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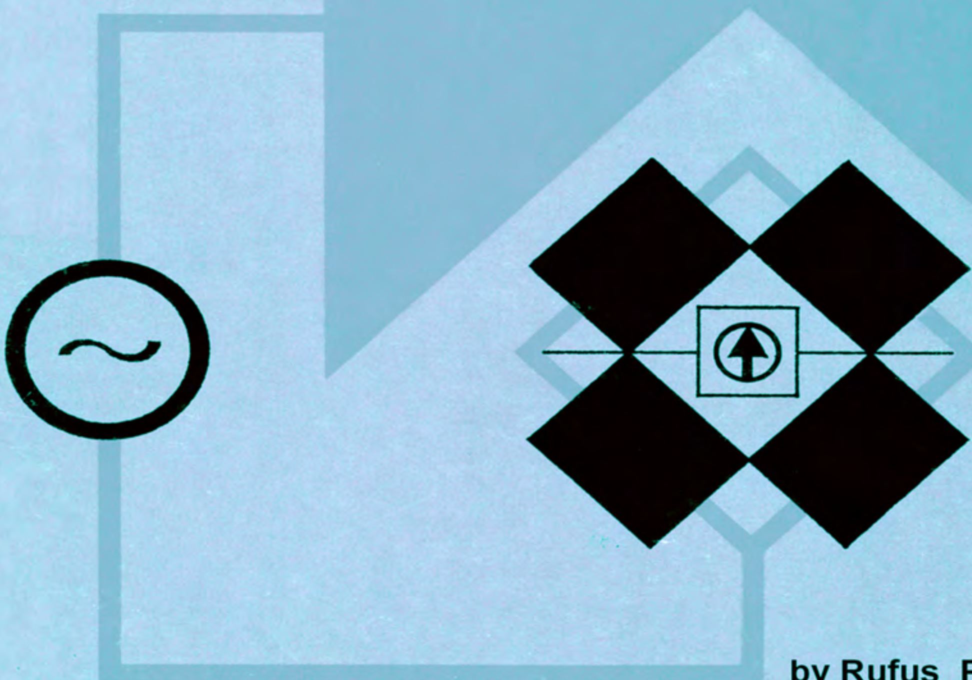
BRIDGES and other NULL DEVICES



by Rufus P. Turner

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Preface

Bridge circuits find widespread usage in modern electronics. Introduced in 1833 as a precision resistance-measuring device, the bridge gradually encompassed measurement of other properties: capacitance, self-inductance, mutual inductance, reactance, impedance, frequency, distortion, and tube and transistor coefficients. Today, bridges—in one form or another—are as familiar in the repair shop as they are in the research laboratory.

Supplementary to bridges, a family of bridge-like null devices also emerged concurrently with the growth of electronics. Like bridges, some of these devices enable measurement of resistance, capacitance, inductance, or frequency; but they find use principally as selective circuits—alone or in amplifiers and test instruments.

Bridges and nonbridge null circuits alike have uses other than the original one of checking components. For example, they are essential parts of modern industrial, military, and communications equipment. Here they provide automatic response or precise control in such roles as voltage regulation; power measurement; frequency control; filtering; phase shifting; sensing of temperature, pressure, strain, humidity, and magnetic fields; carrier suppression; and modulation.

This book describes principal members of the numerous classes of bridges and nonbridge null circuits, bringing the explanations together in this one place for convenient study or reference. The arrangement and development of the material favor either a consecutive study of the text or spot reference to parts of it. Our intended reader is the practical man, whether technician or engineer.

RUFUS P. TURNER

Contents

CHAPTER 1

Basic Theory

It is evident everywhere in the engineering world that scientific progress depends on the ability to make accurate measurements. Reciprocally, it is always an essential task of engineering to design more accurate measuring devices to assure further progress. This interrelationship has resulted in the accumulation of an enormous body of exact knowledge through constantly improved observations and measurements.

In contrast to the measuring devices of 150 years ago, some present-day instruments are amazingly accurate. But this degree of perfection did not arise suddenly. It is worth noting that many of the present instruments evolved through a process of continual refinement. Remarkable progress has been made, especially in the field of electrical measurements.

This chapter briefly contrasts two fundamental methods of measurement and introduces the bridge.

1.1 TWO BASIC TYPES OF MEASUREMENT

All measurements may be said to fall into two categories: those which make indirect use of a standard, and those which make direct use of a standard. Measurements of the "indirect" type enjoy great popularity because of their speed, simplicity, and convenience, but those of the "direct" type offer the better accuracy. It should be noted that the words *direct* and *indirect* refer here to the relationship of the test instrument to the standard, not to the method of taking readings. An "indirect" instrument—yardstick, simple thermometer, or simple voltmeter—usually gives direct *readings*.

Instruments which use the indirect approach are calibrated from a standard; e.g., a voltmeter is connected to a precision voltage source, the corresponding deflection noted, and the meter adjusted, if necessary, for agreement between the standard voltage and the indicated voltage. Thereafter, the instrument is used by itself, and it is returned to the standard, if ever, only when the calibration is rechecked. Instruments which use the direct approach, on the contrary, are always used in conjunction with the standard. This provides a continual comparison of measured values to standard values, and is the chief reason why the direct method is the more accurate.

1.2 INDIRECT METHOD: ADVANTAGES AND DISADVANTAGES

Most instruments using the indirect method give direct readings; i.e., the "unknown" values are read immediately from the position of some indicator on a graduated scale. Examples are thermometers, simple weighing scales, clocks, and electric meters.

Advantages of such instruments and of the indirect method are simplicity (because the readings are easily taken) and speed. Speed is achieved because operation is almost instantaneous and usually requires no manipulation other than setting a range control; an operator, for example, need only set a voltmeter to the correct range, connect the meter to the circuit or device under test, and read the voltage from the meter scale. Additionally, a costly standard is not needed.

Disadvantages of the indirect method are those inherent in most direct-reading indicators used away from a standard: (1) a multistage process is employed and performance in any one, or all, of the stages can deteriorate after calibration; thus, in a voltmeter the voltage under test sends a proportional current through the pivoted coil of the instrument in series with a suitable multiplier resistor, and this current, in turn, sets up a magnetic field which causes the coil to rotate and move the pointer over the scale; (2) since the instrument is divorced from the standard, there is no way of knowing, between calibrations, whether the indications are correct; (3) error can result from reading the indication and may be caused by parallax, misinterpretation of the value denoted by a scale division, or lack of skill in estimating the position of an indicator when it stands between divisions; (4) indirect electrical measurements usually draw current from the circuit being measured and so cause loading effects which may seriously impair accuracy.

1.3 DIRECT METHOD: ADVANTAGES AND DISADVANTAGES

Instruments employing the direct method are always used in conjunction with a primary or secondary standard. In some instances, they are operated near a stationary, external standard; in others, the standard is contained in the instrument itself. In some instances, the standard is used intermittently, and in others, continuously. Examples of "direct" instruments are self-calibrating voltmeters, and frequency meters with standard frequency oscillators.

A simplified illustration of a d-c voltmeter with self-contained standard is given in Fig. 1-1. In this circuit, voltmeter M is normally connected to test terminals X-X through push-button switch S1 (normally closed against contact 1) and can indicate an external voltage applied to the test terminals. E_s is an accurately known voltage supplied by a built-in battery or standard cell. When S1 is depressed so as to close temporarily against contact 2, external voltage E_x is removed and standard voltage E_s is applied to the meter. In this way, the calibration of the meter scale may be checked at will against the standard voltage, E_s. Best accuracy results when $E_x = E_s$, i.e., when the deflection is the same for both voltages. But this condition is obtained only occasionally, e.g., when an external voltage must be adjusted to the standard value.

The advantage of the direct method is the accuracy due to correlation with the standard. Disadvantages are: (1) it is slower than the indirect method; (2) it requires somewhat bulkier instrumentation; and (3) its greater accuracy results when the unknown voltage equals the standard voltage, which means that number of standard voltages (or one accurately adjustable standard) will be needed for most accurate measurement of a range of voltages.

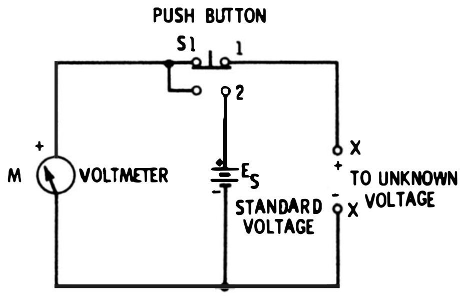


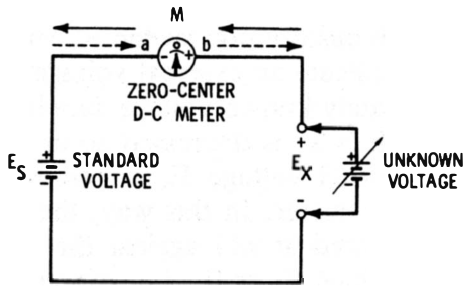
Fig. 1-1. Voltmeter with internal standard.

1.4 THE NULL METHOD

A special version of the direct method keeps the standard in continuous use during a measurement, thus differing from the method described in Section 1.3 in which the standard is engaged only intermittently.

Fig. 1-2 shows a simplified example of this version. Here, M is a center-zero d-c voltmeter or galvanometer. E_s is an accurately known standard voltage, and E_x an unknown voltage to be measured. When $E_x = E_s$, the two voltages cancel each other and the meter reads zero.

When E_x is higher than E_s, terminal *b* of the meter is more positive than terminal *a*, so electrons flow in the direction of the dotted arrows and deflect the meter upscale (positive). When, instead, E_x is lower than E_s, terminal *a* is more positive than terminal *b*, the current reverses to the direction of the solid arrows, and the meter is deflected downscale (negative).



(A) Circuit.

VOLTAGE RELATION	METER DEFLECTION
$E_X > E_S$	NEG
$E_X = E_S$	0
$E_X < E_S$	POS

(B) Response.

Fig. 1-2. Demonstration of null method.

The condition of zero deflection, indicating equality of the two voltages, is termed *null.* Since at null the standard and unknown voltages are balanced, the circuit is said to be *balanced* when it is adjusted to null. In this respect, the arrangement is the electrical analogy of the type of balance in which accurate weights are placed in one pan, and enough material is placed in the other pan to balance the known weight, as indicated by the zero reading of the indicator. Accuracy is increased because, at balance, no current is drawn from the source, E_x.

When E_x and E_s differ greatly, the full-scale range of meter M must be high enough to accommodate the net voltage without damage. As null is approached, however, the meter sensitivity may be increased, with an accompanying large increase in accuracy of the null setting. When E_x is very close to E_s, the lowest range of the meter may be used. Thus, various shunts may be switched across a galvanometer to control its sensitivity.

The null method has several advantages: (1) the standard is continuously referred to; (2) the meter scale needs no special calibration, since no absolute values are read from it; (3) null, the condition of final adjustment, is read as zero deflection—a point which can be easily recognized; (4) accuracy is improved by the absence of "loading" on the source, E_x, at null. Thus, E_x is referred directly to the standard (E_s), and the various links which can introduce error in the indirect method (see Section 1.2) are avoided.

1.5 NULL CIRCUIT WITH VOLTAGE DIVIDERS

One or both of the batteries in the previous circuit (Fig. 1-2) may be combined with a continuously variable voltage divider to make the voltage adjustable. Fig. 1-3 shows a two-divider arrangement.

Here, voltage divider R1 supplies a selected fraction (e_1) of voltage E_s, while R2 similarly supplies a selected fraction (e_2) of voltage E_x. As in the preceding example (Section 1.4), both e_1 and e_2 are positive; therefore, the meter deflects downscale when e_1 is more positive than e_2, upscale when e_1 is less positive than e_2, and reads zero when $e_1 = e_2$.

A disadvantage of this design is that both E_s and E_x are constantly under load and therefore may not be able to put out their true open-circuit voltages. For both E_s and E_x, this is not a *true* null method.

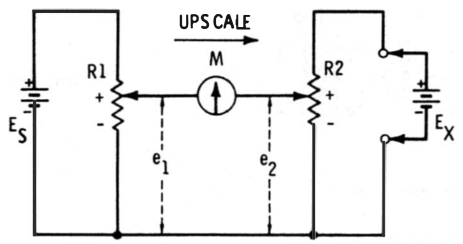


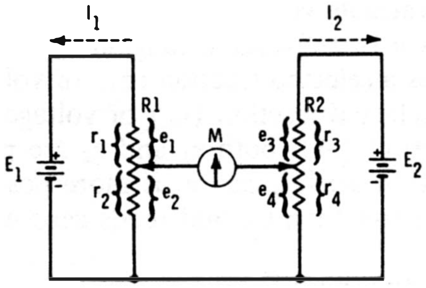
Fig. 1-3. Null circuit with voltage dividers.

1.6 BASIC D-C BRIDGE: CONFIGURATION

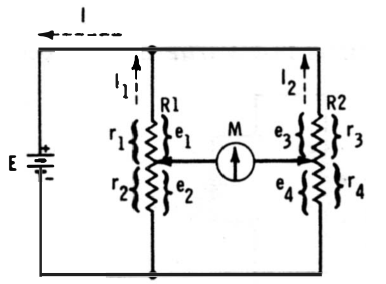
The dual voltage-divider circuit shown in Fig. 1-3 can be used to compare resistances as well as voltages. Thus, in Fig. 1-4A, batteries E1 and E2, which are exactly equal in voltage, force currents I1 and I2, respectively, through voltage dividers R1 and R2. The resultant voltage drops e_1 $(= I_1r_1)$, e_2 $(= I_1r_2)$, e_3 $(= I_2r_3)$, and e_4 $(= I_2r_4)$ are proportional to resistances r_1, r_2, r_3, and r_4, respectively. At null, $e_1/e_2 = e_3/e_4$, so $r_1/r_2 = r_3/r_4$. And from this relationship, any one of the resistances may be determined in terms of the other three: $r_1 = (r_2r_3)/r_4$, $r_2 = (r_1r_4)/r_3$, $r_3 = (r_1r_4)/r_2$, and $r_4 = (r_2r_3)/r_1$. In this way, one unknown resistance may be determined from the values of three accurately known resistances.

Resistance determinations may be made without the two separate batteries; the voltage dividers may be connected in parallel and operated from a single d-c source, as shown in Fig. 1-4B. Here, the current, voltage, and resistance relations are the same as in the preceding example. With Fig. 1-4B, the basic configuration of the bridge emerges.

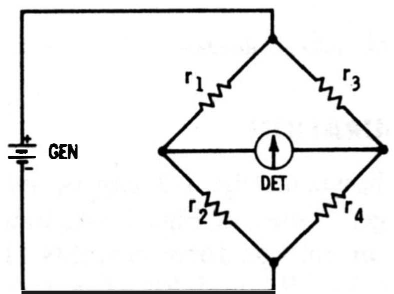
Usually, a bridge circuit is drawn in the diamond shape shown in Fig. 1-4C. Here, r_1, r_2, r_3, and r_4 are separate *arms* of the bridge and correspond to the selected resistances r_1 and r_2 of potentiometer R1, and to r_3 and r_4 of potentiometer R2 in Fig. 1-4A. One arm, such as r_1, may be made continuously variable for balancing the circuit to



(A) Two-battery circuit.



(B) Single-battery circuit.



(C) Conventional representation.

Fig. 1-4. Evolution of bridge.

null, and one (such as r_2) may be the unknown. The two remaining arms, r_3 and r_4, then are accurate resistances whose ratio determines the ratio of the unknown (r_2) to the standard (r_1). Thus, at null the unknown $r_2 = (r_4/r_3)\,r_1$. In bridge parlance, the power source (such as the battery in Fig. 1-4C) is termed the *generator*, and the null indicator (such as the center-zero galvanometer in Fig. 1-4C) is the *detector*.

The basic bridge circuit was invented by S. H. Christie, an Englishman, who described it in a paper in the February 28, 1833, issue of

Philosophical Transactions. But the device attracted little attention until 1843, when it was publicized by Sir Charles Wheatstone, from whom it came to be called the Wheatstone bridge despite Sir Charles's painstaking credit to Christie.

The bridge enables an unknown quantity to be checked directly against a standard which is permanently contained in the circuit. Bridges are used for accurate measurement of such properties as resistance, capacitance, inductance, impedance, and frequency.

1.7 BASIC A-C BRIDGE: CONFIGURATION

As a test instrument, a d-c bridge can be used only for the measurement of resistance. Capacitance, inductance, and impedance measurements, however, require that the bridge be powered by alternating current.

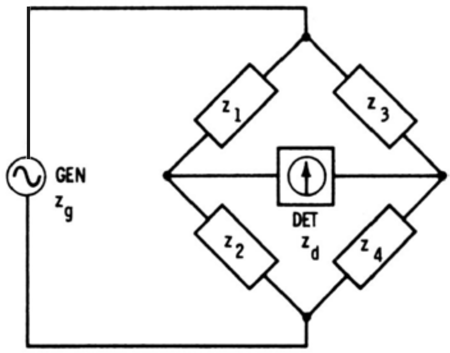


Fig. 1-5. Basic a-c bridge configuration.

Fig. 1-5 shows the basic configuration of an a-c bridge. Note that the circuit is essentially like the d-c bridge, except that an a-c generator (GEN) has replaced the battery, and an a-c null detector (DET) has replaced the center-zero d-c galvanometer. The a-c detector may be a vacuum-tube (or transistorized) voltmeter, oscilloscope, magic-eye tube, or headphones (with or without an amplifier). Also, the resistance arms of the d-c bridge have been replaced with corresponding impedance arms Z_1, Z_2, Z_3, and Z_4. (These impedances are of the form $R + jX$.) Calculations of unknown impedance in terms of standard impedance and ratio arms may be made in the same way described previously for resistance, simply by substituting *z*'s for *r*'s in the formula.

At low (audio) frequencies, an a-c bridge also may be used to measure resistance in the manner described for the d-c bridge, with the same *r* formula being used.

1.8 BRIDGE GENERATOR AND DETECTOR REQUIREMENTS

For the most efficient bridge operation, the generator (whether a-c or d-c) must have good output-voltage regulation, and its maximum output voltage must be held low enough to prevent damage to the bridge arms. Moreover, if the generator is of the d-c type, its output must be free from ripple. If the generator is of the a-c type, its output should exhibit a constant frequency and low distortion.

The detector (whether a-c or d-c) should have provision for adjusting the response, so that its sensitivity may be increased as null is approached, thereby increasing the closeness to which the bridge may be set. Closeness of setting is enhanced also by a very high detector input resistance (impedance) with respect to that of the bridge arms. To sharpen the null response of the bridge, the detector must be sharply tuned to the fundamental frequency if the generator output has significant harmonic content.

In a-c bridges, performance is improved by using an isolating transformer between the generator and bridge, and between the bridge and detector. For best performance, these transformers must be well shielded internally.

For a given combination of resistances or impedances in the arms, the sensitivity of a bridge may be increased by raising the generator voltage (*bridge signal*). However, there is a safe limit beyond which the bridge current becomes excessive and damages the arms. It is for this reason that moderate generator output often is preferred, and the detector sensitivity is increased proportionately, as through a-c or d-c amplification. But some amplifiers tend to become unstable and susceptible to stray pickup when operated at high sensitivity, so a compromise generally must be reached between generator voltage and detector sensitivity.

Maximum sensitivity with a given fixed-output generator and fixed-sensitivity detector is obtained when in the bridge circuit $r_1 = r_2 = r_3 = r_4 = r_g = r_d$ in the d-c bridge, and when $z_1 = z_2 = z_3 = z_4 = z_g = z_d$ in the a-c bridge. (The symbol r_g or z_g is the generator output resistance or impedance; r_d or z_d is the detector input resistance or impedance.)

CHAPTER 2

Resistance Bridges

Although bridges are used today to measure many different quantities, such as resistance, capacitance, inductance, and frequency, the original bridge was strictly a resistance-measuring device. At present, resistance measurement remains a major function of the bridge. Modern resistance bridges as a group cover the range from 0.01 milliohm to 1000 teraohms—a spread of 10^{20} to 1; and, depending on make, model, and technique, their accuracy can be as close as 0.0001 percent of the indicated reistance value.

Described below are representative resistance bridges from the rudimentary slide-wire type to more complicated varieties.

2.1 BASIC SLIDE-WIRE BRIDGE

Fig. 2-1 shows the most rudimentary resistance bridge circuit. In this arrangement, the variable balancing resistor is a single strand of resistance wire (the *slide wire*) tautly stretched between points A and B (or wound around a form having a circular cross section) and provided with a sliding contact (the *slider*). The wire is of uniform cross section and purity, so its resistance is directly proportional to its length.

As the slider is moved along the wire, it divides the latter into two parts: d_1, the length from point A to the slider, which has a resistance of R_3; and d_2, the length from point B to the slider, which has a resistance of R_4. Thus, the resistance increases on one side of the moving slider and decreases on the other side.

The bridge is composed of unknown resistance R1 (connected to terminals X-X), standard resistance R2, and the two sections R_3 and

R_4) of the slide-wire variable resistor. Battery B is the generator, and the center-zero d-c galvanometer, M, is the detector. The on-off switch, S2 permits disconnection of the battery when the bridge is idle. The pushbutton switch, S1, allows the galvanometer to be cut into the circuit momentarily to check the state of balance, thus protecting the galvanometer from continuous exposure to excessive unbalance current.

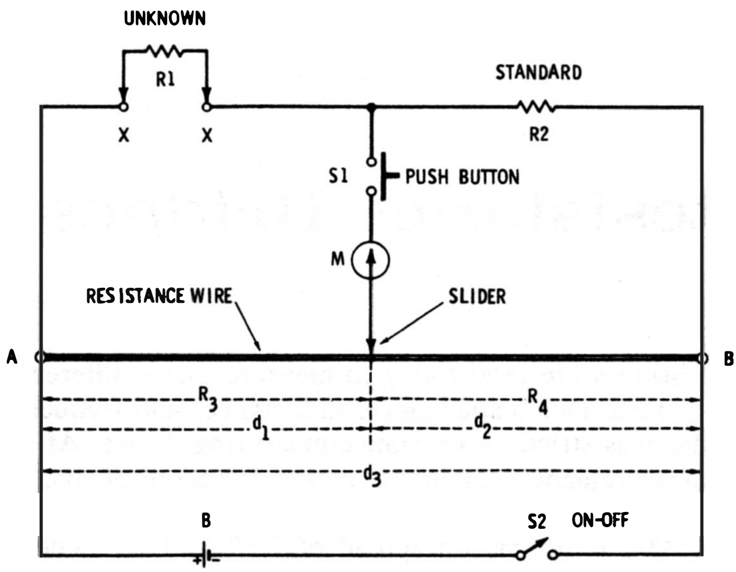


Fig. 2-1. Basic slide-wire resistance bridge.

With the unknown resistance (R1) connected to terminals X-X, the bridge is balanced by moving the slider along the wire until zero deflection is shown on the galvanometer. At this null, $R1/R2 = R_3/R_4$, and from this relationship the unknown may be determined in terms of the standard:

$$R1 = R2(R_3/R_4) \qquad 2\text{-}1$$

From this, it is clear that the slide wire provides the ratio arms of the circuit.

The total resistance of the slide wire is unimportant to the calculation. So also are the two resistance values on each side of the slider. Distances d_1 and d_2 may be measured in inches or centimeters and used in the calculation in place of the actual resistances R_3 and R_4. Thus:

$$R1 = R2(d_1/d_2) \qquad 2\text{-}2$$

For this reason, the basic slide-wire bridge is convenient for emergency measurements of resistance, since it requires only a standard resistor and a length of bare resistance wire stretched along a meter stick or yardstick, in addition to a battery and d-c meter.

Sometimes it is more convenient to read the position of the slider with respect to the total length (d_3) of the wire than to measure d_1 and d_2 separately. In such an instance:

$$R1 = R2[d_1/(d_3 - d_1)] \qquad 2\text{-}3$$

It is apparent from either equation 2-2 or 2-3 that the unknown (R1) is equal to the standard (R2) when null occurs with the slider halfway between A and B (i.e., $d_1 = d_2$, and $d_1/d_2 = 1$). Also, the slider must move to the right of center when R1 > R2, and to the left of center when R1 < R2. In practice, a meter stick or similar linear scale is usually mounted under the wire for reading d_1 and d_2 from the position of the slider.

If the wire is long (say, 1 meter), little difficulty is experienced in measuring resistance over the range 0.01R2 to 100R2, provided a sensitive galvanometer is used. Unless the wire has reasonably high resistance, however, the current may heat it and impair the accuracy of measurement.

2.2 CAREY-FOSTER RESISTANCE BRIDGE

The Carey-Foster circuit (see Fig. 2-2) is a special version of the conventional slide-wire bridge. In this circuit, the slide wire is connected between the unknown (R_x) and the standard (R_s).

The bridge is balanced in three steps: (1) Arms R1 and R2 are selected for an approximate null (i.e., $R1/R2 \cong R_x/R_s$). (2) The null is sharpened next by moving the slider over the length d_1. (3) The

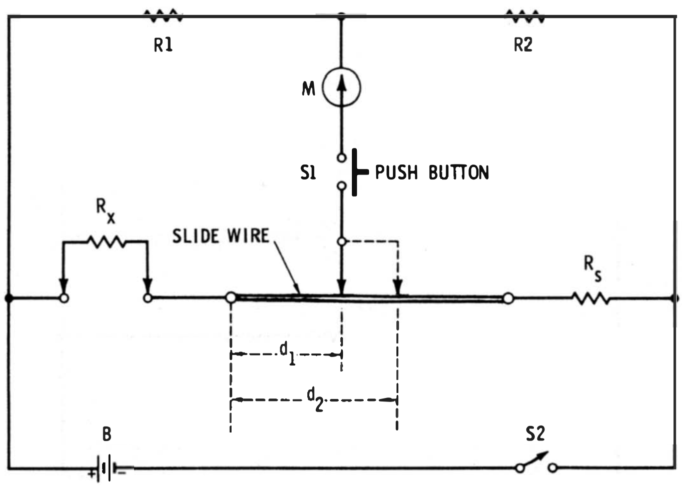


Fig. 2-2. Carey-Foster resistance bridge.

positions of R_x and R_s then are interchanged in the circuit, and the slide wire readjusted over length d_2 for a final null. The unknown resistance then is calculated:

$$R_x = R_s - (d_1 - d_2)r \qquad 2\text{-}4$$

where,

d_1 and d_2 are the lengths along the slide-wire,
r is the *resistance per unit length* of the wire (d_1 and d_2 being in this same unit of length).

Note that R1 and R2 do not appear in the balance equation. Also note that d_2 will often be larger than d_1, and that the laws of algebra must be carefully observed in these cases.

The Carey-Foster bridge is advantageous for accurately measuring an unknown resistance which is close in value to a standard resistance—or, similarly, for comparing or matching two resistances of nearly equal value.

2.3 POTENTIOMETER-TYPE SLIDE-WIRE BRIDGE

The modern slide-wire bridge substitutes a potentiometer for the single-strand slide wire of the basic circuit. Otherwise, the circuit is unchanged. In Fig. 2-3, for example, potentiometer R3 is the "slide-wire" element, and its slider divides the total resistance R3 into the bridge ratio arms r_3 and r_4.

As before, at null the unknown (R_x) is determined in terms of the standard:

$$R_x = R_s(r_3/r_4) \qquad 2\text{-}5$$

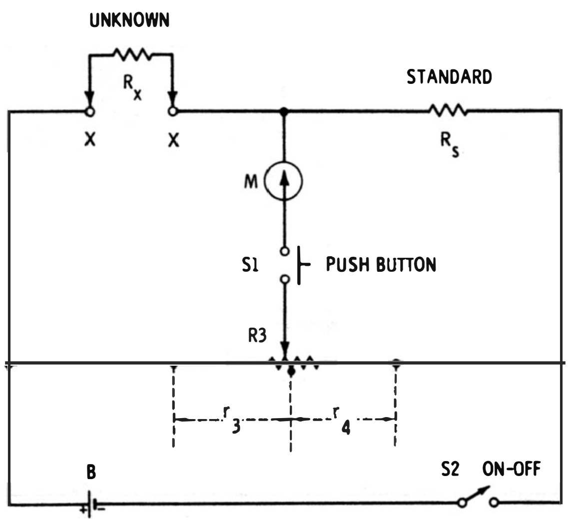


Fig. 2-3. Potentiometer-type slide-wire resistance bridge.

A dial reading directly in ohms, based upon equation 2-1, usually is attached to the potentiometer and calibrated by means of a number of accurately known resistors connected successively to the circuit in place of R_x. When, instead, the dial reads the resistance *setting* (r_3) of the potentiometer (as is usual with modern 10-turn potentiometers), $r_4 = R3 - r_3$, and the unknown resistance is determined from a modification of equation 2-1:

$$R_x = [r_3/(R3 - r_3)]R_s \qquad \text{2-6}$$

where,

R_x is the unknown resistance,
R_s is the standard resistance,
R3 is the total resistance of the potentiometer,
r_3 is the resistance setting of the potentiometer at null.

2.4 PRACTICAL MULTIRANGE SLIDE-WIRE RESISTANCE BRIDGE

Fig. 2-4 shows the circuit of a practical potentiometer-type slide-wire bridge for measuring resistance in seven ranges from 0.1 ohm to 10 megohms. The 10,000-ohm potentiometer, R8, is the balance control, and the dial attached to it reads 0.1 ohm to 10 ohms, with 1 ohm at center scale. This basic range is multiplied by the setting of range switch S3, which selects the appropriate standard resistor (R1 to R7). The unknown resistance is connected to terminals X-X.

The unbalance deflection of d-c microammeter M decreases as the standard resistance increases (i.e., as S3 is moved from the lowest

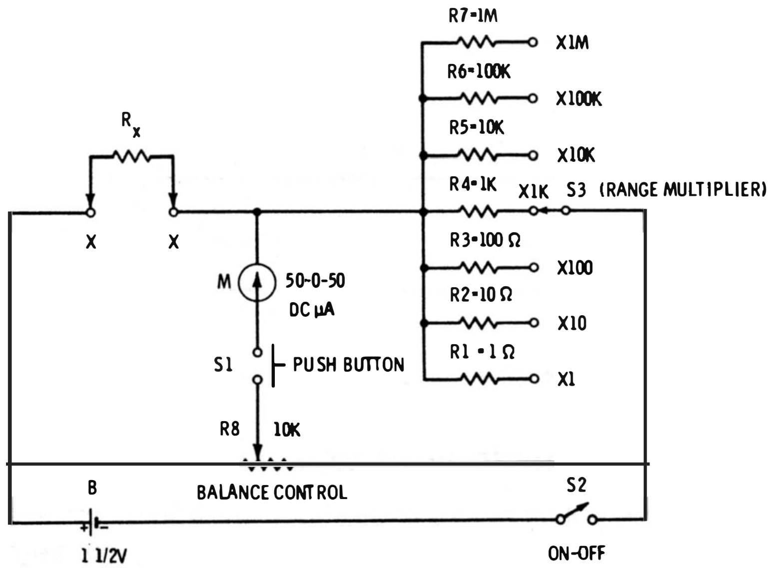


Fig. 2-4. Practical slide-wire resistance bridge.

toward the highest standard resistor), becoming only 1.5 μa in the 1-megohm (×1M) range and difficult to read. However, on the higher ranges (×10K to ×1M), this may be overcome by increasing the voltage of battery B to 6 volts. If a d-c vtvm set to its 1½-volt range is used as the null detector, no difficulty of this kind is experienced, since on the ×1M range of the bridge, the unbalance deflection is over 0.5 volt.

The dial of potentiometer R8 need be calibrated on only one range (say, the ×1K range) by successively connecting a number of accurately known resistors in that range to unknown terminals X-X and successively balancing the bridge. However, the dial should be graduated from 0.1 ohm to 10 ohms—the basic range. If all of the standard resistors are accurate (1% or better), each range will track with the dial calibration.

The slide-wire bridge has the advantage of simplicity, since it requires a minimum of parts. However, with a single-turn potentiometer of commercial grade, its resistance coverage with any given standard resistor (R_s) is restricted, for accurate reading, to 100:1 (i.e., from $0.1R_s$ to $10R_s$—thus, 0.1-10 ohms, 10-1000 ohms, 1K-100K, etc.). Also, the divisions tend to become crowded in some parts of the dial.

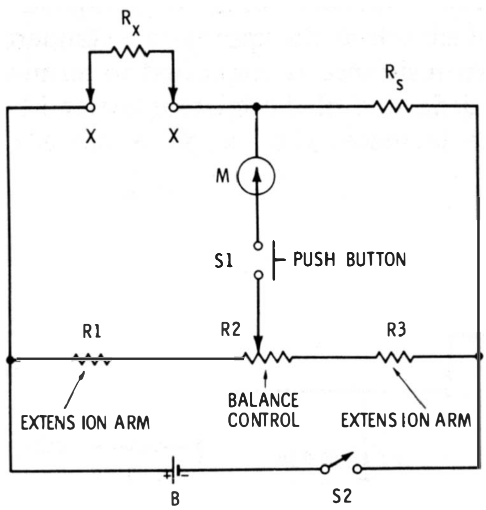


Fig. 2-5. Slide-wire bridge with extension arms.

2.5 SLIDE-WIRE BRIDGE WITH EXTENSION ARMS

In the typical potentiometer-type slide-wire bridge, the ends of the potentiometer are wasted, since the $0.1R_s$ and $10R_s$ null points occur above the low end and below the high end, respectively, of the potentiometer winding.

Extension arms (R1 and R3 in Fig. 2-5)sometimes are used to correct this defect by spreading the useful range of the bridge over the entire potentiometer winding. These arms are limiting resistances chosen in value to place the low and high ends of the measurement range at the extreme settings of the potentiometer.

Although a single standard resistor (R_s) is shown in Fig. 2-5, a set of switched standards may be used in this position, as shown previously in Fig. 2-4.

2.6 BASIC WHEATSTONE BRIDGE

Many of the disadvantages of the slide-wire bridge are resolved by the modern Wheatstone bridge. This is the classic bridge for general-purpose resistance measurement.

Fig. 2-6 shows the basic Wheatstone circuit. In this arrangement, a rheostat, R3, has been substituted for the potentiometer of the slide-wire bridge. A dial attached to R3 reads directly in the resistance setting of this rheostat. Resistors R1 and R2 are the ratio arms of the bridge. The unknown resistance, R_x, is connected to terminals X-X.

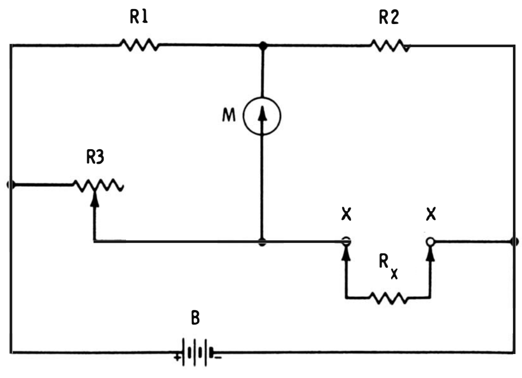


Fig. 2-6. Basic Wheatstone bridge.

At null, $R_x/R3 = R2/R1$, from which:

$$R_x = R3(R2/R1) \qquad 2\text{-}7$$

Thus, the setting of rheostat R3 simply is multiplied by the bridge ratio (R2/R1) to determine the value of R_x. In practice, various values of R1 and R2 are switched into the circuit to provide standard multipliers from ×0.001 to ×1000.

The total resistance of the rheostat usually is 10,000 ohms, but other values, such as 1000 or 5000 ohms, sometimes are used. A logarithmic taper for the rheostat winding provides a dial having uniform spacing throughout its range.

2.7 PRACTICAL WHEATSTONE BRIDGE

Fig. 2-7 shows a typical circuit of a practical bridge. Here, the rheostat has been replaced with a set of resistance decades (R9 to R48). This type of balance control allows closer settings and readings than are possible with a dial-calibrated rheostat; it covers the range from 1 ohm to 11111 ohms in 1-ohm steps.

The bridge ratio is established by resistor R8 in combination with any resistor in the R1-R7 group selected by means of switch S1. At null, the resistance setting of the decades is multiplied by this ratio to obtain the unknown resistance, R_x. The multipliers provided are 0.001, 0.01, 0.1, 1, 10, and 1000; and these multipliers applied to

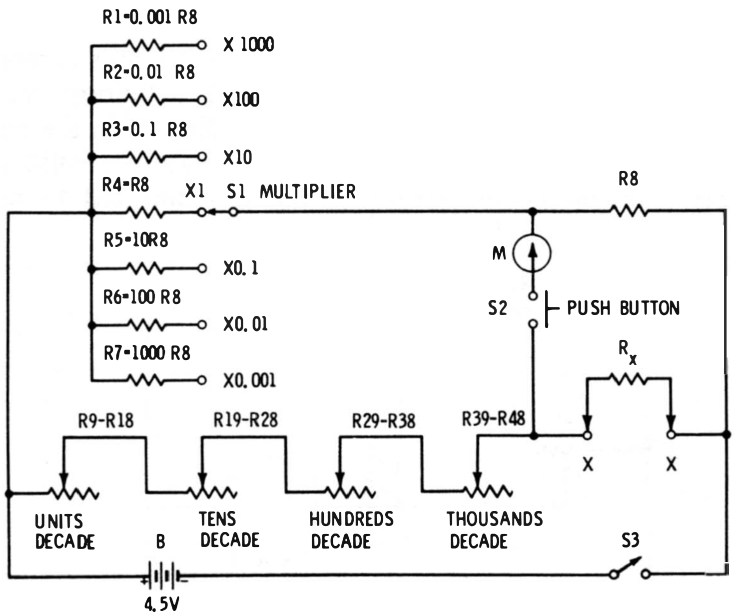


Fig. 2-7. Practical Wheatstone bridge.

the full resistance range of the decades (1-11,111 ohms) gives the bridge a measurement range of 0.001 ohm to 11.111 megohms.

Various commercial bridges use other methods of ratio-arm switching than that shown in Fig. 2-7. In some instruments, for example, the ratio resistors are switched in pairs, rather than against a single resistor such as R8.

Fig. 2-8 shows a portable bridge of the dial-operated decade type. This is a completely self-contained model, having an internal battery and permanent galvanometer. Fig. 2-9 shows a laboratory-type bridge. In this latter model, which requires an external battery and galvanometer, the decades are plug type instead of the dial type of the previous model. The five left-hand decades constitute the balance control, while the two right-hand ones form the ratio arms. Fig. 2-10 shows a Wheatstone bridge which has a special guarded construction to eliminate errors due to leakage current, such as that arising from humidity. In this model, each dial of the rheostat arm is behind the

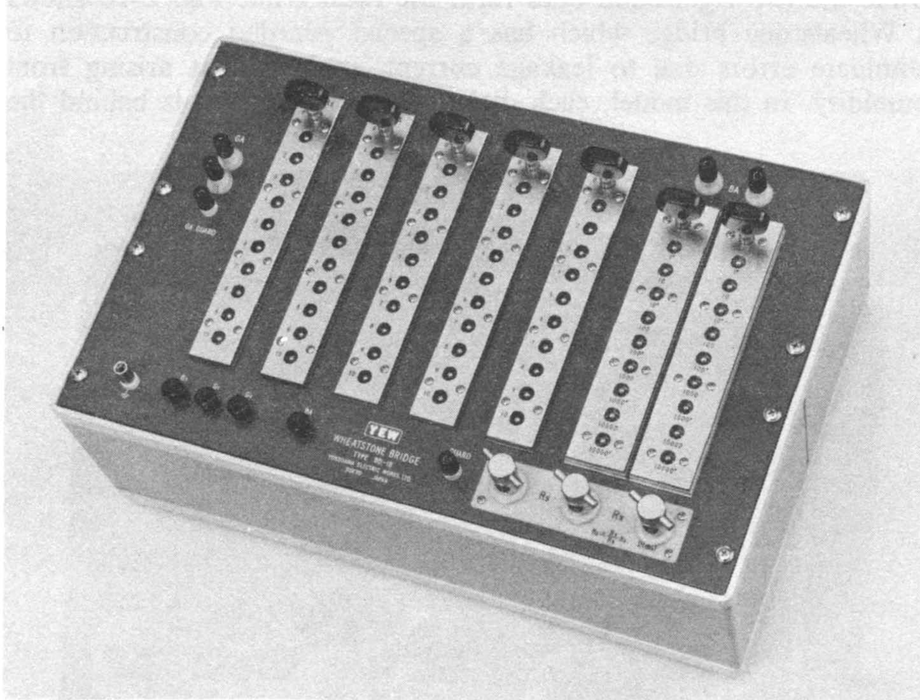
Courtesy Yokogawa Electric Works, Inc.

Fig. 2-8. Dial-type Wheatstone bridge.

panel, so that only the figure indicating the resistance setting shows through a dial window. The multiplier (ratio arms) dial is seen at top center.

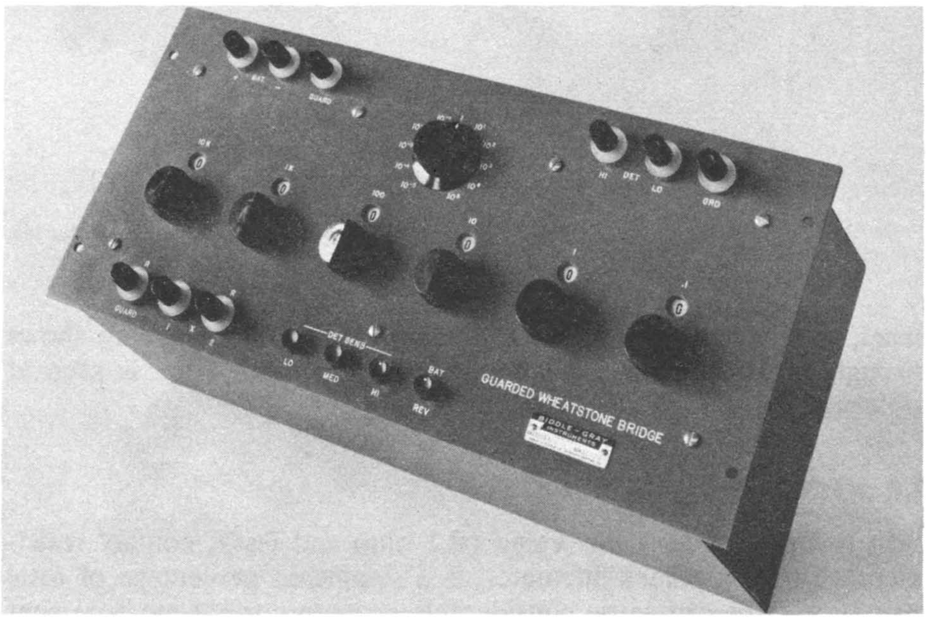
2.8 KELVIN DOUBLE BRIDGE

In resistors of very low value (0.1 ohm and less), contact resistance, which sometimes fluctuates, is a significant percentage of total resistance. This can cause considerable error in current measurement if, for example, an ammeter is connected directly to the current-carrying terminals of an ammeter shunt. To reduce such error, some



Courtesy Yokogawa Electric Works, Inc.

Fig. 2-9. Plug-type Wheatstone bridge.



Courtesy James G. Biddle Co.

Fig. 2-10. Guarded Wheatstone bridge for close measurements.

low-value resistors (such as ammeter shunts) are supplied as four-terminal devices which cannot be measured with a conventional
The Kelvin double bridge (see Fig. 2-11) accommodates low-value, four-terminal resistors. In this circuit, the unknown (R_x) and standard (R_s) both are such resistors. On unknown resistor R_x, c_1 and c_2 are the current terminals, and p_1 and p_2 the potential

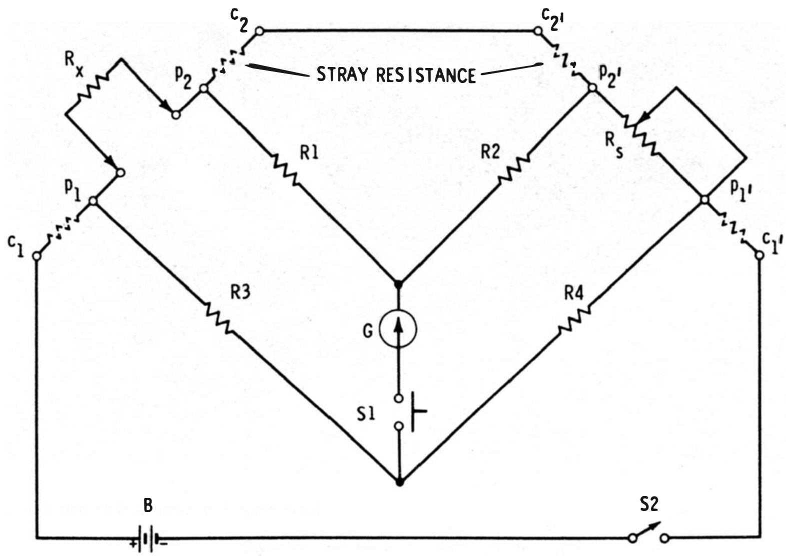


Fig. 2-11. Kelvin double bridge.

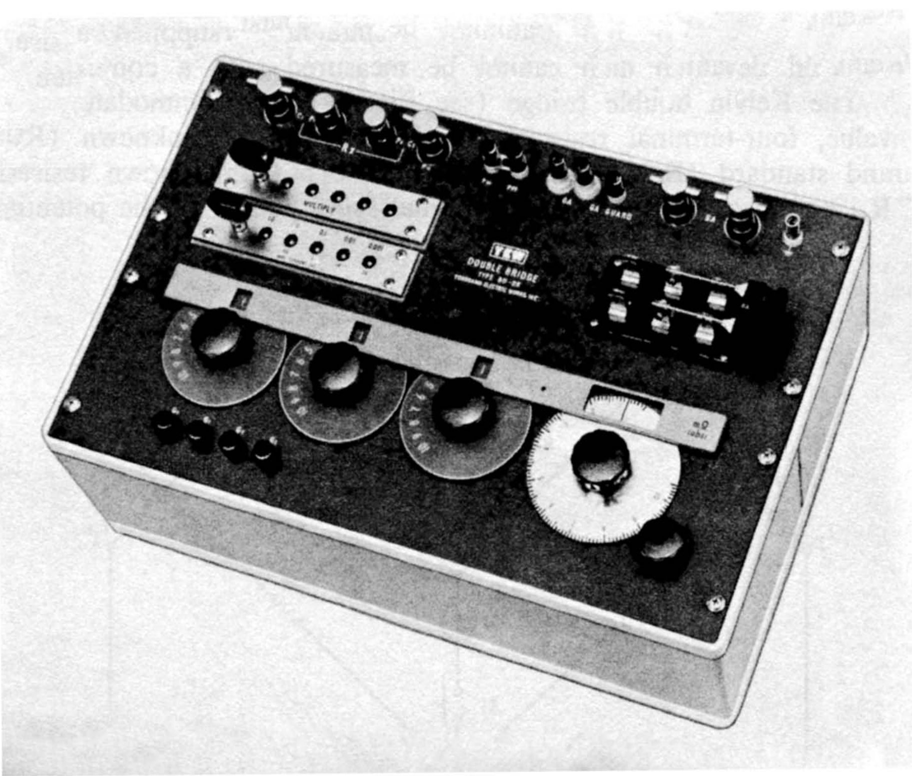
terminals. Similarly, on standard resistor R_s (which is the internal, low-resistance balance control), c_1' and c_2' are the current terminals, and p_1' and p_2' are the potential terminals.

Resistor combinations R1-R2 and R3-R4 provide two pairs of ratio arms, giving the circuit the double-bridge configuration from which its name comes. If these resistances are adjusted so that the ratio arms are similar (i.e., R1/R2 = R3/R4), then at null:

$$R_x = R_s(R1/R2) \quad \text{2-8}$$

The auxiliary ratio arms, R1 and R2, balance out the error due to stray resistance from p_2 to p_2'.

Fig. 2-12 shows a Kelvin double bridge for use with an external battery and galvanometer. Note from this photograph that four terminals are provided in the upper left-hand corner of the panel for connection of the unknown resistor to the bridge. Bridges of this type make possible the accurate measurement of resistance in the milliohm range.



Courtesy Yokogawa Electric Works, Inc.

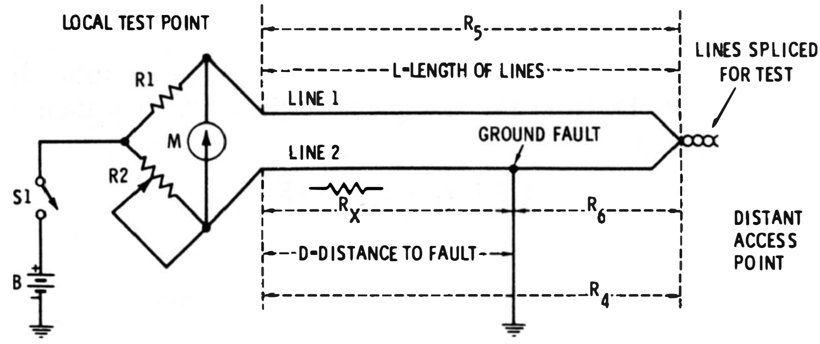
Fig. 2-12. Kelvin double bridge.

2.9 MURRAY LOOP

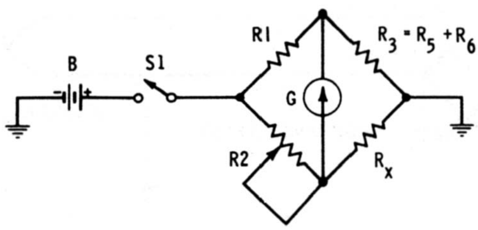
The Murray loop is an adaptation of the Wheatstone bridge for finding the distance from a test point to a remote ground fault on a line. It is particularly useful in telephone work, but it may be used wherever an extra line is available along the same route as the grounded one.

Fig. 2-13A shows the circuit. Here, line 1 is the spare, "clean" line, and line 2 the grounded one. For the test, the two lines are spliced together, by twisting or cord patching, at a point (such as a distant station) beyond the ground. This forms the loop from which the device takes its name. The four bridge arms are (1) resistor R1; (2) resistor R2; (3) the resistance R_3 of line 1 from the test point, around the loop, to the fault; and (4) resistance R_x of line 2 from the test point out to the fault. When R2 is adjusted for null, $R1/R_3 = R2/R_x$. Fig. 2-13B shows the equivalent bridge circuit.

The resistance per unit length of the wire in the lines must be known beforehand, as must also the length from the test point to the distant splice point. The resistance of line 1 then is $R_5 = Lr_1$,



(A) Loop circuit.



(B) Equivalent bridge.

Fig. 2-13. Murray loop.

where L is the line length and r_1 the resistance per unit length of this line. (If L is in miles, r_1 is expressed in ohms per mile; if L is in feet, r_1 is in ohms per foot.) Similarly, the resistance R_4 of the entire length of line 2 is equal to Lr_2, where r_2 is the resistance per unit length of line 2. If both lines are of the same diameter and material, $R_5 = R_4$.

From the above relationships, the unknown distance, D, from the test point to the fault may be determined. At null:

$$D = \frac{LR2(r_1 + r_2)}{r_2(R1 + R2)} \qquad 2\text{-}9$$

If the two lines have the same resistance per unit length, $r_1 = r_2$, and the equation may be simplified:

$$D = \frac{2L(R2)}{R1 + R2} \qquad 2\text{-}10$$

2.10 VARLEY LOOP

The Varley loop is similar to the Murray loop in function and application. In the Varley circuit (Fig. 2-14), however, the ratio arms are kept constant and the bridge balanced by adjusting an extra arm, R4.

With this circuit, as with the Murray loop, the resistance per unit length, r_1 and r_2, of line 1 and line 2, respectively, must be known beforehand. From these values and length L of both lines, the unknown distance, D, from the test point to the fault may then be determined. At null:

$$D = \frac{LR2(r_1 + r_2) - (R1R4)}{r_2(R1 + R2)} \qquad 2\text{-}11$$

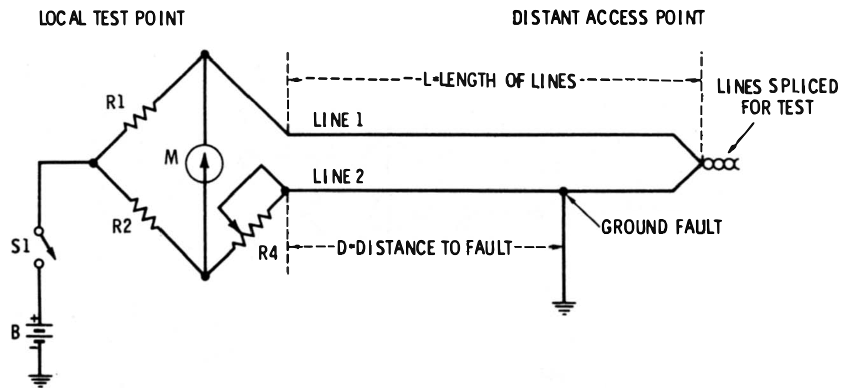


Fig. 2-14. Varley loop.

As in the Murray loop, if the two lines are of the same resistance, per unit length, $r_1 = r_2 = r$, and the equation may be simplified:

$$D = \frac{(2LR2r) - (R1R4)}{r(R1 + R2)} \qquad 2\text{-}12$$

Unlike the Murray loop, the Varley loop (owing to the presence of a third, lumped bridge arm—R4) can be used to measure the total resistance of the loop if the lower terminal of the battery is transferred from ground to the junction of R4 and line 2. Some Wheatstone bridges are equipped with a panel switch for quickly selecting either standard bridge operation or Varley loop operation.

2.11 UNIVERSAL GALVANOMETER SHUNT

A sensitive d-c galvanometer may easily be damaged by the high unbalance current of a bridge. To prevent this, some means usually is provided—internally or externally—for shunting the galvanometer to decrease its sensitivity and for progressively removing the shunt (increasing the sensitivity) as null is approached.

One such accessory for varying the galvanometer sensitivity is the Ayrton-Mather universal shunt, shown in Fig. 2-15. In this arrangement, which is a kind of attenuator, resistors R1 to R4 (totaling resistance R_t) may be separate resistors, as shown, or taps on a single resistor R_t. By means of tap switch S, which selects these resistors, the sensitivity of the galvanometer may be varied at will. In the SHORT position of this switch, the galvanometer has maximum protection.

The total resistance, R_t, is chosen to equal the critical damping resistance of the galvanometer. The damping of the galvanometer thus

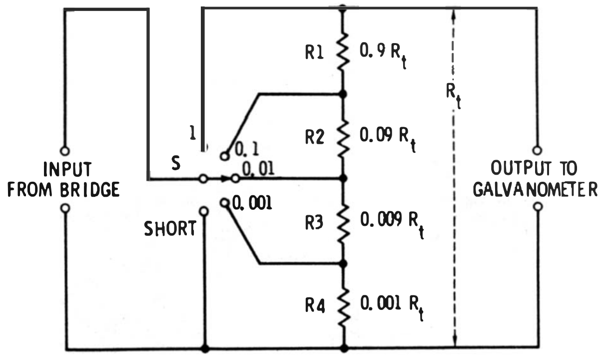


Fig. 2-15. Universal galvanometer shunt.

is constant at all settings of switch S, since a constant resistance (R_t) is always across the galvanometer. A typical value of R_t in commercial shunts is 1000 ohms, $R1 = 900\Omega$, $R2 = 90\Omega$, $R3 = 9\Omega$ and $R4 = 1\Omega$.

2.12 MEGOHM BRIDGE

Very high resistances (several megohms to several teraohms) may be measured with special bridges. While the coverage of a standard bridge may be extended into the megohm range, the very-high-resistance standards required for such an extension are usually less accurate than lower-resistance ones, and the extreme division of voltage between at least two of the bridge arms demands a detector so sensitive as to be unstable and susceptible to noise. For these and other reasons, a special megohm bridge is advisable.

Fig. 2-16 shows the basic circuit of a megohm bridge typified by the General Radio Type 544-B. This arrangement is seen to be similar to the conventional Wheatstone bridge except for guard terminals and paths (to minimize the effects of surface resistance of the unknown and standard resistors and of leakage between the unknown terminals and from terminals to ground) and the inclusion of a sensitive d-c vtvm type of null detector (triode V1) in the bridge.

The dial of balance-control rheostat R2 is direct reading in megohms, and standard resistors R3 and R7 are switched by means of S1 to change range (i.e., to multiply the reading of R2). In the General Radio bridge, these standard resistors extend from 10K to 100M; R1 is 100K; and R2, 12K.

The d-c input voltage often is high (say, 500 volts) to permit testing R_x under actual working conditions and to provide an off-balance output voltage high enough to ensure accurate null adjustment.

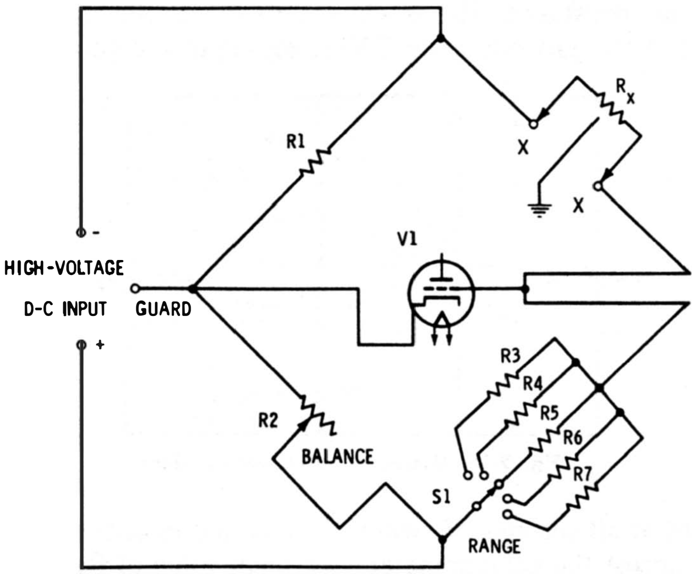


Fig. 2-16. Megohm bridge.

2.13 A-C BRIDGE MEASUREMENT OF RESISTANCE

Where desired, a resistance bridge may be used with an a-c generator and an a-c detector. One reason for choosing the a-c procedure is the easily obtained high detector sensitivity in such high-input-impedance indicators as vacuum-tube millivoltmeters, amplifier-type meters, and oscilloscopes.

From Fig. 2-17, the bridge is seen to be conventional. While, for simplicity, a meter (M) is shown as the null detector in this circuit, the detector may be any one of the devices mentioned earlier. The generator usually is an oscillator (conventional test frequencies are 120, 400, 500, and 1000 Hz). For a sharp null, either the bridge signal must be harmonic-free or the detector must be sharply tuned to the fundamental frequency of the signal.

Problems of isolation and strays are more pronounced here than in d-c bridge circuits, and are solved in various ways. For example, well-shielded isolating transformers (T1 and T2) are inserted between the generator and bridge, and between the bridge and detector. (In some instances, one of these transformers—often T2—is omitted.) In addition, either point X or point Y (but not both) may be grounded. The shortest practicable leads (preferably shielded) must be employed between generator, bridge, and detector. Addi-

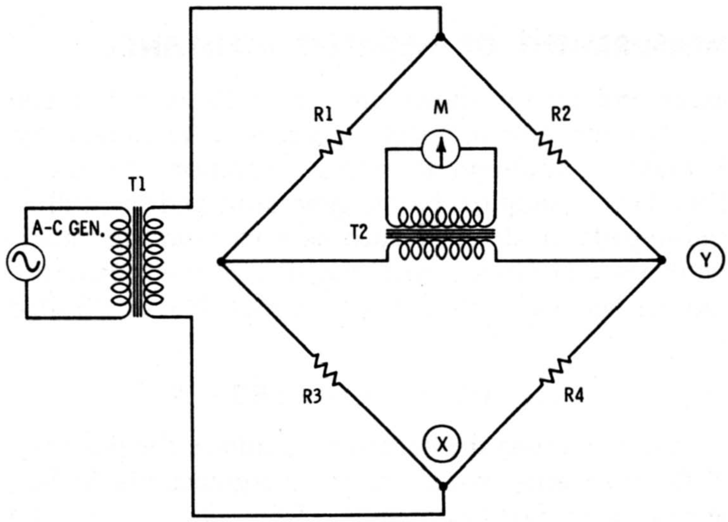


Fig. 2-17. A-c resistance bridge.

tionally, efficient high-frequency operation (20 kHz and beyond) requires that each bridge arm be individually shielded. Reactive components in the bridge arms must be minimal; otherwise, the bridge will be balanced for complex impedance instead of simple resistance. (Standards, ratio arms, and balance control therefore are designed for extremely low inductance and capacitance.) The harmful effects of stray reactance are most pronounced at high frequencies and high resistance values.

A standard Wheatstone bridge may be adapted for a-c measurement of resistance by replacing its battery and galvanometer with an a-c source (of suitable voltage output) and an a-c detector. Isolating transformers enhance the accuracy of measurement and should be used unless they are already provided in the generator and detector.

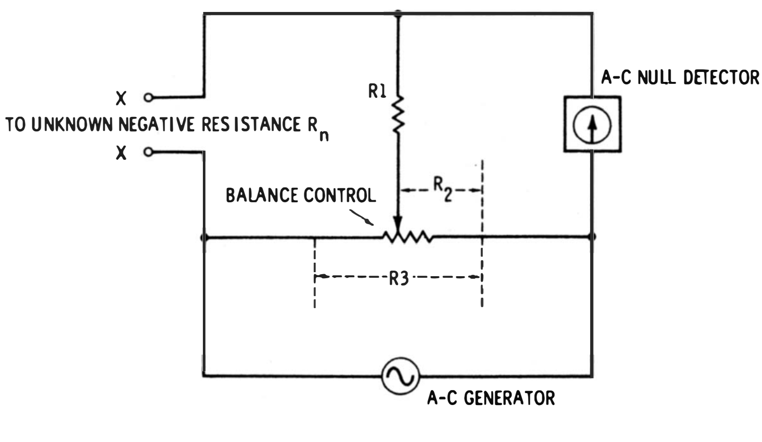


Fig. 2-18. Negative-resistance bridge.

2.14 MEASUREMENT OF NEGATIVE RESISTANCE

Although the circuit shown in Fig. 2-18 is not a conventional bridge, it permits measurement of negative resistance by the null method which is common to bridge operation. An a-c test signal (e.g., 1000 Hz) is supplied by the generator, and the null is indicated by a high-impedance detector such as a vacuum-tube millivoltmeter.

The unknown negative resistance, R_n, is connected to terminals X-X, and the balance-control potentiometer, R3, is adjusted for null. At null:

$$R_n = R1(R3/R_2) + (R3 - R_2) \qquad 2\text{-}13$$

To prevent swamping the negative resistance, the positive resistance (R_t) of the measuring circuit must be significantly higher than the negative resistance. For best results, $(R3 - R_2) + 1/R1 + [1/(R_2 + R_d)]$ must be much higher than $10R_n$.

CHAPTER 3

Capacitance Bridges

The capacitance bridge is similar to the resistance bridge in configuration and operating principle, the chief difference being the presence of capacitors in two or more of the bridge arms. Unlike the resistance bridge, the capacitance bridge is always an a-c device.

For a long time, the capacitance bridge was the only means of measuring capacitance accurately and quickly without calculations. As a group, capacitance bridges cover a range extending from a few tenths of a picofarad to several thousand microfarads.

The descriptions in this chapter progress from the simple slide-wire type of bridge through the more complex types which are in common use. Similarities between these bridges and their resistance-measuring counterparts will be readily apparent.

3.1 BASIC SLIDE-WIRE CAPACITANCE BRIDGE

Fig. 3-1 shows the most rudimentary type of capacitance bridge, the basic slide-wire type. This circuit is seen to be similar to the basic slide-wire resistance bridge described in Chapter 2. Here, C_x is the unknown capacitance, C_s the standard capacitance; the balancing device is the slide wire—a single strand of resistance wire of uniform cross section, stretched taut between terminals A and B or wound around a form having a circular cross section, and provided with a sliding contact. An a-c generator (e.g., an audio oscillator) and an a-c detector (e.g., vtvm or oscilloscope) are connected to the GEN and DET terminals, respectively. Transformer coupling is advisable.

The circuit is balanced by moving the slider between A and B to locate the null. At null:

$$C_x = C_s(R_2/R_1) \qquad 3\text{-}1$$

Thus, the unknown capacitance is found by multiplying the standard capacitance by the resistance ratio established by the position of the slider along the wire. Note that the resistance fraction in equation 3-1 is inverted with respect to its counterpart in the slide-wire resistance-bridge equation 2-1, Chapter 2. This results from the fact that in reality the *reactance*, not the capacitance, of C_x is balanced against the reactance of C_s (i.e., $X_{c_x}/X_{c_s} = R_1/R_2$). It also accounts for the fact that null points for the higher capacitances occur near point A, and those for the lower capacitances near point B (the reverse of the situation in the resistance bridge).

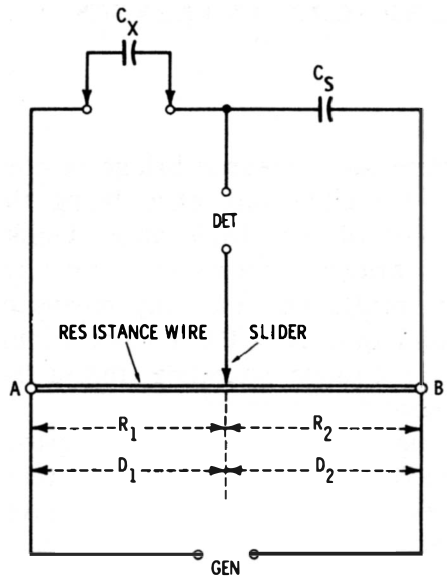


Fig. 3-1. Basic slide-wire capacitance bridge.

Neither the total resistance ($R_1 + R_2$) of the slide wire nor the individual values R_1 and R_2 need be known in order to make a C_x measurement. The calculation may be made in terms of distances D_1 and D_2 measured along the slide wire in centimeters, inches, or arbitrary linear units. Thus:

$$C_x = C_s(D_2/D_1) \qquad 3\text{-}2$$

From the balance equations, it is seen that null occurs halfway between A and B when the unknown capacitance equals the standard capacitance, to the left of this center point when $C_x > C_s$, and to the right of center when $C_x < C_s$. To change the capacitance range, one needs only to change the value of the standard capacitance, C_s.

Like the slide-wire resistance bridge, this type of capacitance bridge has the advantages of simplicity and relative economy. It requires few components and needs no capacitance or resistance calibration if equation 3-2 is used. And it is convenient for emergency measurements, as it requires only a standard capacitor, a length of resistance wire (stretched over a meter stick, yardstick, or other linear scale), and a generator and detector—all usually available in the laboratory or shop. A disadvantage is the failure of the circuit to give a *complete* null (absolute zero balance) unless the losses in the unknown capacitor equal those in the standard capacitor.

3.2 POTENTIOMETER-TYPE SLIDE-WIRE CAPACITANCE BRIDGE

A more compact version of the slide-wire capacitance bridge, like the similar version of the resistance bridge, substitutes a wire-wound potentiometer for the slide wire. Fig. 3-2 shows this arrangement.

In this circuit, the balancing potentiometer has a total resistance R3. At any setting, $R3 = R_1 + R_2$, the resistance between the contact blade and the low end, and the contact blade and the high end, respectively, of the resistance element. Resistances R_1 and R_2 determine the bridge ratio.

At null, the unknown capacitance (C_x) is determined from the standard capacitance (C_s) and the resistance ratio:

$$C_x = C_s(R_2/R_1) \qquad 3\text{-}3$$

The potentiometer dial may be calibrated to read directly in picofarads or microfarads by locating the null points for a number of known capacitances connected successively at C_x. It may also be calibrated with the aid of equation 3-3, in terms of measured R_1 and R_2 values. If the potentiometer dial already reads directly in resistance

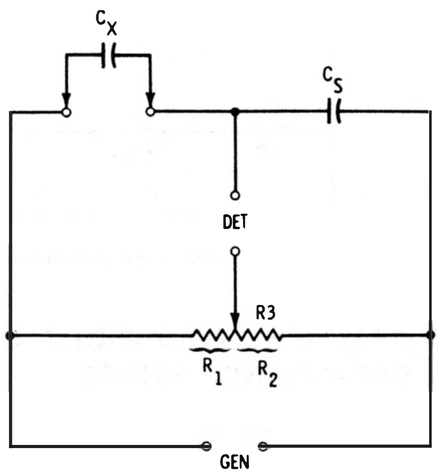


Fig. 3-2. Potentiometer-type slide-wire capacitance bridge.

setting (as the dial of a multiturn potentiometer often does), the unknown capacitance may be calculated with reference to thatsetting:

$$C_x = C_s[(R3 - R_1)/R_1] \qquad 3\text{-}4$$

where,

R_1 is the resistance setting indicated by the dial,
R3 is the total resistance of the potentiometer.

Like the basic slide-wire bridge, the potentiometer-type circuit is simple and economical. However, it has three disadvantages: (1) because of stray capacitances in the circuit, C_x values lower than 100 pf cannot be measured accurately; (2) unless the potentiometer has a special taper, such as logarithmic, the graduations will crowd at the ends of the dial, seriously impairing accuracy, and for that reason the capacitance range with any one standard capacitor C_s should not exceed 0.1 to 10 times C_s, even though the potentiometer can afford a wider range; and (3) a complete null occurs only if losses in the unknown capacitor equal those in the standard capacitor.

The degrading effect of stray capacitances in the useful ranges of the bridge may be reduced somewhat by keeping the potentiometer resistance, R_3, reasonably low—say, 5000 ohms. Also, the wasted ends of the potentiometer in the simple circuit may be resolved by means of extension arms (such as R1 and R3 in Fig. 3-3) which allow the useful 100:1 capacitance range to be spread over the entire resistance range of the potentiometer.

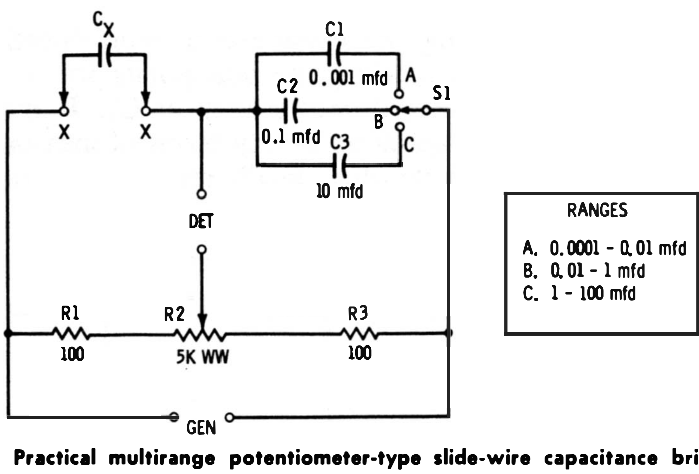


Fig. 3-3. Practical multirange potentiometer-type slide-wire capacitance bridge.

3.3 PRACTICAL POTENTIOMETER-TYPE SLIDE-WIRE CAPACITANCE BRIDGE

Fig. 3-3 shows the circuit of a practical bridge using a 5000-ohm wirewound potentiometer (R2) as the balance control. This bridge covers the range 100 pf to 100 mfd in three steps: 100 pf–0.01 mfd, 0.01–1 mfd, and 1–100 mfd. The capacitance ranges are selected by switching standard capacitors (C1, C2, C3). The 100-ohm noninductive resistors R1 and R3 provide the extension arms explained in Section 3.2.

Any convenient test frequency may be employed, up to about 5 kHz. Best stability is obtained when the generator is coupled to the bridge through a shielded transformer. It is advisable also to insert a transformer between the bridge and the detector; however, this second transformer may be omitted if the first one is used, especially if the lower DET terminal is grounded.

This bridge may be calibrated by balancing it with a number of accurately known capacitors connected successively to terminals X-X, and inscribing the potentiometer dial accordingly. Only one range need be calibrated in this manner; the others will track if accurate values are used at C1, C2, and C3. Thus, the lowest range (0.0001–0.01 mfd) may be calibrated, and the A, B, and C settings of switch S1 used to multiply this basic range by 1, 100, and 10,000 respectively.

The null balance is sharp if losses in the capacitor under test equal those in the standard capacitor (C1, C2, C3). Since high-quality capacitors customarily are used as standards, this usually means that a sharp null is obtained with low-loss test of significant losses in the capacitor under test.

3.4 THE NEED FOR POWER-FACTOR BALANCE

The preceding sections explain that a complete null is impossible with a simple slide-wire bridge unless the losses in the unknown and standard capacitors are equal. The reason for this is the complex nature of the capacitor as an impedance network (resistance and capacitance in combination, the resistance component representing the losses). When the unknown and standard capacitors are of the same kind and magnitude, and when their equivalent impedances are the same in all respects, then a complete null may be obtained. Separate balances are required for the resistive and reactive components.

It is convenient to express the relation between losses (resistive component) and capacitance (reactive component) as the *power factor* of the capacitor. Numerically, power factor $pf = R/Z$ = cosine of the phase angle of the equivalent r-c network. The highest value which power factor can reach is 1 (also expressed as 100%). A low-loss capacitor therefore has a very low power factor. Capacitor quality may be expressed also as the quality factor $Q = X_c/R$ = tangent of the phase angle, or as the dissipation factor $D = R/X_c$ = cotangent of the phase angle. A low-loss capacitor has a high Q and low D.

In somewhat more specific terms, the original statement may be expressed in the following way: *A complete null is obtained only when the power factor, Q, or D of the unknown capacitor equals that of the standard capacitor.* When these factors are unequal, as they almost always are in practice, enough resistance must be added in series (or sometimes in parallel) with the lower-loss capacitor (usually the standard) to increase its power factor (reduce its Q) and thus make it equivalent to the higher-loss capacitor. Fig. 3-4 shows this arrangement. Here, rheostat R1 adds this sort of compensating resistance to standard capacitor C_s. In use, the bridge is

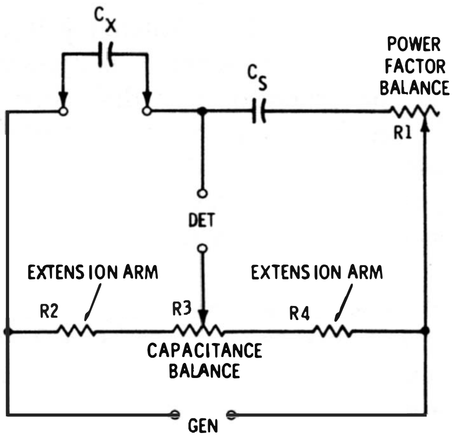


Fig. 3-4. Slide-wire capacitance bridge with power factor balance.

balanced first for capacitance by adjusting potentiometer R3 for minimum deflection of the detector. Then, rheostat R1 is adjusted to improve the null. Additional adjustments of R1 and R3, alternately, will give complete null (dip to zero). The resistance of R1 does not enter into the calculation of unknown capacitance. Nor does the resistance of R3 enter into the calculation of power factor (see the next section).

For complete null, the resistance of R1 must be adjustable to the equivalent series resistance of the unknown capacitance, C_x. A dial attached to this rheostat may be calibrated to read directly the power factor (shown either as a decimal or as a percentage); however, this dial will read correctly only at the calibration frequency. All complete capacitance bridges are equipped with such a power, Q, or D balance.

Under some circumstances, resistive balance may be obtained also by means of a rheostat in *parallel* with the standard capacitor. This method works best when the dissipation factor of the unknown capacitor is higher than 1, particularly when the unknown capacitor has a significant equivalent parallel resistance (as in electrolytic capacitors). It tends sometimes to broaden the capacitance null.

Fig. 3-5 shows how a power-factor balance rheostat is included in a universal bridge so that it may be switched in series with either the unknown or standard capacitor, whichever has the lower power factor—i.e., the lesser resistance as related to capacity. (A universal bridge is a skeleton-type instrument to which unknown *and* standard

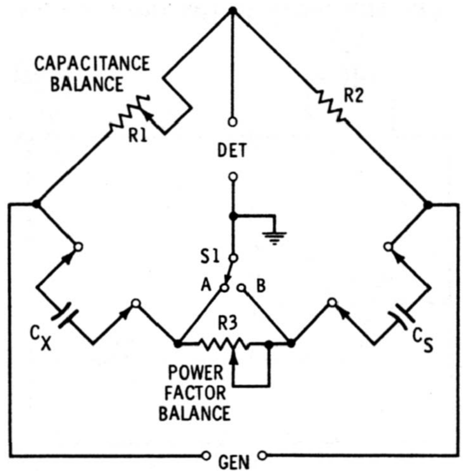


Fig. 3-5. Universal bridge with power factor balance.

resistors, capacitors, or inductors may be connected.) Such a bridge is convenient for checking an unknown (C_x) against any available capacitor (C_s) whose capacitance is known, but not necessarily its power factor. When switch S1 is in position A, the power-factor rheostat (R3) is in series with standard capacitor C_s; when S1 is in position B, rheostat R3 is in series with unknown capacitor C_x.

3.5 CAPACITANCE COMPARISON BRIDGE

The simplest and perhaps most common capacitance bridge merely substitutes capacitance arms for two of the resistance arms in the Wheatstone bridge, retains the other two resistance arms—the *ratio arms*—to complete the basic circuit, and adds a power-factor balance. Because of the similarity of the two bridges, the capacitance type is sometimes familiarly called a "Wheatstone capacitance bridge," but the term *capacitance comparison bridge* appears most often in technical literature.

The comparison bridge overcomes many of the shortcomings of the slide-wire bridge, principally the dial crowding and lower-capacitance limit of the latter instrument. Fig. 3-6 shows the basic circuit of the comparison bridge. Here, the bridge ratio (R3/R2) is established by fixed resistor R3 and the capacitance-balance rheostat, R2.

A second rheostat, R1, provides the power-factor balance. At null:

$$C_x = C_s(R3/R2) \qquad 3\text{-}5$$

The circuit of a practical multirange capacitance comparison bridge is shown in Fig. 3-7. This bridge covers the range from 10 pf to 1000 mfd in four steps: 10–1000 pf, 1000 pf–0.1 mfd, 0.1–10 mfd, and 10–1000 mfd. The similarity to the basic circuit (see Fig. 3-6) is

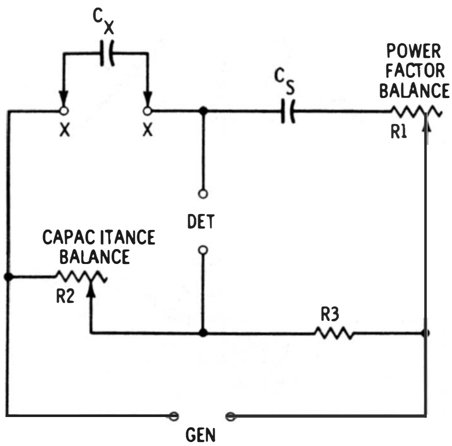


Fig. 3-6. Capacitance comparison bridge.

easily seen, the chief difference between the two circuits being the switched values of standard capacitance C1-C3 and of ratio resistors R4 and R5.

The capacitance ranges are selected by switching the standard capacitance value. The same 1-mfd standard (C3) is used both on the

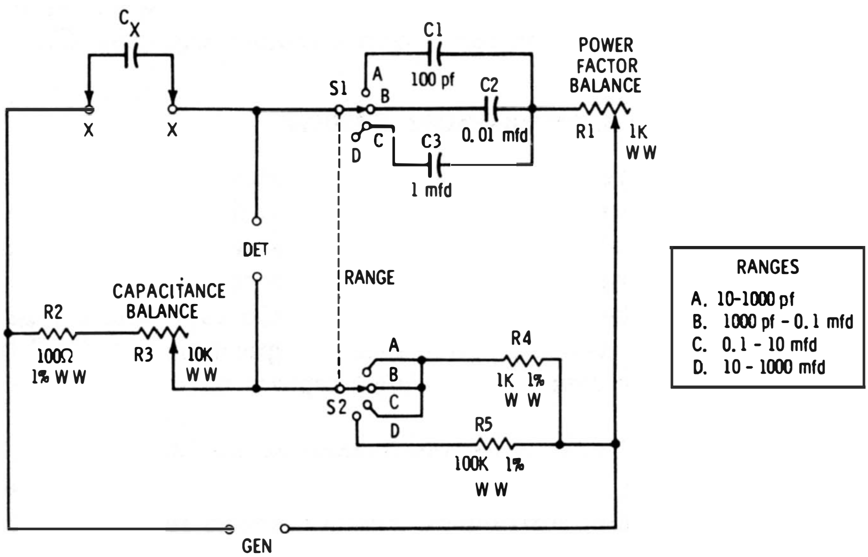


Fig. 3-7. Practical multirange capacitance comparison bridge.

0.1–10- and 10–1000-mfd ranges by switching the ratio resistor from 1000 ohms (R4) to 100K (R5) for the higher range. Otherwise, a 100-mfd standard capacitor would be needed for the higher range, and such a high-capacitance unit is hard to obtain in accurate capacitance and low power factor. The single 1000-ohm ratio resistor (R4) is used on all three low ranges.

The R3 dial may be calibrated to read directly in pf and mfd either by means of equation 3-5 or by actually balancing the bridge for a number of accurately known capacitors connected successively to terminals X-X. Only one range need thus be calibrated; if C1, C2, and C3 are accurate, the other ranges will track automatically. The R1 dial may be calibrated to read directly in percent power factor on the basis of the resistance setting of R1 at complete null:

$$pf(\%) = 0.000628fR1C_x \qquad 3\text{-}6$$

Here, f is in Hz, R1 (the power-factor rheostat setting) in ohms, and C_x (the value of the unknown capacitance under test, read from the R3 dial) in microfarads. However, the power factor is frequency dependent, so the power factor dial will be correct only at the calibration frequency (which, for that reason, should be the normal bridge-signal frequency—for example, 1000 Hz).

3.6 SCHERING BRIDGE

In the Schering bridge, the resistance balance is obtained by means of a variable capacitor in parallel with one of the ratio arms (see C1 in Fig. 3-8). This capacitor should itself have excellent power factor, so in practice it is either an air-type variable or a mica decade.

With an unknown capacitor (C_x) connected to terminals X-X, the bridge is balanced for capacitance by adjustment of rheostat R2,

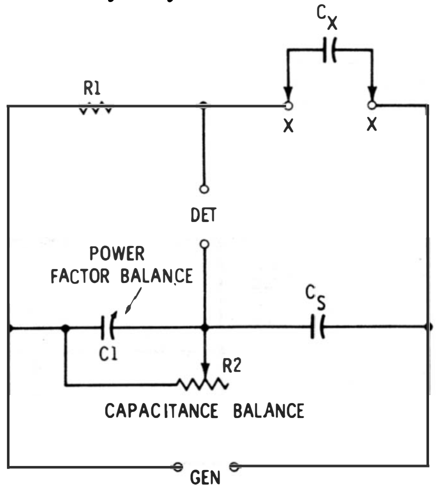


Fig. 3-8. Basic Schering bridge.

and separately for power factor by adjustment of capacitor C1. At null:

$$C_x = C_s(R2/R1) \qquad 3\text{-}7$$

The equivalent series resistance of the unknown capacitor is

$$R_x = R1(C1/C_s) \qquad 3\text{-}8$$

The Schering bridge is often used for testing electrolytic and other capacitors to which a d-c voltage must be applied simultaneously with the a-c bridge signal (See section 3.8). Practical bridges of this type have been used over a wide capacitance range (e.g., 0.1 pf to 1000 mfd). Ranges are changed by switching C_s and R1 together.

3.7 WAGNER GROUND

When the bridge arms are high impedances, stray capacitances can impair the accuracy of measurement. Such capacitances from bridge arms to ground are represented by C1 and C2 in Fig. 3-9.

The *Wagner ground,* which is incorporated into some bridges, permits these objectionable capacitances to be balanced out. This arrangement consists simply of the auxiliary potentiometer R4 and spdt switch S1. A Wagner ground may be added to a bridge which does not already have it.

In operation, the bridge is balanced first in the normal manner; i.e., with switch S1 set to position A, rheostats R1 and R2 are adjusted for the best null that can be obtained. The switch then is thrown to position B, and potentiometer R4 adjusted for null. Next, with the

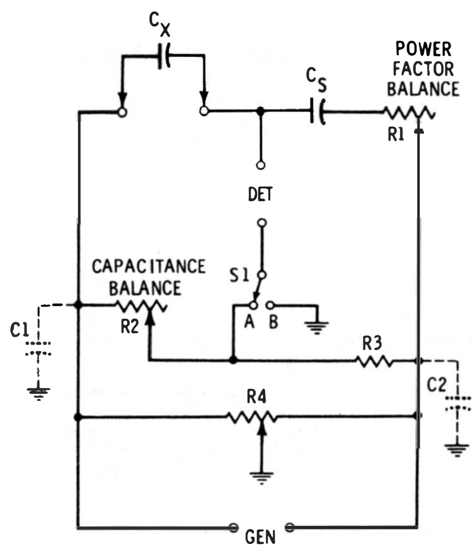
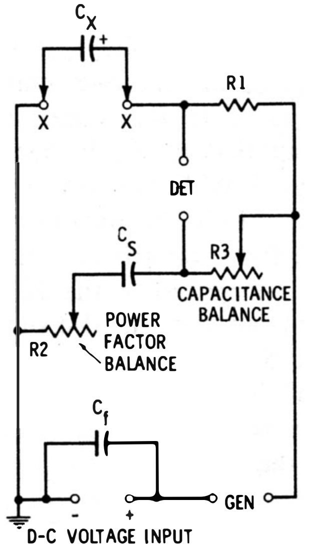


Fig. 3-9. Capacitance bridge with Wagner ground.

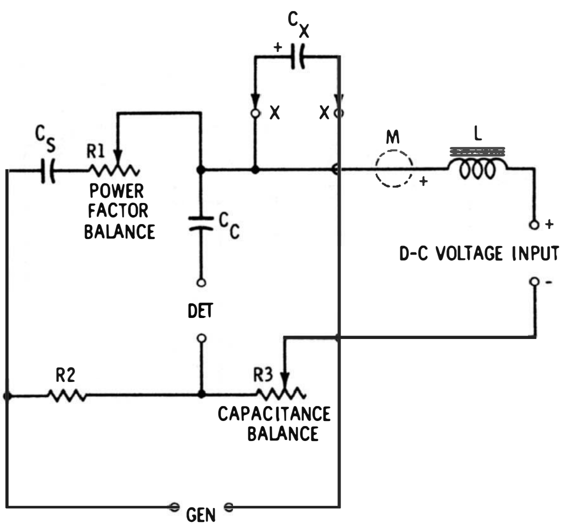
switch again at A, R1 and R2 are adjusted for deeper null; and then with S1 once more at B, R4 is adjusted for deeper null. Working back and forth in this way between the R1-R2 and R4 adjustments produces a final balance which cannot be improved by further adjustments and which frees the bridge from the effects of the stray capacitances.

3.8 BRIDGE OPERATION WITH D-C POLARIZING VOLTAGE

The capacitance of an electrolytic capacitor is not the same with and without a d-c polarizing voltage. The direct application of an a-c



(A) Internal d-c supply.



(B) External d-c supply.

Fig. 3-10. Capacitance bridge with d-c polarizing voltage.

signal to an unpolarized electrolytic capacitor, besides giving false readings, is also damaging because of polarity reversal. Moreover, the capacitance varies with the magnitude of the d-c voltage. The same is true of a capacitor having a nonlinear ceramic dielectric.

Such capacitors must be measured with their normal d-c working voltage applied simultaneously with the a-c bridge signal voltage. Fig. 3-10 shows how the a-c and d-c voltages may be applied to a typical capacitance bridge. In Fig. 3-10A, the d-c voltage is in series with the generator a-c voltage. Capacitor C_f offers a low-impedance path to the a-c signal.

Note that a capacitor (C_x, C_s) sets up a d-c block in each branch of the bridge circuit, preventing a short circuit of the d-c supply. However, the d-c voltage is impressed upon standard capacitor C_s, as well as upon unknown capacitor C_x, and the standard capacitors must be rated to withstand safely the maximum d-c voltage which may be applied to the bridge.

In Fig. 3-10B, the d-c voltage is applied across unknown capacitor C_x. This arrangement is often used with bridges having no provision for a polarizing voltage, since an external d-c supply may easily be connected to the X-X terminals. Choke coil L (30 henrys or higher) prevents the d-c supply from short-circuiting the a-c bridge signal. A d-c milliammeter, M, may be inserted to indicate leakage current of the unknown capacitor. Capacitor C_c provides d-c blocking to protect the null detector. In this arrangement, the output capacitance of the d-c power supply must be negligible with respect to the capacitance being measured; otherwise, it must be subtracted from each measured C_x value. The d-c voltage must be free from ripple, or the ripple frequency will interfere with the a-c signal and obscure the bridge balance.

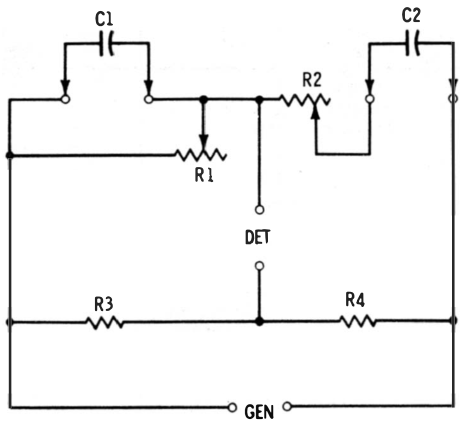


Fig. 3-11. Typical Wien bridge containing capacitance and resistance components in parallel.

3.9 WIEN BRIDGE

In the Wien bridge (Fig. 3-11), one of the two impedance arms (C1R1) contains capacitance and resistance in parallel, and the other one (C2R2) contains capacitance and resistance in series. The branch containing these arms is balanced against the branch containing resistance ratio arms R3 and R4, which form a voltage divider. Resistances R3 and R4 may either be equal or in any convenient ratio.

At one frequency, null is obtained by adjusting R1 and R2. At null, the values of C1 and C2 may be determined with the aid of the following simplified Wien-bridge equations:

$$C1 = \sqrt{\frac{0.0253(R1R4 - R2R3)}{f^2R1^2R2R3}} \qquad 3\text{-}9$$

where,

C1 is in farads,
f is in Hz,
R is in ohms.

$$C_2 = \sqrt{\frac{0.0253\ R3}{f^2(R1R4 - R2R3)R2}} \qquad 3\text{-}10$$

Thus, the two capacitance values are found in terms of frequency and resistance. This is important whenever frequency and resistance values can be known more accurately than standard-capacitor values. If one is interested only in determining the value of C1, the capacitance of C2 need not even be known. Also, C1 and C2 both may be unknown and their values determined with equations 3-9 and 3-10 after a single bridge balance. While the capacitance calculation is somewhat involved, equations 3-9 and 3-10 may be simplified to some extent when the frequency is fixed and R3 and R4 provide a fixed ratio.

Because the Wien bridge is frequency sensitive, the generator frequency must always be stable. As long as this stability requirement is met, various frequencies may be employed, but different settings of controls R1 and R2 are required when the generator frequency is changed.

3.10 CAPACITANCE BRIDGE WITH TRANSFORMER RATIO ARMS

If the secondary winding of the bridge-generator transformer is accurately center tapped, its two halves can supply the ratio arms of the bridge, and the other two arms will consist of the unknown capacitance and standard capacitance. This arrangement is shown in Fig. 3-12.

Fig. 3-12A illustrates the basic principle. The secondary of transformer T has an equal number of turns (N_x, N_s) on each side of the center tap; therefore, equal voltages (E_x, E_s) are developed across the two halves of this winding. However, these voltages are of opposite phase, and because of this no current will flow through the detector if C_x and C_s are equal—that is, if the bridge is balanced. If C_x differs from C_s, the bridge may be balanced by varying C_s. It may be balanced also by leaving C_s fixed and varying the number of turns (N_s) in the lower half of the secondary, as shown in Fig. 3-12B. This latter adjustment may be accomplished with a series of taps. For example, with a 100-turn lower secondary section, each 10th turn may be tapped up to the 90th turn and these taps

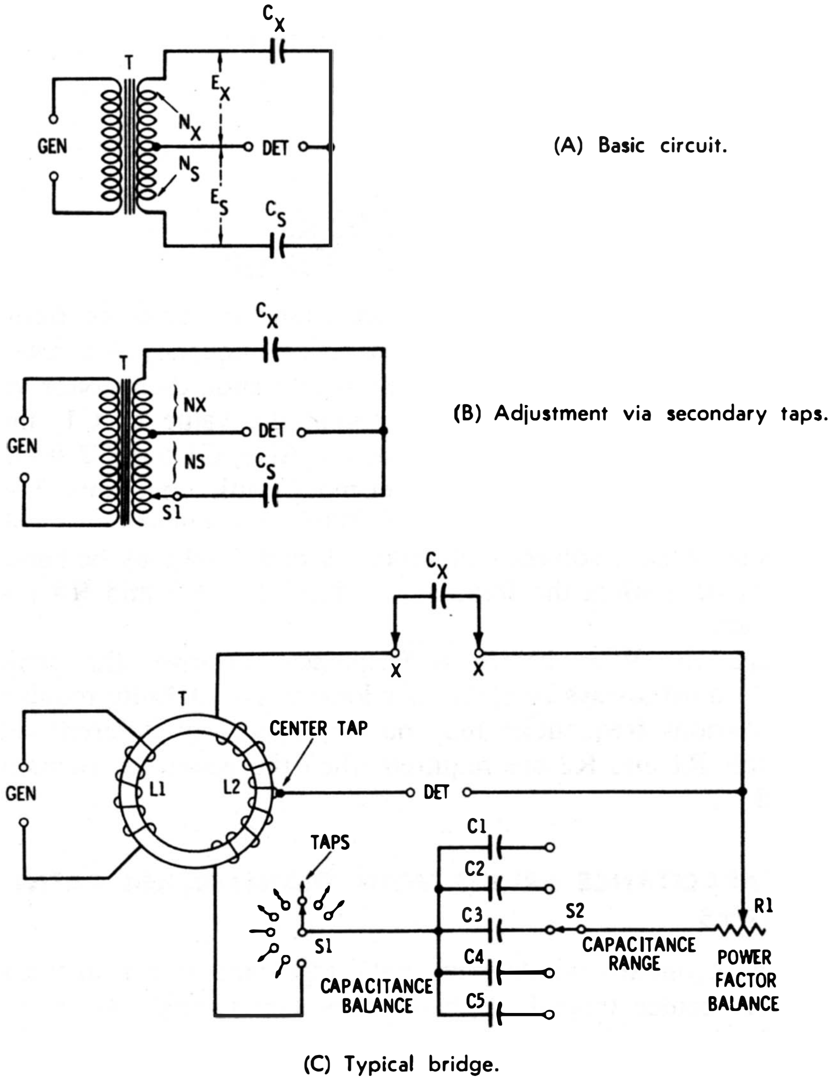


Fig. 3-12. Capacitance bridge with transformer ratio arms.

connected to one selector switch, and each of the remaining 10 turns may be tapped and connected to a second selector switch. The settings of the two switches then will allow N_s to be set to any value from 1 to 100 in one-turn steps, while N_x remains constant at 100. (Some commercial bridges of this type use 1000-turn secondary halves.) The value of standard capacitor C_s then may be changed to change capacitance ranges. Whichever balance method is employed, $C_x/C_s = N_s/N_x$, and from this relationship:

$$C_x = C_s(N_s/N_x) \qquad 3\text{-}11$$

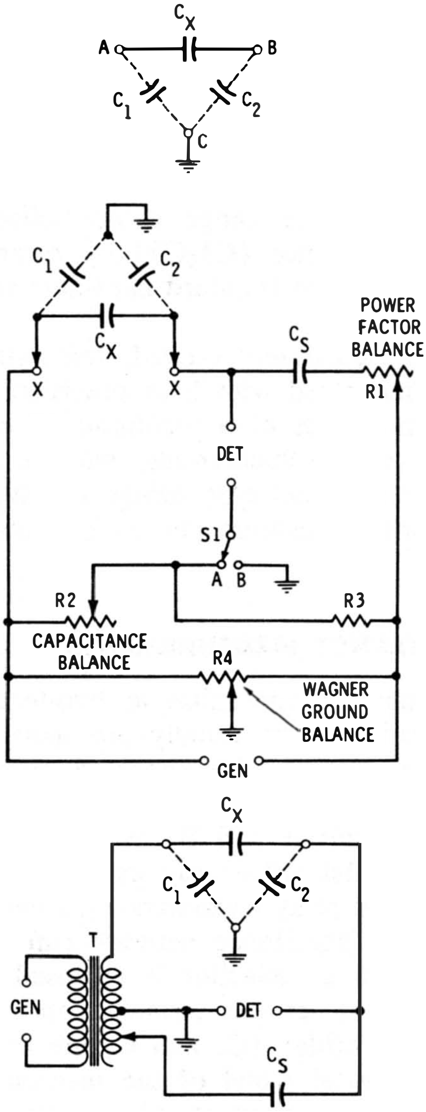
Fig. 3-12C shows the configuration of a typical bridge employing transformer ratio arms. In this arrangement, high-efficiency toroidal core construction is employed in transformer T. The number (N_s) of lower-half secondary turns is varied by means of taps (a single switch, S1, is shown, but two or more decade switches might be included for small-step adjustment), and this constitutes the capacitance balance. The basic capacitance range is multiplied by suitably changing the standard capacitance (C1-C5) by means of switch S2. Rheostat R1 in series with the standard capacitor serves as the power-factor balance.

Because a transformer may be accurately tapped, this type of bridge allows capacitance to be measured with high precision over a wide range—down to a few millionths of a picofarad. The circuit is insensitive to generator voltage fluctuations, and its ratio arms are low impedances. The transformer-type bridge is advantageous for three-terminal capacitance measurements, as is explained in Section 3.11.

3.11 THREE-TERMINAL CAPACITANCE MEASUREMENTS

In practical situations, capacitance never exists in two-terminal simplicity; two or more stray capacitances usually are associated with any capacitance of interest. Fig. 3-13A shows a common example of this situation. Here, capacitance C_x is not the only capacitance between capacitor terminals A and B (as might incorrectly be assumed), but is in parallel with stray capacitance C_1 between terminal A and ground, and stray capacitance C_2 between terminal B and ground. This three-capacitance network can result from a number of conditions: when a capacitor is enclosed in a metal can (C_1 and C_2 are capacitances between terminals and can); when a capacitor is connected to a bridge (C_1 and C_2 are capacitances between terminals and the metal panel of the instrument); when an interelectrode capacitance of a vacuum tube is measured (C_x may be the grid-plate capacitance, C_1 the grid-cathode capacitance, and C_2 the plate-cathode capacitance). In every instance, the main capacitance (C_x) is termed the *direct capacitance*.

Several techniques are available for measuring direct capacitances in the presence of stray capacitances. Fig. 3-13B shows a common one. Here, the two "stray" capacitances are returned to the ground point of the Wagner ground in the conventional capacitance bridge. (See Section 3.7 for a description of the Wagner ground.) The Wagner-ground balance then nullifies the effect of C_1 and C_2, and the bridge measures the direct capacitance, C_x, alone. In this way, the error which would be caused by C_1 and C_2 in series across C_x is avoided.



(A) Three-terminal capacitance.

(B) Conventional bridge.

(C) Transformer bridge.

Fig. 3-13. Three-terminal capacitance measurement.

In the instance of a shielded capacitor with the shield not connected to either terminal A or B, the junction point of C_1 and C_2 is the shield. In a triode tube, C_1 and C_2 represent the interelectrode capacitances not under measurement, and the junction point of C_1 and C_2 is the electrode (e.g., cathode) which is common to these two capacitances.

Fig. 3-13C shows the setup with a transformer-type capacitance bridge. Here, the C_1-C_2 junction is returned to the transformer center tap, which may be grounded. This places C_1 across the upper-transformer ratio arm, where its impedance is so high with respect to the arm that it causes no change in the secondary voltage, and places C_2 across the detector, where it can have no effect on the bridge balance.

3.12 SUBSTITUTION METHOD

The difficulty of accurately measuring capacitances lower than 100 pf with some bridges, because of stray capacitances, has been mentioned in previous sections. Low values may be measured satisfactorily, however, with any bridge by using the *substitution method,* a process which automatically compensates for strays.

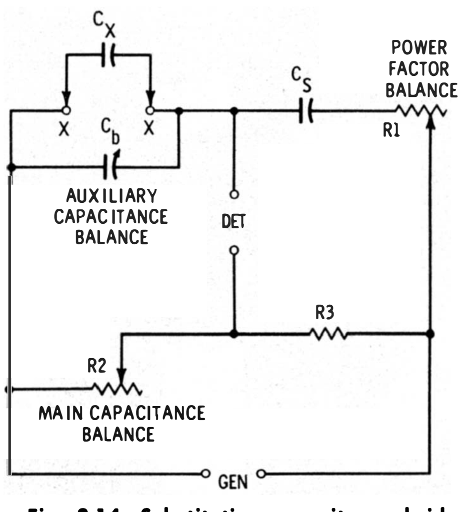


Fig. 3-14. Substitution capacitance bridge.

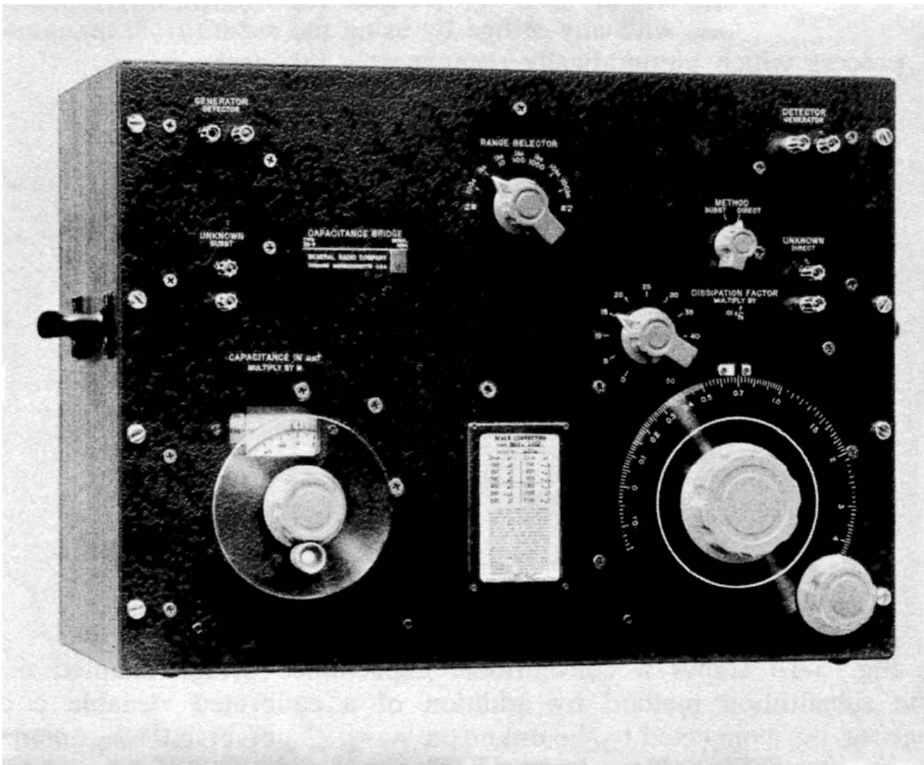
Fig. 3-14 shows a conventional capacitance bridge adapted to the substitution method by addition of a calibrated variable capacitor, C_b, connected to the unknown terminals, X-X. This capacitor is usually a 100- or 1100-pf air type having a dial reading directly in picofarads. In all other respects, this bridge is identical to the one shown in Fig. 3-6.

A four-step test procedure is employed:

1. With no unknown capacitor (C_x) connected to terminals X-X, variable capacitor C_b is set to its maximum capacitance (C_1), and the bridge then balanced in the regular manner by adjustment of R2 and R1. Hence, R2 is not disturbed.
2. The unknown capacitor next is connected to terminals X-X by the shortest possible leads. This adds capacitance and accordingly unbalances the bridge.
3. The bridge is rebalanced by setting variable capacitor C_b to a lower capacitance (C_2) and readjusting power-factor balance R_1.
4. Finally, the unknown capacitance is calculated from the two settings of the variable capacitor:

$$C_x = C_1 - C_2 \tag{3-12}$$

In this way, capacitances of less than 0.1 pf are measured with good accuracy. If long connecting leads are unavoidable, the bridge may be initially balanced in Step 1 with the leads connected to



Courtesy General Radio Co.

Fig. 3-15. Laboratory-type capacitance bridge with provision for making substitution measurements.

X-X and in the same position they will have when connected to the capacitor, C_x. Then the capacitor is connected for the remaining steps.

A practical advantage of the substitution method is the ease with which it may be used with any capacitance bridge. Thus, a dial-calibrated variable air capacitor may be connected externally to the unknown terminals of any bridge, provided that short rigid leads are used and that the four-step procedure described above is employed.

In commercial substitution bridges, the variable capacitor dial sometimes is graduated with zero at the maximum-capacitance setting of this capacitor, and with maximum capacitance at the zero-capacitance setting. The initial balance with R_2 (see Step 1 above) then becomes a zero adjustment, and after rebalance (Step 3) the dial reads C_x directly in pf, no calculation being required.

Fig. 3-15 shows a laboratory-type capacitance bridge employing the substitution method in addition to the direct method. The dial in the lower left corner of the panel operates the variable capacitor and permits the determination of C_1-C_2 (equation 3-12) from 0.1 to 1050 pf. An external generator and detector are required.

CHAPTER 4

Inductance Bridges

Inductance bridges may be regarded as forming the second large class of a-c bridges. In general configuration, the inductance bridge resembles the capacitance bridge and the resistance bridge. It differs from the capacitance bridge, however, in the presence of inductors in one or more of its arms and also in the respect that it can, in some versions, compare unlike impedances, i.e., inductance with capacitance.

As a class, inductance bridges cover the range from 0.1 nanohenry to 10,000 henrys with accuracies reaching $\pm 0.1\%$ or better. Like the resistance bridge and capacitance bridge, the inductance bridge may be either rudimentary or complex to suit individual demands.

This chapter describes bridges for the measurement of self-inductance and mutual inductance. Representative types in each category are shown.

4.1 BASIC SLIDE-WIRE INDUCTANCE BRIDGE

Fig. 4-1 shows the most rudimentary inductance bridge. In this arrangement, the variable balancing resistor is a single strand of resistance wire (the *slide wire*) tautly stretched between points A and B (or wound around a form having a circular cross section) and provided with a sliding contact (the *slider*). The wire has a uniform cross section and is of constant composition, so its resistance is directly proportional to its length.

As the slider is moved along the wire, it divides the latter into two parts: D_1, the length from point A to the slider, which has a resistance of R_1; and D_2, the length from point B to the slider, which has a resistance of R_2. Thus, the resistance increases on one side of the moving slider and decreases on the other side.

The bridge is composed of unknown inductor L_x (connected to terminals X-X), standard inductor L_s, and the two sections (R_1 and R_2) of the slide-wire variable resistor, which provide the ratio arms.

The bridge is balanced by moving the slider along the wire until the detector response is minimum. At this null point, $L_x/L_s = R_1/R_2$, and from this relationship the unknown inductance may be determined in terms of the standard:

$$L_x = L_s(R_1/R_2) \qquad 4\text{-}1$$

The total resistance of the slide wire is unimportant to the calculation. So also are the actual resistance values on each side of the slider. In fact, distances D_1 and D_2 may be measured in inches, centimeters, or arbitrary linear units and used in the calculation in place of resistances R_1 and R_2:

$$L_x = L_s(D_1/D_2) \qquad 4\text{-}2$$

For this reason, the basic slide-wire bridge is convenient for emergency measurements of inductance, since it requires only a standard inductor and a length of bare resistance wire stretched along a meter stick or yardstick, in addition to a generator (e.g., an audio oscillator) and a detector (e.g., a-c vtvm, oscilloscope, or high-resistance headphones).

Sometimes, it is more convenient to read the position of the slider with respect to the total length (D_3) of the wire than to measure D_1 and D_2 separately. In such an instance:

$$L_x = L_s[D_1/(D_3 - D_1)] \qquad 4\text{-}3$$

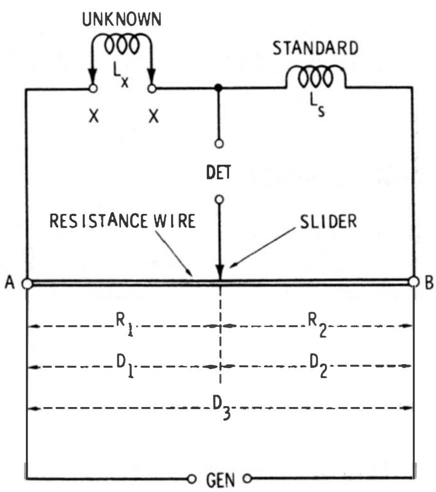


Fig. 4-1. Diagram of a basic slide-wire inductance bridge.

It is apparent from either equation 4-2 or 4-3 that the unknown (L_x) is equal to the standard (L_s) when null occurs with the slider halfway between A and B (i.e, when $D_1 = D_2$, making $D_1/D_2 = 1$). Also, the slider must move to the right of center when $L_x > L_s$, and to the left of center when $L_x < L_s$.

If the wire is long (say, 1 meter or more), little difficulty is experienced in measuring inductances over the range $0.01L_s$ to $100L_s$, provided a sensitive detector is used. Unless the wire has reasonably high resistance (large ratio of length to thickness), however, the current may heat it enough to impair the accuracy of measurement.

4.2 POTENTIOMETER-TYPE SLIDE-WIRE INDUCTANCE BRIDGE

A more compact version of the slide-wire inductance bridge, like the similar version of the resistance and capacitance bridges, substitutes a wirewound potentiometer for the slide wire. Fig. 4-2 shows this arrangement.

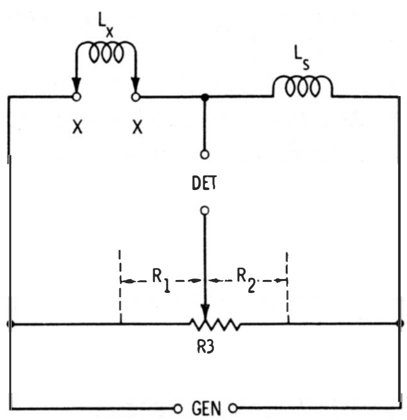


Fig. 4-2. Potentiometer-type slide-wire inductance bridge.

In this circuit, the balancing potentiometer has a total resistance R_3. At any setting, $R3 = R_1 + R_2$, the sum of the resistances between the contact blade and the low end, and the contact blade and the high end, respectively, of the resistance element. Resistances R_1 and R_2 determine the bridge ratio.

At null, the unknown inductance (L_x) is determined from the standard inductance (L_s) and the resistance ratio:

$$L_x = L_s(R_1/R_2) \qquad 4\text{-}4$$

The potentiometer dial may be calibrated to read directly in microhenrys, millihenrys, or henrys by locating the null points for a number of known inductances connected successivly at L_x. It may also be calibrated with the aid of equation 4-4, in terms of measured R_1 and R_2 values. If the potentiometer dial already reads directly in resistance setting (as the dial of a multiturn potentiometer often does), the unknown inductance may be calculated with reference to that setting:

$$L_x = L_s[R1/(R3 - R1)] \qquad 4\text{-}5$$

where

R_1 is the resistance setting indicated by the dial,

R3 is the total resistance of the potentiometer.

Like the basic slide-wire inductance, the potentiometer-type circuit is simple and economical. However, it has three disadvantages: (1) because of stray reactances in the circuit, L_x values lower than 100 μh often cannot be measured accurately; (2) unless the potentiometer has a special taper, such as logarithmic, the graduations will crowd at one end of the dial, seriously impairing readability and accuracy—and for that reason, the inductance range with any standard inductor L_s should not exceed 0.1 to 10 times L_s, even though the potentiometer affords a wider range; and (3) a complete null occurs only if losses in the unknown inductor equal those in the standard inductor.

The degrading effect of stray reactances in the useful ranges of the bridge may be reduced somewhat by keeping the potentiometer resistance, R_3, reasonably low—say, 5000 ohms. The wasted ends of the potentiometer in the simple circuit may be saved by means of extension arms (such as R1 and R3 in Fig. 4-3) which allow the useful 100:1 inductance range to be spread over the entire resistance range of the potentiometer.

4.3 PRACTICAL POTENTIOMETER-TYPE SLIDE-WIRE INDUCTANCE BRIDGE

Fig. 4-3 shows the circuit of a practical bridge using a 5000-ohm wirewound potentiometer (R2) as the balance control. This bridge covers the range from 100 μh to 100 hy in three steps: 100 μh–10 mh, 10 mh–1 hy, and 1–100 hy. The inductance ranges are selected by switching standard inductors (L1, L2, L3). The 100-ohm non-inductive resistors, R1 and R3, provide the extension arms mentioned in Section 4.2.

Any convenient test frequency may be employed, up to about 5 kHz. Best stability is obtained, however, when the generator is coupled to the bridge through a shielded transformer. It is advisable also to insert a transformer between the bridge and detector, but this second transformer may be omitted if the first one is used, especially if the lower DET terminal is grounded.

This bridge may be calibrated by balancing it with a number of accurately known inductors connected successively to terminals

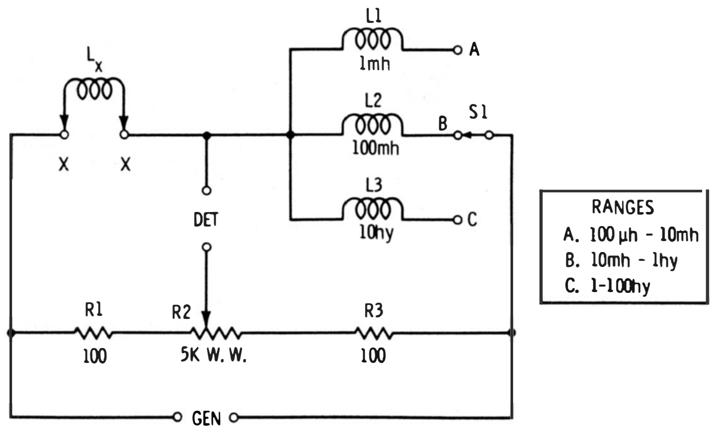


Fig. 4-3. Practical potentiometer-type slide-wire inductance bridge.

X-X, and inscribing the potentiometer dial accordingly. Only one range need be calibrated in this manner; the others will track if the values of L1, L2, and L3 are all accurate. Thus, the lowest range (0.0001–0.01 hy) may be calibrated and the A, B, and C settings of switch S1 then used to multiply this basic range by 1, 100, and 10,000, respectively.

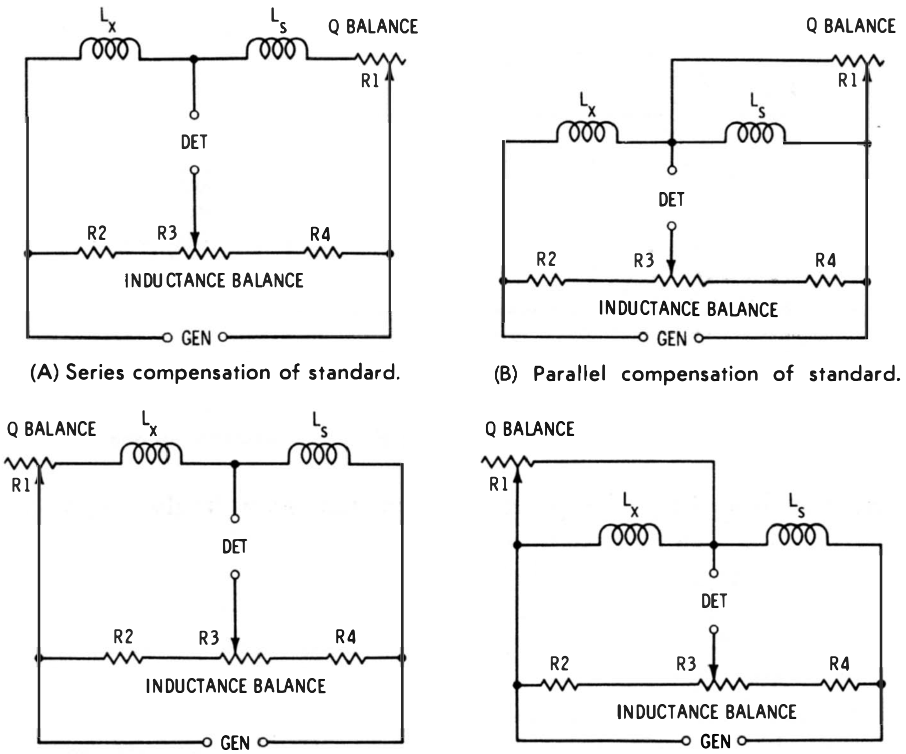
The null balance is sharp if losses in the inductor under test (L_x) equal those in the standard inductor (L_s). Since low-loss inductors customarily are used as the standards, this usually—but not always—means that a sharp null indicates low loss in the inductor under test. However, a broad null can also reveal that the standard is inferior to the test inductor.

4.4 THE NEED FOR Q BALANCE

The preceding sections explain that a complete null is impossible with a simple slide-wire bridge unless the losses in the unknown and standard inductors match. The reason for this is the complex nature of the practical inductor as an impedance network (inductance and resistance in combination, the resistive component representing the losses). In order to obtain a complete null in most practical situations, separate balances are required for the resistive and reactive components, and this makes the bridge somewhat more complicated.

It is convenient to express the relationship between losses (resistive component) and inductance (reactive component) as the *figure of merit* or *Q* of the inductor. Numerically, Q = XL/R = tangent of the phase angle of the equivalent LR network. A low-loss inductor has a high Q.

In somewhat more specific terms, the original statement may be rephased thus: *A complete null is obtained only when the Q of*



(A) Series compensation of standard.

(B) Parallel compensation of standard.

(C) Series Compensation of unknown.

(D) Parallel compensation of unknown.

Fig. 4-4. Slide-wire inductance bridge with Q balance.

the unknown inductor equals that of the standard inductor. When the Q's are unequal, as they almost always are in practice, enough resistance must be added in series (sometimes in parallel) with the lower-loss inductor to reduce its Q to that of the higher-loss inductor. Addition of resistance to one or the other of the inductors will complete the balance of the bridge.

Fig. 4-4 shows how a Q-balance rheostat may be added to the slide-wire inductance bridge. In Figs. 4-4A and 4-4B, the standard inductor (L_s) is the higher-Q component, and the resistance-balance rheostat (R1) is added in series (Fig. 4-4A) or parallel (Fig. 4-4B) with L_s. In Figs. 4-4C and 4-4D, the unknown inductor is the higher-Q component, and the resistance-balance rheostat (R1) is added in series (Fig. 4-4C) or parallel (Fig. 4-4D) with L_x. In use, the bridge is balanced first for inductance by adjusting potentiometer R3 for minimum response of the detector. Then, rheostat R1 is adjusted to improve the null. Additional adjustments of R1 and R3, alternately, will give complete null (zero detector output). The resistance of R1 does not enter into the calculation of unknown inductance. Nor does the resistance of R3 enter into the calculation of Q (see the next section).

For complete null, the resistance of R1 must be adjustable to the equivalent resistance of the opposite inductor. A dial attached to this rheostat may be calibrated to read directly in Q; however, this dial will read correctly only at the calibration frequency. All complete inductance bridges are equipped with such a Q balance.

Fig. 4-5 shows how a Q-balance rheostat is included in a *universal bridge* so that it may be switched in series with either the unknown or standard inductor, whichever has the higher Q (See Section 3.4 and Fig. 3-5 for an explanation of the universal bridge.) This bridge is convenient for checking an unknown inductor (L_X) against any available inductor (L_S) whose inductance, but not necessarily its Q, is known. When switch S1 is in position A, the Q-balance rheostat (R3) is in series with standard inductor L_S; when S1 is in position B, rheostat R3 is in series with unknown inductor L_X.

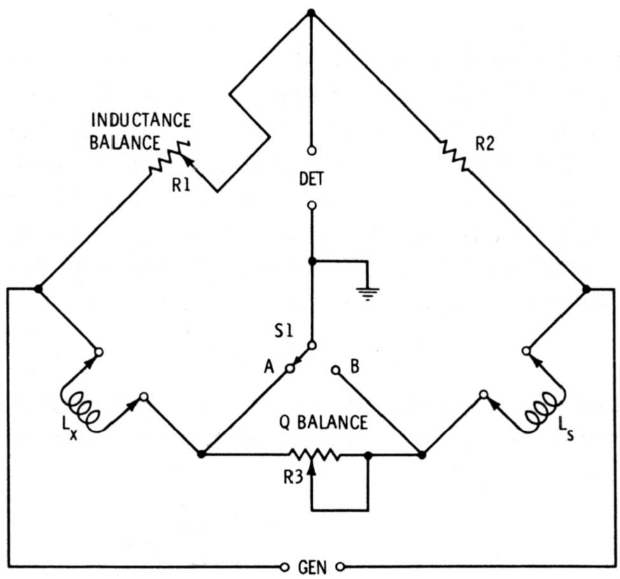


Fig. 4-5. Universal bridge with Q balance.

4.5 INDUCTANCE COMPARISON BRIDGE

A common inductance bridge merely substitutes inductance arms for two of the resistance arms in the Wheatstone bridge, retains the other two resistance arms—the *ratio arms*—to complete the basic circuit, and adds a Q balance. Because of the similarity of the two bridges, the inductance type is sometimes familiarly called a "Wheatstone inductance bridge," but the term *inductance comparison bridge* appears most often in technical literature.

The inductance comparison bridge, like the capacitance comparison bridge (Section 3.5), overcomes many of the shortcomings

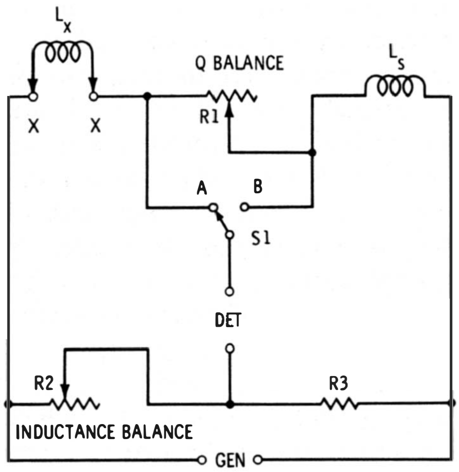


Fig. 4-6. Inductance comparison bridge.

of the slide-wire bridge, principally the dial crowding and lower-inductance limit of the latter instrument. Fig. 4-6 shows the basic circuit of the comparison bridge. Here, the bridge ratio (R2/R3) is established by fixed resistor R3 and the inductance-balance rheostat, R2. A second rheostat R1 provides the Q balance. With switch S1 at position A, rheostat R1 is in series with standard inductor L_s; with switch S1 at position B, rheostat R1 is in series with unknown inductor L_x. At null:

$$L_x = L_s(R2/R4) \qquad 4\text{-}6$$

The circuit of a practical multirange inductance comparison bridge is shown in Fig. 4-7. This bridge covers the range from 10

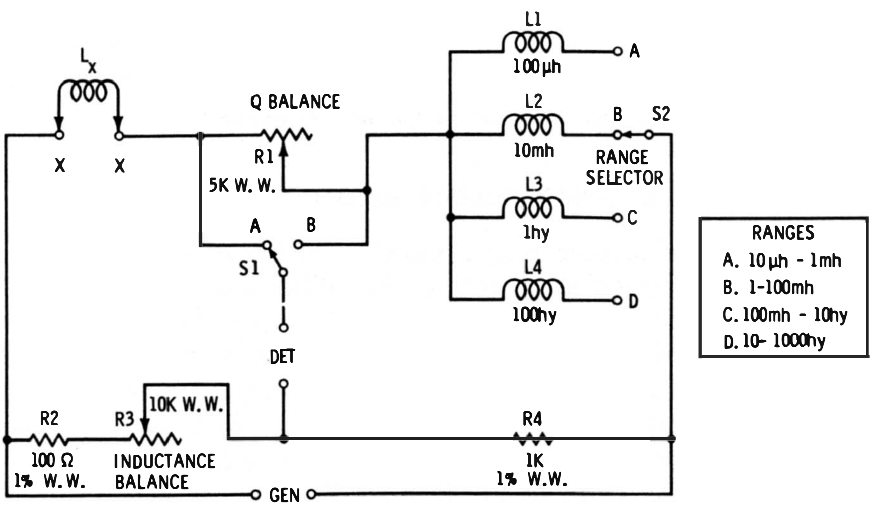


Fig. 4-7. Practical multirange inductance comparison bridge.

μh to 1000 hy in four steps: 10 μh–1 mh, 1–100 mh, 100 mh–10 hy, and 10–1000 hy. The similarity of this circuit to the basic circuit (Fig. 4-6) is easily seen, the chief difference being the switched values of standard inductance (L1 to L4 in Fig. 4-7).

The R3 dial may be calibrated to read directly in microhenrys, millihenrys, or henrys—either by means of the relationship $L_x = L_s(R3/R4)$, where L_s is either L1, L2, L3, or L4 depending on the setting of range switch S2, or by balancing the bridge for a number of accurately known inductors connected successively to terminals X-X. Only one range need thus be calibrated. If L1, L2, L3, and L4 are accurate, the other ranges will track automatically. If the R3 dial is graduated according to the lowest (A) range, the settings of switch S2 then may be used to multiply the dial settings by 1, 100, 10K, and 1M when S2 is set to A, B, C, and D, respectively.

Rheostat R1 provides the resistive (Q-balance) adjustment. When switch S1 is thrown to position A, this rheostat is connected in series with the standard inductor, whereas when switch S1 is at position B, R1 is in series with the unknown inductor. The R1 dial may be graduated to read directly in equivalent series resistance R_x (where $R_x = R$, the resistance setting of the Q-balance rheostat at null). However, it must be remembered that the dial setting shows the resistance of either the unknown or standard, depending on the position of switch S1. While the 5000-ohm total resistance shown for rheostat R1 will suffice for a great many test inductors, a higher resistance (for very-high-loss inductors) or a lower resistance (for closer reading of low values) may be needed in some instances and be obtained with an external rheostat.

4.6 WAGNER GROUND

When the bridge arms contain high impedances (high inductances and resistances), stray capacitances can impair the accuracy of measurement. Such capacitances from bridge arms to ground are represented by C_1 and C_2 in Fig. 4-8.

As in the capacitance bridge, a *Wagner ground* permits these objectionable capacitances to be balanced out. This arrangement consists simply of the auxiliary potentiometer R5 and spdt switch S2. A Wagner ground may be added to a bridge which does not already come equipped with one.

In operation, the bridge is balanced first in the normal manner; i.e., with switch S2 set to position A, rheostats R1 and R3 and switch S1 are adjusted for the best null that they will afford. Switch S2 then is thrown to position B, and potentiometer R5 adjusted for null. Next, with switch S2 again at A, R1 and R3 are adjusted

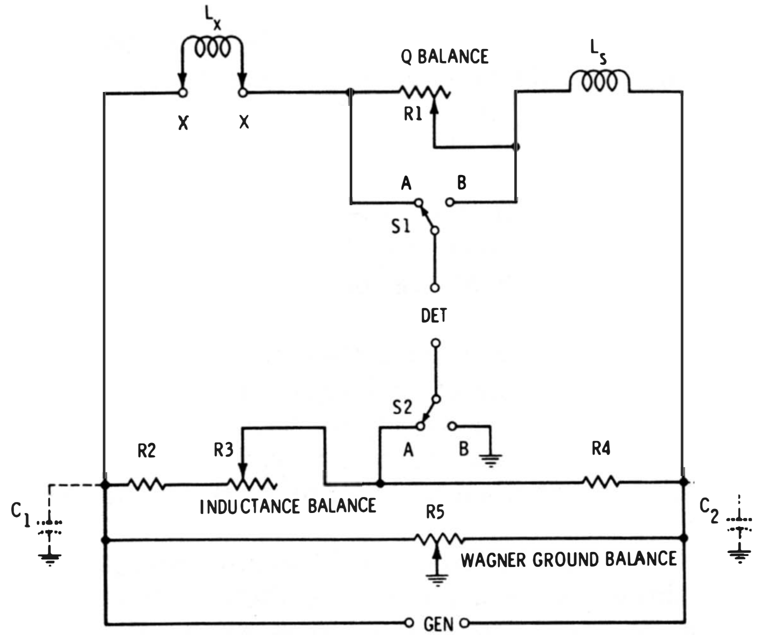


Fig. 4-8. Inductance bridge with Wagner ground.

for deeper null; and then with S2 once more at B, R5 is adjusted for still deeper null. Working back and forth in this way between the R1-R3 and R5 adjustments produces a final balance which is independent of the error-producing effects of the stray capacitances.

4.7 INDUCTANCE BRIDGES WITH CAPACITANCE STANDARDS

The bridge circuits described in the foregoing sections compare an unknown inductance (L_x) with a standard inductance (L_s). This method is readily apparent, for it involves the basic a-c bridge comparison of two similar impedances. Bridges of another group, however, compare an unknown inductance with a standard capacitance (C_s). This latter procedure is advantageous, since highly accurate standard capacitors are often more readily obtainable than are standard inductors; and generally the capacitors do not have the bulk of inductors, especially at the higher values; nor are they ordinarily susceptible to interfering hum fields.

The Hay, Maxwell, Anderson, and Owen bridges use capacitance standards. These bridges are separately described in the next four sections.

4.8 HAY BRIDGE

Fig. 4-9 shows the basic circuit of the Hay bridge. Here, unknown inductance L_x is compared with standard capacitance C_s. The bridge is balanced in the conventional manner by adjusting inductance-balance rheostat R2 and Q-balance rheostat R3 for null. At null:

$$L_x = \frac{C_s R1R2}{1 + (R3^2 \omega^2 C_s^2)} \qquad 4\text{-}7$$

where,

C_s is in farads,
L is in henrys,
R is in ohms,
ω equals $2\pi f$ (f is in Hz)

Since R3 appears in the denominator of the fraction, the inductance balance of this bridge is not independent of the Q balance. Also, since ω also appears in the equation, the balance is frequency

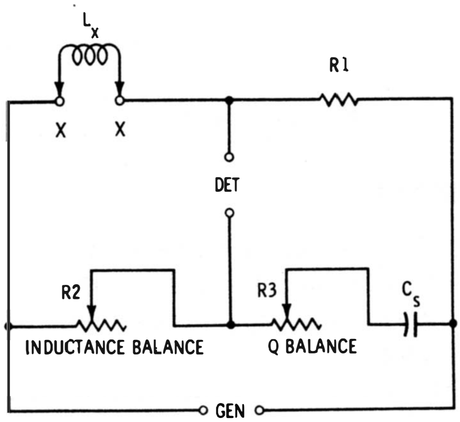


Fig. 4-9. Hay bridge.

dependent. However, if the Q of the test inductor, L_x, is higher than 10, the frequency may be ignored, and equation 4-7 may be simplified, with an error of less than 1 percent:

$$L_x = C_s(R1R2) \qquad 4\text{-}8$$

At null, the equivalent resistance of the unknown inductor also may be determined:

$$R_x = \frac{\omega^2 C_b^2 R1R2R3}{1 + (\omega^2 C_b^2 R3^2} \qquad 4\text{-}9$$

From this, the Q of the unknown inductor may be determined:

$$Q_x = (\omega L_x)/R_x \qquad 4\text{-}10$$

Again, note that the settings of both the inductance-balance and Q-balance rheostat enter into the calculation, and that the R_x de-

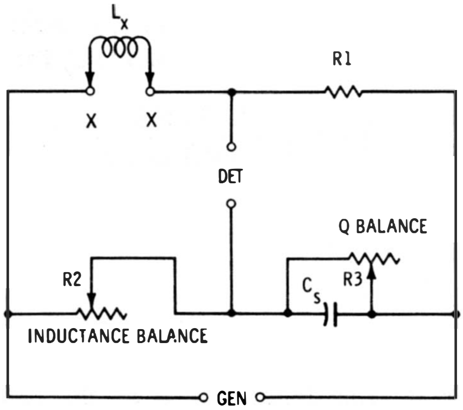


Fig. 4-10. Maxwell bridge.

termination, like that of L_x, is frequently dependent. However, if the Q of inductor L_x is higher than 10, equation 4-9 may be simplified:

$$R_x = (R_1R_2)/R_3 \qquad 4\text{-}11$$

4.9 MAXWELL BRIDGE

The basic circuit of the Maxwell bridge is shown in Fig. 4-10. This circuit compares unknown inductance L_x with standard capacitance C_s. It differs from the Hay bridge, just described, in the parallel connection of the Q-balance rheostat (R3) and standard capacitor (C_s).

In this circuit, the inductance balance and Q balance are independent of each other and each is independent of the frequency. At null:

$$L_x = C_s(R1R2) \qquad 4\text{-}12$$

where,

C_s is in farads,
L_x is in henrys.
R is in ohms.

The equivalent resistance (R_x) of the unknown inductor is calculated:

$$R_x = R1(R2/R3) \qquad 4\text{-}13$$

4.10 ANDERSON BRIDGE

Fig. 4-11 shows the basic circuit of the Anderson bridge. This is a six-impedance network (L_x, C_s, R1, R2, R3, R4) in which unknown inductance L_x is compared with standard capacitance C_s. This circuit is somewhat harder to adjust than is the usual four-impedance bridge, but it offers a wider range.

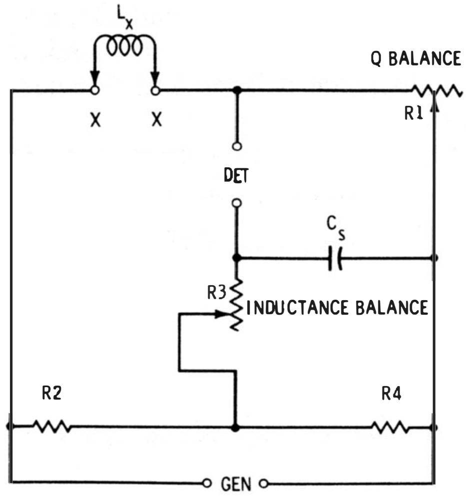


Fig. 4-11. Anderson bridge.

In the Anderson bridge, the inductance balance and Q balance are independent of each other and are independent of the frequency. At null:

$$L_x = C_s[R3\ (1 + R2/R4) + R2] \qquad 4\text{-}14$$

where,

C_s is in farads,
L_x is in henrys,
R is in ohms.

The equivalent resistance (R_x) of the unknown inductor is calculated:

$$R_x = (R1R2)/R4 \qquad 4\text{-}15$$

4.11 OWEN BRIDGE

Fig. 4-12 shows the basic circuit of the Owen bridge. In this arrangement, a variable capacitor (C_r) is employed for the Q balance.

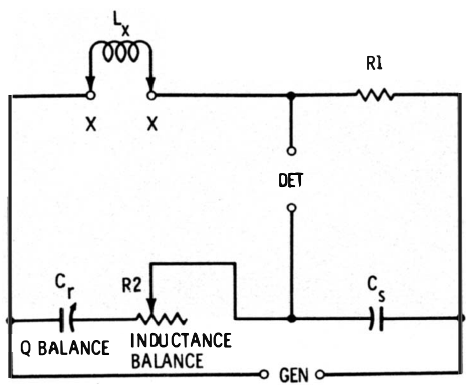


Fig. 4-12. Owen bridge.

(If a suitable variable capacitor is not available, a variable resistor may be connected in series with inductor L_x, and a fixed capacitance used at C_r.) The inductance balance and Q balance are independent of each other and of the frequency.

The Owen bridge provides an extremely wide inductance range for a narrow range of C_s and R values. At null:

$$L_x = C_s(R1R2) \qquad 4\text{-}16$$

where,

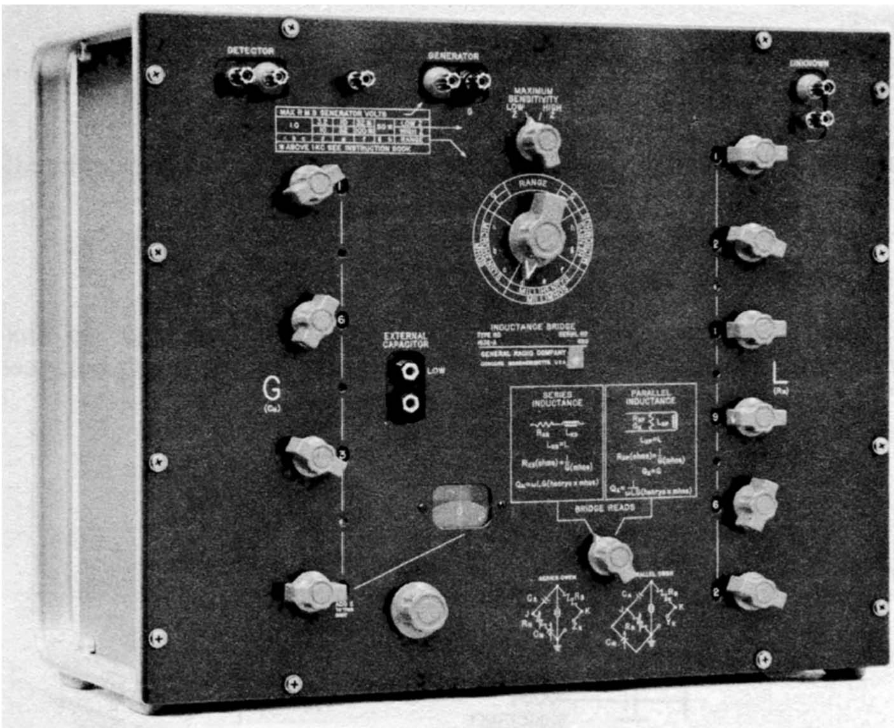
C_s is in farads,
L_x is in henrys,
R is in ohms.

The equivalent resistance (R_x) of the unknown inductor may be calculated:

$$R_x = R1(C_s/C_r) \qquad 4\text{-}17$$

where C_r and C_s are *both* in the same unit of capacitance.

Fig. 4-13 shows a laboratory-type inductance bridge (General Radio Type 1632-A) employing the Owen circuit. This instrument covers the range from 0.0001 μh to 1111 hy and provides for either



Courtesy General Radio Co.

Fig. 4-13. Laboratory-type inductance bridge employing Owen circuit.

series or parallel connection of the Q-balance capacitor to the inductance-balance rheostat. An external generator and detector are required.

4.12 INDUCTANCE BRIDGE WITH D-C POLARIZATION

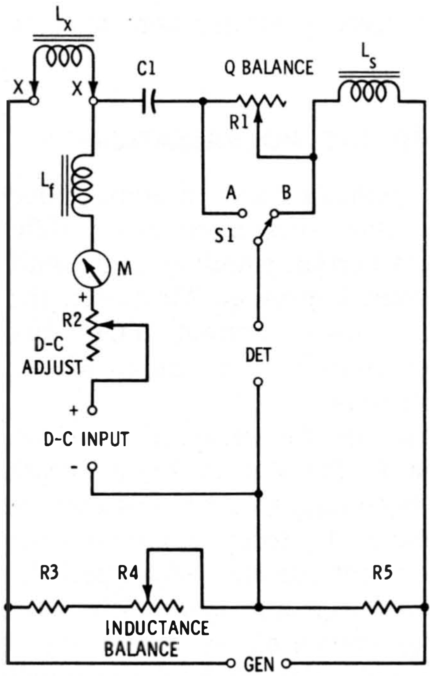
The inductance of an iron-core inductor (and of some inductors having cores of magnetic material other than iron) has a different value when the coil is carrying direct current simultaneously with the a-c signal than when no direct current is present. Moreover, the inductance varies with the amount of direct current. This situation necessitates that the inductance of such coils be measured while they are conducting rated d-c operating current.

Fig. 4-14 shows two bridge circuits for measuring inductance under conditions of d-c polarization. In Fig. 4-14A, the conventional inductance comparison bridge has been equipped for this application by applying direct current to inductor L_x through filter choke L_f, d-c ammeter or milliammeter M, current-adjusting rheostat R2, and extension-arm resistor R3. In all other respects, the circuit is identical to the inductance comparison bridge shown earlier in Fig. 4-6. The direct current is set to the required operating level by adjustment of rheostat R2, and is indicated by current meter M. Filter choke L_f, whose inductance must be many times that of unknown inductor L_x and which must be able to carry the maximum direct current safely and without saturating, prevents short circuit of the a-c bridge signal by the d-c supply. Similarly, blocking capacitor C1 prevents short circuit of the d-c supply by the bridge circuit. In this arrangement, the direct current passes through the inductance-balance rheostat (R4) and extension-arm resistor (R3).

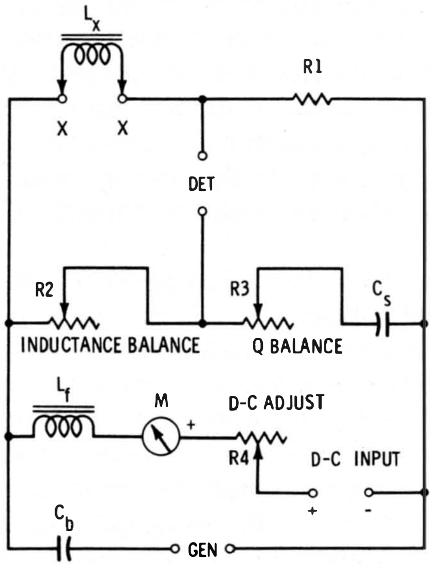
In Fig. 4-14B, the Hay bridge has been adapted by applying the d-c voltage in parallel with the a-c signal voltage. The direct current is set to the required operating level by means of rheostat R4 and direct-current meter M. Filter choke L_f, whose inductance must be much higher than that of unknown inductor L_x and which must be able to carry the maximum direct current without saturating, prevents short circuit of the a-c signal by the d-c supply. Similarly, capacitor C_b prevents short circuit of the d-c supply by the a-c genertaor. The direct current passes through bridge resistor R1, so this resistor must be rated to carry this current safely.

4.13 MUTUAL-INDUCTANCE MEASUREMENT

Mutual inductance, as well as self-inductance, may be measured by the bridge method. Fig. 4-15 shows two bridge circuits used for this purpose.



(A) Inductance comparison bridge.

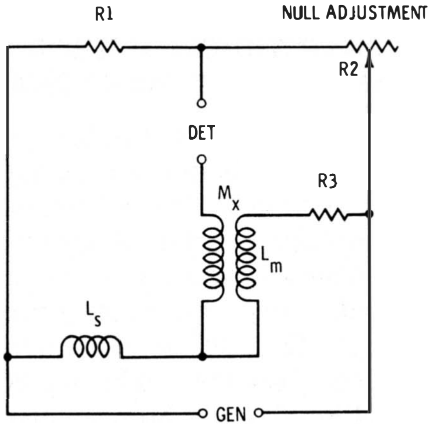


(B) Hay bridge.

Fig. 4-14. Inductance bridges with d-c polarization.

In the comparison bridge shown in Fig. 4-15A, unknown mutual inductance M_x is compared with standard self-inductance L_s. Here, coil L_m of the mutual inductor constitutes the lower right arm of the bridge. (Inductance L_m must be known.) The bridge is balanced by adjustment of rheostat R2. At null:

(A) L_x-M_x comparison bridge.



(B) Carey-Foster mutual-inductance bridge.

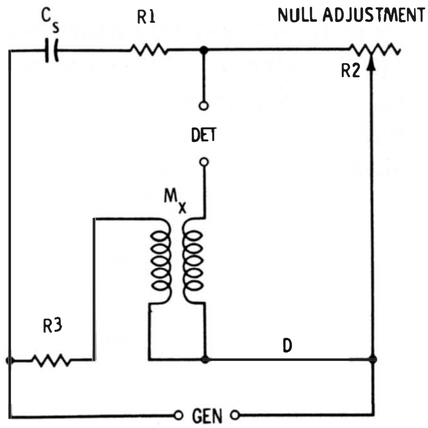


Fig. 4-15. Mutual-inductance bridges.

$$M_x = \frac{(L_m R1/R2) - L_s}{1 + (R1/R2)} \qquad 4\text{-}18$$

where,

L_m, L_s, and M_x are in henrys,
R1 and R2 are in ohms.

In the Carey-Foster mutual-inductance bridge in Fig. 4-15B, unknown mutual inductance M_x is compared with standard capacitor C_s. In this circuit, the lower right arm (D) of the bridge has zero resistance, which frees the null from frequency dependence. The bridge is balanced by adjustment of rheostat R2. At null:

$$M_x = C_s(R2R3) \qquad 4\text{-}19$$

where,

C_s is in farads
M_x is in henrys,
R2 and R3 are in ohms.

4.14 SUBSTITUTION METHOD OF INDUCTANCE MEASUREMENT

The same kind of trouble experienced in measuring small capacitances (see Section 3.12, Chapter 3) is encountered in measuring small inductances. That is, residual inductances in the bridge wiring and components, being of approximately the same magnitude as the small unknown inductance, restrict the smallest value which can be measured. They also make erroneous those values which can be measured. This difficulty is overcome by the *substitution method*, which resembles the similar method of capacitance measurement.

Fig. 4-16 shows a bridge employing the substitution method. This circuit employs a calibrated variable inductor (L_s), which may be several series-connected inductance decades or a continuously variable laboratory inductor (either one will be direct reading in henrys, millihenrys, or microhenrys). Unknown inductor L_x is connected to terminals X-X by the shortest possible leads. The bridge is operated as follows:

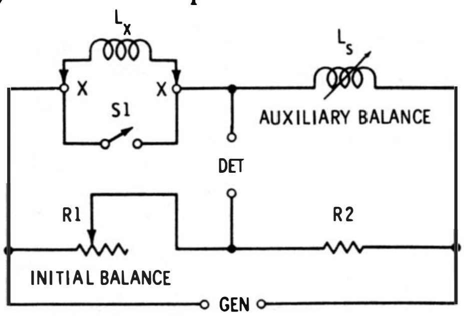


Fig. 4-16. Substitution inductance bridge.

1. Switch S1 first is closed to short-circuit unknown inductor L_x. The standard inductor, L_s, is set to its lowest value (recorded here as L_1), and the bridge is balanced by adjustment of rheostat R1. This rheostat then need not be disturbed.
2. Next, switch S1 is opened. This inserts L_x into the circuit and unbalances the bridge.
3. Variable inductor L_s then is adjusted to restore the bridge balance, and its new inductance setting recorded as L_2.
4. Finally, the unknown inductance is calculated from the two settings of the variable inductor:

$$L_x = L_2 - L_1 \qquad \text{4-20}$$

where the L's are in the same units: henrys, millihenrys, or microhenrys.

Inductance values of 0.1 μh or lower may be accurately measured by means of the substitution method. This method automatically takes into account the residual inductances of the circuit.

CHAPTER 5

Combination Bridges

For reasons of economy, convenience, and versatility, many users prefer bridges which measure more than one quantity. Because of this demand, a number of bridges are in current use which measure resistance and capacitance; or resistance, capacitance, and inductance.

This chapter describes representative *combination bridges* which perform these joint functions, and calls attention to the special features of some of these instruments.

5.1 UNIVERSAL BRIDGE

The *universal bridge* has been mentioned earlier (Sections 3.4 and 4.4). This instrument was the forerunner of all combination bridges, having appeared some time ago and offering measurements of resistance, capacitance, and inductance.

Fig. 5-1 shows the circuit of the universal bridge (also called *general-purpose bridge*). This arrangement is seen to be a skeleton circuit which supplies the ratio arms R1 and R_a, the latter collectively referring to resistors R2 and R6, balancing arm R7, unknown terminals X-X, standard terminals S-S, detector terminals DET, generator terminals GEN, and changeover switch S2. When generator, detector, unknown, and standard are properly connected to this skeleton, a bridge of the desired sort results. Typical uses are described below.

Resistance

1. Connect the unknown resistance (R_x) to terminals X-X.
2. Connect a jumper between terminals S-S.

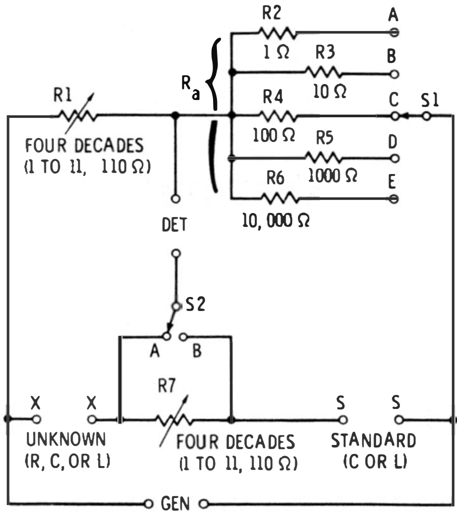


Fig. 5-1. Universal bridge.

3. Throw switch S2 to position A; this connects R7 as the lower right arm of the bridge, where it acts as the resistance standard.
4. For d-c resistance, connect a battery to the GEN terminals, and a center-zero d-c meter to the DET terminals.
5. For a-c resistance, connect an audio oscillator to the GEN terminals, and an oscilloscope, a-c vtvm, or high-resistance headphones to the DET terminals.
6. Balance the bridge by adjustment of R1, R7, and S1. At null:

$$R_x = R7(R1/R_a) \tag{5-1}$$

where,

R_a is the resistor (R2 to R6) selected by S1.

Capacitance

1. Connect the unknown capacitance (C_x) to terminals X-X.
2. Connect a standard capacitor (C_s) to terminals S-S.
3. Set switch S2 to position A. This places R7 in series with the standard capacitor, as the power-factor balance.
4. Connect an audio oscillator to the GEN terminals.
5. Connect an oscilloscope, a-c vtvm, or high-resistance headphones to the DET terminals.
6. Balance the bridge by means of R1 and S1.
7. Improve the null by adjusting R7, readjusting R1 if necessary. If no improvement results and null is imperfect, throw switch S2 to position B, thus placing R7 in series with the unknown capacitor, and readjust R7. At null:

$$C_x = C_s(R_a/R1) \tag{5-2}$$

where,

R_a is the resistor (R2 to R6) selected by S1.

Inductance

1. Connect the unknown inductor (L_x) to terminals X-X.
2. Connect a standard inductor (L_s) to terminals S-S.
3. Set switch S2 to position A. This places R7 in series with the standard inductor, as the Q balance.
4. Connect an audio oscillator to the GEN terminals.
5. Connect an oscilloscope, a-c vtvm, or high-resistance headphones to the DET terminals.
6. Balance the bridge by means of R1 and S1.
7. Improve the null by adjusting R7. If no improvement results and null is imperfect, throw switch S2 to position B, thus placing R7 in series with the unknown inductor, and readjust R7. At null:

$$L_x = L_s(R1/R_a) \qquad 5\text{-}3$$

where,

R_a is the resistor (R2 to R6) selected by S1.

The plug-in nature of the universal bridge gives this instrument its great flexibility. Many circuits can be set up with it. The elimination of selector switches for changing the values of standards eliminates much of the trouble resulting from the stray reactance introduced by such switches. Perhaps the single disadvantage of this bridge is its lack of compactness: a reasonable supply of standards must be stored.

5.2 RESISTANCE-CAPACITANCE BRIDGE

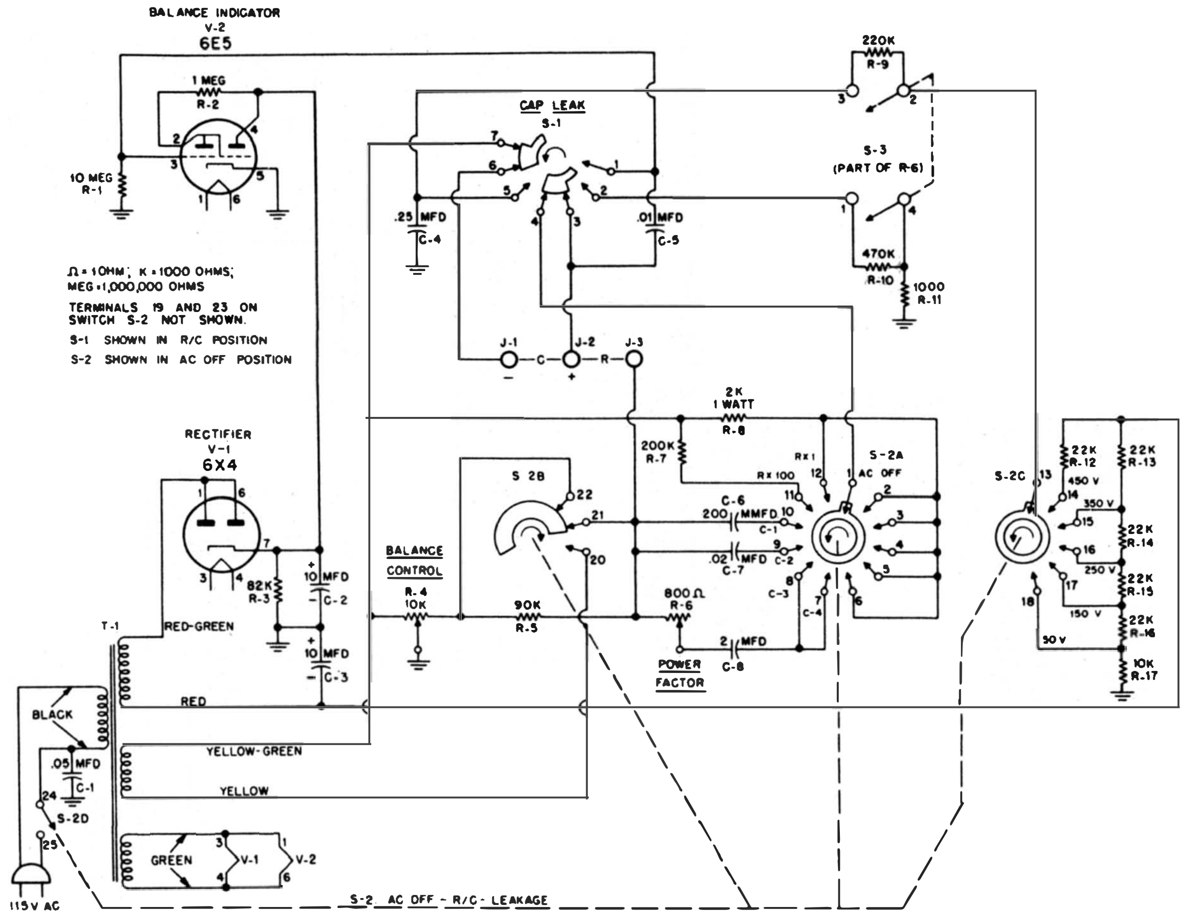
Among radio-tv servicemen and electronics maintenance technicians, a bridge for resistance and capacitance measurements is popular. Figs. 5-2 and 5-3 show an inexpensive bridge of this type (Knight KG-670) which is available either as a factory-built instrument or in kit form. This bridge covers the resistance range from 100 ohms to 5 megohms in two steps: 100–50,000 ohms, and 10,000 ohms–5 megohms; and it covers the capacitance range from 10 pf to 1000 mfd in four steps: 10 pf–0.005 mfd, 0.001–0.5 mfd, 0.1–50 mfd, and 20–1000 mfd. Additionally, it checks the leakage of capacitors at any of the following d-c voltages: 50, 150, 250, 350, or 450.

Fig. 5-3 shows the bridge circuit. This is a potentiometer-type slide-wire circuit in which the balance potentiometer (R4) supplies the two ratio arms of the bridge. The third arm of the bridge contains the unknown; and the fourth arm contains a standard capacitor (C6, C7, or C8) for capacitance measurements, or a standard resistor (R7 or R8) for resistance measurements. The power-factor balance rheostat (R6) is always in series with the standard capacitor, and its dial reads directly in power factor from 0 to 50 percent. The bridge signal is a 60-Hz voltage delivered by a special secondary winding on the power transformer, T1. The detector is a 6E5 magic-eye tube (V2). D-c voltage for the leakage test and for the magic-eye tube is supplied by voltage-doubler rectifier tube V1, and the step-type voltage divider (resistors R12-R17 and selector switch S2C). During the leakage test, the eye tube acts as a simple current indicator.

Courtesy Allied Radio Corp.

Fig. 5-2. Resistance-capacitance bridge.

An unknown capacitor is connected to terminals J1 and J2. An unknown resistor is connected to J2 and J3 (J2 is the common terminal for resistance and capacitance and is the positive terminal for the leakage test). J1 is negative. Switch S1 (lower left corner of the front panel in Fig. 5-2) selects either the resistance-capacitance or leakage function of the bridge. The three-section selector switch, S2a-S2b-S2c, in the upper right corner in Fig. 5-2 selects the various resistance, capacitance, and leakage-voltage ranges.



Courtesy Allied Radio Corp.

Fig. 5-3. Circuit of the resistance-capacitance bridge.

5.3 COMBINATION INDUCTANCE BRIDGE

Fig. 5-4 shows the circuit of an inductance bridge which is either a Maxwell or a Hay bridge, depending on the setting of a dpdt changeover switch, S3. An inductance range of 10 μh to 100 hy is covered in four steps: 10 μh–1 mh, 1–100 mh, 0.1–10 hy, and 1–100 hy. Only two standard capacitors are required for the four inductance ranges, C1 being used for the two lower ranges, and C2 for the two higher ones. The resistance in the No. 2 arm of the bridge is switched along with the standard capacitance. However, only three resistors (R1, R2, R3) are required, R2 being used on ranges B and C.

When switch S3 is set to its MAX position, the Q-balance rheostat (R5) is connected in parallel with the standard capacitor (C1 or C2), and the circuit is that of the Maxwell bridge (for comparison, see Fig. 4-10, Chapter 4). This arrangement is suitable for Q's higher than 10. When switch S3 is set to its HAY position, the Q-balance rheostat is in series with the standard capacitor, and the circuit is that of the Hay bridge (for comparison, see Fig. 4-9, Chapter 4). This arrangement is suitable for Q's up to 10.

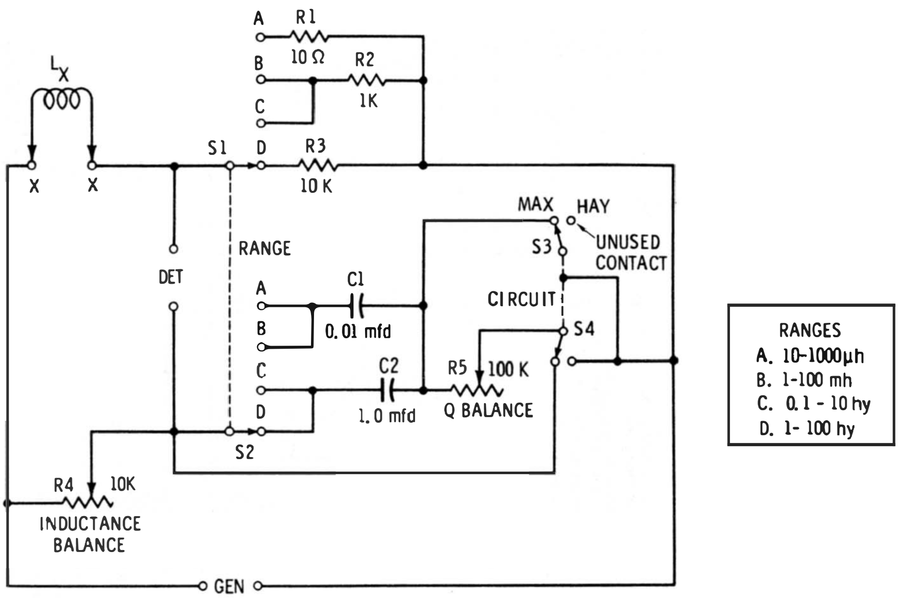
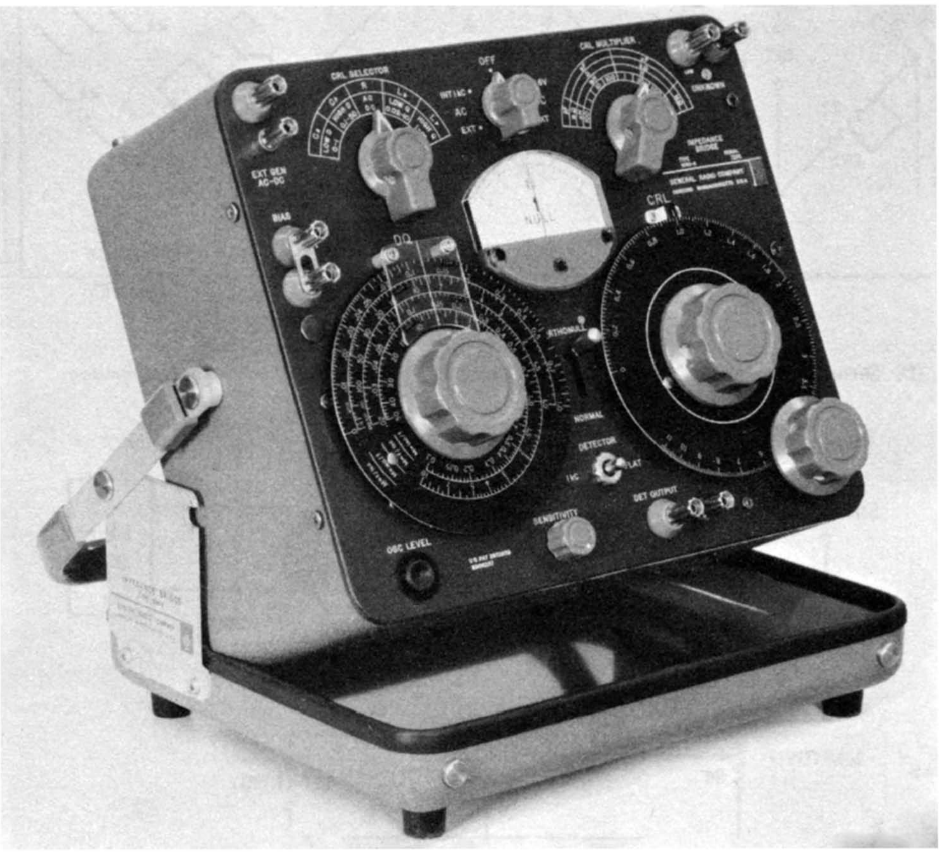


Fig. 5-4. Combination inductance bridge.

5.4 IMPEDANCE BRIDGE

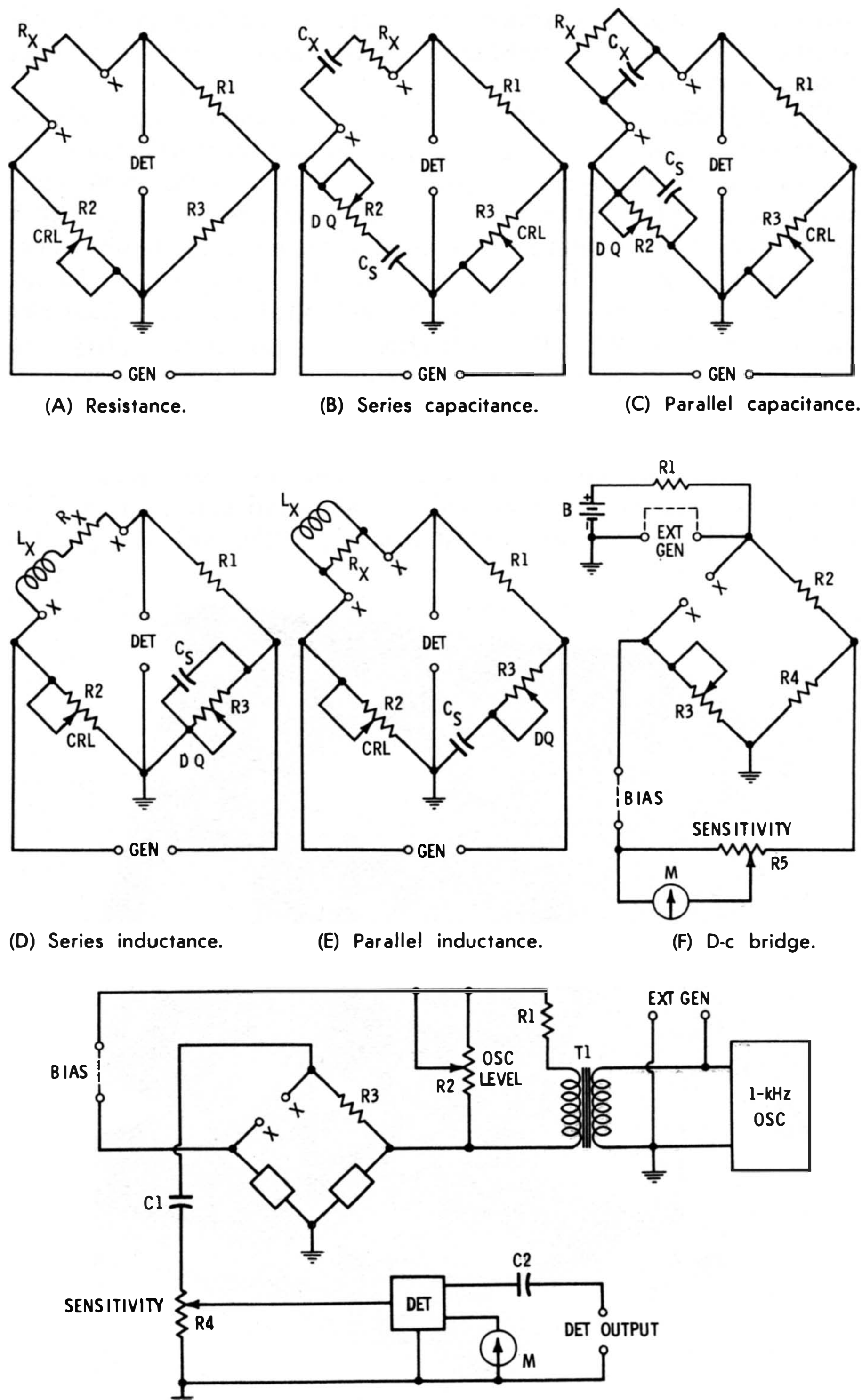
An *impedance bridge* is a general-purpose instrument for measuring capacitance, inductance, and a-c and d-c resistance. This all-around instrument accomplishes its several purposes by means of a switching arrangement which, by properly rearranging the components, sets up the resistance-, capacitance-, and inductance-measuring circuits.

Fig. 5-5 shows one such bridge (General Radio Type 1650-A) which provides the following ranges of measurement selected by the two switches in the upper left and right corners of the front panel: capacitance, 1 pf–1100 mfd (seven steps); inductance, 1 μh–1100 hy (seven steps); and resistance, 1 milliohm–11 megohms (eight steps). The lower right (CRL) dial reads direct capacitance, resistance, and inductance, and its reading is multiplied by the upper-right selector switch (CRL MULTIPLIER). The lower left (DQ) dial reads the dissipation factor (D) of capacitors or Q of inductors, depending on the setting of the upper-left selector switch (CRL SELECTOR). In capacitor tests, dissipation factor at 1 kHz may be measured between 0.001 and 1 when the resistance-balance rheostat is switched in series with the standard capacitor, and between 1 and 50 when this rheostat is in parallel with the standard



Courtesy General Radio Co.

Fig. 5-5. General Radio Type 1650-A impedance bridge.



(A) Resistance.

(B) Series capacitance.

(C) Parallel capacitance.

(D) Series inductance.

(E) Parallel inductance.

(F) D-c bridge.

(G) A-c bridge.

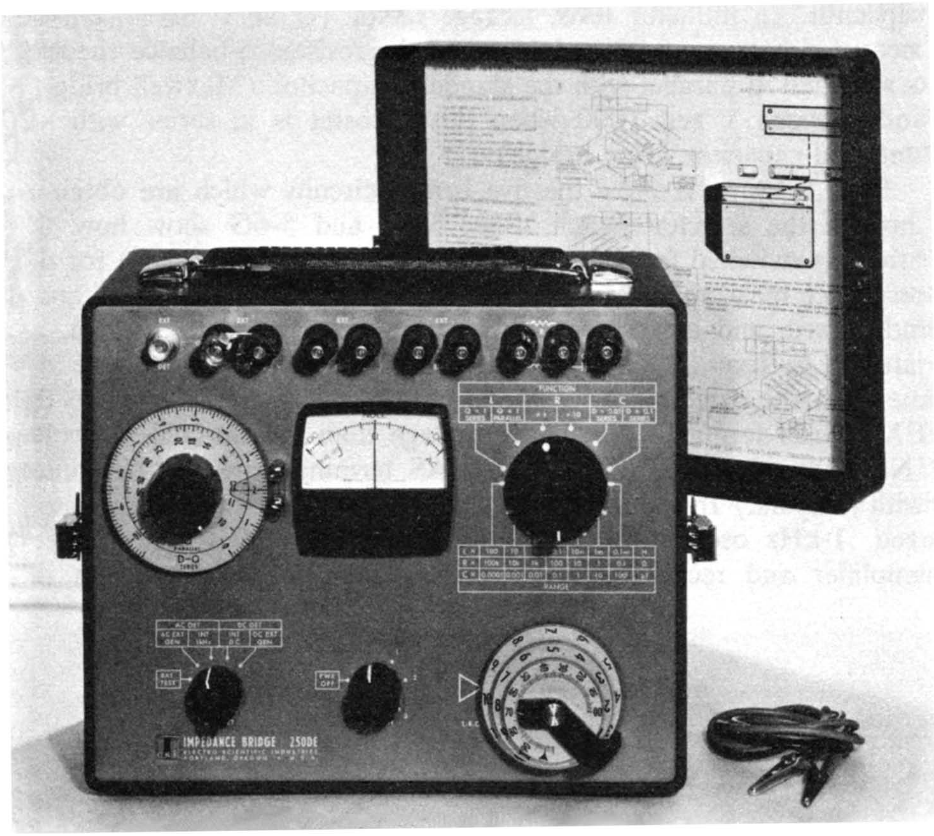
Fig. 5-6. Setups of the impedance bridge.

capacitor. In inductor tests, storage factor (Q at 1 kHz) may be measured between 0.02 and 10 when the resistance-balance rheostat is switched in parallel with the standard capacitor (Maxwell bridge), and between 1 and 1000 when this rheostat is in series with the standard capacitor (Hay bridge).

Figs. 5-6A to E show the five bridge circuits which are obtained through the selector switch. Figs. 5-6F and 5-6G show how the bridge is switched from the d-c Wheatstone type (Fig. 5-6F) for d-c resistance measurements to the a-c type (Fig. 5-6G) for capacitance, inductance, and a-c resistance measurements. In the d-c bridge, an internal battery (B) acts as the generator, and a center-zero d-c meter as the detector. An external battery may be connected to the EXT GEN terminals for higher voltage than the internal six volts. (Normally, the EXT GEN and BIAS terminals are short-circuited with jumpers.) In the a-c bridge (Fig. 5-6G), an internal, transistorized, 1-kHz oscillator acts as the generator, and a transistorized amplifier and rectifier (DET) driving the d-c meter acts as the

Courtesy Hewlett-Packard Co.

Fig. 5-7. Portable impedance bridge with automatic DQ balance.



Courtesy Electro Scientific Industries

Fig. 5-8. Wide-range, high-accuracy portable impedance bridge.

detector. (The amplifier may be peaked at 1 kHz or operated broad.) An external generator may be connected to the EXT GEN terminals for a different frequency, and a d-c supply may be connected to the BIAS terminals for d-c polarization of a capacitor or inductor. (Normally, the EXT GEN terminals are left open, and the BIAS terminals short-circuited with a jumper.) An external a-c detector (null indicator) may be connected to the DET OUTPUT terminals. In the a-c bridge, internal battery B powers both the oscillator and amplifier.

The circuits shown in Figs. 5-6A to E are those obtained by selector switching in all impedance bridges. Individual instruments differ in ranges of measurmeent, type of internal generator and detector, type of CRL and DQ readout, and whether an internal battery or external power line is used.

Fig. 5-7 shows an impedance bridge (Hewlett-Packard Model 4260A) which, in addition to the usual circuits (Wheatstone, capacitance comparison, Maxwell, Hay), embodies the following features:

1. A counter-type digital CRL readout (upper right indicator) with automatic placing of the decimal point by the range switch (center panel) and accordingly with elimination of calculations.
2. Selectable, all-electronic, automatic adjustment of the DQ balance in proper step with the manually operated CRL adjustment for rapid initial balancing for either C or L.

Either an internal or external a-c and d-c generator may be employed, as well as an internal or external detector. This bridge provides the following ranges: resistance, 10 milliohms–10 megohms (seven steps); capacitance, 1 pf–1000 mfd (seven steps); inductance, 1μh–1000 hy (seven steps); dissipation factor, 0.001–0.12 (series capacitance) and 0.05–50 (parallel capacitance); and Q, 0.02–20 (series inductance) and 8–1000 (parallel inductance).

The impedance bridge shown in Fig. 5-8 (Electro Scientific Model 250DE) provides the following ranges: resistance, 0–12 meg (eight steps); capacitance, 0–1200 mfd (seven steps); inductance, 0–1200 hy (seven steps); dissipation factor, 0–1.05 (two steps); and Q, 0–10.5 (series inductance) and 10–1000 (parallel inductance). The accuracy of measurement is superior to that of most portable impedance bridges: resistance, 0.1%; capacitance, 0.2%; and inductance, 0.3%. A solid-state ac-dc generator and detector are employed internally. The main-balance control is a special arrangement of fixed-decade steps and variable resistance, adjusted by means of the concentric CRL dial system in the lower right corner of the front panel.

CHAPTER 6

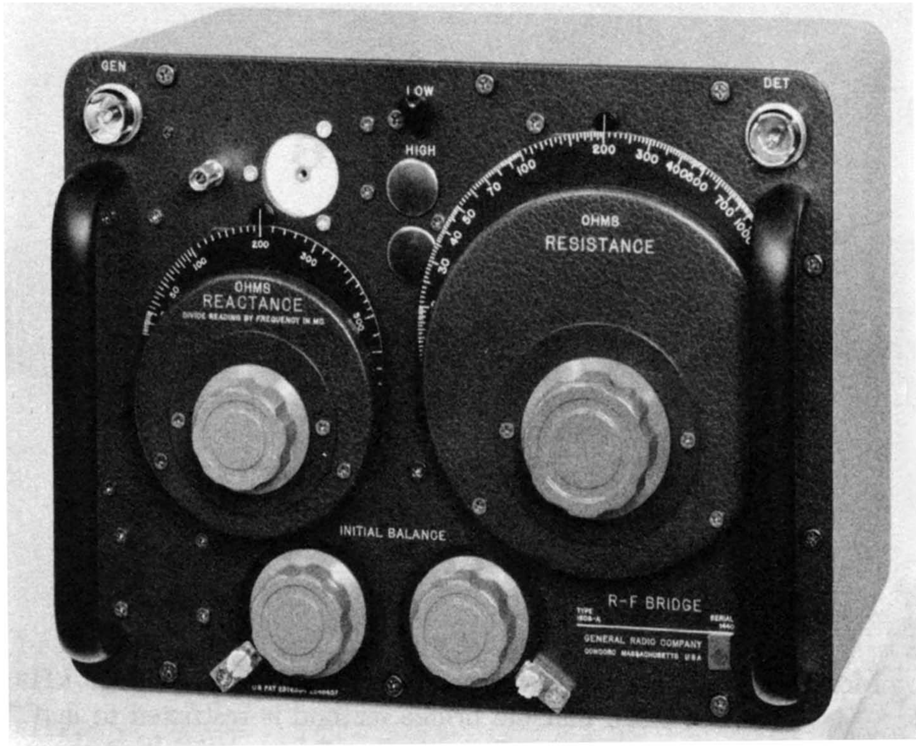
Radio-Frequency Bridges

Most bridges operate at frequencies between 60 Hz and 20 kHz. But this does not mean that the bridge method is restricted to audio frequencies. On the contrary, specially designed and built bridges are operated at frequencies as high as the vhf end of the radio spectrum, in the conventional configuration, and into the microwave spectrum in special configurations. Thus, the *radio-frequency bridge* allows networks, antennas, transmission lines, and circuit components to be checked at their actual operating frequencies.

Radio-frequency bridges, or r-f instruments employing the bridge principle, are used to advantage in the laboratory, shop, and field, and generally their adjustment is no more difficult than that of low-frequency bridges.

6.1 REACTANCE/RESISTANCE RADIO-FREQUENCY BRIDGE

Fig. 6-1 shows a laboratory-type r-f bridge (General Radio Type 1606-A) which, like lower-frequency bridges, has separate balances for reactance (X) and resistance (R). The R and X values measured with this instrument may be used directly, or they may be employed in the calculation of r-f impedance (Z). An external generator and external detector are required. The former must be a well-shielded r-f oscillator or signal generator, and the latter a well-shielded multiband radio receiver or heterodyne detector. This bridge may be operated at frequencies from 400 kHz to 60 MHz. Resistance values between 0 and 1000 ohms may be read directly from the right-hand dial (Fig. 6-1); the left-hand dial indicates reactance values up to



Courtesy General Radio Co.

Fig. 6-1. Professional radio-frequency bridge.

±5000 ohms at 1 MHz. The latter dial is direct reading only at 1MHZ; at all other frequencies, its readings must be divided by the frequency in MHz.

Fig. 6-2 shows the basic circuit of the bridge. This is a version of the Schering bridge (see Section 3.6, Chapter 3). The use of variable

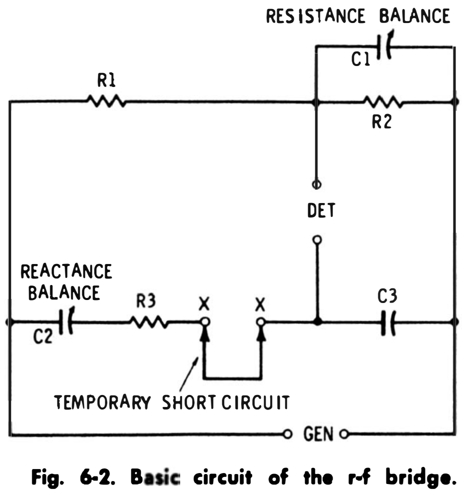


Fig. 6-2. Basic circuit of the r-f bridge.

air capacitors for each balance (C2 for reactance and C1 for resistance) is an advantage, since strays, which are particularly disturbing at radio frequencies, may be minimized more easily in variable air capacitors than in the usual balancing rheostats and potentiometers. A substitution method is employed. In this procedure, the unknown terminals (X-X) first are short-circuited by means of a short, stout jumper, as shown in Fig. 6-2, and the bridge balanced by adjustment of capacitors C1 and C2. (The capacitor settings are recorded as C_{1a} and C_{2a}.) Then, the unknown impedance is connected to the X-X terminals in place of the jumper, and the bridge is rebalanced. (The new settings of the capacitors are recorded as C_{1b} and C_{2b}.) The unknown resistance (R_x) and unknown reactance (X_x) may be determined in terms of these settings. At null:

$$R_x = R1\frac{C_{1b} - C_{1a}}{C3} \qquad \text{6-1}$$

and

$$X_x = \frac{1}{\omega}\left[\frac{1}{C_{2b} - C_{2a}}\right] \qquad \text{6-2}$$

Fig. 6-3 shows the complete bridge circuit. In this arrangement, the resistance-balance capacitor (C2 corresponds to C1 in the basic circuit (Fig. 6-2). Similarly, the reactance-balance capacitor (C6) corresponds to C2 in the basic circuit. Separate capacitors (C1 and C7) are provided for initial resistance and reactance balance, respectively. This enables the main dials (of C2 and C6 to be set to zero for the initial balance, and direct readings of resistance and reactance then to be obtained. Switch S1 enables the initial reactance balance to be made at either the high end or the zero end of the C6 dial. The internal compensating capacitors (C3 and C4) preadjust the circuit for this HIGH/LOW feature.

Note the elaborate shielding of the bridge, including the concentric DET and GEN jacks, J1 and J2. This shielding reduces the effects of strays and protects the bridge from external fields. The generator coupling transformer (T1) is a wide-band component, operating efficiently over the 400 kHz-60 MHz range. The high permeability of the ferrite core of this transformer permits the use of just a few turns in both the primary and secondary coils, and this markedly reduces internal capacitances.

6.2 SWR BRIDGE

A simplified version of the radio-frequency bridge is used for measurement of *voltage standing-wave ratio* (vswr, or swr). This type of measurement finds wide application in transmitter practice, and this kind of instrument is standard equipment in communications shops and transmitting stations. Because of its low cost and comparative simplicity, it is popular with amateurs and experimenters who may either build it or buy it factory-assembled or in kit form.

Fig. 6-4A shows the circuit of the swr meter. This instrument is convenient for checking antennas, networks, and components at radio frequencies up to 100 MHz. The device under test (load) is connected

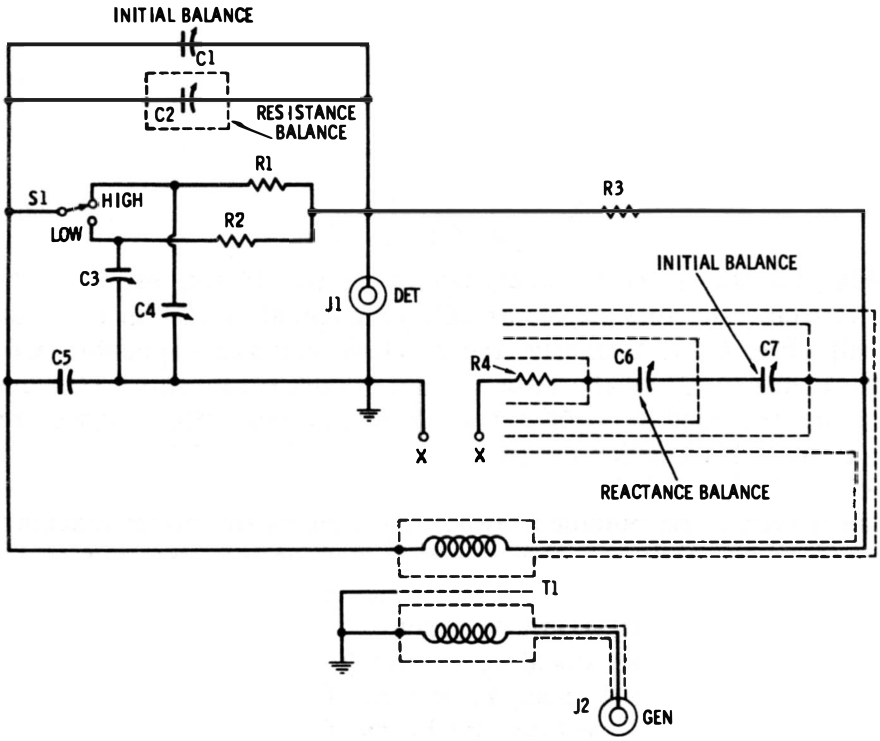


Fig. 6-3. Complete circuit of the r-f bridge.

to coaxial jack J2 by the shortest possible leads, and the transmitter or signal generator is connected to coaxial jack J1. Meter M reads the input signal when switch S1 is thrown to position A, and the signal across the external load when S1 is thrown to B. This meter is previously calibrated, in combination with diodes X1 and X2 and resistors R1, R2, R3, and R7, to read r-f volts directly. The bridge arms consist of noninductive resistors R4 and R5 in one line, and noninductive resistor R6 and the device (RL) under test (connected to jack J2) in the other line; see equivalent circuit, Fig. 6-4B).

All resistors must be 1-watt, 1% noninductive units. The resistance of R6 must equal exactly the impedance of the transmission line, network, antenna, or component connected to J2 for test; otherwise, the bridge will not null. (Common impedance values met in practice are 52, 75, 300, and 600 ohms.) When R4, R5, and R6 have the exact specified values, the bridge will null at the selected "J2 impedance." That is, meter M will read zero when S1 is at position B, and the bridge is driven by an r-f input voltage applied through jack J1. Layout of the instrument must be carefully planned to minimize stray capacitance, inductance, and coupling, and fully shielded to prevent r-f pickup from the fields always present around transmitters.

When the instrument is in use, an r-f signal at the desired frequency is applied to the circuit through jack J1; a noninductive dummy resistor equal to RL (the impedance of the device being

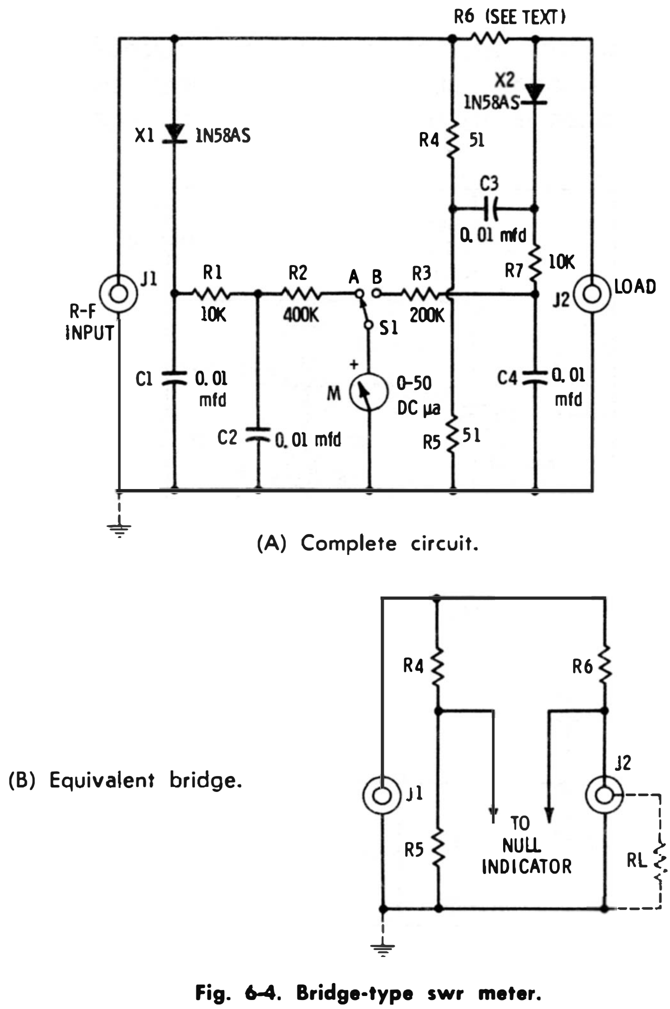


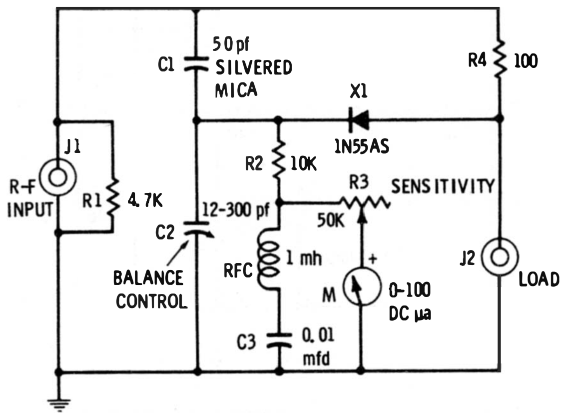
Fig. 6-4. Bridge-type swr meter.

tested) is connected by the shortest possible leads temporarily to J2; a noninductive resistor equal to the impedance or resistance R_L also is connected in the position of R6; and switch S1 is thrown to position A. Next, the output of the signal generator is adjusted for full-scale deflection of meter M, and the voltage indicated by M is recorded as E_1. Switch S1 then is thrown to position B, whereupon the meter reading should fall to zero; since $R6 = R_L$. Then, the dummy resistor is removed from J2, and the device under test (RL) is connected in its place. Any reading of the meter at this point is recorded as E_2. Finally, the standing-wave ratio is calculated:

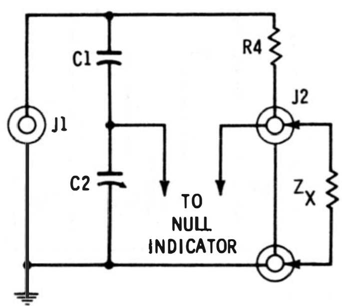
$$\text{swr} = \frac{E_1 + E_2}{E_1 - E_2} \qquad 6\text{-}3$$

6.3 R-F IMPEDANCE BRIDGE

Fig. 6-5A shows the circuit of an r-f impedance bridge suitable for measuring the impedance of amateur and Citizens-band antennas, transmission lines, filters, and circuit components between 20 and 600 ohms at frequencies up to 50 MHz. After the instrument is cali-



(A) Complete circuit.



(B) Equivalent bridge.

Fig. 6-5. R-f impedance bridge.

brated, the impedance in ohms may be read directly from the dial of variable capacitor C2 (the balance control). The null detector is a diode-type r-f voltmeter consisting of a miniature 1N55AS germanium diode (X1), resistor R2, rheostat R3, r-f choke RFC, and bypass capacitor C3. Rheostat R3 serves as a sensitivity control. Since M is used only as a null indicator, no voltage or current calibration is required. An r-f signal voltage from an oscillator or signal generator is applied to the circuit through jack J1. Finally, the unknown impedance is connected by the shortest possible leads to jack J2.

The arms of the bridge are capacitors C1 and C2 in one line, and resistor R4 and the unknown impedance (Z_x, connected to jack J2) in the other line (see equivalent circuit, Fig. 6-5B). Variable capacitor C2 is adjusted to balance the bridge. At null, the unknown impedance may be calculated in terms of resistance R4 and capacitances C1 and C2:

$$Z_x = C1R4/C2 \qquad 6\text{-}4$$

The simplest way to calibrate the instrument is to balance the bridge separately with as many noninductive resistors (between 20 and 600 ohms) as are available connected successively to jack J2 with short leads, and to inscribe the dial of variable capacitor C2 with these resistance values.

After initial calibration, use of the instrument is simple:

1. An r-f test signal of desired frequency is injected into the circuit via jack J1.
2. The device under test (antenna, transmission line, filter, or circuit component) is connected by the shortest possible leads to jack J2.
3. Variable capacitor C2 is adjusted to balance the bridge, and rheostat R3 is adjusted to increase the sensitivity of the meter as null is approached.
4. At null, the unknown r-f impedance is read directly from the calibrated dial of C2.

When the bridge is used in the adjustment of an impedance Z (e.g., adjustment of a network or load device), the C2 dial is set to the desired impedance value, and the external impedance (Z) is then adjusted for null, as indicated by meter M.

6.4 BRIDGE-TYPE UHF ADMITTANCE METER

Fig. 6-6 shows the basic arrangement of a coaxial-type admittance meter (General Radio Type 1602-B) which gives direct readings of conductance and susceptance of antennas, transmission lines, and components at test frequencies from 40 to 1500 MHz. This instru-

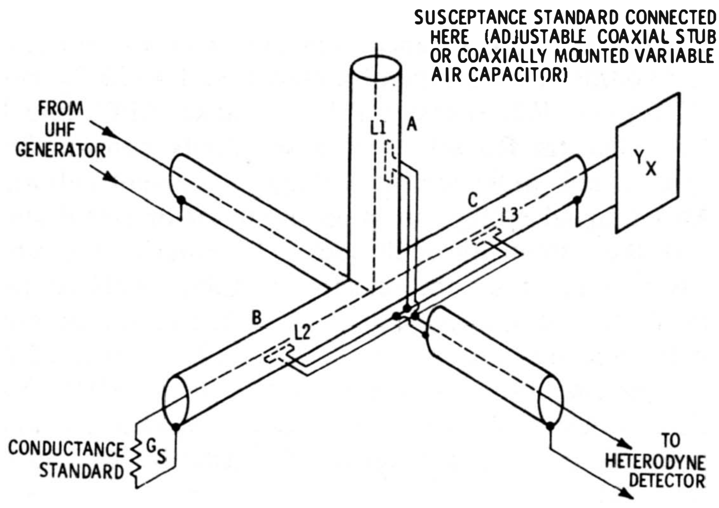


Fig. 6-6. Bridge-type uhf admittance meter.

ment requires an external uhf signal generator and external heterodyne detector.

The instrument consists of a system of joined coaxial lines. The unknown admittance (Y_x) is connected to line C, a conductance standard (G_s) to line B, and a susceptance standard to line A. The conductance and susceptance standards screw directly on the ends of line B and line A, respectively.

Current in each of the three lines is sampled by means of a rotatable loop (L1, L2, L3). The three loops are connected in parallel and to the detector. A dial attached to L1 read directly in susceptance (millimhos), one attached to L2 reads directly in conductance (millimhos), and one attached to L3 reads a multiplying factor (from 1.1 to infinity). With the bridge energized by the generator, the loops are adjusted for balance, as indicated by the detector. At null, the susceptance and conductance of the device under test are read from the settings of the L1 and L2 dials, respectively, and these values are multiplied by the setting of the L3 dial.

If a suitable constant-impedance adjustment line (such as General Radio Type 874-LK20L) is connected between the unknown (Y_x) and line C, and this line is set to one-quarter wavelength at the generator frequency, the two main dials will read impedance components of the device. That is, the L1 dial will read reactance and the L2 dial series resistance.

CHAPTER 7

Frequency-Measuring Devices

A few bridge circuits are frequency sensitive, i.e., their null setting changes with generator frequency. This necessitates that a single test frequency be accurately maintained if direct-reading dials are used on such bridges. It also means that a frequency term often appears in formulas for calculating inductance, capacitance, and equivalent resistance.

While frequency sensitivity can be somewhat of a nuisance in capacitance, resistance, and inductance bridges, this property may be utilized for the measurement of frequency if all of the bridge arms are filled and the bridge is balanced for frequency. In such an instance, the balance adjustment may be made direct reading in frequency instead of in R, C, or L. The resistance, capacitance, and/or inductance values in the bridge arms usually can be known with high precision, so that the bridge becomes a useful dveice for accurate and simple measurement of frequency.

Measurements of this sort are usually limited to frequencies no higher than 20 kHz, since small stray reactances blunt the null response and may even bypass the signal around the bridge at higher frequencies.

Aside from conventional bridges, some nonbridge null circuits function reliably for frequency measurement and sometimes are preferred to bridges (see Chapter 9).

7.1 SERIES-TYPE RESONANCE BRIDGE

Fig. 7-1 shows the circuit of a *resonance bridge* of the series type. This bridge may be balanced only at the frequency determined by inductance L1 and capacitance C1 in one arm of the circuit. At this resonant frequency of the L1-C1 combination, the inductive reactance cancels the capacitive reactance, and only the inherent resistances of the inductor and the capacitor remain in the upper left arm of the bridge, making the circuit a four-arm resistance bridge at that one frequency. At null, the unknown frequency is:

$$f_x = 1/(2\pi\sqrt{LC}) \qquad 7\text{-}1$$

where,

C is the value of C1 in farads,
f is in Hz,
L is the value of L1 in hy
π equals 3.1416.

Frequency coverage may be provided by making C1 variable for tuning, and by switching L1 to appropriate values to change ranges. (Conversely, a d-c–tuned variable inductor might be employed for

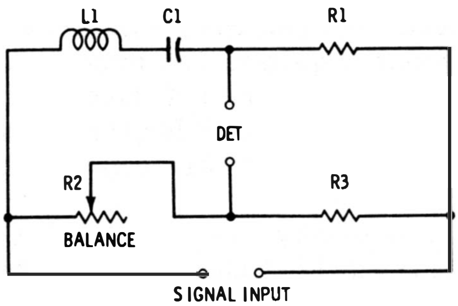


Fig. 7-1. Series-type resonance bridge.

tuning, and capacitors switched to change ranges.) Since high capacitances are needed in the audio spectrum, a variable C1 would, for practical purposes, be a capacitor decade.

For sharp null response, both L1 and C1 must be high-Q components. This calls for a low-resistance inductor of toroidal construction and for a mica capacitor.

7.2 SHUNT-TYPE RESONANCE BRIDGE

Because the impedance arm in Fig. 7-1 is a series-resonant circuit, the current through this arm and the one containing resistor R1 is maximum at the resonant frequency and can reach high levels. In some inductors, especially when high input-signal voltage and low R1 resistance are unavoidable high current can cause distortion.

In Fig. 7-2, the impedance arm has been changed to a shunt connection of inductance and capacitance (parallel-resonant circuit).

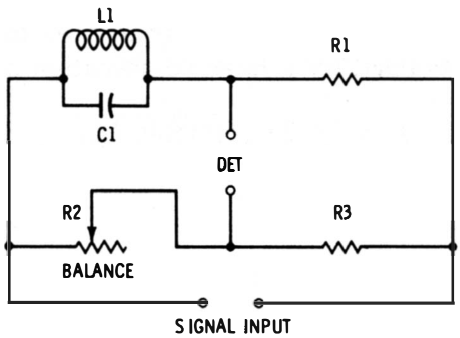


Fig. 7-2. Shunt-type resonance bridge.

Here, at resonance the current through this arm and the one containing resistor R1 is minimum, approaching zero for high-Q capacitors and inductors.

In all other respects, the shunt-type bridge is identical to the series type. The balance equation is the same (see equation 7-1).

7.3 WIEN BRIDGE

The Wien bridge (Fig. 7-3) is very desirable for frequency measurements, since it uses only capacitors and resistors and can be made very compact, and since its balance equation can be greatly simplified.

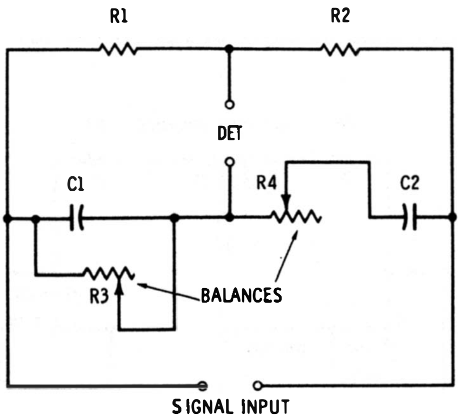


Fig. 7-3. Wien bridge.

In this circuit, there are two resistance arms (R1 and R2) and two adjustable impedance arms (C1-R3 and C2-R4). One impedance arm must have its resistance and capacitance in parallel (like C1-R3) and the other one must have its resistance and capacitance in series (like C2-R4). Alternatively, resistors R3 and R4 may be fixed, and capacitors C1 and C2 variable.

Both variable elements must be adjusted to balance the bridge, and null may be obtained for a large combination of R and C values:

$$f_x = 1/(2\pi\sqrt{R3R4C1C2}) \qquad 7\text{-}2$$

where,

C is in farads,
f_x is in Hz,
R is in ohms,
π equals 3.1416.

If a dual rheostat is used for simultaneous adjustment of the two variable resistances so that R3 equals R4 at all settings, if resistance R2 is made twice R1, and if the capacitors are matched so that C1 equals C2, the balance equation can be simplified to:

$$f_x = 1/(2\pi R3C1) \qquad 7\text{-}3$$

This reuslts in a simplified circuit which has often been used as a highly practical audio-frequency meter.

Fig. 7-4 shows the complete circuit of a Wien-bridge audio-frequency meter which covers the range from 20 Hz to 20 kHz in three steps: (A) 20–200 Hz, (B) 200–2000 Hz, and (C) 2–20 kHz. The single tuning control is the dual 10,000-ohm wirewound balancing rheostat (R4-R5). The three ranges are selected by double-pole, three-position switch S1-S2 which switches capacitors in identical pairs (C1-C4, C2-C5, and C3-C6). The dial attached to rheostat R4-R5 is calibrated to read directly in Hz on the lowest (20–200-Hz)

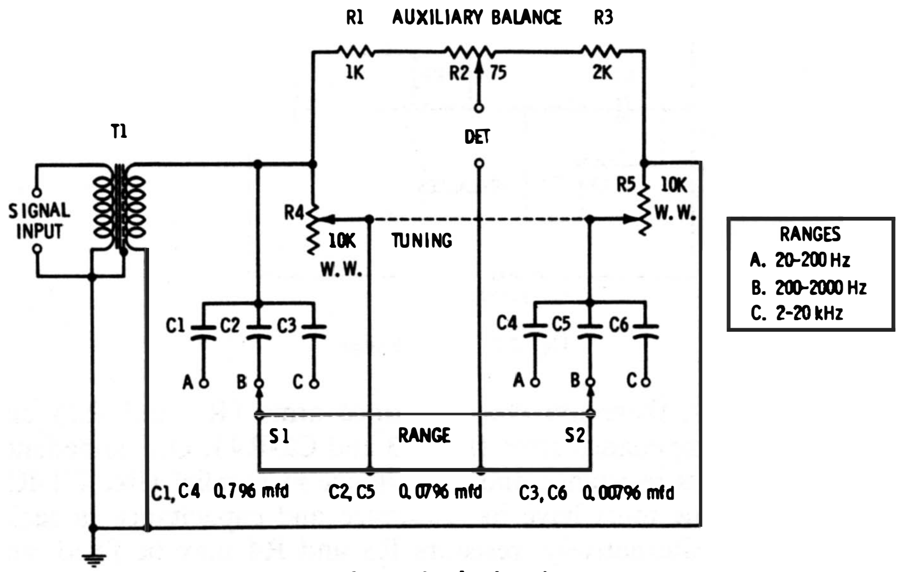


Fig. 7-4. Complete Wien-bridge frequency meter.

range, and the switch settings then multiply this range by 1 (range A), 10 (range B), and 100 (range C). Potentiometer R2 is an auxiliary control which allows the null to be sharpened without upsetting the frequency reading of the R4-R5 dial. Resistors R1 and R3 provide the 2:1 ratio required for simplification of the bridge balance. These resistors and all of the capacitors must be accurately rated (1% or better).

A shielded input transformer (T1) is required. The detector may be an oscilloscope, a-c vtvm, magic-eye tube, high-impedance a-c null detector, or high-impedance headphones with or without an amplifier.

The instrument is calibrated best by feeding as many accurate frequencies as are available between 20 and 200 Hz into the circuit via the SIGNAL INPUT terminals, and balancing the bridge by adjustment of rheostats R4-R5 and potentiometer R2, with switch S1-S2 in its position A. The rheostats and potentiometer R2 should be adjutsed alternately for the deepest obtainable null. At each null, the corresponding frequency is inscribed on the R4-R5 dial. Only this one range need be calibrated; if capacitors C1-C6 are accurate, the two higher frequency ranges will track automatically.

7.4 SUPPLEMENTARY USES OF FREQUENCY BRIDGES

Besides frequency identification, there are numerous other applications of frequency-sensitive bridges—all of which exploit the ability of this type of bridge to null at one frequency. Thus, the bridge can function as a band-elimination filter (notch filter), heterodyne suppressor in radiophone reception, interference eliminator in radiotelegraph reception, fundamental remover in audio distortion measurements, and tuning network in audio amplifiers and oscillators.

In Chapter 10 will be found detailed explanations of some of the supplementary applications of frequency-sensitive bridges.

CHAPTER 8

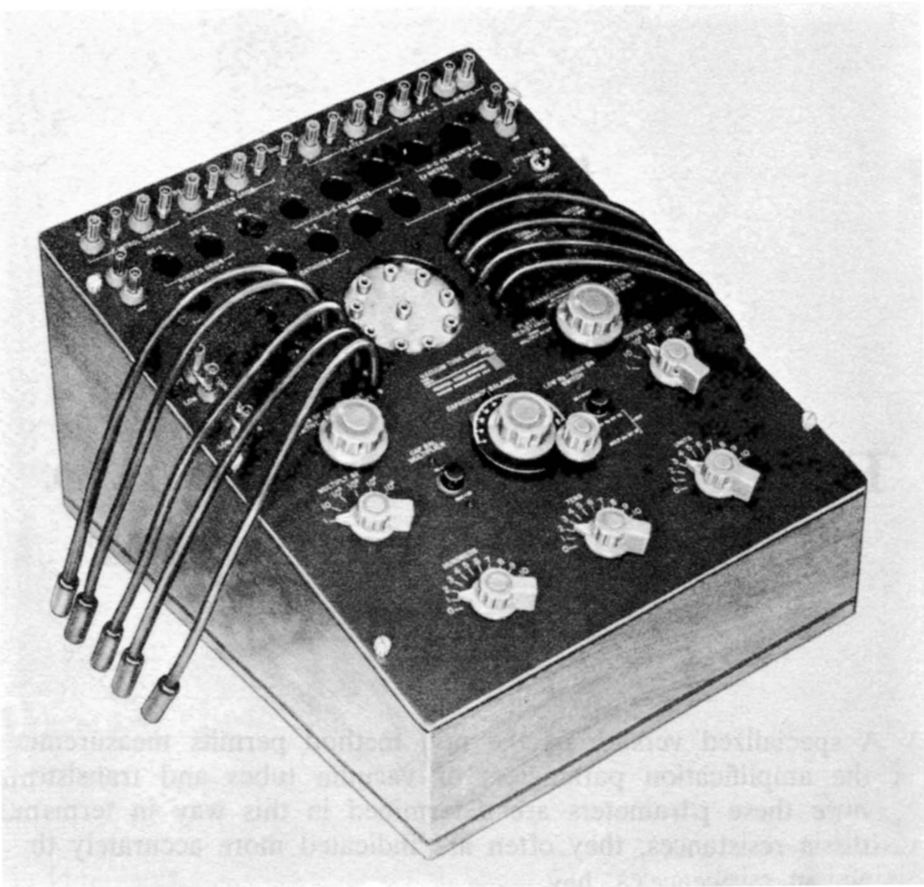
Bridge Measurement of Tube and Transistor Parameters

A specialized version of the null method permits measurement of the amplification parameters of vacuum tubes and transistors. Because these parameters are determined in this way in terms of precision resistances, they often are indicated more accurately than when meters alone are used.

Some bridges used for tube and transistor testing are hybrids. All, however, exploit the advantages inherent in the null method. Of these instruments, the classic vacuum-tube bridge is seldom found outside the laboratory, meter-type tube testers commonly being used for nonengineering testing in the field. But bridge-type transistor testers may be found both in the laboratory and in the field.

8.1 VACUUM-TUBE BRIDGE

Fig. 8-1 shows a vacuum-tube bridge (General Radio Type 1161-B). In instruments of this type, the tube under test provides one or two of the arms of a bridge, which then is balanced in the presence of a selected combination of applied operating and signal voltages. The tube parameter of interest (such as plate resistance, amplification factor, or transconductance) is read from the values of the bridge resistances at null. Under certain conditions, the bridge balance control is direct reading in the parameter. The a-c and d-c operating power for the tube is obtained from an external supply, and an external generator and detector are used.



Courtesy General Radio Co.

Fig. 8-1. Professional vacuum-tube bridge.

By means of selector switches and/or patch cords, various circuits are set up in the bridge for testing the tube for the desired parameters. Typical circuits of this sort are described in the following Sections 8.2, 8.3, 8.4, and 8.5, and illustrated in the corresponding figures. While, for simplicity, a triode tube is shown in each of these circuits, multielement tubes also may be tested, provided proper voltages are supplied.

8.2 TUBE PLATE RESISTANCE

Fig. 8-2 shows the circuit for measuring the dynamic plate resistance (R_p) of a vacuum tube. Here, tube V1 (under test) is supplied with grid voltage by variable d-c supply E_c, and with plate voltage by variable d-c supply E_b. Plate current (I_p) is adjusted to the desired operating level, as indicated by d-c milliammeter M1, by varying E_b. Plate-to-ground and grid-to-ground d-c voltage may be

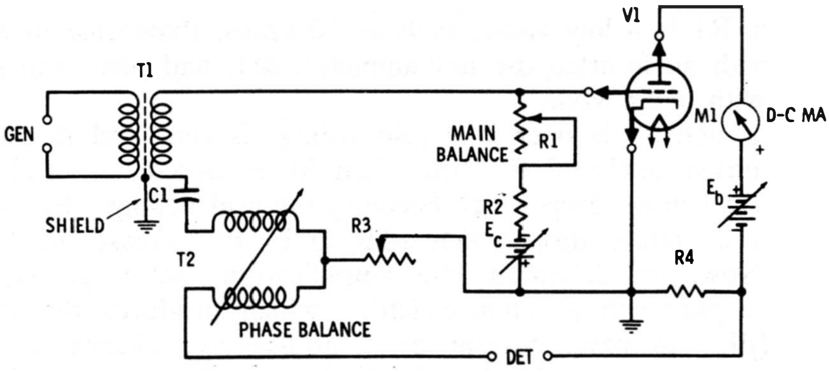


Fig. 8-2. Circuit for tube plate resistance.

read with a high-resistance d-c voltmeter or d-c vtvm. The effect of tube capacitance in the bridge is compensated by adjustment of a variometer—the *phase balance*, T2.

The input signal (normally 1 kHz) is applied through the shielded transformer, T1. The output signal is indicated by an a-c null detector connected to the DET terminals.

Rheostat R3 is set to a resistance equal to 10 times the amplification factor of the tube, and the bridge is balanced by adjustment of rheostat R1. If R4 is precisely 1000 ohms, then at null the plate resistance may be found in terms of R1:

$$R_p = 100R1 \qquad 8\text{-}1$$

8.3 TUBE AMPLIFICATION FACTOR (D-C)

Fig. 8-3 shows the circuit for measuring the static amplification factor (d-c μ) of a vacuum tube. Here, tube V1 is supplied with the desired level of plate voltage by variable d-c supply E_b, and with grid voltage by the voltage drop produced across resistor R1 by current flowing from d-c signal source E_i through the R1-R2 line.

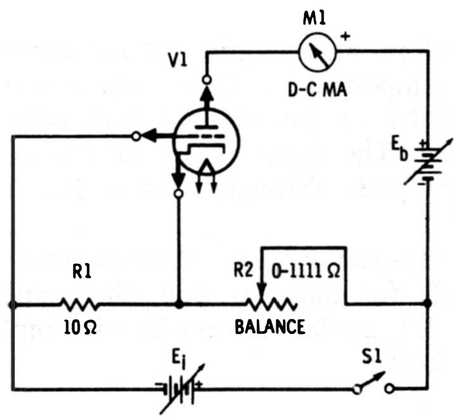


Fig. 8-3. D-c circuit for tube amplification factor.

Resistance Rl is a low value, such as 10 ohms. Plate current may be read with an inserted d-c milliammeter, M1, and plate and grid voltages with a d-c vtvm.

When switch S1 is open, the grid voltage is zero and the plate voltage consist solely of E_b. But when S1 is closed, E_t produces across R1 a voltage drop which becomes the grid voltage (E_g), and across R2 a voltage drop which adds to E_b to increase the plate voltage. Now, by definition, the amplification factor, μ, is the ratio of a plate-voltage change (dE_p) which produces the same change (dI_p) in plate current as a grid-voltage change (dE_g). But because of the amplification provided by a tube, the grid voltage is μ times as effective as the plate voltage in producing a given plate-current change.

Operation of the circuit consists accordingly of comparing these elements. This consists of opening switch S1 and noting the static plate current, I_p. Then S1 is closed, whereupon I_p increases because of the application of grid voltage. Rheostat R2 next is adjusted to bring I_p back to its original "zero" value (a kind of null adjustment). At this point, $R2 = \mu R1$, and:

$$\mu = R2/R1 \qquad 8\text{-}2$$

8.4 TUBE AMPLIFICATION FACTOR (A-C)

The static value of μ afforded by the d-c bridge test described in the preceding section is often considered an inadequate indicator of tube performance, since, except in steady-state d-c amplifiers and certain control circuits, tubes usually operate under dynamic conditions—that is, with a-c signals. A dynamic test of amplification factor thus may be considered more suitable. Fig. 8-5 shows a circuit for the bridge measurement of this parameter.

This arrangement is similar to the d-c circuit of Fig. 8-4, but here the a-c signal voltage, supplied through shielded transformer T2, develops an a-c grid voltage across 10-ohm resistor R1, and an a-c component of plate voltage across rheostat R2. Balance is indicated by an external a-c null detector connected to the DET terminals. (The input circuit of this detector must provide a low-resistance path, through resistor R_d, for the d-c plate current of tube V1.)

Plate rheostat R2 and phase-balance variometer T1 are adjusted alternately for null. At null, the amplification factor is the ratio of R2 to R1, a relationship which is simplified because the resistor R1 is equal to 10:

$$\mu = 0.1R2 \qquad 8\text{-}3$$

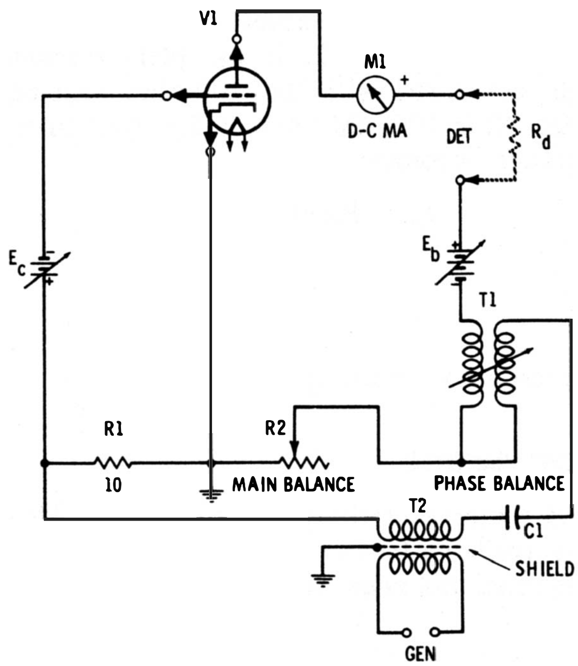


Fig. 8-4. A-c circuit for tube amplification factor.

8.5 TUBE TRANSCONDUCTANCE

Fig. 8-5 shows the circuit for bridge measurement of dynamic transconductance (g_m) of tube V1. Here, the desired level of plate voltage is provided by variable d-c source E_b, and grid voltage by variable d-c source E_c. An a-c grid-signal voltage results from the voltage drop across R2, produced by generator current flowing through R2. Capacitor C1 provides d-c blocking in the grid circuit.

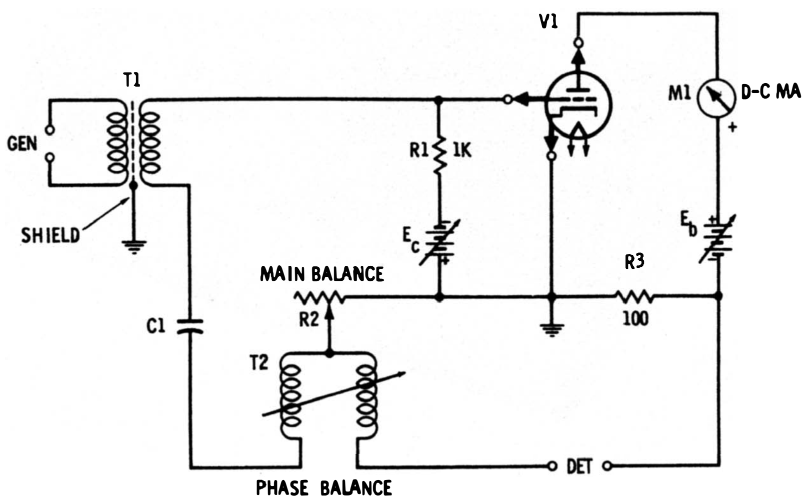


Fig. 8-5. Circuit for tube transconductance.

The bridge is balanced by alternate adjustment of rheostat R2 and phase-balance variometer T2. If the plate resistance (R_p) of the tube is high, compared with the 100-ohm resistance, R3, the setting of rheostat R2 is 100,000 times μ/R_p. And since $g_m = \mu/R_p$, the transconductance becomes:

$$g_m = R2/100{,}000 \qquad 8\text{-}4$$

where,

g_m is in mhos,
R2 is in ohms.

To convert to micromhos, multiply g_m by 1,000,000.

8.6 TRANSISTOR BRIDGE

Fig. 8-6 shows a representative transistor bridge (Heathkit Model IM-30) which is available factory wired or as a kit. This instrument operates from internal batteries or from an external power supply.

Courtesy Heath Company

Fig. 8-6. Bridge-type transistor tester.

Its center-zero meter is used as the null detector and also as a voltmeter and current meter for checking various transistor voltages and currents.

Selector switches and key switches enable the various functions of the instrument to be obtained, and the dial of the balance control is graduated directly in d-c alpha and d-c beta, the amplification factors of transistors.

Transistor bridge circuits are described in Sections 8.7 and 8.8.

8.7 D-C AMPLIFICATION FACTOR OF TRANSISTOR

The current amplification factor (β—beta or h_{fe}) of a transistor is comparable to the voltage amplification factor (μ) of a vacuum tube. Beta is the ratio of collector-current change to the base-current change which produces it (i.e., $\beta = dI_c/dI_b$). Static (d-c) beta and dynamic (a-c) beta both may be measured with bridge-type circuits. Fig. 8-7 shows a d-c circuit.

In this arrangement, PNP transistor X1 is supplied with base current from variable d-c source E_b through limiting resistor R1, and with collector current from variable d-c source E_c. The polarity of each of these sources must be reversed for an NPN transistor. If desired, collector current may be read with d-c milliammeter M3, and collector voltage with high-resistance d-c voltmeter M2. D-c meter M3 is a center-zero galvanometer or other d-c null detector.

Base-current flow develops a voltage drop (E_1) across resistor R2, and collector-current flow develops a voltage drop (E_2) across

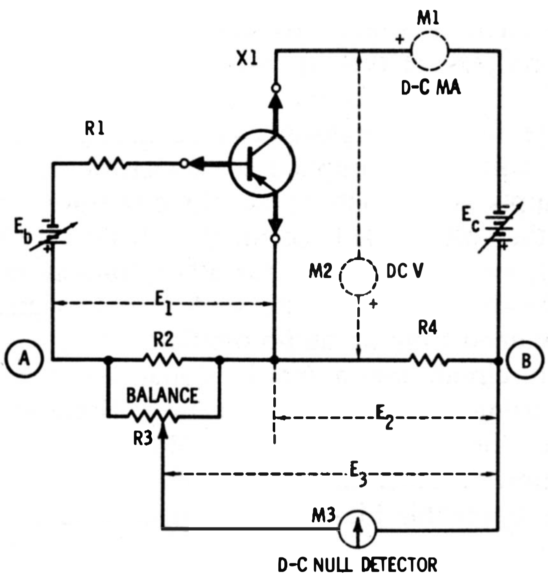


Fig. 8-7. Circuit for d-c amplification factor of transistor.

resistor R4. If either of these resistors is varied and the other held constant, E_1 can be made equal to E_2, whereupon voltage E_3 between points A and B will be zero. At this null point, beta is equal to E_2/E_1, and therefore is proportional to R4/R2. Since it is impossible to vary R2 without disturbing the operating point of the transistor, a potentiometer, R3, whose resistance is much higher than that of R2 is provided for sampling the voltage across R2 and balancing it against E_2. In this way, R2 and R4 may be kept fixed and their ratio changed for various beta ranges.

At null ($E3 = 0$), the amplification factor may be determined in terms of the base and collector voltages:

$$\beta = E_2/E_1 \qquad 8\text{-}5$$

The dial of potentiometer R3 may be graduated for direct reading of beta according to this ratio. It may be calibrated also on the basis of the resistance ratio (i.e., the ratio of R4 to the resistance of R2 and R3 in parallel). The dial may also be graduated in alpha, α (current amplification factor of the common-base configuration) on the basis of $\alpha = \beta/\beta + 1$.

8.8 A-C AMPLIFICATION FACTOR OF TRANSISTOR

Just as the dynamic (a-c) value of the tube amplification factor is sometimes preferred to the static value (see Section 8.4), so is the a-c beta of a transistor sometimes preferred to the d-c value. Fig. 8-8 shows a bridge circuit for measuring the a-c amplification factor.

This arrangement is similar to the d-c circuit described in the preceding section, except that an a-c test signal is supplied through transformer T1, and an external a-c detector (oscilloscope, a-c vtvm, magic-eye tube, or high-impedance headphones) is required. D-c operating voltage is supplied to transistor X1 under test by variable d-c supply B. A double-pole, three-position reversing switch (S1-S2) turns the voltage OFF (center), ON positive for NPN transistors (lower), or ON negative for PNP transistors (upper). C1 and C2 are d-c–blocking capacitors. A 1-kHz signal is desirable and may be obtained from an audio oscillator; however, many simple versions of this circuit use a 60-Hz signal taken from the power line, with T1 being a filament transformer. Because R3 is a very high resistance, the base-current generator is essentially constant current in nature. Another high resistance, R2, determines the d-c base bias. It is upon this bias that the much smaller a-c test signal is superimposed.

When the circuit is in operation, the transistor amplifies the a-c base current, and the resulting a-c collector current develops an

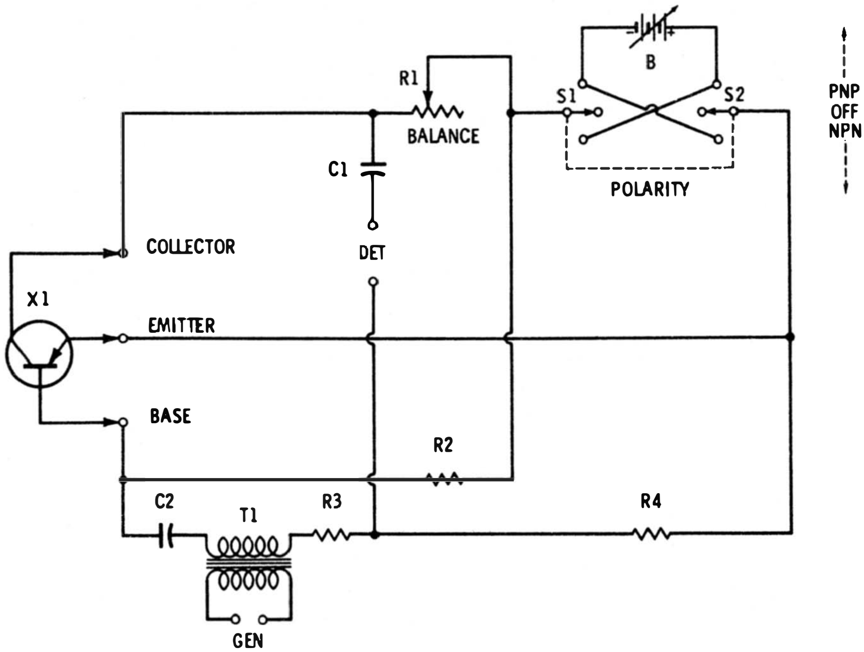


Fig. 8-8. Circuit for a-c amplification factor of transistor.

a-c component across rheostat R1; this component is opposite in phase to the a-c drop across resistor R4. When R1 is adjusted to make these out-of-phase voltages equal in amplitude, the output signal at the DET terminals is minimum; i.e., the bridge is balanced. At this null point, the current amplification is proportional to the ratio of the two resistances:

$$\beta = R4/R1 \qquad 8\text{-}6$$

The dial of rheostat R1 accordingly may be graduated to read directly in beta. It may be calibrated to read also in alpha, α (current amplification factor of the common-base configuration) on the basis of $\alpha = \beta/\beta + 1$.

Typical circuit constants are: B, 1½-12 volts; C1, 0.05 mfd; C2, 0.1 mfd; R1, 5K 1%; R2, 1K wirewound; R3, 100K; and R4, 250K.

CHAPTER 9

Nonbridge Null Devices

A small group of extremely useful null circuits are not bridges at all in the conventional sense. They utilize polarities and amplitudes in a-c circuits in such a way that their sum becomes zero (i.e., the circuits null) at one discrete value of circuit component values and/or frequency and voltage. Thus, in some applications they provide the same end result that bridges do.

Nonbridge null circuits offer many of the advantages peculiar to bridges—namely, accuracy of measurement, direct reference to a standard, and high resolution. Occasionally they are simpler than bridges. Sometimes they offer additional advantages, such as a common ground between input and output.

This chapter describes the principal members of the nonbridge group.

9.1 D-C POTENTIOMETER

The *d-c potentiometer* is a device for the accurate measurement of voltage by balancing an unknown d-c voltage against the precise voltage supplied by a standard cell. Since the cell voltage is on the order of 1.018 volt, most unknown voltages must be stepped down to this value by means of a voltage divider in order to obtain a balance.

Fig. 9-1 shows the basic circuit of a potentiometer. The basis of this arrangement is the voltage divider (potentiometer) consisting of a length of resistance wire, AB, provided with a sliding contact which moves over a calibrated scale. (Like the scale of the slide-wire bridges described in Chapters 2, 3, and 4, this scale may be graduated in inches, centimeters, ohms, or arbitrary linear units.) In some potentiometers, the slide wire is simply a straight length of wire stretched tautly between points A and B, as shown in Fig. 9-1, and situated over or alongside a scale; in more compact instruments, however, the wire is wound in a ring or spiral around a rotatable drum of insulating material. Meter M is a sensitive center-zero d-c galvanometer. The standard cell is represented by voltage E_1, and an external battery by voltage E_5. These two voltages are of the same polarity at point A. Resistor R1 limits the galvanometer current to a safe value and can be removed from the circuit as null is approached. Rheostat R5 allows the current from the external battery to be preadjusted.

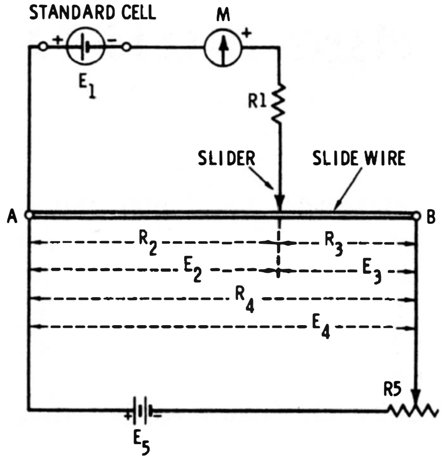


Fig. 9-1. Basic circuit of d-c potentiometer.

Moving the slider along the wire changes the values of R2 and R3 and therefore the ratio of R_2 to R_4, the full resistance of the slide wire. Corresponding voltages E_2 and E_3 are proportional to R_2 and R_3, respectively. When a voltage E_4 is impressed across the slide wire (i.e., between points A and B), the voltage (E_2) selected by the slider is proportional to the R_2/R_4 ratio: $E_2 = E_4(R_2/R_4)$.

The potentiometer is balanced first by setting the slider to a trial point along the slide wire, as shown in Fig. 9-1, and adjusting rheostat R5 for null, as indicated by meter M. At this point, voltage E_2 then is equal to the voltage of the standard cell (E_1), since the two voltages now cancel each other exactly. This completes the initial adjustment. The standard cell then is removed and the unknown voltage (E_x) connected in its place, the polarity being the same as that of the cell. Unless $E_x = E_1$, the circuit then will become unbalanced, and the slider must be moved to a new point to obtain a second null. This will change resistances R_2 and R_3, and at this new null setting, the unknown voltage is:

$$E_x = E_1(R_{2b}/R_{2a}) \qquad 9\text{-}1$$

where,

E_1 is the standard cell voltage,
R_{2a} is the initial value of R_2 at initial balance,
R_{2b} is the final value of R_2 at rebalance.

The slide wire may be calibrated to read directly in volts over a somewhat narrow range. Higher voltages may be stepped down through an external, precision voltage divider (*volt box*) to fall within the range of the potentiometer. The unknown voltage then becomes:

$$E_x = e(1/n) \qquad 9\text{-}2$$

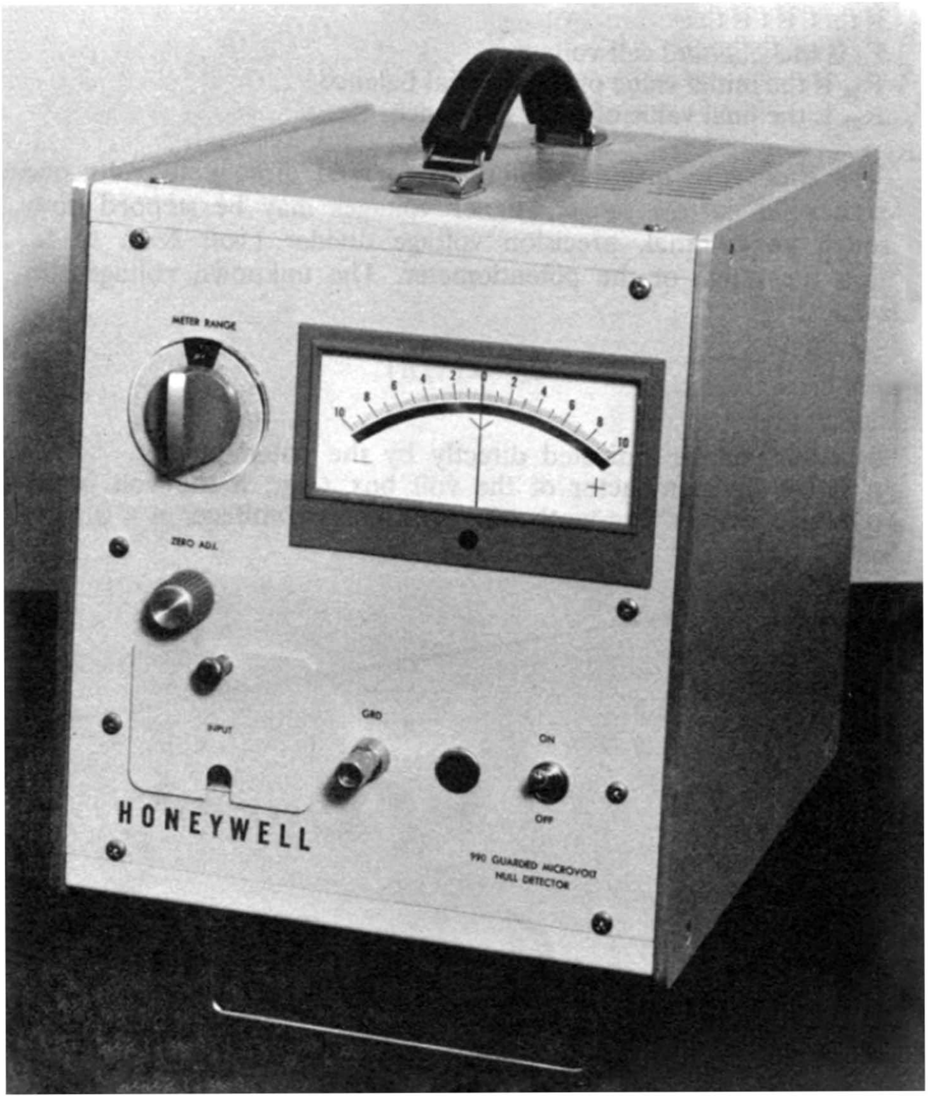
where,

e is the voltage indicated directly by the potentiometer,
n is the division factor of the volt box (e.g., if the volt box is set to deliver one-tenth of the unknown voltage, n = 0.1).

Courtesy Honeywell, Inc.

Fig. 9-2. Laboratory-type d-c potentiometer.

Fig. 9-2 shows a laboratory-type potentiometer (Honeywell Model 2784) which offers direct readings of voltage from zero to 11.1100 volts in four ranges: 0–0.0111100, 0–0.111100, 0–1.11100, and



Courtesy Honeywell, Inc.

Fig. 9-3. Sensitive d-c null detector used with potentiometer.

0–11.1100. The slide-wire dial (see center of panel) has 550 scale divisions for close resolution. Accuracy to ± (0.002% + 10 μv) is provided. Fig. 9-3 shows the a-c–operated, guarded microvolt null detector (Honeywell Model 3990) which is the companion unit for this potentiometer.

9.2 POTENTIOMETRIC VOLTMETER

The potentiometer principle described in the preceding section may be used in a d-c voltmeter which, using a precision variable resistor and the null method, can furnish higher accuracy than is obtainable with the meter method of measurement. The reason is that, at null, the instrument draws no current from the unknown-voltage source, and therefore offers zero loading (infinite resistance).

Fig. 9-4 shows the basic circuit of this type of instrument. Here, the unknown voltage (E_x) is applied to the potentiometer, and the resultant output voltage (E_o) is proportional to the position of the wiper along the resistance element (i.e., upon the resistance ratio $R_3/R1$). Meter M is a center-zero d-c galvanometer, and E_s is a standard voltage derived from a standard cell, mercury battery, or voltage-regulated supply.

The applied and standard voltages are applied to the meter in polarity, so that the deflection is zero when the potentiometer is adjusted to make $E_o = E_s$. At this null point, the unknown voltage may be determined from the resistance ratio and standard voltage:

$$E_x = E_s(R_3/R1) \qquad 9\text{-}3$$

In practice, the potentiometer is calibrated to read directly in volts. High resolution is obtained through the use of a series of decade resistors or a precision, multiturn potentiometer. Alternating voltages may be measured if a suitable rectifier is operated ahead of the INPUT terminals, and corrections made for the rectifier performance.

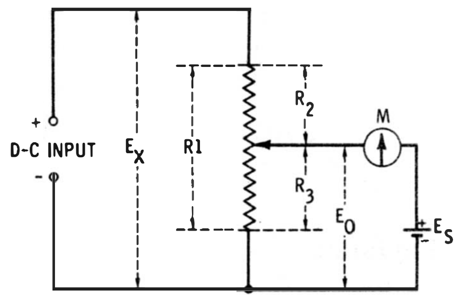


Fig. 9-4. Basic potentiometer voltmeter circuit.

9.3 FELICI MUTUAL-INDUCTANCE BALANCE

Fig. 9-5 shows a null circuit (*Felici mutual-inductance balance*) for checking an unknown mutual inductance (M_x) in terms of a known standard one (M_s). The two double-coil inductor units are so situated that no mutual inductance exists between them, and the secondary windings must be connected in phase opposition. Con-

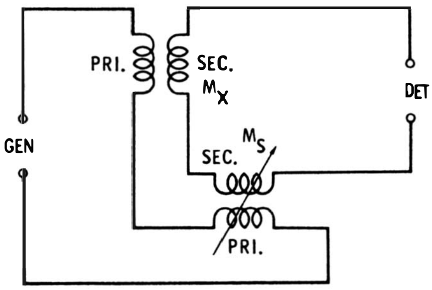


Fig. 9-5 Felici mutual-inductance balance.

tinuously variable mutual-inductance standards are available. One such device (General Radio Type 107-N) provides a range of 0-110 mh at ± 2.5% accuracy (related to full scale).

The circuit is balanced by adjustment of standard inductor M_s. At null:

$$M_x = M_s \qquad 9\text{-}4$$

Unlike the Carey-Foster circuit (Section 4.13, Chapter 4), the Felici arrangement is not a true bridge.

9.4 BRIDGED-T R-C NETWORK

The resistance-capacitance null circuit shown in Fig. 9-6 is termed a *bridged T* from the bridging effect of resistor R1 across the T network C1-C2-R2. With a given set of R and C values, the phase relations at one frequency are such that a signal applied to the input (GEN) terminals is canceled at the output (DET) terminals.

This circuit does not give a true null; i.e., the output voltage at balance does not fall completely to zero. But it does reach a low minimum in most cases, especially when R1 is very high with respect to R2. If C1 is kept equal to C2, then at balance:

$$f = \frac{\sqrt{1/(R1R2)}}{2\pi C} \qquad 9\text{-}5$$

where,

C equals C1 or C2 in farads,
f is in Hz,
R is in ohms,
π equals 3.1416.

The circuit is used sometimes to check a high resistance (R1) in terms of a known low resistance (R2), especially at high frequencies. In this application, C1 and C2 are varied simultaneously and have the same capacitance at all settings. At balance:

$$R1 = \frac{1}{R2(2\pi fC)^2} \qquad 9\text{-}6$$

The advantage of the bridged-T circuit is its relative simplicity and its provision of a common ground for the generator, network, and detector. This latter feature removes the need for a shielded

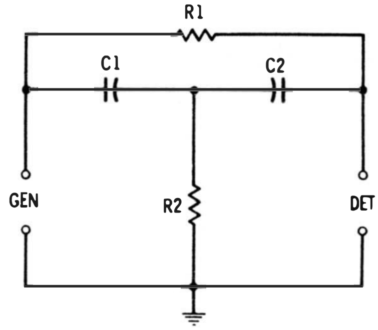


Fig. 9-6. Bridged-T r-c network.

transformer at either the input or output and reduces cross coupling and pickup. It also permits operation of the network at radio frequencies.

9.5 BRIDGED-T L-C NETWORK

Another bridged-T circuit is shown in Fig. 9-7. In this one, an inductor (L) bridges the T network (C1-C2-R2) in the same manner that a resistor bridges the T in the preceding section. In

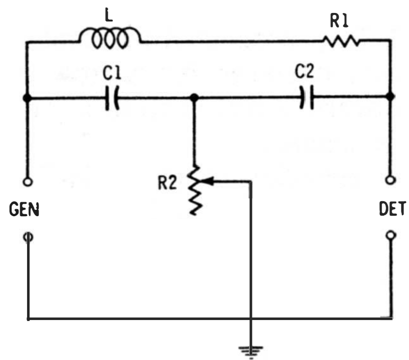


Fig. 9-7. Bridged-T l-c network.

the inductive leg, R1 is the equivalent series resistance of the inductor and has a low value in the high-Q inductors preferred for this circuit.

The capacitances C1 and C2 are made equal and may be varied simultaneously to tune the network. (Alternatively, the capacitors may be maintained equal and fixed, and the inductance varied.) To balance the network, the capacitance (or inductance) and resistance

R2 must be adjusted alternately. Unlike the R-C bridged-T network, this L-C version gives a complete null. At balance:

$$f = 1/(2\pi\sqrt{0.5LC}) \qquad 9\text{-}7$$

where,

C equals C1 or C2 in farads,
f is in Hz,
L is in hy,
π equals 3.1416.

$$R_2 = 1/(4\pi^2 f^2 R1) \qquad 9\text{-}8$$

$$C1 \text{ or } C2 = 1/(8\pi^2 f^2 L) \qquad 9\text{-}9$$

$$L = 1/(2\pi^2 f^2 C) \qquad 9\text{-}10$$

The bridge-T circuit is used for the measurement of inductance and capacitance at frequencies as high as the lower vhf spectrum, and it is used also as a band-rejection filter (notch filter). An advantage of the bridge-T network is its provision of a common ground for generator, network, and detector. This removes the need for a shielded transformer at either the input or the output, thus reducing cross coupling and pickup and permitting operation at high frequencies.

9.6 TWIN-T (PARALLEL-T) NETWORK

This resistance-capacitance null network (Fig. 9-8) takes its name from the fact that it consists of two T's (R1-R2-C3 and C1-C2-R3) connected in parallel. It gives a complete null, and its passband is reasonably narrow if high-Q capacitors are used and if the network is driven from a low impedance and loaded with a very high impedance.

The behavior of the twin-T network is similar to that of the Wien bridge (Section 7.3), but its selectivity is significantly better.

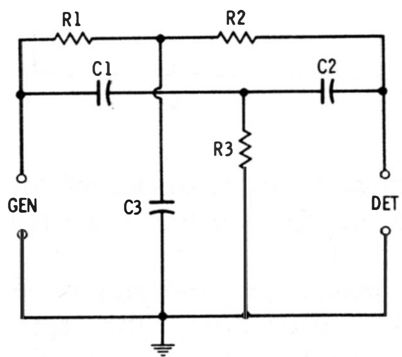


Fig. 9-8. Twin-T (parallel-T) network.

The twin-T, in common with other frequency-selective networks, may be balanced at only one frequency at a time, and this frequency is determined by the resistance and capacitance values, as in the Wien bridge. Conversely, at a fixed frequency, resistance and capacitance values may be determined with the aid of the twin T.

In the twin T, as in the Wien bridge, there are many combinations of resistance and capacitance which will result in balance at a given frequency. The balance equation is greatly simplified when the network is symmetrical—i.e., when C1 = C2 = ½C3, and R1 = R2 = 2R3. Under these conditions, at null:

$$f = 1/(2\pi R1C1) \qquad 9\text{-}11$$

where,

C1 is in farads,
f is in Hz,
R1 is in ohms,
π equals 3.1416.

In the Wien bridge, two capacitances or two resistances must be varied simultaneously to tune the circuit over a range of null frequencies. In the twin-T network, three capacitances (C1, C2, and C3) or three resistances (R1, R2, and R3) must be varied simultaneously. This sometimes makes the continuously variable twin T difficult to actualize, since three-gang rheostats or variable capacitors with a high degree of tracking are hard to make and therefore expensive.

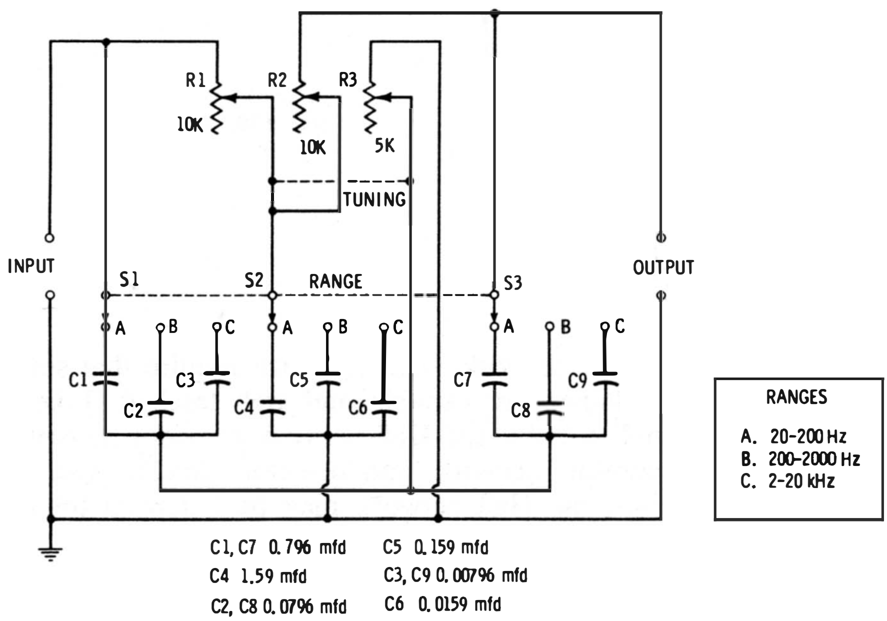


Fig. 9-9. Twin-T a-f meter.

The twin T, like the bridged-T network, provides a common ground for generator, network, and detector, thus removing the need for a shielded transformer at either the input or output and thereby reducing cross coupling and pickup, as well as permitting operation at radio frequencies.

The twin T is widely used as a band-rejection filter (notch filter) and is used occasionally to check capacitance or impedance (shunted across C3 or R3) at high frequencies. It is advantageous as a null-type frequency meter similar to those described in Section 7.3. Fig. 9-9 shows the circuit of a twin-T audio-frequency meter which covers the range from 20 to 20,000 Hz in three steps: 20–200 Hz, 200–2000 Hz, and 2–20 kHz. Tuning is accomplished with a three-gang wirewound rheostat having two 10,000-ohm sections (R1 and R2) and one 5000-ohm section (R3). Ranges are changed by the three-pole, three-position switch (S1-S2-S3), which selects closely matched capacitors (C1 to C9) in threes. A high-impedance detector must be used (e.g., oscilloscope, a-c vtvm, or magic-eye tube).

9.7 HALL NETWORK

The Hall network (Fig. 9-10) is a resistance-capacitance circuit in which continuously variable tuning is obtained with a single potentiometer, R2. In this respect, this circuit is superior to the

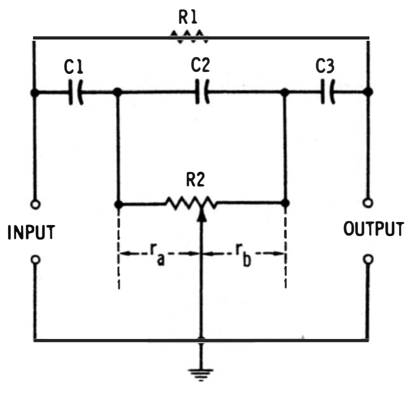


Fig. 9-10. Hall network.

Wien bridge and twin-T network, both of which require that several components be simultaneously variable and well tracked. Like the bridged-T and twin-T circuits, the Hall network provides a common ground for the generator, network, and detector (load). Also, like the preceding circuits, the Hall network may be balanced for only one frequency at a time.

Three capacitors (C1, C2, and C3) must be switched together to change tuning ranges, and the relationship C1 = C3 = 10C2 must be maintained in each range. At null:

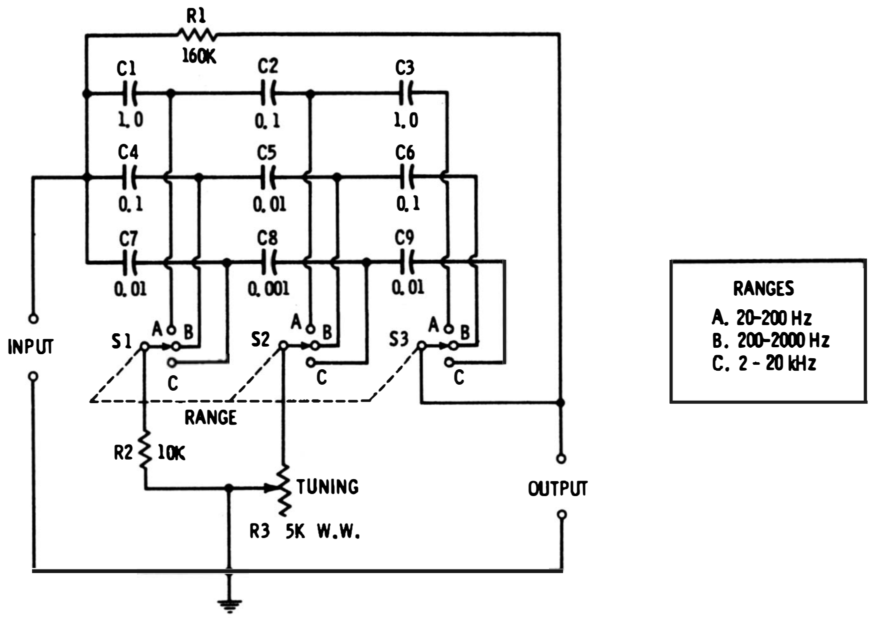


Fig. 9-11. Tunable Hall network.

$$f = \frac{\sqrt{\dfrac{r_a r_b}{C_1(2C_2 + C_1)}}}{2\pi} \qquad 9\text{-}12$$

where,

C is in farads,
f is in Hz,
r is in ohms,
π equals 3.1416.

The Hall network finds its principal use as a band-rejection filter (notch filter), its response curve being similar to that of the twin-T network. It may also be used, similar to the Wien bridge and parallel T, as an audio-frequency meter. Fig. 9-11 shows a continuously tunable Hall network which covers the range from 20 to 20,000 Hz in three steps: 20–200 Hz, 200–2000 Hz, and 2–20 kHz. The 5000-ohm wirewound rheostat, R3, is the tuning control. The ranges are changed by the three-pole, three-position switch, S1-S2-S3, which selects the capacitors in threes: C1-C2-C3, C4-C5-C6, and C7-C8-C9.

CHAPTER 10

Supplementary Uses of Bridge and Null Circuits

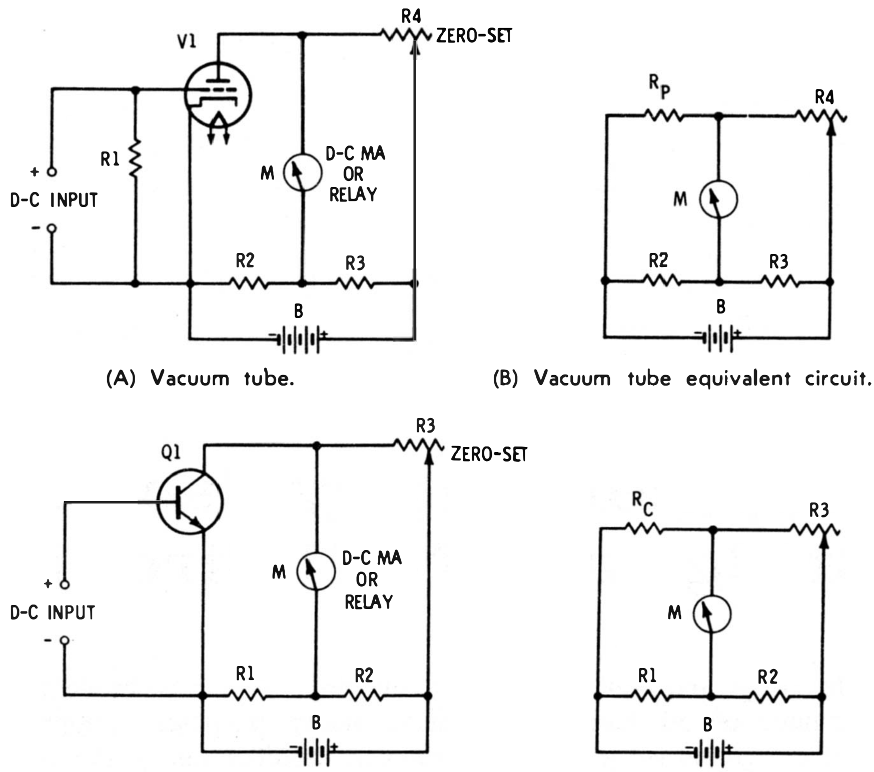
The bridge and nonbridge null circuits are widely used in electronic equipment of all kinds. They serve many purposes, ranging in broad categories from signal control and modification to the supply of power. So commonplace have some of these applications become that users take them for granted. In some instances, the contribution made by the null circuit has been noteworthy: economies and simplifications have been effected which might not otherwise have been so readily achieved.

Selected supplementary applications are given in this chapter. These have been chosen for their representative characteristics and complete practicality.

10.1 METER OR RELAY ZERO-SET

Fig. 10-1 shows the use of a bridge circuit to balance the static current out of a milliammeter, microammeter, or relay operated in the plate circuit of a vacuum tube or in the collector circuit of a transistor. This circuit is common in vacuum-tube voltmeters, where it is used to set the meter initially to zero.

In Fig. 10-1A, the four bridge arms are made up of resistors R2, R3, R4, and the d-c resistance (R_p) of the tube (see equivalent circuit, Fig. 10-1B). The bridge is balanced (meter or relay zeroed) when rheostat R4 is set to make $R4/R_p = R3/R2$. Battery B supplies both the plate voltage and bridge voltage. The resistance of R2 and R3 form a bleeder across the battery, and their values are chosen low enough to make the bleeder current at least five times the maximum



(A) Vacuum tube. (B) Vacuum tube equivalent circuit.

(C) Transistor. (D) Transistor equivalent circuit.

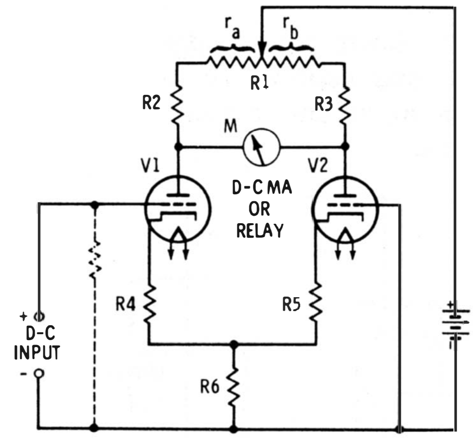
Fig. 10-1. Zero-setting circuit.

plate current of the tube. Also, the resistance of rheostat R4 must be at least 10 times the resistance of the meter or relay. These precautions preserve the sensitivity of the circuit to signal voltage.

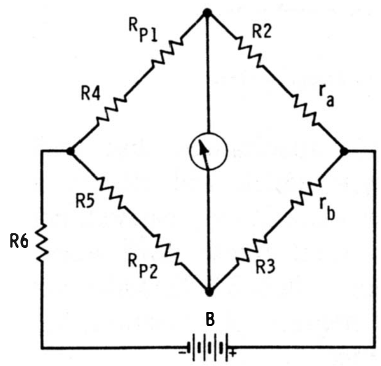
In Fig. 10-1C, the four bridge arms are composed of resistors R1, R2, R3, and the d-c collector resistance (R_c) of the transistor (see equivalent circuit, Fig. 10-1D). The bridge is balanced (meter or relay zeroed) when rheostat R3 is set to make $R3/R_c = R2/R1$. Battery B supplies both the collector voltage and bridge voltage (the battery and meter must be reversed for a PNP transistor). The resistance of R1 and R2 form a bleeder across the battery, and their values are chosen, as in the tube circuit, to make the bleeder current at least five times the maximum collector current of the transistor. Also, the resistance of rheostat R3 must be at least ten times the resistance of the meter or relay. These precautions preserve the sensitivity of the circuit to signal voltage.

10.2 BRIDGE-TYPE METER OR RELAY

For stability, modern d-c vacuum-tube voltmeters and vacuum-tube relays use a balanced circuit. As shown in Fig. 10-2A, this is a bridge circuit in which two arms of the bridge are supplied by the d-c plate resistance of matched tubes (R_{p1} and R_{p2} of tubes V1 and V2, respectively). The circuit is initially balanced (i.e., static plate current is zeroed out of the meter or relay) by adjustment of potentiometer R1. This changes the ratio of r_a to r_b (the resistances determined by the setting of the potentiometer contact) which are in separate arms of the bridge.



(A) Basic circuit.



(B) Equivalent circuit.

Fig. 10-2. Bridge-type vtvm (relay).

After initial balance, the bridge appears as shown in the equivalent circuit, Fig. 10-2B. Resistance R_{p1}, the d-c plate resistance of tube V1, changes in response to a d-c signal voltage applied to the grid of V1, and this unbalances the bridge, deflecting the meter or closing the relay. Tube V2 acts as an automatic compensator to keep the bridge balanced (meter zeroed) in spite of drifts of temperature and electrode voltage. These changes are experienced by both tubes and occur in the same direction and magnitude, since the tubes are closely matched. When, for example, plate resistance R_{p1} of tube V1 changes as the result of any of these causes (not because a signal voltage is applied to its grid), plate resistance R_{p2} of tube V2 changes in the same way and the bridge accordingly remains balanced.

A similar circuit may be arranged for a transistorized voltmeter or relay.

10.3 PHASE SHIFTER

Fig. 10-3 shows the circuit of a bridge-type phase shifter employing only resistance and capacitance. A maximum phase shift of 90 degrees between input and output voltages theoretically is possible in each R-C leg.

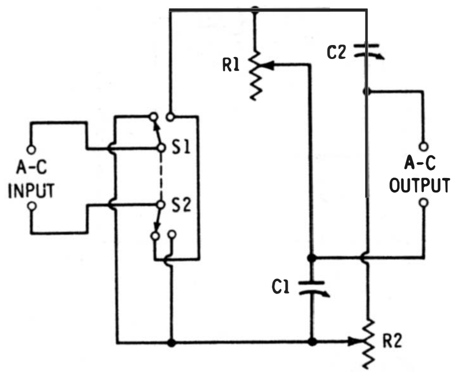


Fig. 10-3. Phase shifter.

Through judicious choice of capacitances and resistances, one may design a shifter of this type which will closely approach the 90 degree performance, so that variation of capacitance and resistance in both legs will give a total phase shift approaching 180 degrees. Then, throwing the dpdt switch S1-S2 will reverse the input and provide an additional 180 degrees of variation, to give a total phase-shift range of 0-360 degrees.

10.4 BALANCED MODULATOR

A four-diode semiconductor bridge makes a simple and convenient balanced modulator which gives good results if the diodes are closely matched for forward and reverse conduction and have better-than-average temperature stability.

Fig. 10-4 shows one type of bridge modulator circuit. For simplicity, transformers, capacitors, and compensating resistors have been omitted. Since the bridge is automatically in balance because the internal resistances of the diodes are all equal, the carrier voltage is nulled and cannot appear at the output terminals. However, diode nonlinearity results in modulation of the carrier by the modulating voltage, and the product of this modulation does appear at the output. This product consists of the upper sideband ($f_c + f_m$) and lower sideband ($f_c - f_m$).

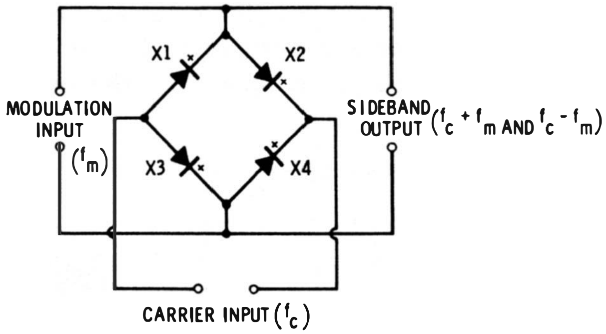


Fig. 10-4. Balanced modulator.

10.5 D-C—A-C CONVERTERS

Amplification of a d-c voltage is often obtained by (1) converting the direct current to a proportionate a-c voltage, (2) amplifying the latter with an a-c amplifier, and (3) rectifying the output of the amplifier. In this way, the stability of an a-c amplifier may be used to advantage, and this stability is usually superior to that of a d-c amplifier. A *d-c–to–a-c converter* changes the direct current to alternating current for presentation to the amplifier.

Fig. 10-5 shows two bridge-type converter circuits. In Fig. 10-5A, two closely matched semiconductor diodes (X1 and X2) are used as the variable elements. Potentiometer R1 is the balance control. The four arms of this bridge are r_a, r_b, X1 + R2, and X2 + R3. When the bridge has been initially balanced (with a-c power supplied, the D-C SIGNAL INPUT terminals open, and R1 adjusted for zero reading of an a-c null detector connected to the A-C SIGNAL OUTPUT terminals), no a-c voltage appears at either the D-C SIGNAL INPUT terminals or at the A-C SIGNAL OUTPUT terminals, if the diodes are perfectly matched and the circuit is adequately shielded. When a d-c signal of the polarity shown is applied, diode X1 conducts forward current; this produces a d-c voltage drop across R2 and unbalances the bridge, causing an a-c voltage proportional to the d-c signal voltage to appear at the output terminals. A d-c signal of opposite polarity causes a d-c voltage to appear across R3, resulting in an a-c output 180 degrees out of phase with the first.

In Fig. 10-5B, two varactors (X1 and X2) are the d-c–sensitive elements of the bridge. The four arms of this bridge are C1, C2, X1, and X2. The bridge is automatically in balance if the capacitors and varactors have been closely matched for capacitance; otherwise, one of the capacitors may be made variable for use as a balance control. At balance, no r-f voltage reaches either the D-C SIGNAL INPUT terminals or the R-F SIGNAL OUTPUT terminals. When a d-c signal is applied, it changes the varactor capacitance by an amount proportional to the d-c voltage and causes an r-f voltage proportional to the d-c voltage to appear at the output terminals.

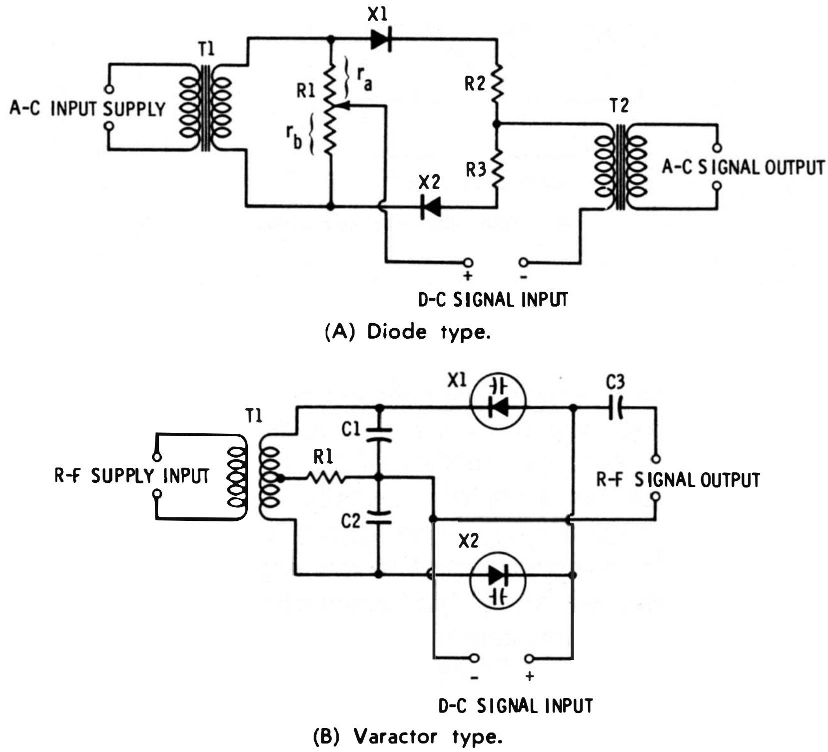
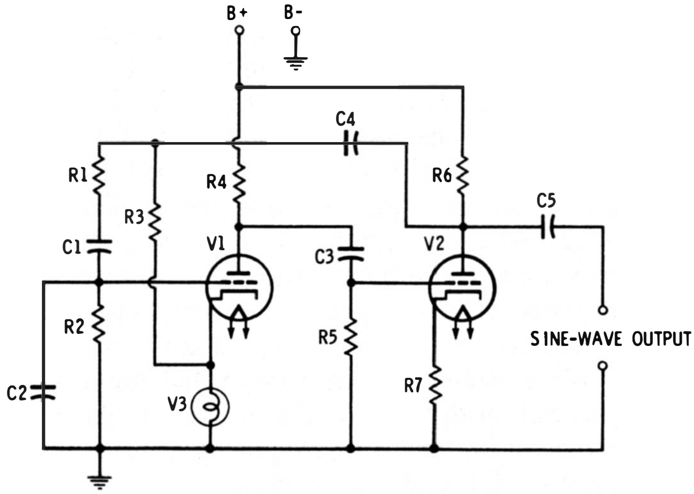


Fig. 10-5. D-c to a-c converters.

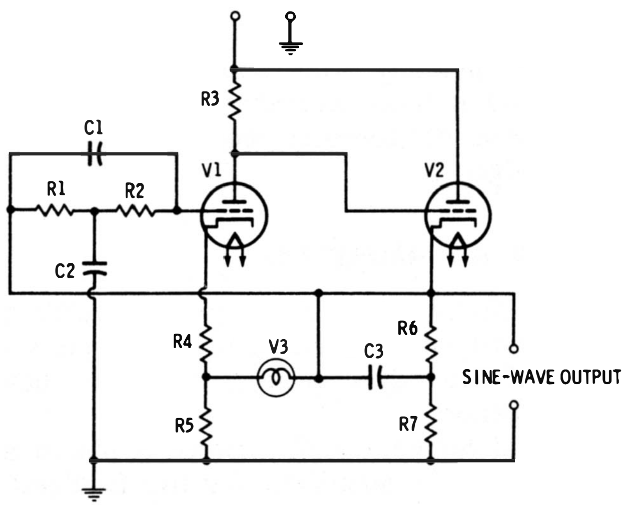
10.6 R-C—TUNED OSCILLATORS

Null circuits provide a convenient means for the resistance-capacitance tuning of audio and supersonic oscillators. Moreover, the selectivity of these circuits reduces distortion in the output of such oscillators to an extremely small amount. Virtually all types of null circuits have been used for this purpose.

Fig. 10-6 shows two circuits of R-C—tuned oscillators. Fig. 10-6A employs a Wien bridge. In this circuit, the four bridge arms are C1 + R1, C2 + R2, R3, and V3 (a tungsten-filament lamp). See Section 7.3, Chapter 7, for a description of the frequency-sensitive Wien bridge. This oscillator circuit is a two-stage amplifier in which positive feedback through the bridge produces oscillation, and nega-



(A) Wien bridge type.



(B) Bridged-T type.

Fig. 10-6. R-c tuned oscillators.

tive feedback through resistor R3 and a voltage-sensitive resistor (the filament of lamp V3) stabilizes the circuit and keeps the output constant over the tuning range. At the frequency to which the bridge is tuned, the phase is correct for oscillation and so is the amplitude of the positive feedback with respect to that of the negative feedback. Oscillation therefore occurs at that one frequency; at other frequencies, the negative feedback predominates and prevents oscillation, and this action results in the low distortion which is characteristic of the Wien-bridge oscillator. For tuning, C1 and C2 may be varied simultaneously, and R1 and R2 switched together to change frequency ranges. Or, R1 and R2 may be varied simultaneously, and C1 and C2 switched together to change ranges. The former method has proved to be the more practicable in most cases.

Fig. 10-6B employs a bridged-T network which is a variation of the type described in Section 9.4, Chapter 9. Its operation is similar to the Wien-bridge oscillator described previously.

Most commercial bridged-T oscillators are tuned by varying R1 and R2 simultaneously, and the frequency ranges are changed by switching capacitors C1 and C2 in pairs.

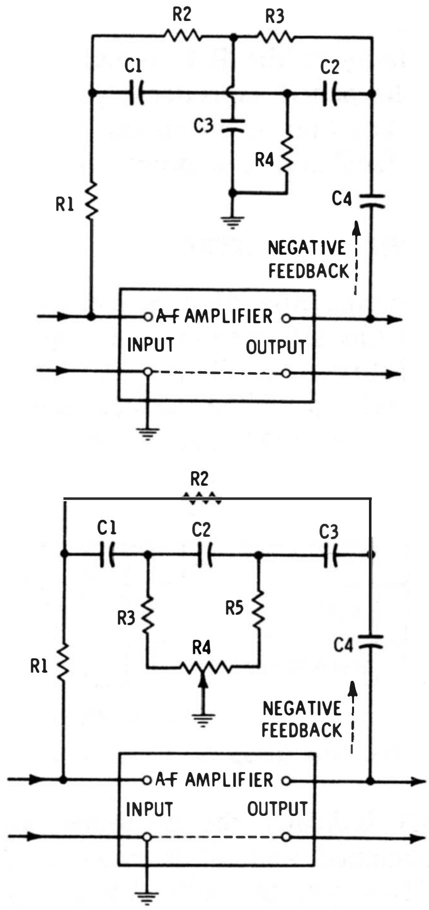
Other null networks used to tune oscillators are the twin T and the L-C–type bridged T. There is some objection to the twin T, except for single-frequency applications, because its requirement that three components be varied simultaneously for tuning and the other three switched together for range changing renders it somewhat inconvenient for variable-frequency use. The L-C–type bridged T makes possible a very low distortion oscillator, but the large values of inductance and capacitance it requires at audio frequencies preclude its use in smoothly variable instruments. The R-C network shown in Fig. 10-3 is used occasionally in bridge-type phase-shift oscillators, but it does not compete significantly with the Wien bridge and R-C–type bridged T.

10.7 R-C–TUNED A-F AMPLIFIERS

Resistance-capacitance null circuits make possible the tuning of tube- and transistor-type audio-frequency amplifiers without the use of inductors or of heterodyne circuits. Fig. 10-7 shows two tuned amplifier arrangements.

In this system of tuning, a null network is placed in a negative-feedback loop around the amplifier. Negative feedback then cancels the gain of the amplifier on all frequencies except the null frequency of the network. The amplifier thus is able to *transmit* that frequency only. It is in this way that the entire system (amplifier plus network) becomes a bandpass filter, although the network alone is a band-rejection filter.

In Fig. 10-7A, a twin-T network is used. (See Section 9.6, Chapter 9, for a description of the twin T.) In this network, C1 = C2 = ½C3, R1 = R2 = 2R3, and the pass frequency is $f = 1/(2\pi R_1 C_1)$. Continuously variable tuning may be obtained with



(A) Twin-T type.

(B) Hall type.

Fig. 10-7. R-c tuned a-f amplifiers.

a three-gang rheostat for the resistances, and the frequency range may be changed by switching capacitors C1, C2, and C3 together (see Section 9.6, Chapter 9). Resistor R1 decreases the loading effect of the network on the high-impedance input of the amplifier, and capacitor C4 provides d-c blocking.

In Fig. 10-7B, a Hall network is used as the tuning circuit. (See Section 9.7, Chapter 9, for a description of the Hall network.) The advantage of this null circuit is its use of a single potentiometer (R4) for tuning. The frequency range may be changed by switching capacitors C1, C2, and C3 together. Resistor R1 decreases the loading effect of the network on the input circuit of the amplifier, and capacitor C4 provides d-c blocking. Resistors R3 and R5 limit the extremes of potentiometer R4.

Sometimes a preamplifier stage is operated ahead of a tuned amplifier, and a cathode follower or emitter follower after the amplifier to isolate the null network from the effects of the signal source and the load.

The selectivity of the R-C–tuned amplifier is good. Nevertheless, it can be sharpened considerably by providing untuned positive feedback, in addition to the tuned negative feedback, and increasing this positive feedback to a point just short of oscillation.

10.8 DISTORTION METERS

Total harmonic distortion is measured with the setup shown in Fig. 10-8. In this arrangement, an amplifier (B) under test is driven by a low-distortion oscillator (A) at the desired test frequency and is terminated with the correct load, R_L. Shunting this load are an rms voltmeter (M1) and a null device (C) tunable to the test frequency. At the output of the null device is a second rms voltmeter, M2.

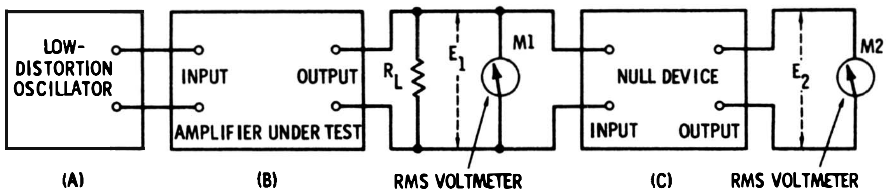


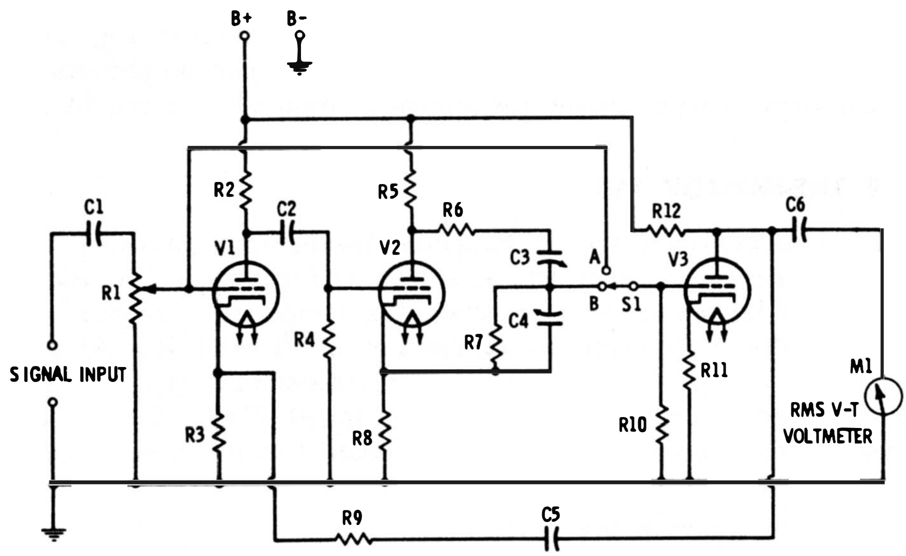
Fig. 10-8. Setup for harmonic distortion measurement.

Meter M1 indicates the combined voltage E_1 due to the test-signal fundamental and all harmonics created by distortion in the amplifier. However, the null device rejects only the fundamental voltage, and passes any remaining voltage (E_2) to meter M2. (If the amplifier output signal contained no harmonics, M2 would read zero.) Voltage E_2 therefore is proportional to the total harmonic content. For economy, a single meter is often used with a switch for connecting it to the input of the null device for reading E_1, and to the output for E_2. The total harmonic (distortion) percentage is determined from the two voltages:

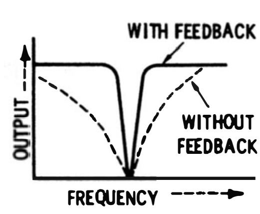
$$D\% = 100(E_2/E_1) \qquad 10\text{-}1$$

Suitable null circuits for this application are the resonance bridge (see Sections 7.1 and 7.2, Chapter 7) and the L-C–type bridged T (see Section 9.5, Chapter 9). The bridged T offers the advantage of a common ground between its input and output terminals. However, neither the resonance bridge nor the bridged T may be smoothly tuned over a wide range of frequencies. The resistance-capacitance bridge and nonbridge null circuits may not be used *directly* for accurate measurements by this method, because they attenuate harmonics, as well as the fundamental, and do this at a nonuniform rate.

For continuous tuning of a distorted meter, a Wien bridge may be used, provided it is included in an amplifier around which a large amount of negative feedback is provided. This gives the simplicity and compactness of the R-C circuit, while affording continuously variable tuning. Fig. 10-9A shows the basic circuit of a tube-type instrument of this type. Here, the tunable network (C3-C4-R6-R7) is inserted between two amplifier stages. Continuously variable tuning is provided by the dual variable capacitor (C3-C4), and ranges are changed by switching identical resistors R6 and R7 together. Negative feedback around the entire circuit is supplied by the C5-R9 path. Fig. 10-9B shows how feedback sharpens the response of the bridge. When switch S1 is in position A, the rms voltmeter (M1) reads the full input voltage E_1 (funda-



(A) Basic circuit.



(B) Response.

Fig. 10-9. R-c tuned variable-frequency distortion meter.

mental plus harmonics); when S1 is at B, M1 reads the total harmonic voltage, E_2. The harmonic percentage may be determined from these voltages by use of equation 10-1. Alternately, E_1 may be preset to a reference point on the meter scale by adjustment of gain-control potentiometer R1, whereupon the E2 deflection may be read directly as total distortion percentage.

Still another method of measuring distortion consists of a *wave analysis.* Here, the fundamental-frequency voltage (E_f) and each harmonic voltage (E_{h2}, E_{h3}, E_{h4} . . . E_{hn}) are separately measured. The strength of any one of the harmonics then may be compared with that of the fundamental. Or the total distortion may be calculated as the vector sum of the harmonic voltages compared with the fundamental voltage:

$$D\% = \frac{100\sqrt{E_{h2}^2 + E_{h3}^2 + E_{h4}^2 \ldots + E_{hn}^2}}{E_f} \qquad 10\text{-}2$$

A highly selective tuned bandpass filter is needed for tuning successively to the fundamental and each of the harmonics. Continuously variable, tuned amplifiers, such as those shown in Fig. 10-7, will be adequate in this filter function if harmonic amplitudes no smaller than 1/100th of the fundamental amplitude are required.

10.9 THERMISTOR BRIDGE

A *thermistor* is a special temperature-sensitive resistor. It can be made one arm of a bridge, as shown in Fig. 10-10 (the thermistor might be mounted in the end of a temperature probe).

The bridge is balanced by adjustment of rheostat R3. At null, the output voltage (E_o) is zero for the temperature (t_1) to which the thermistor is exposed during the adjustment: R1 = (R2R3)/R4. When the thermistor subsequently is exposed to a different tempera-

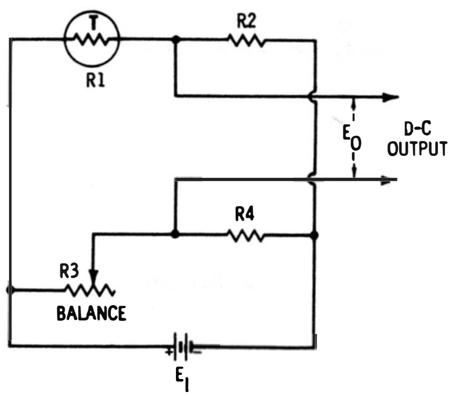


Fig. 10-10. Thermistor bridge.

ture (t_2), its resistance (R1) changes proportional to the new temperature, and the bridge unbalances. The unbalance output voltage (E_o) then is proportional to the temperature change ($t_2 - t_1$). The d-c output of the thermistor bridge may be amplified to increase the sensitivity of the circuit to small temperature changes. The amplifier can drive a recorder for a continuous record of temperature variation.

Another way to use this setup is to balance the bridge at each temperature of interest, determining the thermistor resistance from the relationship $R1 = (R2R3)/R4$, and then to refer the R1 value to a temperature chart based on the thermistor used, to obtain the temperature in degrees. Rheostat R3 may be calibrated in degrees.

10.10 STRAIN-GAUGE BRIDGE

Fig. 10-11 shows the circuit of a bridge in which a *strain gauge* is one arm. (A strain gauge is a stress-sensitive resistor.) The active strain gauge (R1) is cemented to the structure to be strained. An inactive gauge (i.e., one of the same type as R1 but not mounted on the structure) sometimes is used for the R3 arm.

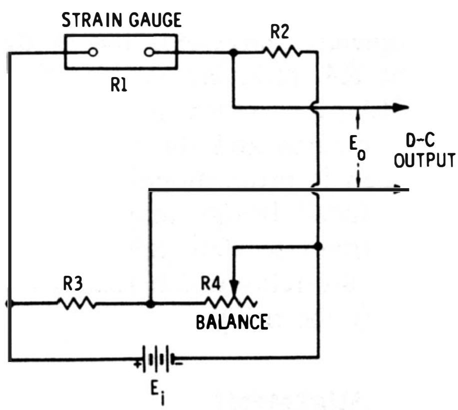


Fig. 10-11. Strain-gauge bridge.

The bridge is balanced under zero-stress conditions by adjustment of rheostat R4, and the output voltage (E_o) accordingly is zero: $R1 = (R2R3)/R4$. When the active strain gauge subsequently is strained, its resistance changes and the bridge unbalances. The unbalance voltage, E_o, then is proportional to the strain suffered by the structure to which the gauge is attached. The d-c output of the strain-gauge bridge may be amplified to increase the sensitivity of the circuit to small strain values. The amplifier can drive a suitable recorder, if a permanent record is desired.

10.11 PHOTOCELL BRIDGE

A bridge circuit is convenient for balancing the dark current out of the output of a photoconductive-type photocell. Fig. 10-12 shows the simple circuit. In this arrangement, the photocell internal resistance (R_1) forms one arm of the bridge.

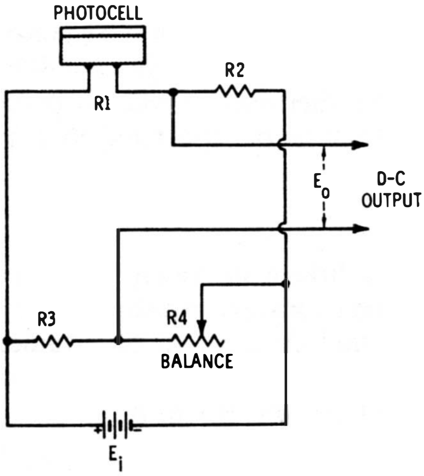


Fig. 10-12. Photocell bridge.

With the cell completely darkened, the bridge is balanced by adjustment of rheostat R4, reducing to zero the dark output voltage (E_o): $R4 = (R2R3)/R1$. When the cell subsequently is illuminated, its resistance decreases and the bridge unbalances. The unbalance voltage, E_o, then is proportional to the light intensity. The d-c output of the photocell bridge may be amplified to increase the sensitivity of the circuit to dim light. The amplifier can drive a recorder, or actuate a d-c relay which requires higher current than that delivered directly by the bridge.

10.12 PRESSURE MEASUREMENT

One type of pressure sensor uses a strain gauge cemented to an elastic metal disc. The disc is mounted in a suitable tube or chamber, where it is subjected to the pressure under study. This pressure distends the disc, stretching the gauge and causing a proportionate change in gauge resistance.

The resistance is measured with a d-c bridge in which the pressure sensor is one arm. This resistance is converted to pressure units by reference to a prior calibration of the strain gauge. The circuit is essentially the same as shown in Fig. 10-11. The dial of rheostat R4 may be calibrated to read directly in pressure units, thus eliminating calculations or references to tables.

10.13 GAS SNIFFER

An instrument for detecting flammable or explosive gases uses a d-c bridge circuit similar to Fig. 10-10. In the gas "sniffer," however, the thermistor in the upper left arm in Fig. 10-10 is replaced by a gas sensor. The latter consists essentially of an open filament which is exposed by blowing across it some of the air suspected of containing dangerous gas.

Resistance R2 is chosen, with respect to battery voltage E_i, so that the current through the sensor filament is of the correct value to keep the filament just warm. The bridge is balanced in the absence of gas by adjustment of rheostat R3. A d-c milliammeter or microammeter at the output terminals then reads zero. When a flammable gas subsequently is blown across the filament, some of the gas burns and the heat changes the resistance of the filament. This causes the bridge to unbalance, giving an output voltage which is proportional to the generated heat and being therefore indicative of the concentration of the gas. In some sniffers, a red area is provided on the meter scale to indicate the danger level.

10.14 TEMPERATURE MEASUREMENT WITH POTENTIOMETER

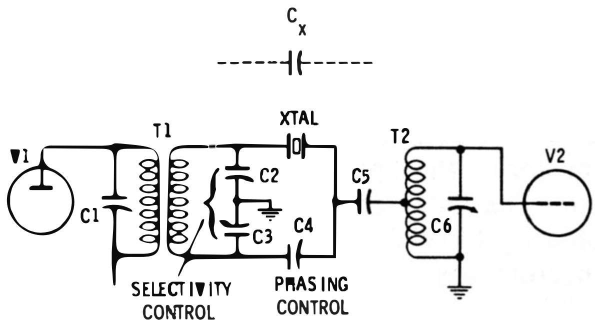
The d-c millivolt potentiometer (see Section 9.1, Chapter 9) is widely used for precise measurements of temperature, both in the laboratory and in the field.

In this application, a thermocouple is the temperature sensor, and its d-c output in millivolts (as indicated by the potentiometer at balance) is referred to a thermocouple table to determine the corresponding temperature in degrees. In portable potentiometric temperature instruments intended for rapid readings, the slide-wire dial is graduated in degrees for direct reading.

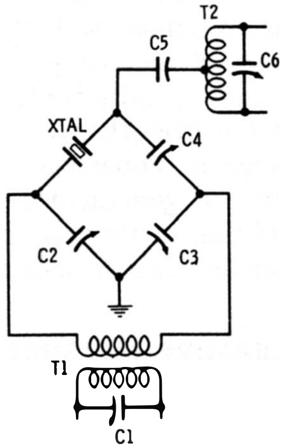
10.15 CRYSTAL FILTER

The crystal filter, which affords high selectivity in communications receivers, employs an intermediate-frequency bridge circuit. Fig. 10-13A shows a representative circuit, and Fig. 10-13B the equivalent bridge. In a receiver, the filter is inserted between two i-f stages or between the converter (first detector) and first i-f stage.

In Fig. 10-13A, the dual variable capacitor, C2-C3, is the tuning control, and variable capacitor C4 is the phasing control. The function of the latter is to balance out the capacitance of the crystal holder; if this operation were not performed, the holder capacitance would transmit the unwanted signals through the system. The phasing control allows adjustment of the filter to place a rejection notch in



(A) Circuit.



(B) Equivalent bridge.

Fig. 10-13. Crystal filter.

its response at the proper point for eliminating interfering signals or heterodynes.

At balance, governed by the setting of phase control C4, $C_x/C2 = C4/C3$, and $C4 = (C3C_x)/C2$, where C_x is the capacitance of the crystal holder. But since C2 = C3 at all settings of the dual capacitor, 2-C3, then $C4 = C_x$.

10.16 MAGNETIC-FLUX MEASUREMENT

A d-c bridge may be used for the measurement of magnetic flux if one of the bridge arms consists of a sensor whose resistance is proportional to the strength of an applied magnetic field. Such a sensor may be a semiconductor-type *magnetoresistor* or a small coil of bismuth wire. The circuit is similar to Fig. 10-12, with the photocell replaced by the flux sensor, and a d-c milliammeter or microammeter used as the output indicator.

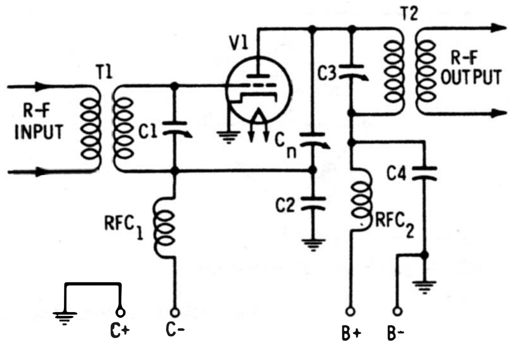
The bridge is balanced initially by adjustment of rheostat R4, with all flux absent. When the sensor subsequently is exposed to a magnetic field, its resistance changes, the bridge accordingly unbalances, and the unbalance current (indicated by the output meter) is proportional to the flux density. The meter scale or the dial of rheostat R4 therefore may be calibrated to read directly in flux units.

10.17 NEUTRALIZATION OF R-F AMPLIFIER

Feedback through the grid-plate capacitance of a tube can cause a radio-frequency amplifier to oscillate. This nuisance, which is pronounced in triodes, may be overcome by neutralization of the amplifier stage. Neutralization requires feeding back some of the r-f energy from the output in proper phase to counteract the effects of the tube capacitance. In Fig. 10-14A, C_n is the *neutralizing capacitor* used for this purpose. When its capacitance is adjusted to the correct point, oscillation ceases (or if the stage is temporarily disabled, all transmission through it due to the tube capacitance is canceled).

The neutralization circuit is a four-capacitor bridge, as shown in Fig. 10-14B. Here, the bridge arms are neutralization capacitor

(A) Circuit.



(B) Equivalent bridge.

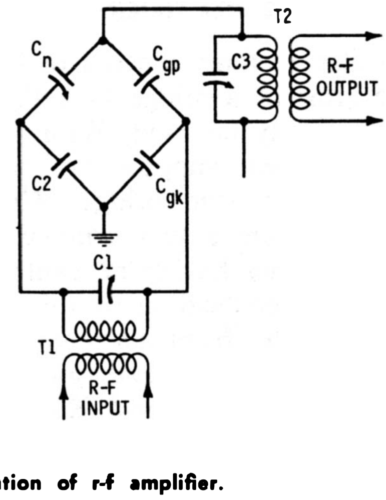


Fig. 10-14. Neutralization of r-f amplifier.

C_n, bypass capacitor C2, tube grid-plate capacitance C_{gp}, and tube grid-cathode capacitance C_{gk}. (Actually, the input capacitance C_i of the stage is in parallel with C_{gk} and must be added to the latter.) Adjustment of neutralizing capacitor C_n balances the bridge. At null:

$$C_n/C2 = C_{gp}/C_{gk} \qquad 10\text{-}3$$

R-f transmission through the grid-plate capacitance thus is reduced to zero.

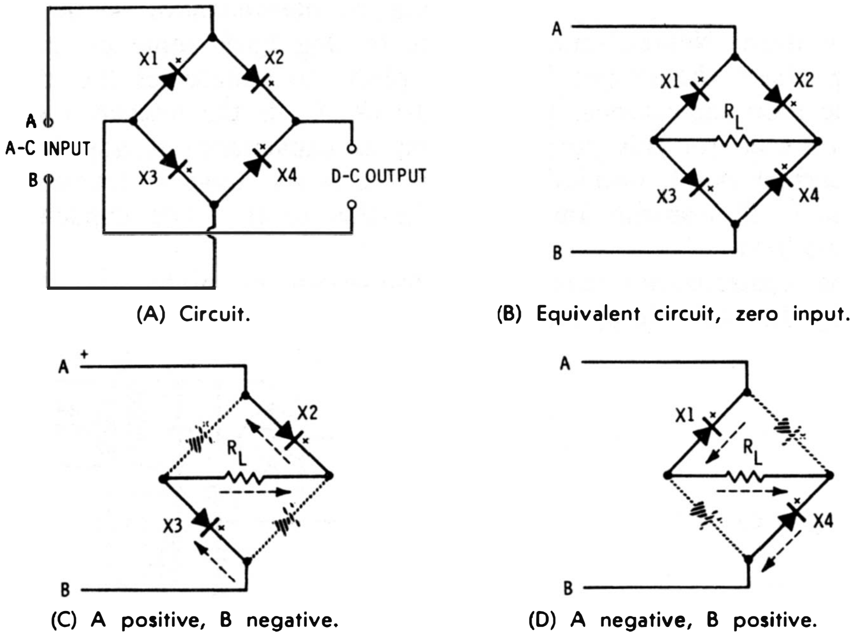


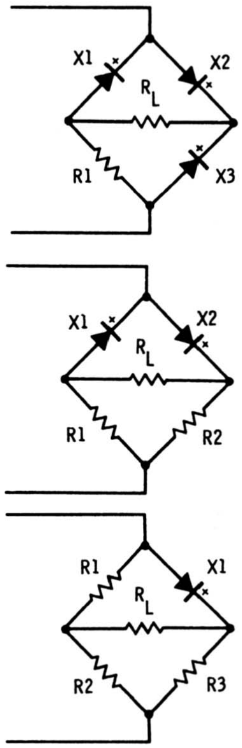
Fig. 10-15. Full-bridge rectifier.

10.18 BRIDGE RECTIFIER

A bridge rectifier consists of four diodes connected so that they are the arms of a bridge (see Fig. 10-15A). Either tubes or semiconductors may be used. With *filament*-type tubes, however, three separate filament supplies are necessary, since only two cathodes are common in the bridge. The bridge rectifier gives full-wave rectification from a two-terminal a-c supply, such as a power line or a transformer having no center tap.

Full-wave rectification by this circuit (see equivalent circuit, Fig. 10-15B) results from alternate conduction by pairs of the bridge arms. When, for example, a-c input terminal A is positive (Fig. 10-15C), diodes X2 and X3 conduct (being forward connected at this polarity) and diodes X1 and X4 block (being reverse connected). Electron current therefore flows from left to right through the load (R_L). When terminal B is positive on the succeeding half-cycle of a-c input, diodes X1 and X4 conduct and diodes X2 and X3 now block. Again, however, electron current flows through the load from left to right. Thus, the d-c output flows in the same direction during each half-cycle of a-c input—or, full-wave rectification results.

For simplicity and economy, and sometimes for emergency repair, diodes occasionally are replaced with resistors in some of the bridge arms. In Fig. 10-16A, for example, resistor R1 has been substituted



(A) Three-quarter bridge.

(B) Half bridge.

(C) Quarter bridge.

Fig. 10-16. Fractional-bridge rectifiers.

for one diode. This configuration thus is termed a *three-quarter bridge*. In the *half bridge* (Fig. 10-16B), resistors R1 and R2 replace two of the diodes. In the *quarter bridge* (Fig. 10-16C), only one diode (X1) remains. Each of these circuits may be analyzed in the same way as the full bridge in Figs. 10-15B, C, D. Neither of the fractional bridges is as efficient as the full bridge, the quarter bridge being the least efficient. A further disadvantage of the quarter bridge is the presence of R1 and R2 always across the a-c supply as a load.

The circuits shown here are for single-phase operation. Somewhat more complicated arrangements are needed for polyphase operation and can be found in electrical handbooks.

10.19 VOLTAGE-SENSITIVE BRIDGES

If a *varistor*—a voltage-sensitive resistor (e.g., thermistor, *Thyrite* resistor, filament-type lamp, or forward-connected diode)—is used as one of the arms of a d-c bridge, null will occur at only one input voltage for any combination of resistances in the other arms of the bridge. Thus, the output of the bridge will be zero for some discrete value of input voltage.

This phenomenon is utilized in voltage-sensing devices and in control equipment in which an output voltage is required to fall to zero at a prescribed value of input voltage. The voltage-sensitive bridge is used occasionally as a frequency multiplier, since in response to an applied a-c voltage the bridge output voltage goes through zero four times during each input half-cycle.

10.20 BRIDGE AS SIMPLE ANALOG COMPUTER

There are several ways of using null circuits as analog computors (electronic slide rules). Fig. 10-17 shows a Wheatstone bridge adapted to this purpose.

In this circuit, R1 equals 1 ohm and is a precision resistor. Each of the other arms contains a variable resistor. At null, R4 = (R2R3)/R1 = (R2R3/1 = R2R3. Since R4 thus equals the resistance of R2 times that of R3, multiplication can be performed by setting R2 equal to the multiplicand, setting R3 equal to the multiplier, balancing the bridge by adjustment of R4, and reading the product from the null setting of R4. Similarly, since R4/R3 = R2, division can be performed by setting R4 equal to the dividend, setting R3 equal to the divisor, balancing the bridge by adjustment of R2, and reading the quotient from the null setting of R2.

The degree of accuracy depends upon the closeness with which resistances R2, R3, and R4 can be read. Series-connected decade boxes in each of these arms will permit readings in 1-ohm steps. Next to be preferred are precision 10-turn potentiometers. Operation of the circuit is independent of the voltage of battery B. However, means must be provided for reducing this voltage when the

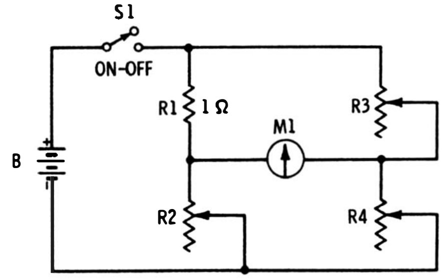


Fig. 10-17. Simple analog computer.

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BRIDGES AND OTHER NULL DEVICES

by Rufus P. Turner

The bridge-type circuit is found with increasing frequency in modern electronic apparatus and test equipment. This is due basically to certain unique properties which make it particularly useful, but which are not always clearly understood by the technician. Chief among these properties is the null principle. It, in general, makes accurately calibrated indicating instruments unnecessary in the bridge-type measuring circuit (except in a few cases where such instruments play an accessory role).

Bridges and Other Null Devices is intended to fill the "information gap" on bridge-type instruments by presenting, in concise form, the essential data on every significant bridge design being manufactured. It is not intended in any way to be a detailed operating manual for any instrument. However, any reader interested in bridge-type instruments will be much better equipped, after reading this book, to choose the correct bridge instruments for his needs before considering a purchase. In addition, the book will expand his knowledge of the capabilities of bridge instruments in general.

The first chapter discusses the basic theory of bridge circuits. Subsequent chapters cover different general types of bridges. The final chapter shows how bridge design is applied to specific instruments serving widely varied purposes.

ABOUT THE AUTHOR

Rufus P. Turner is the author of over 2500 articles and 20 books. He earned his B.A. degree from California State College at Los Angeles, and his M.A. and Ph.D. degrees from the University of Southern California. He is licensed as a registered professional engineer in California and Massachusetts and has had wide experience in the semiconductor field. Other SAMS books by Dr. Turner are *Diode Circuits Handbook*, *Technical Writer's & Editor's Stylebook*, and *abc's of Varactors*.

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