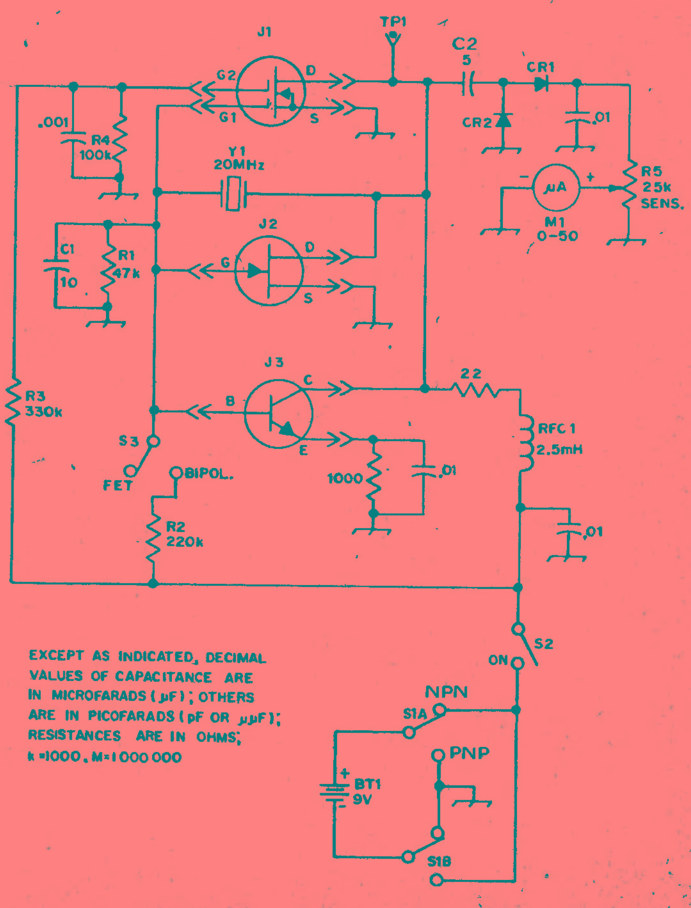
When testing FETs the bias switch, S3, is placed in the FET position, thus removing R2 from the circuit. However, when testing bipolar transistors the switch position must be changed to BIPOL so that forward bias can be applied to the base of the bipolar transistor under test. R1 is always in the circuit, and serves as a gate-leak resistor for FETs being evaluated. It becomes part of the bias network when bipolars are under test. C1 is used for feedback in combination with the internal capacitances of the transistors being checked. Its value may have to be changed experimentally if crystals for lower frequencies are utilized in the circuit. Generally speaking, the lower the crystal frequency, the greater the amount of capacitance needed to assure oscillation. Use only that amount necessary to provide quick starting of the oscillator.

Components R3 and R4 are used as a voltage divider to provide bias for dual-gate MOSFETs. C2 is kept small in value to minimize loading of the oscillator by the low-impedance voltage doubler, CR1 and CR2. Rectified rf from the oscillator is monitored on M1. Meter deflection is regulated manually by means of control R5. S1 is used to select the desired supply voltage polarity – negative ground for testing n-channel FETs and npn bipolars, and a positive ground when working with p-channel and pnp devices.

When testing MOSFETs that are not gate protected (3N140 for one), make certain that the transistor leads are shorted together until the device is seated in the test socket. Static charges on one's hands can be sufficiently great to damage the insulation within the transistor. Use a single strand of wire from some No. 22 or 24 stranded hookup wire, wrapping it two or three times around the pigtails of the FET as close to the transistor body as possible. After the FET is plugged into the

Fig. 1 – Schematic diagram of the transistor tester. Capacitors ara disk ceramic or mica. Resistors ara 1/2 or 1/4-watt composition exeect for R5. Estimated cost for this tester (all parts new) is $15. Numbered components not appearing in parts list ara so designated for text discussion.

BT1 – Small 9-V transistor-radio battery.
CR1, CR2 – 1N34A germanium diode or equiv.
J1 – Four-terminal transistor socket.
J2, J3 – Three-terminal transistor socket.
M1 – Microampere meter. Calectro D1-910 used here.
R5 – 25,000-ohm linear-taper composition control with switch.
RFC1 2.5-mH rf choke.
S1 – Two-pole doubla-throw miniature toggle.
S2 – Part of R5.
S3 – Spst miniature toggle.
Y1 – Surplus crystel (see tex).



socket, unwrap the wire and perform the tests. (It's not a bad idea to have an earth ground connected to the case of the tester when checking unprotected FETs.) Put the shorting wire back on the FET leads before removing the unit from the tester.

The meter indication is significant in checking any type of transistor. If the device is open, shorted, or extremely leaky, no oscillation will take place, and the meter will not deflect. The higher the meter reading, the greater the vigor of the transistor at the operating frequency. High meter readings suggest that the transistor is made for vhf or uhf service, and that its beta is medium to high. Lower readings may indicate that the transistor is designed for hf use, or that it has very low gain. Transistors that are known to be good but will not cause the circuit to oscillate are most likely made for low-frequency or audio applications.

A TESTER FOR CRYSTALS AND BIPOLAR TRANSISTORS

The circuit of Fig. 1 is intended primarily to test surplus crystals and bipolar transistors. It uses a Pierce oscillator. Battery polarity can be switched to allow testing of npn or pnp transistors. Crystal quality is indicated on M1. The greater the crystal activity, the higher the meter reading. A suitable transistor for use at Q1 (when testing crystals) is the 2N4124, MPS3563, or HEP53. All three have f_T ratings well into the vhf spectrum, and each has reasonably high beta. The two characteristics make the devices ideal as general-purpose oscillators.

This tester will work well from the upper hf range down to at least 455 kHz. S1 is used to change the value of feedback capacitance. The lower the frequency of operation, the greater the amount of capacitance required.

A transistor can be checked by plugging the unknown type into the panel socket while using a crystal of known frequency and condition. Both testers can be used as calibrators by inserting

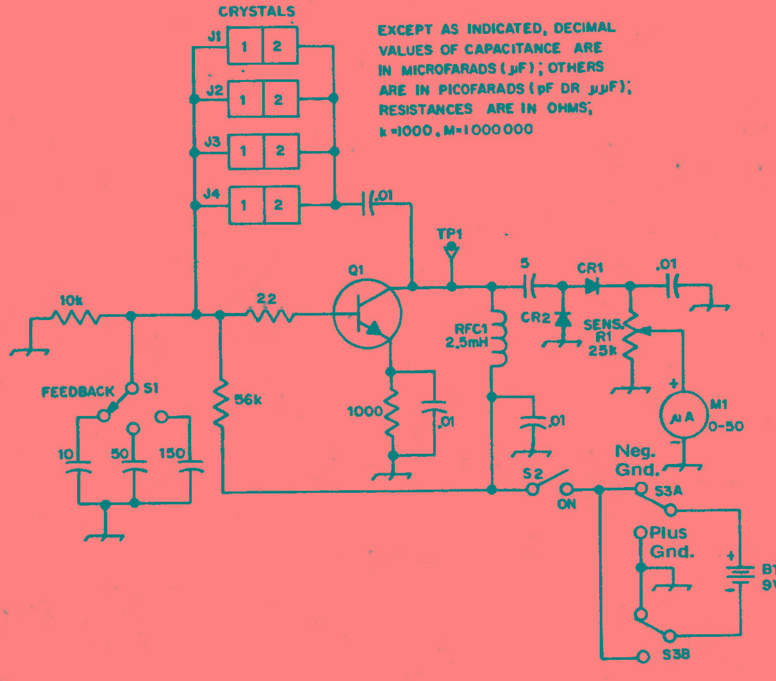


Fig. 1 — Schematic diagram of the No. 2 tester. Capacitors are disk ceramic. Fixed-value resistors are 1/2 or 1/4-watt composition. Estimated cost for this tester (all new parts) is $13.

BT1 — Smell 9-V trensistor-radio battery.
CR1, CR2 — 1N34A germanium diode or equiv.
J1-J4, incl. — Crystal socket of builder's choice.
M1 —Microampere meter. Calectro D1-910 used here.
R1 — 25,000-ohm linear-taper composition control with switch.
RFC1 — 2.5-mH rf choke.
S1 — Single-pole three-position phenolic rotary wafer type, miniature.
S2 — Part of R1.
S3 — Double-pole double-throw miniature toggle.
Q1 — Vhf npn bipolar, 2N4124, MPS3563, HEP53.

crystals for band-edge checking. The frequencies of unknown crystals can be checked by listening to the output from the test oscillators on a calibrated receiver or while using a frequency counter connected to the designated test point

Four crystal sockets are provided in the model shown here. J1 through J4 provide for testing of FT-243, HC-6/U, HC-17, and HC-25 crystals, the most popular holder styles in use today. Other types can be added by the builder if desired.

DIODE NOISE GENERATORS

A noise generator is a device for creating a controllable amount of rf noise ("hiss"-type noise) evenly distributed throughout the spectrum of interest. The simplest type of noise generator is a diode, either vacuum-tube or crystal, with dc flowing through it. The current is also made to flow through a load resistance which usually is chosen to equal the characteristic impedance of the transmission line to be connected to the receiver's input terminals. The resistance then substitutes for the line, and the amount of rf noise fed to receiver input is controlled by varying the dc through the diode.

The noise generator is useful for adjusting the "front-end" circuits of a receiver for best noise figure. A simple circuit using a crystal diode is shown in Fig. 17-51. The unit can be built into a small metal box; the main consideration is that the circuit from C1 through P1 be as compact as possible. A calibrated knob on R1 will permit resetting the generator to roughly the same spot each time, for making comparisons. If the leads are short, the generator can be used through the 144-MHz band for receiver comparisons.

To use the generator, screw the coaxial plug onto the receiver's input fitting, open S1, and measure the noise output of the receiver by connecting an audio-frequency voltmeter to the receiver's af output terminals. An average-reading voltmeter is preferable to the peak-reading type, since on this type of noise the average-reading meter will give a fair approximation of rms, and the object is to measure noise *power*, not voltage.

In using the generator for adjusting the input circuit of a receiver for optimum noise figure, first make sure that the receiver's rf and af gain controls are set well within the linear range of response, and turn off the automatic gain control. With the noise generator connected but S1 open, adjust the receiver gain controls for an output reading that is far enough below the maximum obtainable to ensure that the receiver is operating linearly. This is your reference level of noise. Then close S1 and adjust R1 for a readily perceptible increase in output. Note the *ratio* of the two readings – i.e., the number of dB increase in noise when the generator is on. Then make experimental adjustments of the receiver input coupling, always with the object of obtaining the largest number of dB increase in output when the generator is switched on.

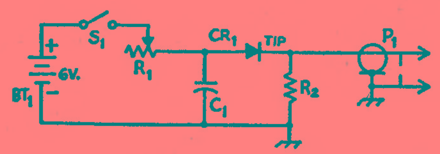


Fig. 17-51 – Circuit of a simple crystal-diode noise generator.
BT1 – Dry-cell battery, any convenient type.
C1 – 500-pF ceramic, disk or tubular.
CR1 – Silicon diode, 1N21 or 1N23. Diodes with "R" suffix have reversed polarity. (Do not use ordinary germanium diodes.)
P1 – Coaxial fitting, cable type.
R1 – 50,000-ohm control, ccw logarithmic taper.
R2 – 51 or 75 ohms, 1/2-watt composition.
S1 – Spst toggle (may be mounted on R1).

A simple crystal-diode noise generator is a useful device for the receiver adjustment, especially at vhf, and for comparing the performance of different receivers checked with the same instrument. It does not permit actual measurement of the noise figure, however, and therefore the results with one instrument cannot readily be compared with the readings obtained with another. In order to get a quantitative measure of noise figure it is necessary to use a temperature-saturated vacuum diode in place of the semiconductor diode. Suitable diodes are difficult to find.

RF PROBE FOR ELECTRONIC VOLTMETERS

The rf probe shown in Figs. 17-52 to 17-55, inclusive, uses the circuit discussed earlier in connection with Fig. 17-15.

The isolation capacitor, C1, crystal diode, and filter/divider resistor are mounted on a bakelite 5-lug terminal strip, as shown in Fig. 17-55. One end lug should be rotated 90 degrees so that it extends off the end of the strip. All other lugs should be cut off flush with the edge of the strip. Where the inner conductor connects to the terminal lug, unravel the shield three-quarters of an inch, slip a piece of spaghetti over it, and then solder the braid to the ground lug on the terminal strip. Remove the spring from the tube shield, slide it over the cable, and crimp it to the remaining quarter inch of shield braid. Solder both the spring and a 12-inch length of flexible copper braid to the shield.

Next, cut off the pins on a seven-pin miniature shield-base tube socket. Use a socket with a cylindrical center post. Crimp the terminal lug previously bent out at the end of the strip and insert it into the center post of the tube socket from the top. Insert the end of a phone tip or a

Fig. 17-52 — Rf probe for use with an electronic voltmeter. The case of the probe is constructed from a 7-pin ceramic tube socket and a 2 1/4-inch tube shield. A half-inch grommet at the top of the tube shield prevents the output lead from chafing. A flexibla copper-braid grounding lead and alligator clip provide a low-inductance return path from the test circuit.

pointed piece of heavy wire into the bottom of the tube socket center post, and solder the lug and tip to the center post. Insert a half-inch grommet at the top of the tube shield, and slide the shield over the cable and flexible braid down onto the tube socket. The spring should make good contact with the tube shield to insure that the tube shield (probe case) is grounded. Solder an alligator clip to the other end of the flexible braid and mount a phone plug on the free end of the shielded wire.

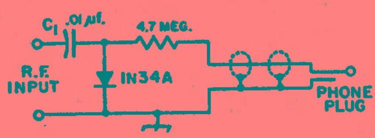


Fig. 17-53 — The rf probe circuit.

Fig. 17-54 — Inside the probe. The 1N34A diode, calibrating resistor, and input capacitor are mounted tight to the terminal strip with shortest leads possible. Spaghetti tubing is placed on the diode leads to prevent accidental short circuits. The tube-shield spring and flexible-copper grounding lead are solderad to the cable braid (the cable is RG-58/U coax). The tip can be either a phone tip or a short pointed piece of heavy wire.

Mount components close to the terminal strip, to keep lead lengths as short as possible and minimize stray capacitance. Use spaghetti over all wires to prevent accidental shorts.

The phone plug on the probe cable plugs into the dc input jack of the electronic voltmeter and rms voltages are read on the voltmeter's negative dc scale.

The accuracy of the probe is within ±10 percent from 50 kHz to 250 MHz. The approximate input impedance is 6000 ohms shunted by 1.75 pF (at 200 MHz).

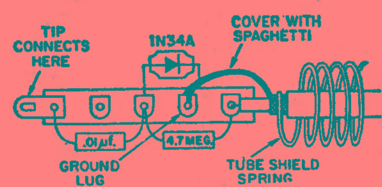


Fig. 17-55 — Component mounting details.

RF IMPEDANCE BRIDGE FOR COAX LINES

The bridge shown in Figs. 1 through 3 may be used to measure unknown complex impedances at frequencies below 30 MHz. Measured values are of equivalent series form, $R + jX$. The useful range of the instrument is from about 5 to 400 ohms if the unknown load is purely resistive, or 10 to 150 ohms resistive component in the presence of reactance. The reactance range is from 0 to approximately 100 ohms for either inductive or capacitive loads. Although the instrument cannot indicate impedances with the accuracy of a laboratory type of bridge, its readings are quite adequate for the measurement and adjustment of antenna systems for amateur use, including the taking of line lengths into account with a Smith chart or Smith transmission-line calculator.

The bridge incorporates a differential capacitor, C1, to obtain an adjustable ratio for measurement of the resistive component of the load. The capacitor consists of two identical sections on the same frame, arranged so that when the shaft is rotated to increase the capacitance of one section, the capacitance of the other section decreases. The capacitor is adjusted for a null reading on M1, and its settings are calibrated in terms of resistance at J3 so the unknown value can be read off the calibration. A coil-and-capacitor combination is used to determine the amount and type of reactance, inductive or capacitive. L1 and C2 in the bridge circuit are connected in series with the load. The instrument is initially balanced at the frequency of measurement with a purely resistive load connected at J3, so that the reactances of L1 and of C2 at its midsetting are equal. Thus, these reactances cancel each other in this arm of the bridge. With an unknown complex-impedance load then connected at J3, the setting of C2 is varied either to increase or decrease the capacitive reactance, as required, to cancel any reactance present in the load. If the load is inductive more capacitive reactance is required from C2 to obtain a balance, indicated by a null on M1, with less reactance needed from C2 if the load is capacitive. The settings of C2 are calibrated in terms of the value and type of reactance at J3. Because of the relationship of capacitive reactance to frequency, the calibration for the dial of C2 is valid at only one frequency. It is therefore convenient to calibrate this dial for equivalent reactances at 1 MHz, as shown in Fig. 4. Frequency corrections may then be made simply by dividing the reactance dial reading by the measurement frequency in megahertz.

Construction

In any rf-bridge type of instrument, the leads must be kept as short as possible to reduce stray reactances. Placement of component parts, while not critical, must be such that lead lengths greater than about 1/2 inch (except in the dc metering circuit) are avoided. Shorter leads are desirable, especially for R1, the "standard" resistor for the bridge. In the unit photographed, the body of this resistor just fits between the terminals of C1 and J2 where it is connected. C1 should be enclosed in a shield and connections made with leads passing through holes drilled through the shield wall. The frames of both variable capacitors, C1 and C2, must be insulated from the chassis, with insulated couplings used on the shafts. The capacitor specified for C1 has provisions for insulated mounting. C2 is mounted on 1-inch ceramic insulating pillars.

Band-switching arrangements for L1 complicate the construction and contribute to stray reactances in the bridge circuit. For these reasons plug-in coils are used at L1, one coil for each band over which the instrument is used. The coils must be adjustable, to permit initial balancing of the bridge with C2 set at the zero-reactance calibration point. Coil data are given in Table I. Millen 45004 coil forms with the coils supported inside provide a convenient method of constructing these slug-tuned plug-in coils. A phenolic washer cut to the proper diameter is epoxied to the top or open end of each form, giving a rigid support for mounting of the coil by its bushing. Small knobs for 1/8-inch shafts, threaded with a No. 6-32 tap, are screwed onto the coil slug-tuning screws to permit ease of adjustment without a tuning tool. Knobs with setscrews should be used to prevent slipping. A ceramic socket to mate with the pins of the coil form is used for J2.

Calibration

The resistance dial of the bridge may be calibrated by using a number of 1/2- or 1-watt 5-percent-tolerance composition resistors of different values in the 5- to 400-ohm range as loads. For this calibration, the appropriate frequency coil

Fig. 1 — An *RCL* bridge for measuring unknown values of complex impedances. A plug-in coil is used for each frequency band. The bridge operates et an rf input level of about 5 volts; pickup-link assemblies for use with a grid-dip oscillator are shown. Before meesurements ara made, the bridge must be balanced with a nonreactive load connected at its measurament terminals. This load consists of a resistor mounted inside a coaxial plug, shown in front of the instrument at the left. The aluminum box measures 4 1/4 X 10 3/4 X 6 1/8 inches and is fitted with a carrying handle on the left and and self-sticking rubber feet on tha right end and bottom. Dials are Millen No. 10009 with skirts reversed and calibrations added.

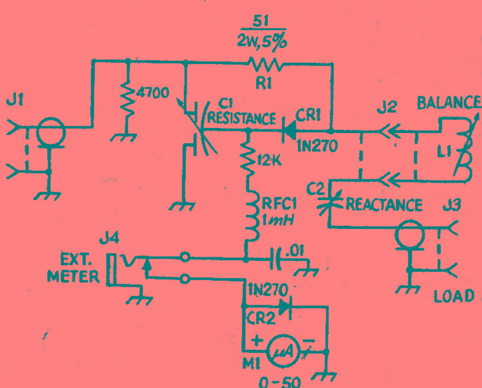


Fig. 2 — Schematic diagram of the impedance bridge. Capacitance is in microfarads; resistances are in ohms. Resistors are 1/2-W 10-percent tolerance unless otherwise indicated.

C1 — Differantial capacitor, 11-161 pF per section (Millen 28801).
C2 — 17.5-327 pF with straight-line capacitance charecteristic (Hammarlund RMC-325-S).
CR1, CR2 — Germanium diode, high back resistance.
J1, J3 — Coaxial connectors, chassis type.
J2 — To mate plug of L1, ceramic.
J4 — Phone jack, disconnecting type.
L1 — See taxt and Tabla I.
M1 — 0-50 μA dc (Simpson Model 1223 Bold-Vue, Cat. No. 15560 or equiv.).
R1 — For text reference.
RFC1 — Subminiature rf choka (Miller 70F103AI or equiv.).

TABLE 17-I

Coil Data for RF Impedance Bridge

Band	*Nominal Inductance Range, μH*	*Frequency Coverage, MHz*	*Coil Type or Data*
80	6.5–13.8	3.2–4.8	28 turns No. 30 enam. wire close-wound on Miller form 42A000CBI.
40	2.0–4.4	5.8–8.5	Miller 42A336CBI or 16 turns No. 22 enam. wire close-wound on Miller form 42A000CBI.
20	0.6–1.1	11.5–16.6	8 turns No. 18 enam. wire close-wound on Miller form 42A000CBI.
15	0.3–0.48	18.5–23.5	4 1/2 turns No. 18 enam. wire close-wound on Miller form 42A000CBI.
10	0.18–0.28	25.8–32.0	3 turns No. 16 or 18 enam. or tinned bus wire spaced over 1/4-inch winding length on Miller form 42A000CBI.

must be inserted at J2 and its inductance adjusted for the best null reading on the meter when C2 is set with its plates half meshed. For each test resistor, C1 is then adjusted for a null reading. Alternate adjustment of L1 and C1 should be made for a complete null. The leads between the test resistor and J3 should be as short as possible, and the calibration preferably should be done in the 3.5-MHz band where stray inductance and capacitance will have the least effect.

If the constructional layout of the bridge closely follows that shown in the photographs, the calibration scale of Fig. 4 may be used for the reactance dial. This calibration was obtained by connecting various reactances, measured on a laboratory bridge, in series with a 47-ohm 1-W resistor connected at J3. The scale is applied so that maximum capacitive reactance is indicated with C2 fully meshed. If it is desired to obtain an individual calibration for C2, known values of inductance and capacitance may be used in series with a fixed resistor of the same approximate value as R1. For this calibration it is *very important* to keep the leads to the test components as short as possible, and calibration should be performed in the 3.5-MHz range to minimize the effects of stray reactances. Begin the calibration by setting C2 at half mesh, marking this point as 0 ohms reactance.

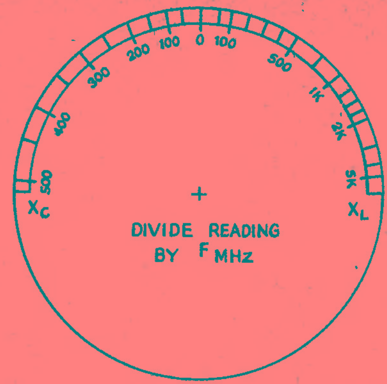
With a purely resistive load connected at J3, adjust L1 and C1 for the best null on M1. From this point on during calibration, do not adjust L1 except to rebalance the bridge for a new calibration frequency. The ohmic value of the known reactance for the frequency of calibration is multiplied by the frequency in MHz to obtain the calibration value for the dial.

Using the Impedance Bridge

This instrument is a low-input-power device, and is *not* of the type to be excited from a transmitter or left in the antenna line during station operation. Sufficient sensitivity for all measurements results when a 5-V rms rf signal is applied at J1. This amount of voltage can be delivered by most grid-dip oscillators. In no case should the power applied to J1 exceed 1 watt or calibration inaccuracy may result from a permanent change in the value of R1. The input impedance of the bridge at J1 is low, in the order of 50 to 100 ohms, so it is convenient to excite the bridge through a length of 52- or 75-ohm line such as RG-58/U or RG-59/U. If a grid-dip oscillator is used, a link coupling arrangement to the oscillator coil may be used. Fig. 1 shows two pick-up link assemblies. The larger coil, 10 turns of 1 1/4-inch-dia stock with turns spaced at 8 turns per inch, is used for the 80-, 40- and 20-meter bands. The smaller coil, 5 turns of 1-inch-dia stock with turns spaced at 4 turns per inch, is used for the 15- and 10-meter bands. Coupling to the oscillator should be as light as possible, while obtaining sufficient

Fig. 3 — All components except the meter are mounted on the top of the box. C1 is visible inside the shield at the left, with C2 at the right and J2 mounted between them. J1 is hidden beneath C1 in this view; a part of J3 may be seen in the lower right corner of the box. Components for the dc metering circuit are mounted on a tie-point strip which is affixed to the shield wall for C1; all other components are interconnected with very short leads. The 4700-ohm input resistor is connected across J1.

Fig. 4 — Calibration scale for the reactence dial associated with C2. See text.



sensitivity, to prevent severe "pulling" of the oscillator frequency.

Before measurements are made, it is necessary to balance the bridge. Set the reactance dial at zero and adjust L1 and C1 for a null with a nonreactive load connected at J3. The bridge must be rebalanced after any appreciable change is made in the measurement frequency. A 51-ohm 1-W resistor mounted inside a PL-259 plug, as shown in Fig. 17-24, makes a load which is essentially nonreactive. After the bridge is balanced, connect the unknown load to J3, and alternately adjust C1 and C2 for the best null.

The calibration of the reactance dial is shown in Fig. 4. The measurement range for capacitive loads may be extended by "zeroing" the reactance dial at some value other than 0. For example, if the bridge is initially balanced with the reactance dial set at 500 in the X_L range, the 0 dial indication is now equivalent to an X_C reading of 500, and the total range of measurement for X_C has been extended to 1000.

A LOW-POWER RF WATTMETER

The wattmeter shown in Fig. 1 can be used with transmitters having power outputs from 1- to 25-watts within the frequency range of 1.8 to 30 MHz. For complete details, see *QST* for June, 1973. A bridge circuit based on a version of the one shown in Fig. 17-16C is used to measure the forward and reflected power on a transmission line.

The rf wattmeter.

It will be necessary to have a nonreactive 50-ohm dummy load for initial adjustment of the power meter. Connect the dummy load to one port of the instrument and apply rf power to the remaining port. S1 should now be thrown back and forth to determine which position gives the highest meter reading. This will be the FORWARD position. Adjust the sensitivity control for full-scale reading of the meter. Now, move the switch to the opposite (RELFECTED) position and adjust the trimmer nearest the transmitter input port for a null in the meter reading. The needle should drop to zero. It is recommended that these adjustments be made in the 10- or 15-meter band. Next, reverse the transmitter and load cables and repeat the nulling procedure while adjusting the trimmer on the opposite side of the pc board. Repeat these steps until a perfect null is obtained in both directions. The switch and the coax connectors can now be labeled, TRANSMITTER, LOAD, FORWARD, and REFLECTED, as appropriate.

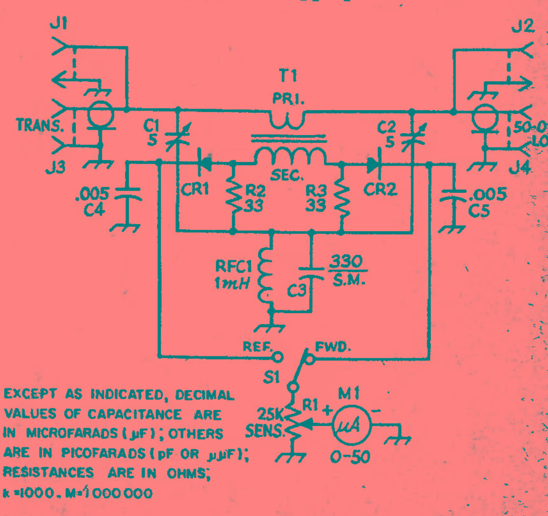
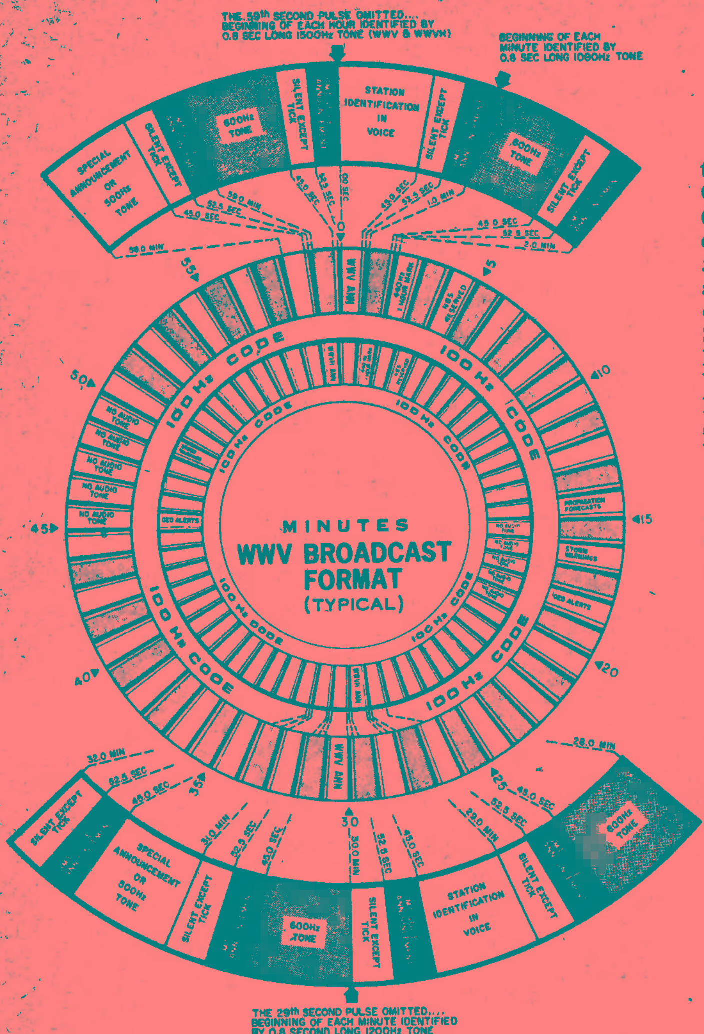


Fig. 1 — Schematic diagram of the wattmeter.
C1, C2 — 0.5- to 5-pF trimmer.
CR1, CR2 — 1N34A or equivalent.
M1 — 50-μA panel meter.
R1 — Lineer-taper, 1/4 or 1/2 watt, 25,000 ohm.
R2, R3 — 33-ohm, 1/2-W composition resistor (metched pair recommended).
RFC1 — 1-mH rf choke.
S1 — Spdt toggle.
T1 — 60 turns No. 28 enam. wire, close wound on Amidon T-68-2 toroid core (secondary). Primary is 2 turns of smell-diameter hookup wire over T1 secondary.



STANDARD FREQUENCIES AND TIME SIGNALS

The National Bureau of Standards maintains two radio transmitting stations, WWV at Ft. Collins, Co., and WWVH near Kekaha, Kauai (Hawaii), for broadcasting standard radio frequencies of high accuracy. WWV broadcasts are on 2.5, 5, 10, 15, 20, and 25 MHz, and those from WWVH are on 2.5, 5, 10, 15, and 20 MHz. The broadcasts of both stations are continuous, night and day. Standard audio frequencies of 440, 500, and 600 Hz on each radio-carrier frequency by WWV and WWVH. The duration of each tone is approximately 45 seconds. A 600-Hz tone is broadcast during odd minutes by WWV, and during even minutes by WWVH. A 500-Hz tone is broadcast during alternate minutes unless voice announcements or silent periods are scheduled. A 440-Hz tone is broadcast beginning one minute after the hour by WWVH and two minutes after the hour by WWV. The 440-Hz tone period is omitted during the first hour of the UT day.

Transmitted frequencies from the two stations are accurate to ±2 parts in 10^{11}. Atomic frequency standards are used to maintain this accuracy.

Voice announcements of the time, in English, are given every minute. WWV utilizes a male voice, and WWVH features a female voice to distinguish between the two stations. WWV time and frequency broadcasts can be heard by telephone also. The number to call is (303) 499-7111, Boulder, CO.

All official announcements are made by voice. Time announcements are in GMT. One-second markers are transmitted throughout all programs except that the 29th and 59th markers of each minute are omitted. Detailed information on hourly broadcast schedules is given in the accompanying format chart. Complete information on the services can be found in NBS Special Publication 236, *NBS Frequency and Time Broadcast Services*, available for 25 cents from the Superintendent of Documents, U. S. Government Printing Office, Washington, D.C. 20402.

Geophysical Alerts

"Geoalerts" are broadcast in voice during the 19th minute of each hour from WWV and during the 46th minute of each hour from WWVH. The messages are changed each day at 0400 UT with provisions to schedule immediate alerts of outstanding occuring events. Geoalerts tell of geophysical events affecting radio propagation, stratospheric warming, etc.

Propagation Forecasts

Voice broadcasts of radio propagation conditions are given during part of every 15th minute of each hour from WWV. The announcements deal with short-term forecasts and refer to propagation along paths in the North Atlantic area, such as Washington, D.C. to London, or New York to Berlin.

CHU

CHU, the Canadian time-signal station, transmits on 3330.0, 7335.0 and 14,670.0 kHz. Voice announcements of the minute are made each minute; the 29th-second tick is omitted. Voice announcements are made in English and French.