SECTION 2

ADVANCED PRACTICAL RADIO ENGINEERING

TECHNICAL ASSIGNMENT

RADIO FREQUENCY MEASUREMENTS PART I

249 E

- TABLE OF CONTENTS -

RADIO FREQUENCY MEASUREMENTS

PART I

INTRODUCTION.—In previous assignments were discussed the use of ammeters, voltmeters and ohmmeters, as well as Wheatstone bridges for the purpose of measuring various electrical characteristics of circuit components. These measurements were to be made at d-c, or at most, at low a-c frequencies. This assignment will concern itself with the problems and methods of measuring electrical quantities at high (r-f) frequencies.

Measurements at high frequencies do not differ in kind nor in method from those at low frequencies. Much the same type of measuring equipment is employed: Wheatstone bridges, voltmeters, ammeters, etc. But the operating principles are often different: for example, vacuum tube voltmeters and thermocouple ammeters replace the ordinary iron vane 60 cycle voltmeters and ammeters, and other modifications in assembled equipment are employed.

The most important difference, however, is that at high frequencies certain effects appear, *that were unnoticed at the lower frequencies because their magnitudes were less than the sensitivity of the measuring device and less than the precision of measurement, whereas at the higher frequencies their magnitudes are appreciable.*

Before proceeding with a discussion of measuring technique, it will be necessary to analyze these effects in order that a proper appreciation be had of the difficulties involved in making high frequency measurements. These effects will be divided into two groups: Residuals and Strays.

Residuals, in this assignment, refer to the inductance and capacitance effects in resistors, inductance and resistance effects in capacitors, and capacitance and resistance effects in inductances.

Strays, in this assignment, refer to random coupling, either of a capacitive or inductive nature, between circuit components in the test setup, or between a circuit component and ground. A study of these effects will shed considerable light upon the reasons why measurements at radio, and particularly at ultra-high, frequencies are so difficult, and how the obstacles to measurement can be obviated.

RESIDUALS

RESISTORS.—An ordinary wire-wound resistor consists of a helical coil of resistance wire wound on a ceramic form. Suppose it has a resistance of 100 ohms. Since it is in the form of a coil, it evidently also has some inductance, and—like all coils—has some distributed capacitance. To a first approximation the resistor may be represented by the circuit shown in Fig. 1. Here R increases with frequency at about the .5 power, (square root), so that R is not of a constant value as indicated in Fig. 1.

The quantities L and C are called the *residuals* in the resistor, and will change the impedance of the unit markedly at the higher frequencies.

Suppose L has a value of 20 μh,

and C of 10 μμf. At any frequency, the circuit of Fig. 1 has a certain impedance that can be expressed as a resistance and reactance in series, as shown in Fig. 2. In terms of R, L, and C and the frequency f, or angular velocity $\omega = 2\pi f$, it will be found that

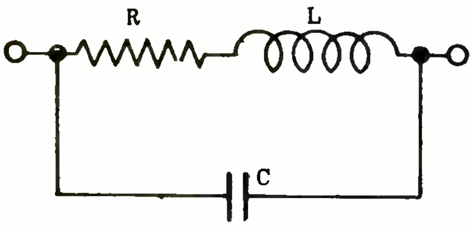


Fig. 1.—Equivalent circuit of a resistance.

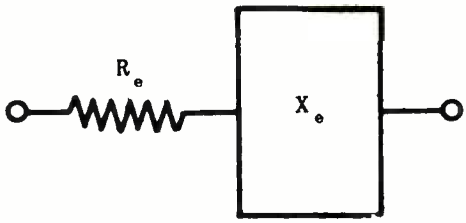


Fig. 2.—Equivalent series circuit of Fig. 1.

$$R_e = \frac{R}{(1 - \omega^2 LC)^2 + (\omega CR)^2}$$

$$X_e = \frac{\omega[L(1 - \omega^2 LC) - CR^2]}{(1 - \omega^2 LC)^2 + (\omega CR)^2}$$

Thus at 1 mc ($\omega = 2\pi 10^6$)

$$R_e = \frac{100}{(1 - 4\pi^2 \times 10^{12} \times 20 \times 10 \times 10^{-18})^2 + (2\pi 10^6 \times 100 \times 10 \times 10^{-12})^2}$$

$$= 101.6 \text{ ohms}$$

$$X_e = \frac{2\pi \times 10^6\ [20 \times 10^{-6}\ (1 - 4\pi^2\ 10^{12} \times 20 \times 10 \times 10^{-18}) - 10 \times 10^{-12} \times 10^4]}{(1 - 4\pi^2 \times 10^{12} \times 20 \times 10 \times 10^{-18})^2 + (2\pi\ 10^6 \times 100 \times 10\ \times 10^{-12})^2}$$

$$= 126 \text{ ohms}$$

This means that the resistor appears to have a value of 101.6 ohms instead of 100 ohms, or is 1.6% high, in series with an inductive reactance of 126 ohms. The effect of C at this frequency is to raise both the resistance and inductive reactance by small amounts, and the error if its presence were ignored would be small.

At ultra-high frequencies, however, the effect of C is of paramount importance. For example, at 100 mc,

$$R_e = \frac{100}{(1 - 4\pi^2 \times 10^{16} \times 20 \times 10 \times 10^{-18})^2 + (2\pi \times 10^8 \times 100 \times 10^{-11})^2}$$

$$= .0164 \text{ ohm}$$

and

$$X_e = \frac{2\pi\ 10^8\ [20 \times 10^{-6}(1 - 4\pi\ \ \times 10^{16} \times 20 \times 10 \times 10^{-18}) - 10 \times 10^{-12} \times 10^4]}{(1 - 4\pi^2 \times 10^{16} \times 20 \times 10 \times 10^{-18})^2 + (2\pi \times 10^8 \times 100 \times 10^{-11})^2}$$

$$= -161.2 \text{ ohms}$$

In other words, the resistor acts like a capacitor whose capacitive reactance is -161.2 ohms, in series with a practically negligible resistance of .0164 ohm. Physically, it is evident from Fig. 1 that C practically shorts out the L-R path, which has a relatively high impedance and thus passes very little current, so that the circuit looks practically like a pure capacitor.

This example illustrates a typical problem in high frequency work: a circuit component that is for all practical intents and purposes a resistor at low frequencies looks like an inductance and resistance in series at medium frequencies, and like a pure capacitor at ultra-high frequencies.

Before one can measure unknown resistors, capacitors, and inductances, one must be sure of the standard impedances he is employing in his measuring instrument.

There are two ways to handle residuals:

1. To incorporate or balance them with other circuit components so that the measurements are unaffected by their presence, and
2. To minimize their magnitude by proper design.

The latter method is preferable, since then the circuit component is more nearly that which, it was intended to be, e.g., a pure resistor, or pure capacitance, etc. However, after all steps have been taken to minimize the residuals, the small magnitudes still remaining can often be taken into account in the measuring instrument. For example, the residual inductance of a resistor or a capacitor can be balanced out in certain forms of Wheatstone bridges *if the inductance remains fixed and independent of the setting of the resistor or capacitor.* Hence, in one design, a compensating variable inductance is employed in conjunction with the variable resistance standard, and furnishes a fixed residual inductance independent of the resistance standard.

One of the earliest methods of minimizing inductance in a resistance was to use an Ayrton-Perry type of winding. Two coils are wound in opposite directions on the form so that the turns of one cross those of the other. The two windings are connected together at each end, so that they are in parallel. The currents flow through the two windings in opposite directions around the form, hence the magnetic flux set up by one winding is balanced by that of the other. The resulting flux set up is very small, and since the inductance of a circuit depends upon the amount of flux set up, it is evident that the residual inductance of an Ayrton-Perry type resistor will be small. A representative value is 0.3 μh. (This does not include the inductance of any connecting leads.)

Assuming a 100 ohm resistor with a shunt capacity of 10 μμf, as before, but with a series inductance of 0.3 μh, at 1 mc,

$$R_e = \frac{100}{(1 - 4\pi^2 \times 10^{12} \times .3 \times 10 \times 10^{-18})^2 + (2\pi 10^6 \times 100 \times 10 \times 10^{-12})^2}$$

$$= 100 \text{ ohms}$$

for all practical intents and purposes.

$$X_e = \frac{2\pi \times 10^6 \; [.3 \times 10^{-6} \; (1 - 4\pi^2 10^{12} \times .3 \times 10 \times 10^{-18}) - 10 \times 10^{-12} \times 10^4]}{(1 - 4\pi^2 \times 10^{12} \times .3 \times 10 \times 10^{-18})^2 + (2\pi 10^6 \times 100 \times 10 \times 10^{-12})^2}$$

$$= +1.256 \text{ ohms}$$

which is very small and should have a negligible effect upon the measurements.

At 100 mc

$$R_e = \frac{100}{(1 - 4\pi^2 \times 10^{16} \times .3 \times 10 \times 10^{-18})^2 + (2\pi \times 10^8 \times 100 \times 10^{-11})^2}$$

$$= 233 \text{ ohms}$$

which is a prohibitively large increase. The reactance is

$$X_e = \frac{2\pi \times 10^8 [.3 \times 10^{-6} \; (1 - 4\pi^2 \times 10^{16} \times .3 \times 10 \times 10^{-18}) - 10 \times 10^{-12} \times 10^4]}{(1 - 4\pi^2 \times 10^{16} \times .3 \times 10 \times 10^{-18})^2 + (2\pi \times 10^8 \times 10^2 \times 10^{-11})^2}$$

$$= -228 \text{ ohms}$$

and is also large and capacitive in nature. It is evident from these calculations that the resistor will be satisfactory in the standard broadcast range, but unsatisfactory in the ultra-high frequency range. Its superiority over the ordinary wire-wound resistor is obvious, however.

By reducing both residuals L and C, and further, by arranging through skillful design that $R = \sqrt{L/C}$, it is possible to make a resistance which varies but little over a wide frequency, and whose equivalent series reactance is very small. For example, in Fig. 3 is shown a Type 663 resistor manufactured by the General Radio Company. It consists of a very fine manganin wire A stretched across two flat metal plates B-B, to which it is soldered at each end. A thin piece of mica separates and insulates the wire from the two plates, except at the soldered ends. The assembly is then rigidly clamped together with a top piece of insulating material, and the extended sides of the two metal plates form slotted terminals.

Current flows in the two plates in a direction opposite to that in the wire, and since the spacing is small, the flux-canceling effect of the opposite current flows is very great, i.e., the circuit has very little inductance. For example, a 100 ohm resistor of this form has an inductance of only 0.39 μh. The close proximity of the wire to the plates enables the latter to conduct heat away from the wire, so that the wattage rating is considerable considering the fineness of the wire, [on the order of magnitude of one mil (.001 inch)]. Thus, a 100 ohm resistor can pass 0.06 amperes with a temperature rise of only 40° C. This corresponds to a wattage of $(.06)^2 \times 100 = 0.36$ watt.

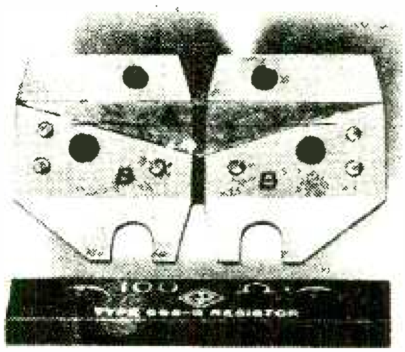


Fig. 3.—Type 663 General Radio resistor having low values of residuals.

The fineness of the wire makes the skin effect less than 1% for frequencies below at least 50 mc, so that R of Fig. 1 can be regarded as practically independent of frequency and of a value equal to its d-c resistance.

The distributed capacity varies somewhat with the method of mounting, but a representative value is 3.9 μμf. If L is .039 μh, then

$$\sqrt{\frac{L}{C}} = \sqrt{\frac{.039 \times 10^{-6}}{3.9 \times 10^{-12}}} = 100$$

This ratio is the optimum value for a 100 ohm resistor. If the values of R = 100, L = .039 μh, and C = 3.9 μμf be substituted in the previous equations, the following values are obtained at 60 mc.

R_e = 102.3 ohms and X_e = -.327 ohms

At 100 mc

R_e = 106.2 ohms and X_e = -1.531 ohms

This resistor is very satisfactory as a resistance standard up to at least 60 mc for use as a circuit element in bridges and similar equipment.

Variable resistors, as in the form of decade boxes, are not satisfactory above a few megacycles. This is because the inductance and the stray capacity of the wiring in the box is usually excessive, and moreover varies with the setting of the dial, so that corrections for these residuals are not so easy to make.

It is possible to build a variable resistance whose inductance remains constant and independent of the resistance setting, but there is still the element of shunt capacity to consider and, as the previous numerical examples have shown, the shunt capacitance, if large—say 20 μμf, or greater—can render the resistor useless in the u.h.f. range. Hence in any measuring instrument to be used at high frequencies, the calibrated *variable* unit is not a resistance, but a *capacitor*, because this can be built to have very small residuals that are substantially independent of its setting. Resistors, where required in the device, are of the fixed type, whose residuals can be made very small. Similarly, calibrated inductances are avoided as far as possible, because of their high residuals (shunt capacity and series resistance). Calibrated variable inductors are particularly unsatisfactory in this respect. Inductances are used only when their value does not have to be known, as in substitution methods to be described.

CAPACITORS.—A variable air capacitor is the best continuously adjustable impedance standard available for r-f measurements. Its residuals are low, and the calibration of its dial can be made mechanically to be very accurate and reproducible. However, in the r-f and u.h.f. range its residuals must at least be noted in order to avoid errors in measurement. A General Radio 722-N Precision Capacitor for use at radio frequencies is shown in Fig. 4. It is mounted in a rigid cast frame, and is constructed of aluminum alloys that give it the strength of brass with the weight and temperature coefficient of aluminum. The worm in the worm drive is cut directly on the shaft to avoid slight eccentricities and consequent errors in adjustment. The capacitor rotor shaft is mounted on ball bearings. These are not suitable for electrical connections, however, so that a brass drum on the rotor and a phosphor-bronze brush on the frame are used for this purpose.

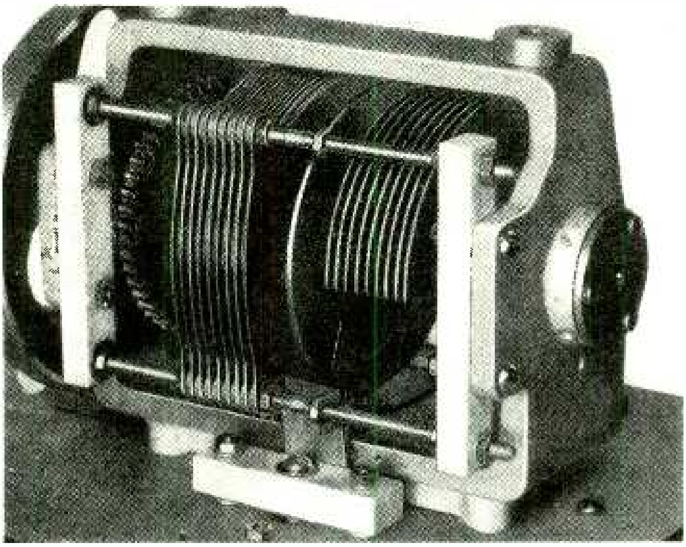


Fig. 4.—A variable tuning capacitor.

Electrically the capacitor is equivalent to the circuit shown in Fig. 5. Here R and L correspond to metallic losses and magnetic flux set up by the conduction currents flowing in the metal parts of the capacitor; C represents the static or low-frequency capacity of the capacitor; and G represents the dielectric losses in the insulating supports that hold the stator plates in place in the frame.

At low frequencies the reactance of C is high, so that *series* residuals such as R and L have negligible effect, whereas *shunt* residual G may be important. As one goes up into the r-f range, the losses represented by R begin to exceed the dielectric losses represented by G, and the positive reactance of L begins to be appreciable. As is evident from the figure, this positive inductive reactance, $2\pi fL$, cancels some of the negative capacitive reactance $1/2\pi fC$, so that the *net* negative reactance is less. This in turn means that C appears to have a larger value than its static value, and this is characteristic of a circuit approaching its series resonant frequency (here L and C are in series). The apparent capacity is

$$C_e = \frac{C}{1-\omega^2 LC}$$

The loss factors R and G may be regarded at any particular frequency as equivalent to a single resistance in series with C_e, and of a value approximately

$$R_e = R + \frac{G}{(\omega C_e)^2}$$

provided the residuals are small compared to C in effect. Actually R increases with frequency owing to skin effect: above about 1 mc it increases as the square root of the frequency.

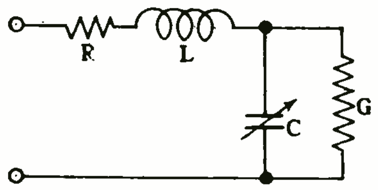


Fig. 5.—Electrical circuit of variable capacitor.

As an example of the error that may be incurred, consider a typical precision air capacitor operated at a frequency above its normal range, say at 10 mc. Let its residual inductance L be .065 μh, its residual resistance R = .065 ohms at 10 mc, and its dial setting represent 1,000 μμf. Assume that its dielectric loss is negligible at this frequency in comparison with its metallic losses, i.e., G = 0. then

$$R_e = R = .065$$

$$C_e = \frac{1000}{1 - 4\pi^2 \times 10^{14} \times .065 \times 10^{-6} \times 10^{-9}} = 1{,}347\ \mu\mu f$$

Thus, a capacity of presumably 1,000 is actually 1,347 μμf, an error of

$$\frac{1347 - 1000}{1000} = 34.7\%!$$

The reactance of the capacitor is

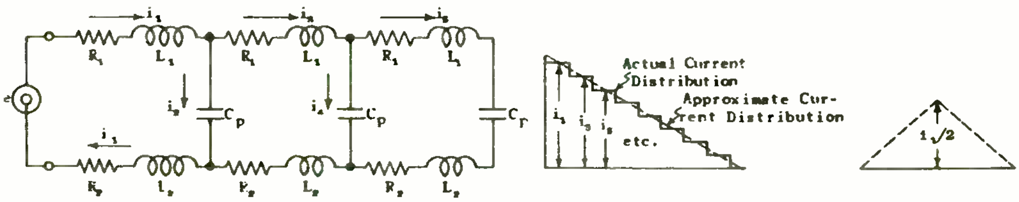
$$X_e = \frac{1}{2\pi f C_e} = \frac{1}{2\pi 10^7 \times 1347 \times 10^{-12}} = 11.8 \text{ ohms}$$

In comparison with this R_e is negligible, so that the most important error is that in C_e owing to the inductive residual L.

To reduce both R and L, a very simple and effective means is employed. The residuals are present mainly in the rotor shaft and in the stator rod washers. If current is fed in at one end of these, then there is in effect a kind of transmission line as shown in Fig. 6(A). Here a voltage e is impressed at one end of the assembly, and causes a current i_1 to flow. The capacity between each pair of capacitor plates of the stack is denoted by C_p, and between each pair of plates there is a portion of the rotor rod resistance and inductance. R_1 and L_1, and the stator rod resistance and inductance, R_2 and L_2. Of the inflowing current i_1, a fraction i_2 flows across through the capacity C_p of the first pair of plates encountered. This leaves a smaller amount of i_3 to continue to the next pair of plates, where a current i_4 is diverted, leaving a still smaller portion i_5 to continue on. Thus, the current distribution in the shaft is as shown in (B), both in actual (solid line) and in approximate (dotted line) form.

Corresponding to this distribution and value of R_1 and L_1 between pairs of plates there is an equivalent R and L for the capacitor as a whole, where the R and L represent the apparent resistance and inductance *if the current were uniform along the entire length and equal to the maximum value i.* Suppose, however, that instead of feeding the capacitor in the normal manner from one end, it is fed from the center of the assembly to each end. The current distribution is now approximately as shown in (C). The peak current in *either* direction is now $i_1/2$, one-half of its previous value, or, in other words, the two halves of the capacitor are now *fed in parallel.* The equivalent resistance and inductance of each half is half of their previous values, or R/2 and L/2; and since two halves are involved in parallel, the total resistance and inductance is R/4 and L/4. *Thus the residuals have been reduced to one-quarter of their original values.*

If the capacitor is fed at two equidistant points along its length, R and L are reduced to one-sixteenth their original values, and so on for a greater number of points of entry. However, a practical feature comes in, and that is that the number of interconnections becomes so great that the inductance and resistance of these nullifies the advantages of multiple feed, and so center feeding of the assembly, or a subdivision of 2:1 is the most that is attempted. In Fig. 4 is shown the actual construction for the General Radio Type 722-N capacitor. A heavy strip feeds the stator stack, and a brass disc with a wide brush contactor feeds the rotor stack. The residuals are thereby actually reduced to about 1/3 (instead of the theoretical 1/4). Thus, R is about .024 ohms at 10 mc, and L is 0.024 µh for the particular capacitor mentioned above.



(A) (B) (C)

Fig. 6.—Circuit equivalent to rotor shaft of variable capacitor and current distribution in the circuit.

The effective capacity for a dial setting of 1,000 µµf and 10 mc is

$$C_e = \frac{1000}{1 - 4\pi^2 \times 10^{14} \times .024 \times 10^{-6} \times 10^{-9}}$$

$$= 1{,}103 \ \mu\mu f$$

which corresponds to an error of

$$\frac{1103 - 1000}{1000} = 10.3\%$$

The percentage error is less for smaller dial settings, since C is smaller. Thus, for a dial setting of 100 μμf, (a magnitude of capacity more often to be encountered at 10 mc),

$$C_e = \frac{100}{1 - 4\pi^2 \times 10^{14} \times .024 \times 10^{-6} \times 10^{-10}} = 101\ \mu\mu f$$

or an error of only 1%. This capacitor is satisfactory for use up to about 30 mc, in that its readings, while in error at these higher frequencies, can be corrected to obtain the true values by means of the approximate equivalent circuit shown in Fig. 5, valid up to that frequency.

Another example of an air capacitor for u.h.f. use is the following:

Range: 5 μμf to 140 μμf

Residuals: R = 0.005 ohm at 1 mc, about 0.1 at 400 mc

Note: Owing to skin effect, R increases as the square root of the frequency from its 1 mc value.

Residuals: G = 100 μmhos at 100 mc

L = .0055 μh

The student can check its performance at various frequencies.

STRAYS

After the residuals in component parts have been minimized as far as possible, and then corrections made for what remains of these, one is still faced with the problems that arise when the components are combined into a measuring device. Each component has a certain amount of stray capacity to ground, and each connecting lead, as well as ground leads, have a certain amount of inductance. The presence of these stray effects can cause considerable error at radio frequencies because capacitive reactance is low and inductive reactance is high at these frequencies. Thus stray capacity can act as an appreciable shunt across a component, or as an appreciable coupling means between two components. On the other hand, if a ground lead has appreciable inductive reactance, the device may not be grounded (at radio frequency) as supposed, and errors owing to difference in ground potential can occur.

One particularly bad feature of stray capacity is that a somewhat different arrangement in the test layout may change the stray capacities considerably and thus alter the test results. For this reason test instruments assembled into a fixed form, such as a Wheatstone bridge, or a Q-Meter, are more accurate than a bread-board or bench layout even when precision components are employed. Even in a given layout of parts, the *stray capacities to the observer* will vary as he moves about, and impair the accuracy of the test. (The observer may be considered as a movable extension of ground.)

To obviate these strays, shielding is employed. To illustrate this, and r-f bridge will be used as an example. Consider the bridge shown in Fig. 7. There are two fixed resistors R_A and R_B of the high-frequency type described above. The d-c bridge was explained in an earlier assignment. The a-c bridge shown here operates on exactly the same principle, but because the actuating source (marked Gen.) is a.c.—at any desired radio frequency—a complication enters into the question of balance.

For a-c balance two conditions are necessary:

1. The resistive components of the bridge must be in balance.

2. The reactive components of the bridge must be independently in balance.

Thus, if the unknown device has both resistance and capacity, for example, the bridge to which it is connected must have means for adjusting an element to balance the resistive component of the unknown, and another element to balance the capacitive reactance of the unknown.

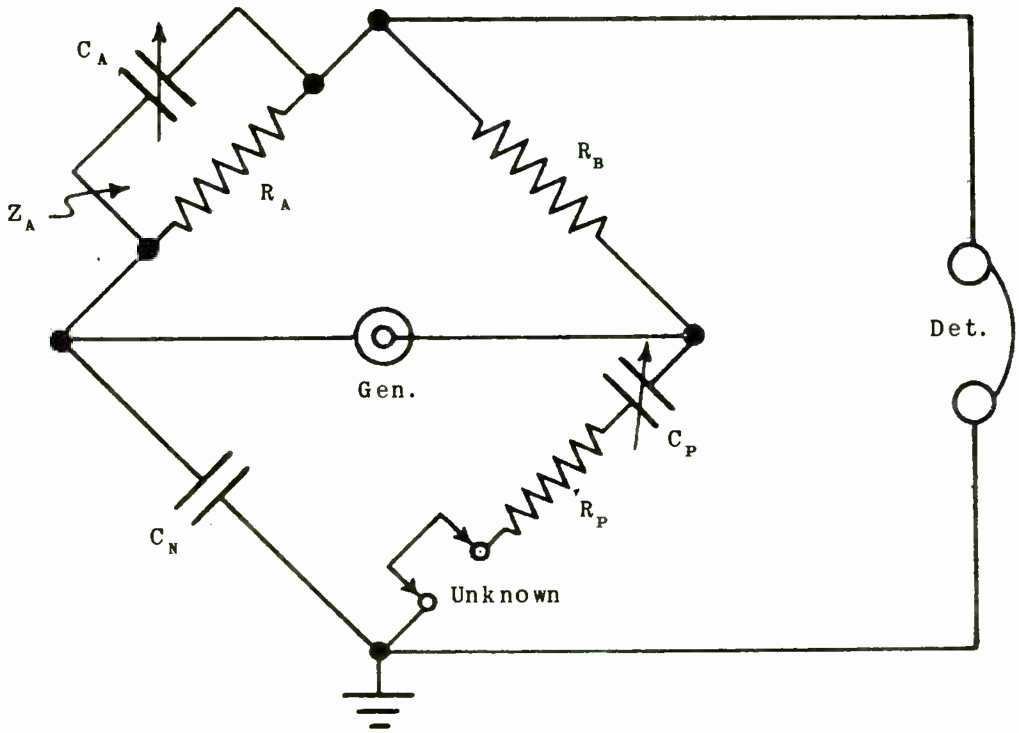


Fig. 7.—A-C bridge circuit.

In the bridge shown in Fig. 7, C_A is used to balance the resistive component of the unknown, (call it R_x), and C_p is used to balance the reactive component, X_x. Before describing the use of shielding, it will be necessary to discuss the method of measurement, known as the *series-substitution* method.

First the unknown terminals are short-circuited, and the bridge is balanced internally (initial balance). The capacitor C_p is adjusted in conjunction with capacitor C_A until a minimum indication is obtained on the detector. As mentioned previously, the generator is a source of r.f., and is preferably modulated with an audio tone, say 400 cycles. The detector can then be an ordinary radio receiver whose input terminals are connected to the bridge at the points indicated. The output of the receiver will then be the 400 cycle tone that has modulated the r-f carrier of the generator, and when this output is a minimum, the bridge is balanced. Call the settings of the capacitors C_{A1} and C_{p1}.

Now substitute the unknown impedance for the short-circuit. Suppose it has a certain capacitive reactance and a certain resistance. For a balance, both C_A and C_p will have to be reset to a new pair of values, (call them C_{A2} and C_{p2}). Then

$$R_x = R_B \frac{(C_{A2} - C_{A1})}{C_N}$$

$$X_x = \frac{1}{\omega}\left(\frac{1}{C_{p2}} - \frac{1}{C_{p1}}\right)$$

Thus, R_x is measured in terms of a *fixed* resistance R_B, a fixed capacitor C_N, and the difference in readings of a *calibrated variable capacitor* C_A. X_x is measured in terms of the frequency and the difference in readings of a *calibrated variable capacitor* C_p. The capacity of the leads connecting the bridge to the unknown impedance, for example, will not have any effect on the measurement so long as they are left connected to the bridge during the initial balance, *and the short circuit includes these leads.* The accuracy depends essentially upon how accurately *changes in capacity* in C_A and C_p are calibrated. Note also that measurements are made in terms of fixed resistances and variable capacitors. This bears out the previous statement in the assignment that variable capacitors can be built with smaller residuals than variable resistors, and are therefore normally preferred in measuring devices as adjustable calibrated standards.

There is, however, a source of error elimination illustrating very strikingly the use of shields. This error is the stray capacity of the capacitor C_p to ground, a capacity that shunts the unknown terminals and affects the accuracy of the measurement of an impedance connected to those terminals. This is obviated by surrounding the capacitor C_p in a metallic box or shield, in which the only openings are those necessary to permit an insulated extension of the capacitor shaft and also the connecting wires to pass through. Hence the inner capacitor can have only a fixed capacity to the shield, in which it is located in a fixed position. The shield is surrounded by another shield, to which it has a fixed capacity, and this shield is surrounded by a third shield, which is then connected to a proper point in the circuit.

The arrangement is shown in Fig. 8. In the actual bridge the capacitor C_p of Fig. 7 is actually two capacitors in series, C_p and C'_p of Fig. 8. This enables the main calibrated capacitor C_p to be set initially at the zero end of the scale for inductance measurements, or at the maximum end of the scale for capacitive measurements, and C'_p to be adjusted for the initial balance. Then, when the unknown impedance is connected, balance is restored by *adjusting* C_p. The dial reading on the latter then either gives the change in capacity directly if the initial setting of C_p was zero, or the difference of the reading from maximum gives the change in capacity. In either case the bridge can be made direct-reading at some particular frequency, actually at 1 mc.

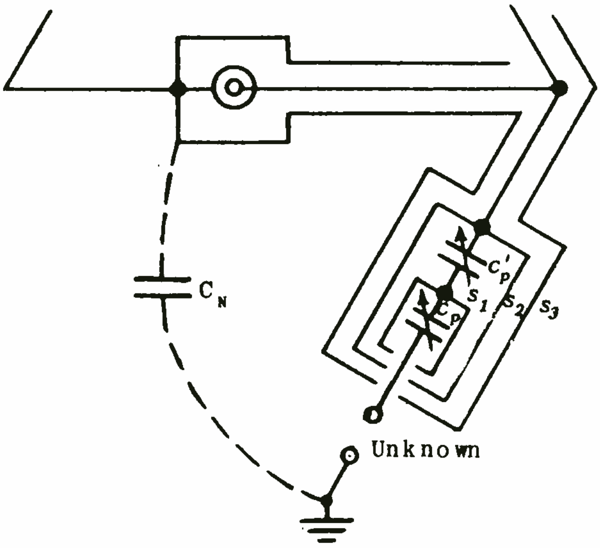


Fig. 8.—Circuit showing details of C_p of Fig. 7.

Shield S_1 surrounds C_p, and prevents C_p from having capacity to various parts of C'_p. Instead S_1 has a fixed capacity to shield S_2 surrounding C'_p. But S_2 is connected to the right-hand terminal of C'_p. Hence, the capacities of C_p to S_1 and S_1 to S_2 in series are to a definite point of C'_p, i.e., its right-hand terminal. This prevents interlocking of the settings of the two capacitors, that is, two different pairs of dial settings that give the same balance condition.

Note that the shields have prevented interlocking between the two capacitors, and also have prevented

C_p from having stray capacity to ground and hence across the unknown impedance. Instead, this capacity has been made a *definite* value across C_p'. However, S_2 would have a certain capacity to ground and thus across the unknown. To prevent this, S_2 is surrounded by shield S_3, which is connected to the left-hand side of the generator. Thus, the capacity between S_2 and S_3 is placed across the generator, and is equivalent to an additional load on the generator besides that of the bridge proper. Such a load can easily be carried by the generator, and has no effect on the measurements.

There is finally the capacity of S_3 to ground. As is clearly shown in Fig. 8, this is simply the capacity C_N that constitutes the lower left-hand arm of the bridge shown in Fig. 7. In short, a stray capacity from C_p to ground has been converted by the set of three shields into additional, fixed capacity in the lower right-hand arm of the bridge, plus a capacity across the generator (of no consequence to the bridge proper), plus an amount of capacity that furnishes practically all that is required for the lower left-hand arm of the bridge. In actual practice a small trimmer capacitor is employed to adjust the capacity of this arm to a certain desired value. The trimmer thus helps to take care of variations in the S_3 shield capacity from one instrument to another.

The above example illustrates one of the most complete and ingenious utilizations of shields. It is to be noted that one of the most important functions of a shield is to replace the *variable stray capacity* of a component to ground by a *fixed capacity to the shield.* If the shield is grounded, then the component has a fixed capacity to ground; if not, then the component has a fixed capacity to whatever the shield is connected. Variable quantities are thus replaced by fixed quantities, whose magnitudes can be measured, and whose effects can therefore be taken into account. Although usually the presence of a shield increases the capacity associated with the enclosed component, this is ordinarily not as objectionable as a capacity that is variable not only with the test setup, but with the position of the observer in the setup.

A shield has another important function, that of shielding the enclosed component from stray electric and magnetic fields. As an example, consider the coil L in Fig. 9(A). It has a certain amount of stray capacity to ground which can be lumped, as a good approximation, into two capacities C_1 and C_2 to ground, as shown. Suppose there is a wire W that forms part of a grounded circuit energized by some a-c generator (not shown) at some particular frequency. This wire has a certain amount of stray capacity to its surroundings; in particular it has stray capacity to L, which may be represented by C_1' and C_2' as shown. Current will flow through C_1' and C_1 in series, also current will flow through C_2' and C_2, and unless

$$\frac{C_2'}{C_1'} = \frac{C_2}{C_1}$$

in which case a balanced bridge condition obtains, some unbalance current will flow through L, producing a voltage across its terminals 1-2. If L is a circuit of its own, this voltage will be impressed upon the circuit and cause interference there. Note that the circuit may be a measuring device, or a radio receiver, or audio amplifier, etc., the mechanism of extraneous pickup from the neighboring circuit W is the same in all cases.

If either terminal of L is grounded, then the corresponding capacity is shorted out and the unbalance current will be greater, i.e., pickup will be greater. Thus, if terminal 1 were grounded, C_1 would be shorted out, and a larger unbalance current from C_2' through L to ground would flow. The voltage across L would be correspondingly greater. This merely illustrates the fact that a circuit balanced to ground, such as L ungrounded, and $C_1 = C_2$, is less troubled by pickup than one unbalanced to ground, such as terminal 1 of L grounded.

To eliminate the pickup, a shield is placed around the coil, with openings just large enough to allow the leads to pass through. The circuit now appears as in Fig. 9 (B). The shield is usually grounded at one point, as shown. The capacity of W to the shield, C, is a capacity to ground, since all points of the shield are practically at one potential if it is made of good conducting material, and the size is small compared to a wave length of the disturbing frequency, i.e., the frequency of W is not too high. The capacity of the shield to ground, C_3, is shorted out by the ground connection. Note that the capacities of L to the shield, namely C_1 and C_2 become capacities to ground. There is thus no tendency for W to couple to L, nor for L to couple to W, as far as capacitive

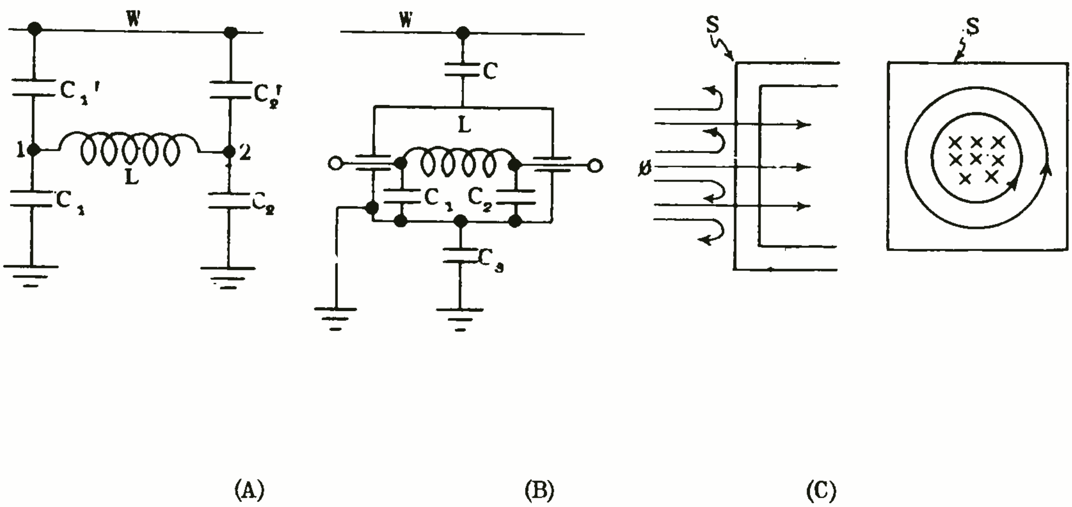


Fig. 9.—Circuit and action of currents in shielding circuits.

coupling is concerned.

A shield also protects the enclosed component from magnetic coupling. Thus let ϕ in Fig. 9 represent a-c flux at some instant proceeding into the side S of a shield can. Since it is alternating, it induces alternating voltages in circular swirls in S, as indicated to the right in (C). Since S is a conductor, these voltages cause currents to flow in S, and these, called eddy currents, produce a magnetomotive force *in a direction opposite to* ϕ. As a result, most of ϕ is repelled back in the direction from which it came. Only enough flux penetrates the shield to induce sufficient eddy voltages to produce enough eddy currents to repel the rest of the flux back from the shield.

The lower the impedance of the circular paths in the shield, mainly resistive, the less eddy voltage is required to produce the required eddy currents. Hence, the penetration of flux in a shield is less if the shield has high conductivity. At the higher frequencies the currents are confined to the surface of the metal (skin effect) so that the surface may be silver-plated to increase the conductivity and thus decrease the penetration of the flux.

The higher the frequency, however, the less flux is required to induce the requisite amount of eddy voltage, since the latter is proportional to the rate of change of the flux and hence to the frequency. As a consequence, there is very little penetration of flux into a shield can at radio frequencies; shielding is very effective at those frequencies. This is fortunate, for coupling, whether magnetic or electrostatic, is very effective at these frequencies, too. Note also that the flux produced within a shield can cannot pass out through the can because of the eddy currents it produces. Hence, a component within a shield cannot affect an external component.

It is apparent from the above that shielding as a means of preventing pickup and coupling between units is an important tool not only for measuring instruments, but for all sorts of radio equipment, such as transmitters, receivers, etc. In the case of measuring equipment, the further advantage of shielding is that it permits unavoidable capacities to be made fixed in value and to be located as desired in the circuit. The Wheatstone bridge example given above brings out this additional advantage very strikingly.

CONNECTIONS AND GROUNDS.—The series inductance of leads is a serious problem, not only in the case of connections between components, but connections to ground as well. Coaxial transmission line connectors between parts, such as between the generator and the bridge, or between the latter and the detector generally have less reactance than ordinary leads. Generally one ground connection point is preferred, and grounds from all components of the test setup are run to this point. The connectors are often made of copper strip about 1 inch wide, and as short as possible. In the laboratory, ground consists of a copper sheet or foil that is used to cover the top of the test table. Its area should be great enough to enable practically all test instruments used in the layout to be placed on it. In general, it is unnecessary to connect the sheet to a water pipe or similar ground because the reactance of the ground lead generally exceeds that of the capacity of the sheet itself to the earth. In field tests "ground" is preferably some large metal structure such as the frame of a piece of equipment, relay rack, etc.

FREQUENCY MEASUREMENTS

We are now ready, in the light of the above discussion, to study the use of measuring instruments in the determination of the various constants of circuit components. The measuring instrument may be of a very elementary type, or a precision device such as an r-f bridge, or a Heterodyne Frequency Meter, etc. The component to be measured may be a coil, a capacitor, a combination of the two, an antenna, etc.

One of the simplest, yet most important measurements, is that of the resonant or fundamental frequency of an L-C circuit. The coil and capacitor may be physically separate, or the capacity may be that distributed in the coil. The apparatus required is a source or generator of a-c power (often called a driver) that is adjustable in frequency, and a means of checking the frequency such as a wavemeter. If the driver is calibrated in frequency, then the wave meter may not be necessary.

THE DRIVER.—The driver is ordinarily a low power vacuum tube oscillator, of which an elementary form is that shown in Fig. 10. The arrangement oscillates at approximately the resonant frequency of L and C. By varying C, the frequency of oscillation can be changed to cover a range determined by the range of variation of C. To cover a greater range, more than one coil will be necessary, and these may be switched into the circuit, or a

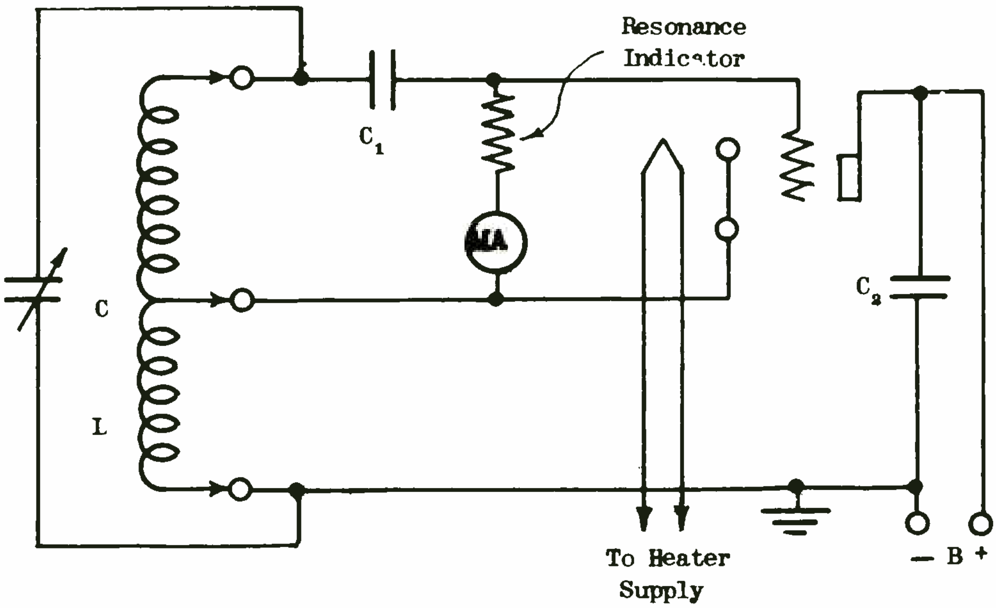


Fig. 10.—Low power driver.

plug-in system of coils may be used. The tube, in conjunction with the d-c plate supply, furnishes the losses of the system, i.e., those in L and C and in the grid circuit. The less the losses, the stronger is the oscillation, the greater is the grid excitation, and hence the greater is the reading of the d-c grid milliammeter.

This meter may therefore be used as a resonance indicator: a neighboring L-C circuit tuned to the oscillation frequency will extract appreciable power from the oscillator, decrease its amplitude of oscillation, and thus cause the grid milliammeter to "dip." The meter should ordinarily have a full-scale range of one to one and one-half milliamperes.

More elaborate commercial drivers are known as signal generators. These will be discussed in greater detail later. They are used for a variety of purposes, particularly to test radio receivers. They may be of a relatively crude type sufficient for servicing radio receivers, or of a laboratory type employing precision parts. These are in all cases calibrated in frequency, so that a wave meter is not required. The elementary form shown in Fig. 10 is not presumed to be thus calibrated.

The method of testing a given L-C combination is evident from the above discussion. The test setup is shown in Fig. 11. The resonant circuit is brought near the oscillator to couple it to the latter. The frequency of the oscillator is then varied until a dip in the grid meter occurs. This indicates that the oscillator is generating the resonant frequency of the L-C circuit under test. To be sure that the dip is caused by this circuit and not by some neighboring circuit, place the hand on the coil, thus increasing its capacity to ground and lowering its resonant frequency. If the grid current increases, then the dip was evidently caused by the circuit under test. Note that this

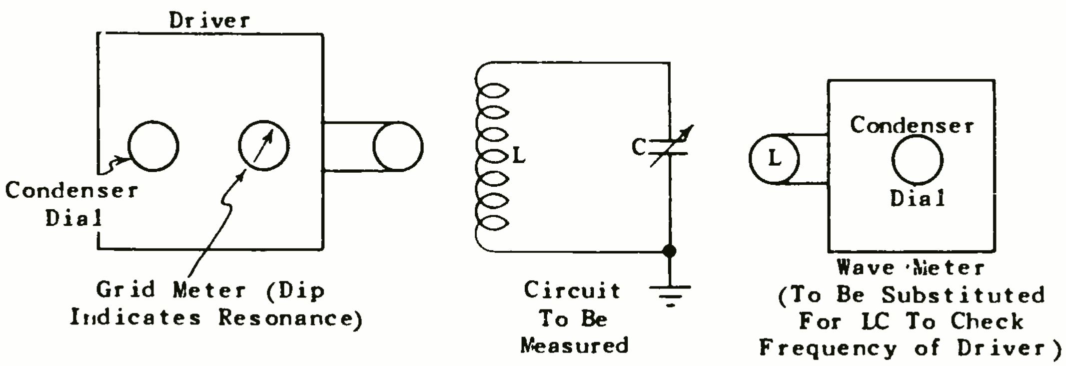


Fig. 11.—Testing L-C circuits for frequency of operation.

auxiliary test indicates that the test setup is susceptible to stray capacity, i.e., is not adequately shielded. This is true, but is not of such importance here if the frequency is not too high, because the precision of measurement is not very great, although satisfactory for a great many purposes. It is in general advisable to ground the circuit under test as shown because this generally reduces some of the stray effects, and moreover, tests the circuit more nearly under the same conditions as it will subsequently operate.

The frequency of the driver must now be checked by means of the wavemeter, (if the driver itself is not calibrated). The wavemeter replaces the circuit under test and is thus coupled to the driver. The capacitor of the wavemeter is adjusted until the driver grid meter dips once more, and the frequency read off the wavemeter capacitor dial.

Before discussing the frequency or wavemeter, a few remarks concerning the test procedure just described will be of value. The driver can be very easily constructed by the student. The capacitor C can be of the ordinary radio receiver type, and should preferably be equipped with a dial having Vernier control. The inductance L depends upon the frequency range to be covered. C_1 is a feedback coupling capacitor (50 to 200 μμf is a normal value) to couple the grid of the tube to the tank (L-C) circuit, and C_2 is a bypass capacitor, used to bypass the r-f energy around the power supply. If the oscillations are too feeble, a larger value of C_1 is indicated, whereas if C_1 is too large, the circuit will block periodically or "squedge", i.e., oscillate intermittently. The tube can be an ordinary receiver type, such as the 6J5 triode.

When using the driver, the circuit to be measured should first be coupled rather closely so that a pronounced dip is obtained. This will make it easy to select the correct coil and to "find" the approximate resonant setting. When an approximate adjustment has been determined the coupling should be gradually weakened until the dip can just be clearly distinguished at the point of exact resonance. Weak coupling should *always* be used when employing a driver for work requiring an adjustment to some exact frequency, for a strong coupling produces a strong reaction and tends appreciably to shift the frequency of the driver.

THE WAVEMETER.—The wavemeter in its fundamental form is a very simple device. It consists simply of a series circuit of inductance and capacity that can be tuned to resonance with any other circuit by means of a variable capacitor and usually a number of coils that may be changed by either a plug-in or switching arrangement. In most cases a radio frequency milliammeter or other resonance indicator is incorporated in the device to indicate when it is in resonance with the other circuit. In other frequency meters the milliammeter is not used, resonance being indicated in the driving circuit. The fundamental circuit is shown in Fig. 12. When the meter is coupled to a radio frequency oscillator or transmitter and tuned to resonance with the driving source, the condition of resonance is indicated by the maximum reading on the milliammeter.

A simpler type uses a small

flashlight bulb in place of the milliammeter. Maximum brilliancy of the bulb indicates resonance.

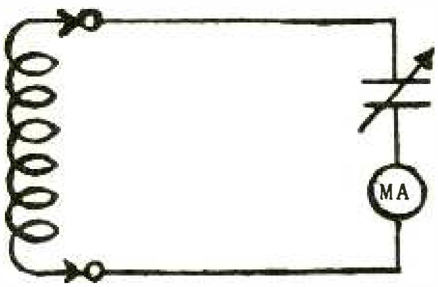


Fig. 12.—Simple wavemeter circuit.

This type of indicator is not as precise as the former, but is more accurate than might at first be supposed. The brilliancy of the light goes up rapidly with a small increase in current, so that the indication is fairly sharp. However, it requires about two watts of power to light it to full brilliancy, and if the exciting source is one of low power, the lamp will fail to indicate. In this case the reaction of the wavemeter on the source, such as the dip in the latter's grid meter, may be used to indicate resonance.

The precision of measurement for the type described above may be between ± 2 and ± 3%, the lower value being generally for the medium frequencies, such as the standard broadcast and short wave portions of the spectrum. A more precise form of wavemeter is the General Radio Type 724-B shown in Fig. 13. Its accuracy is ± 0.25% between 50 kc and 50 mc ± 1.0% between 16 kc and 50 kc. Note the seven plug-in coils (one mounted in the instrument), to cover the large frequency range.

In order that a wavemeter mav

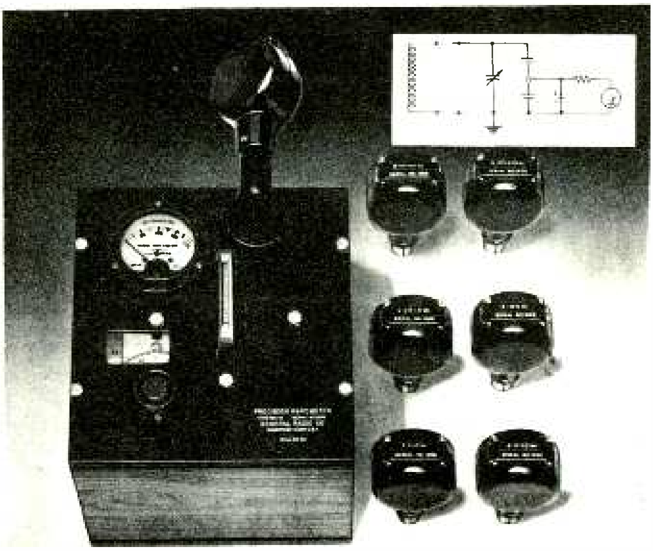
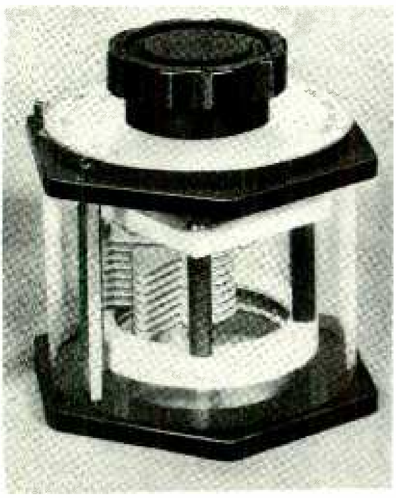


Fig. 13.—Type 724.B Precision Wavemeter.

have such high accuracy, it is necessary that it constitute a low-loss circuit, so that it has a very sharp resonance peak. The ordinary thermogalvanometer used as an r-f meter, or the flashlight bulb introduces too much resistance in the L-C circuit to permit such a sharp peak. Hence, in this instrument a much more sensitive diode voltmeter is employed and is connected essentially across the capacitor. At resonance maximum voltage occurs across both the coil and capacitor, and is so indicated by the meter.

A high-frequency type of precision capacitor having low residuals is employed. The reduction in residual resistance is not so important in view of the higher losses in the associated inductance, but the reduction in residual inductance in the capacitor permits of a larger inductance unit for the highest frequency range. This is important, because even so the inductance for the highest frequency range is often but a thick short bar of metal essentially across the capacitor terminals.

Another type of wavemeter, of about ± 2% accuracy, that covers in one range the span from 55 to 400 mc, is shown in Fig. 14(A) and (B).



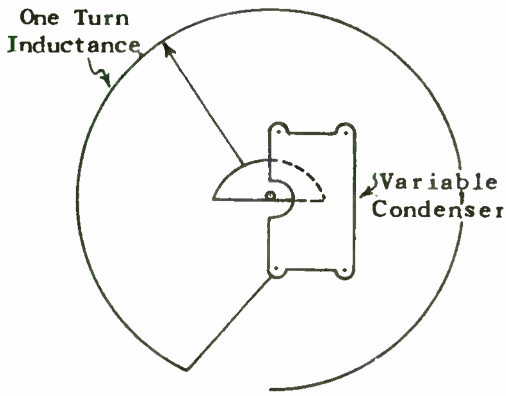


Fig. 14.—A simple wavemeter with an accuracy of about 2%.

The principle of operation is shown in (B). A one turn silver coil is employed in conjunction with a high-frequency type of variable air capacitor. A contact arm in the form of a silver spring on the capacitor rotor slides along the silver coil, changing the amount of inductance associated with the capacitor. The simultaneous variation of L and C permits a much greater range of frequencies to be covered. A flashlight bulb in an aperiodic (non-resonant) circuit loosely coupled to the above resonant circuit acts as the resonance indicator.

Wavemeters are not as accurate as the heterodyne frequency meter to be described further on; nevertheless they are much more convenient to use, and can rapidly determine, for example, the frequency range covered by some experimental oscillator, the preliminary lining up of radio transmitters, etc.

MEASUREMENT OF IMPEDANCE

The above setup not only permits the resonant frequency of an L-C circuit to be measured, but the circuit constants as well. More complicated instruments permit impedances in general to be measured with relative ease.

DISTRIBUTED CAPACITY OF A COIL.—As mentioned previously, all coils have distributed capacity, and will resonate with it at a succession of frequencies, like a multimesh resonant circuit. For frequencies around and below the lowest self-resonant frequency, (of the parallel resonance type), the distributed capacity may be regarded approximately as lumped and connected across the coil terminals, just like the L-C circuit shown in Fig. 11. Hence, the same test setup as in Fig. 11 may be employed to measure the distributed capacity of a coil.

This measurement is an important one. For example, r-f choke coils are often employed to act as a choke in a frequency range that includes their lowest self-resonant or fundamental frequency. At the top end of the range (above the fundamental frequency) the choke exhibits a capacitive reactance (the coil acts like a capacitor!) whereas below the fundamental frequency the choke exhibits an inductive reactance. The magnitude of this reactance corresponds to an apparent inductance that is higher than the low frequency or so-called true inductance of the coil, and is a result of the tuning action of the distributed capacity. In the case of a coil to be used in tuning an r-f receiver stage, or in an oscillator required to operate over as large a frequency range as possible, it is important that the fundamental frequency be as high as possible, i.e., that the distributed capacity be at a minimum. Hence, its measurement is of importance to the radio engineer.

The test setup is similar to that shown in Fig. 11. The coil to be tested is shown as L, and the variable capacitor should in this test be an *accurately calibrated unit*, since the accuracy of the measurement depends upon the accuracy of calibration of this capacitor. The driver should be adjustable from a frequency that is about 1/3 or 1/4 of the fundamental frequency of the coil up to about the coil frequency. Expressed in another way, the λ range of the driver should be several times greater than the fundamental λ of the coil. Finally, the wavemeter should be *accurately calibrated.* The reason for such accuracy throughout is that the distributed capacity to be determined is usually very small, and large errors can be incurred in its determination unless the test components are known to a high degree of precision.

The test is performed exactly as in the previous example. First, set the calibrated variable capacitor to a setting near its maximum scale value. Suppose, for example, that the maximum is 70 μμf, and that 60 μμf is chosen as a start. Tune the driver to the *dip,* then loosen the coupling and tune to resonance accurately. Remove the L-C combination and couple the wavemeter *without changing the driver adjustment.* Tune the wavemeter to resonance with the driver, get an accurate setting, and then take the λ from the calibration curve for the wavemeter coil used. If the meter is calibrated in frequency the conversion can be made by dividing the frequency *in cycles* into 3×10^8, or the frequency *in kilocycles* into 3×10^5. The result is λ in meters.

Repeat the test at other capacity settings lower on the dial, such as, for example, 50 μμf, 45 μμf, 30 μμf, or about five readings in all. Enter the results under three column leadings as follows:

Capacity in μμf	*Wavelength*	*(Wavelength)²*
60	---	---

The third column to the right is obtained from the test readings entered in the second column by squaring them.

Now plot the points of capacity against $(\lambda)^2$ on a sheet of graph paper. A representative plot is shown in Fig. 15. Draw a straight line as shown through all points, intersecting both the ordinate and the abscissa. If the readings have been taken carefully and if the calibration of the wavemeter and capacitor are accurate, all of the points will fall on a straight line. *The point where this straight line crosses the abscissa to the left of the ordinate indicates the distributed capacity of the coil.* In this case the coil will have a distributed capacity of 8 μμf.

The theory of this measurement is very simple. Suppose the coil had no distributed capacity. Then the only capacity across its terminals would be that of the calibrated capacitor. If the latter were set to zero, the resonant frequency would be infinite, and the corresponding λ would be zero.

Since

$$\lambda = 1884\sqrt{LC} \quad \text{(L and C in micro-units)}$$

so that

$$\lambda^2 = (1884)^2\, LC$$

it is evident that if L is kept constant, λ^2 is *directly proportional* to C, where C is the total capacity, is that of the calibrated capacitor alone. The plot of the above equation of direct proportionality between λ^2 and C is a straight line *that passes through the origin* (shown by the broken line in Fig. 15).

If the coil has distributed capacity, then even a zero value of the external capacitor will still

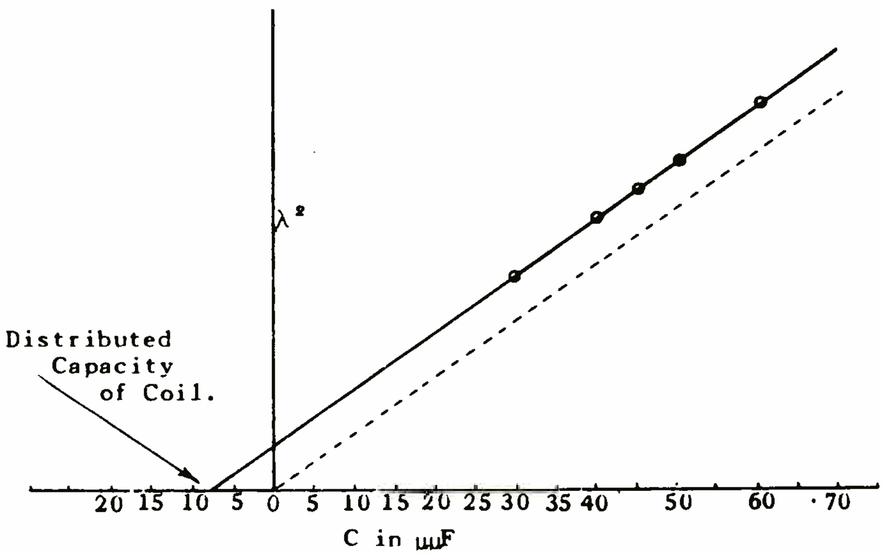


Fig. 15.—Plot of capacity against λ^2 to find coil distributed capacity.

correspond to a λ (hence λ^2) greater than zero. It would be necessary for the external capacitor to develop a *negative* capacity equal to the distributed capacity before λ would be reduced to zero. This means that the actual plot is above that for zero distributed capacity, and hence crosses the abscissa to the left of the origin. In the example, this comes out to be 8 μμf.

APPARENT AND TRUE INDUCTANCE.—The distributed capacity paralleling the coil makes it appear as an apparent higher inductance in the range below the fundamental frequency. Thus, from

$$\lambda = 1884 \sqrt{LC}$$

(L and C in micro-units)

or

$$\lambda^2 = (1884)^2 \ LC$$

$$L = \frac{\lambda^2}{(1884)^2 \ C}$$

The capacity is that of the external capacitor, (call it C_e). In parallel with it is the coil's distributed capacity, (call it C_d). The value of C in the above formula is clearly $C_e + C_d$. Hence, the true value of L is

$$L = \frac{\lambda^2}{(1884)^2 \ (C_e + C_d)}$$

The apparent value of L, or L_a, is that found by using for C simply C_e, the external capacity. Thus

$$L_a = \frac{\lambda^2}{(1884)^2 \ (C_e)}$$

In this expression the denominator involves C instead of $(C_e + C_d)$ or is smaller, hence the fraction, equal to L_a, is larger. In the example given, the apparent inductance is found by dividing

$$\frac{\lambda^2}{(1884)^2}$$

by the corresponding reading of the calibrated capacitor, whereas the true inductance is found by dividing it by the reading of the calibrated capacitor plus 8 μμf distributed capacity.

THE Q METER.—Various other tests can be made with the above setup, but it will be more instructive to study an interesting and important arrangement similar to the above, and assembled in one unit known as the Q Meter. It can be employed to measure the inductance and Q of a coil, the capacity and losses of a capacitor, the resonant frequency of an L-C combination, etc. Indeed, the components that can be measured, or the types of measurements that can be made by a particular instrument often depend as much upon the ingenuity and common sense of the engineer as upon the test instrument. No one assignment, nor even a complete book, can begin to cover the variety of measurements that can and are made, and so it is the aim here to describe some typical measuring instruments and methods so that the engineer can gain an insight into the technic of measurements, and proceed from there on to devise means and methods for special tests that may be required in the course of his work.

It will be of value to review briefly what the Q of a coil is, and why it is important. Consider the series resonant circuit shown in Fig. 16. The resistance R_L is considered to be the high-frequency resistance (including skin effect) of the coil L. The losses in the capacitor may ordinarily be considered negligible in comparison to those of L. If the frequency e_1 is that of resonance for L and C, then the inductive reactance of L is just balanced by the capacitive reactance of C, and the current i is determined solely by R_L, namely

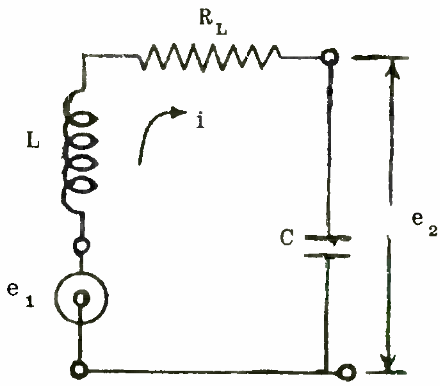


Fig. 16.—Series resonant circuit.

$$i = \frac{e_1}{R_L}$$

This current produces approximately equal and opposite voltage drops across L and C that are usually many times e_1 in value. Thus the voltage across C is

$$e_2 \cong iX_c$$

The drop across L is equal to e_2 plus the voltage drop across R_L, which is e_1, and is small compared to e_2 and at right angles to it, so that the vector sum of the two is practically no greater than e_2. Therefore,

$$e_2 = iX_L$$

But

$$i = e_1/R_L$$

so that

$$e_2 \cong e_1\left(\frac{X_L}{R_L}\right)$$

also

$$= e_1\left(\frac{X_c}{R_L}\right)$$

The quantity X_L/R_L is often called the storage factor for the coil, or its Q. The above equation may therefore be rewritten as

$$\frac{e_2}{e_1} \cong \frac{X_L}{R_L} = Q$$

This shows that the step-up in voltage owing to series resonance is equal to the Q of the coil, which is the Q of the circuit if the capacitor losses are negligible, as is usually the case. Consequently the "figure of merit" of a coil is very properly represented by its Q, and the measurement of the latter quantity is an important one.

One further point is worthy of note. Suppose the *coil* in Fig. 16 had no resistance, i.e., had an infinite Q, but that the capacitor had the losses represented by R_L. Then, since one can also write

$$\frac{e_2}{e_1} = \frac{X_c}{R_L}$$

it is possible to denote X_c/R_L as the Q of the capacitor (since R_L is now assumed to reside in the capacitor). However, it is often customary in such a case to speak instead of the reciprocal of the above, or

$$\frac{R_L}{X_c} = \frac{R_L}{\frac{1}{\omega C}} = \omega CR_L$$

and this is called the *dissipation factor, D,* for the capacitor. One can therefore write

$$\frac{e_2}{e_1} = \frac{1}{D}$$

For the circuit as a whole it is more customary to compare the *total series losses* to the *reactance of the coil,* ($X_L = 2\pi fL$), rather than to the admittance of the capacitor ($1/X_c = 2\pi fC$) so that one usually speaks of the former ratio, and calls it the circuit Q. As stated above, this is practically always the same as the *coil* Q; i.e., the losses in the capacitor are usually relatively negligible.

The analysis for a series resonant circuit can also be applied to a coupled circuit with very little modification, particularly if the primary coil that is coupled to L in Fig. 16 has a constant current excitation, such as that provided by a pentode tube. For example, let the current in such a primary coil be I_p, and the mutual inductanct between this coil and L be M. Then the *series voltage* e_1 induced in the secondary circuit composed by R_L, L and C is simply

$$e_1 = \omega M I_p$$

From then on the analysis is as described previously.

From the foregoing it is evident that in order to find the Q of the circuit it is necessary to measure either the current in the circuit, or the induced voltage, and the voltage across the capacitor or the coil. The arrangement that accomplishes this in a simple operational manner is known as a Q Meter, and the basic circuit is shown in Fig. 17. A driver or oscillator of impedance that is practically a pure resistance, R_G. The driver feeds a very low resistance, R_M, about .04 ohm in value. Practically short-circuit current flows through R_M, because it is extremely low compared to the internal resistance of the driver.

The action can readily be analyzed by means of a network theorem known as Thevenin's Theorem. This theorem states that regardless of where the generator may be in a network, if an additional circuit be connected to two terminals of the network, the latter appears to the attached circuit like a generator whose apparent generated voltage is that developed across the terminals before the circuit was attached, and whose apparent internal impedance is that measured across the terminals before the circuit was attached. Specifically, in Fig. 17 a generator

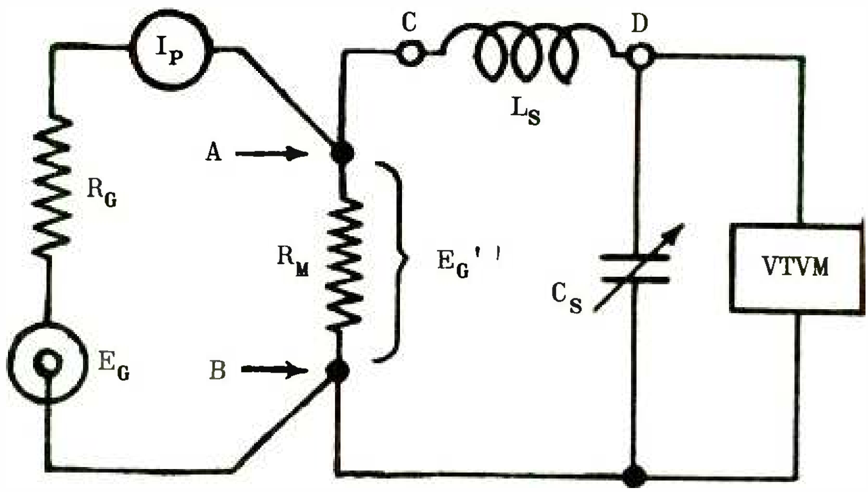


Fig. 17.—Basic circuit of the Q meter.

high quality design and wide frequency range develops the voltage E_G. The driver itself has inherently a very appreciable internal feeds a simple little network composed solely of R_M, to whose terminals A-B is connected a resonant circuit L_s, C_s. To the resonant circuit the combination to the left of A-B appears, by Thevenin's Theorem, as an apparent source having the following characteristics;

1. The apparent internal impedance of this source is that to be measured between terminal A-B when the resonant circuit is disconnected. Clearly this is R_M in parallel with R_G. (In calculating this impedance it is of course assumed that the actual generator is momentarily inactive, i.e., that $E_G = 0$). Since R_G is many times R_M in value, it acts as a *negligibly high shunt to R_M*, so that the apparent internal impedance is for all practical intents and purposes simply R_M, or about .04 ohm, and *practically independent of moderate variations in R_G*. In this way an apparent source resistance is obtained that is extremely low compared to the resistive losses in any ordinary coil to be measured and so may be considered negligible.

2. The apparent generated voltage is that appearing across the terminals A-B when the resonant circuit is disconnected. This is equal to $I_P R_M$. But evidently

$$I_P = \frac{E_G}{R_G + R_M} \cong \frac{E_G}{R_G}$$

since R_M is negligible compared to R_G. Hence, the apparent generated voltage is

$$E_G' = I_P R_M \cong E_G \frac{R_M}{R_G}$$

The simplest means of maintaining E_G' constant under test is to use an r-f milliammeter, such as a thermo-galvanometer in series with R_M, and adjust E_G (by means of a kind of volume control on the oscillator) so that I_P reads a certain value. In this way E_G' can be kept constant regardless of variations either of E_G, the actual generated voltage of the oscillator, or R_G, its actual internal resistance.

Fig. 18 shows the circuit equi-

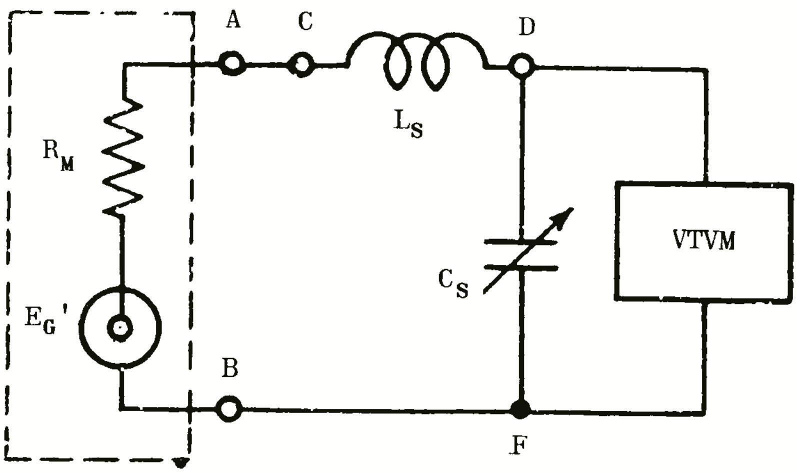


Fig. 18.—Equivalent circuit of Fig. 17.

valent to that in Fig. 17. The apparent source or generator for the resonant circuit is shown enclosed in dotted lines. Its internal resistance is so low—.04 ohm—that the resonant circuit connected across it is relatively a high impedance load that draws very little current from it. Hence, the internal voltage drop in the apparent source impedance R_M is negligibly small when the resonant circuit is connected to it, and consequently the voltage across the latter circuit is practically constant at the value E_G'. Under such simplifying conditions the Q of the resonant circuit can readily be measured.

In the instrument the terminals shown as C–D in Fig. 17 and 18 are normally open. To them is connected the coil L_s whose Q is to be measured. The driver or oscillator is set to the specified frequency by means of a coil-changing band switch and a calibrated tuning dial. The oscillator range is from 50 kc to 74 mc. The current I_p is adjusted to a predetermined value as indicated by a calibration mark on the I_p meter in the driver circuit. This produces a definite voltage, E_G', between terminals A–B across R_M.

The standard calibrated capacitor C_s that is part of the meter is varied over its entire range once while watching the vacuum tube voltmeter (VTVM) for resonance as indicated by a maximum reading. Then C_s is *widely* detuned from resonance to reduce the current in the tuned circuit to practically zero. The VTVM is then adjusted to read zero by turning up its adjusting control from a minimum setting until a slight reading is observed, and then reducing the setting to bring the indicator just back to zero.

Theoretically, the calibration current I_p produces a voltage across A–B just sufficient to produce a deflection of one unit on the 0–250 scale of the VTVM when L_sC_s is considerably detuned to a higher frequency. Actually this deflection is so small that negligible error is introduced by adjusting the VTVM to zero whether the circuit is detuned to a higher or lower frequency.

Capacitor C_s is now varied to obtain exact resonance, as indicated by a maximum reading on the VTVM. The number of scale units on the meter then indicates the Q, or number of times the voltage across C_s at resonance exceeds the (theoretical) one unit reading in the off-resonance condition.

An example of a commercial form of this device is shown in Fig. 19,

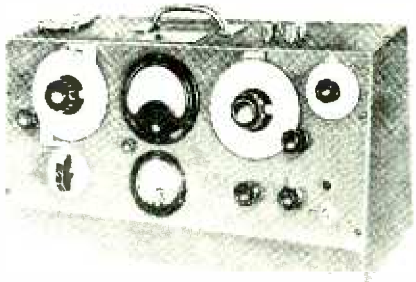


Fig. 19.—160-A Q-meter; Boonton Radio Corp.

and is the Boonton Radio Corporation Type 160-A Q Meter. The two dials at the left determine the oscillator frequency; the top one is the oscillator tank capacitor dial, and the lower one the coil-selector switch (frequency range). The top meter is part of the VTVM and reads Q (essentially volts across C_s). The zero-adjust knob for the VTVM is directly below and to the left of the meter. The bottom meter is the r-f thermocouple meter that determines I_p. (Since I_p must set up an appreciable energizing voltage across .04 ohms for the resonant circuit, it is evident that the oscillator must develop considerable output).

The large dial to the right is the main capacitor dial, and as may be noted, has a fine-tuning adjustment. The capacitor C_s of Figs. 17 and 18 actually consists of two capacitors in parallel: the main or Q tuning capacitor of 0-450 μμf range equipped with the large dial and a smaller vernier capacitor, of 0-6 μμf range, equipped with the smaller dial shown at the top extreme right. Directly under the main capacitor dial is the oscillator output control used to adjust I_p to the correct value, as noted by the lower meter, and under the vernier dial are the pilot light and "ON-OFF" switch.

Various ranges of Q are taken care of in the following manner: it is evident that if I_p were kept constant and hence the voltage E_G' injected via the .04 ohm resistor into the resonant circuit, the deflection on the VTVM for very high Q coils would be off-scale. To prevent this, I_p must be reduced. In the older Model 100-A meter, this was done in two steps: for high Q's around 500, I_p was set to a low calibration line marked 500 on the thermocouple meter, whereas for Q's around 250, it was set to a line marked 250 representing twice the value of current, (four times the deflection, since the thermocouple meter is of the square-law type).

In the later Model 160-A there are several such calibration settings marked "X 1", "X 1.1", "X 1.2" etc. For Q's up to 250, the "X 1", setting is used, and the VTVM reading is multiplied by 1, for Q's exceeding 250, the current I_p is reduced by choosing the appropriate calibration setting "X 1.1", "X 1.2", etc., and the VTVM reading multiplied by the calibration setting. For example, if the I_p meter reading is made 1.2 and the VTVM reads 240, then the Q will be $1.2 \times 240 = 288$. By this means the VTVM is kept reading up scale at all times, where its accuracy is greatest. The accuracy of the Q meter is better than 5% on all frequencies up to 30 mc, for Q values in excess of 50.

MEASUREMENTS WITH THE Q-METER.—The method of measuring the Q of a coil has been described above. However, in actual practice, oddly enough, the Q meter is perhaps used more often to measure the inductance of a coil or the capacity of a capacitor, than the Q of a circuit. The reason is apparent: The Q-meter has all the elements that the previously described test setups have, namely, a driver calibrated both in output and frequency, an external resonance indicator or VTVM (compared to the grid meter dip indication) and in addition, a calibrated capacitor.

1. *MEASUREMENT OF INDUCTANCE*.—Suppose it is desired to measure the inductance of a coil. The coil is connected across terminals C-D, Fig. 18. The meter case should be grounded to avoid voltage pickup by the VTVM and violent off-scale deflection when the operator

touches the terminals while connecting the coil. The oscillator output should be increased gradually to avoid overloading and damaging the thermocouple meter, and should be decreased when switching from one frequency to another owing to the possible large variations in output in the different frequency ranges.

Choose any desired frequency, f, provided it is below the fundamental frequency of the coil, and vary the Q-capacitor until resonance is indicated by a maximum reading on the VTVM. The latter reading indicates the Q of the coil. The inductance, however, can easily be calculated from the usual resonance formula:

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

where C is the circuit capacity as indicated by the dial reading on the Q-capacitor. Note that the dial is calibrated in terms of the total capacity of the circuit, including that of the VTVM, provided the vernier tuning capacitor is set at zero.

2. *MEASUREMENT OF CAPACITANCE.*—When it is desired to measure the capacitance of a capacitor, a coil must first be connected to terminals C-D, Fig. 18, and then the capacitor to be tested is connected across terminals D-F, i.e., across the Q-tuning capacitor. It is not necessary that the inductance of the coil be known, since a substitution method is to be employed. The method is very simple. First, at any desired frequency, the coil is resonated with the Q-capacitor, while the unknown capacity is disconnected. Call the capacity dial setting C_1. Then the unknown capacity is connected across terminals D-F, and the Q-capacitor capacity reduced until resonance once more is obtained. Call its new dial setting C_2. The capacity of the unknown capacitor is simply $C_1 - C_2$.

The accuracy of this measurement depends upon the accuracy of setting of the capacitor. This is ± 1 μμf for the range from 30 to 100 μμf and ± 1% above 100 μμf. For example, if C_1 is 226 μμf, and C_2 is 47 μμf, then the capacity is 226 - 47 = 179 μμf, but C_1 may be in error by (± .01) (226) = ± 2.26 μμf; i.e., C_1 may be 226 + 2.26 or 226 - 2.26 μμf, or in round numbers anywhere from 228 to 224 μμf. C_2, on the other hand, may be 47 ± 1 or between 46 and 48 μμf. The total error may be as great as 2.26 + 1 = 3.26 or about 3 μμf in 179 μμf, an error of 3/179 = 1.7%.

At high frequencies, above about 5 mc, the residual inductance and resistance in the Q-tuning capacitor, as well as in the connecting leads, will reduce the accuracy unless corrections are made. The residual inductance of the Q-tuning capacitor is .015 μh at the binding posts, and correction can be made for this by means of the formula for capacitor residuals given in the first part of this assignment.

3. *METHODS OF MEASURING IMPEDANCES IN GENERAL.*—Any impedance may be represented, *at any given frequency*, by two alternative forms, the series, and the parallel forms. For example, an impedance that has both capacitive reactance and resistance may be represented by a resistance and capacitor of some other appropriate magnitudes in parallel. Either arrangement is equally correct in specifying the impedance. Generally the series form

is preferred for low impedances, and the parallel form for high impedances.

This concept can be profitably carried over into the field of measurement, such as when using the Q-meter. In Fig. 20 are shown a inductance, L, in conjunction with the Q-meter, in order to measure Z. First the meter is resonated with L to the desired frequency of measurement by adjusting Q-tuning capacitor C. Let the resulting readings be Q_1 and C_1. Now connect

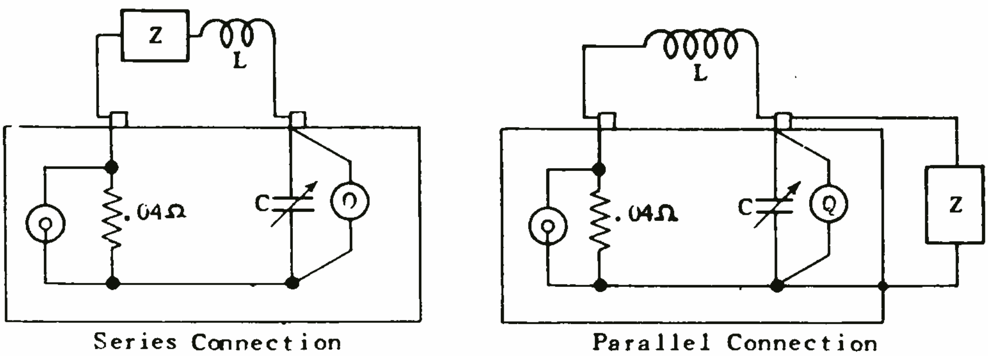


Fig. 20.—Different connections used with Q-meter.

series and parallel connection for the use with the Q-meter and generally employed for low and high impedances, respectively.

Since Z, the unknown impedance, may have a capacitive of inductive reactance, it is necessary to use an Z to the circuit, either in series or in parallel, as desired, and re-resonate the system to the same frequency. Let the new readings be Q_2 and C_2. Then the constants for the unknown impedance Z are as follows:

Series Connection

$$Q_x = \frac{(C_1 - C_2)\, Q_1\, Q_2}{C_1\, Q_1 - C_2\, Q_2}$$

$$R_s = \frac{1.59 \times 10^8 \left(\frac{C_1}{C_2} Q_1 - Q_2\right)}{f C_1\, Q_1\, Q_2}$$

$$X_s = \frac{1.59 \times 10^8\ (C_1 - C_2)}{f\, C_1\, C_2}$$

Parallel Connection

$$Q_x = \frac{(C_2 - C_1)\, Q_1\, Q_2}{C_1\, (Q_1 - Q_2)}$$

$$R_p = \frac{1.59 \times 10^8\, Q_1\, Q_2}{f C_1\, (Q_1 - Q_2)}$$

$$X_p = \frac{1.59 \times 10^8}{f\,(C_2 - C_1)}$$

$$L_s = \frac{2.53 \times 10^{10} (C_1 - C_2)}{f^2 C_1 C_2}$$

$$L_p = \frac{2.53 \times 10^{10}}{f^2 (C_2 - C_1)}$$

$$C_s = \frac{C_1 C_2}{(C_2 - C_1)}$$

$$C_p = C_1 - C_2$$

NOTE.—If C_1 is greater than C_2, X_s is inductive (+), otherwise it is capacitive.

NOTE.—If C_1 is greater than C_2, X_p is capacitive (-), otherwise it is inductive.

In the formulas for Q_x, however, $(C_1 - C_2)$ and $(C_2 - C_1)$ are always considered positive. In the above formulas, f is in kc, L is in microhenries, and C is in micromicrofarads.

RADIO FREQUENCY MEASUREMENTS PART I

EXAMINATION

1. (A) Describe the Ayrton-Perry type of resistor.

(B) How does this construction reduce the residual inductance?

2. (A) Describe the General Radio Type 663 resistor.

(B) How is the residual inductance reduced by this type of construction?

(C) What feature of the construction increases the wattage rating?

2. (Cont'd.)

(D) Why are variable resistors in general unsatisfactory above a few megacycles?

3. A high-frequency resistor of the type described in the text (See Fig. 1) has the following constants:
R = 270 ohms, C = 0.4 $\mu\mu$f.

(A) Calculate the optimum value of inductance, L.

(B) Calculate the resistive and reactive components of this resistance at 60 mc. (Assume that the skin effect is negligible at 60 mc, and that R is still 270 ohms at this frequency).

3. (B) (Cont'd.)

4. (A) What are the residuals in an air capacitor?

(B) Which may be important at very low frequencies? Why?

(C) Which are important at very high frequencies?

(D) What effect do these have upon the effective capacity of the capacitor?

5. A high frequency variable calibrated air capacitor has the following residuals (as per Fig. 5 in the text):

R = 0.1 ohm; G $\cong$ 0; L = .0055 μh; C variable from 5 μμf to 140 μμf. These values are for a frequency of 400 mc. During

5. (Cont'd.)

a series of tests at this frequency, the capacitor dial is set to 10 $\mu\mu f$.

(A) What is the apparent capacity of the capacitor, i.e., what would be the capacity of an ideal, pure capacitor, to have the same reactance at 400 mc as that of the above actual capacitor?

(B) What is the series resistance of the actual capacitor?

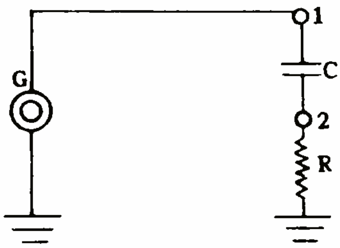
6. A certain capacitor, when measured at 1,000 cycles, shows a capacity of 7 $\mu\mu f$. At 100 mc its apparent capacity is 14 $\mu\mu f$. What is its residual inductance?

7. (A) It is found that the addition of a shield around a certain test component increases its capacity to ground; nevertheless such an addition is found to be advantageous. Explain.

7. (Cont'd.)

(B) Explain how a shield protects the enclosed component from magnetic coupling.

8. Given the circuit shown. The capacitor C has a certain amount of capacity to ground, which may be localized as a certain amount of capacity from terminal 1 to ground, and a certain amount of capacity from terminal 2 to ground. The latter capacity evidently shunts R. This is undesirable for the particular use that this circuit is to have, whereas the capacity from 1 to ground is across generator G, and does no harm. How and where would you shield this circuit, and where would you connect the shield, to minimize any stray capacity across R? Explain the reasons for your choice.



9. The following readings are obtained in measuring the distributed capacity of a coil:

Capacity in μμf	Wavelength
60	149
50	137
40	123
30	108

(A) Find, graphically, the distributed capacity of the coil.

(B) From the above value of distributed capacity, and from any capacitor setting and corresponding frequency, calculate the *true* inductance of the coil.

9. (Cont'd.)

(C) Calculate the *apparent* inductance of the coil.

10. Explain the use of the Q meter in measuring the Q and the inductance of a coil.