CHAPTER 15.2
SUBSTITUTION AND ANALOG MEASUREMENTS

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VOLTAGE SUBSTITUTION

The constant-current potentiometer, which is used for the precise measurement of unknown voltages below 1.5 V, is shown schematically in Fig. 15.2.1. For a constant current, the output voltage V_o is proportional to the resistance included between the sliding contacts. In this circuit all the current-carrying connections can be soldered, thereby minimizing contact-resistance errors. When the sliding contacts are adjusted to produce a null, V_o is equal to the unknown emf and no current flows in the sliding contacts. At null, no current is drawn from the unknown emf, and therefore the measured voltage is independent of the internal resistance of the source.

The circuit of a multirange commercial potentiometer is shown in Fig. 15.2.2. The instrument is standardized with the range switch in the highest range position as shown and switch *S* connected to the standard cell. The calibrated standard-cell dial is adjusted to correspond to the known voltage of the standard cell, and the standardizing resistance is adjusted to obtain a null on the galvanometer. This procedure establishes a constant current of 20 mA through the potentiometer. The unknown emf is connected to the emf terminals, and switch *S* is thrown to the emf position. The unknown emf can be read to at least five significant figures by adjusting the tap slider and the 11-turn 5.5-Ω potentiometer for a null on the galvanometer. The range switch reduces the potentiometer current to 2 or 0.2 mA for the 0.1 and the 0.01 ranges, respectively, thereby permitting lower voltages to be measured accurately. Since the range switch does not alter the battery current (22 mA), the instrument remains standardized on the lower ranges. When making measurements, the current should be checked using the standard cell to ensure that the current has not drifted from the standardized value. The Leeds and Northrup Model 7556-B six-dial potentiometer operates in a similar manner and provides an accuracy of ±(0.001 percent of reading + 0.1 μV).

The constant-resistance potentiometer of Fig. 15.2.3 uses a variable current through a fixed resistance to generate a voltage for obtaining a null with the unknown emf. The constant-resistance potentiometer is used primarily for measurements in the millivolt and microvolt range.

The microvolt potentiometer, or low-range potentiometer, is designed to minimize the effect of contact resistance and thermal emfs. Thermal shielding is used to minimize temperature differences. The galvanometer is connected to the circuit through a special Wenner thermo-free reversing key of copper and gold construction to eliminate thermal effects in the galvanometer circuit.

A typical microvolt potentiometer circuit consisting of two constant-current decades and a constant-resistance element is shown in Fig. 15.2.4. The constant-current decades use Diesselhorst rings, in which the constant current entering and leaving the ring divides two paths. The *IR* drop across the resistance in the isothermal shield increases in 10 equal increments as the dial switch is rotated. The switch contacts are in the constant-current

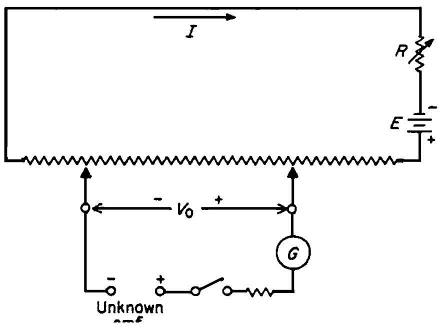


FIGURE 15.2.1 Constant-current potentiometer.

supply circuit, and therefore the effects of their *IR* drops and thermal emfs are minimized. A 100-division milliameter associated with the constant-resistance element provides nearly three additional decades of resolution. Readings to 10 nV are possible with this type of potentiometer.

The direct-current comparator potentiometer used for precise measurement of unknown voltages below 2.1 V is shown in Fig. 15.2.5. Feedback from the zero-flux detector winding is used to adjust the current supply for ampere-turn balance between the primary and secondary windings as in the dc-comparator ratio bridge of Chap. 15.4.

Standardization on the 1X range is obtained in a two-step procedure. First, the external standard cell and galvanometer are connected in series across resistor EH by the selector switch, and the constant current source is adjusted for zero galvanometer deflection. This transfers the standard cell voltage across resistor EH. Second, the seven measuring dials are set to the known standard voltage, and the selector switch is used to connect the galvanometer across the opposing voltages AD and EH. Trimmer turns n_s are used to obtain zero galvanometer deflection. This results in the standard cell voltage being generated across resistor AD with the dials set to the known standard cell value, and therefore calibration of the 1X range.

The unknown emf is measured on the 1X range by using the selector switch to connect the unknown emf and the voltage generated across resistor AD in series opposition across the galvanometer. The seven

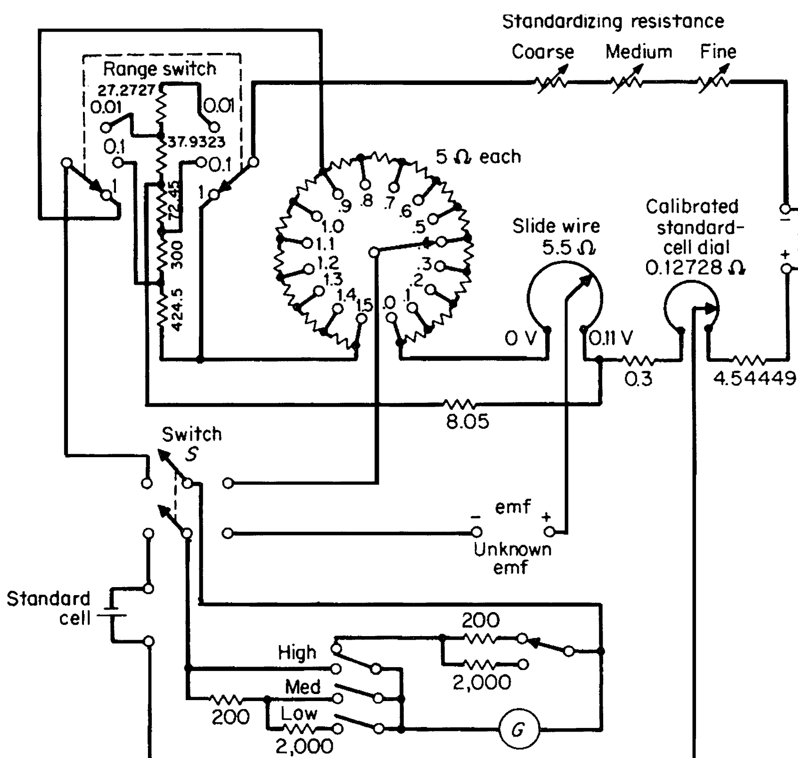


FIGURE 15.2.2 *K*2 potentiometer. (*Leeds and Northrup*)

measuring dials are adjusted to obtain zero deflection. Specified measurement accuracies are ±(0.5 ppm of reading + 0.05 μV) on the 1X range (2.111111 V full scale), ±(1 ppm of reading + 0.01 μV) on the 0.1X range, and ±(2 ppm of reading + 0.005 μV) on the 0.01X range.

DIVIDER CIRCUITS

The volt box (Fig. 15.2.6) is used to extend the voltage range of a potentiometer. The unknown voltage is connected between 0 and an appropriate terminal, for example, ×100. The potentiometer is connected between the 0 and P output terminals. When the potentiometer is balanced, it draws no current, and therefore the current drawn from the source flows through the resistor between terminals 0 and P. The unknown voltage is equal to the potentiometer reading multiplied by the selected tap multiplier. Unlike the potentiometer, the volt box does load the voltage source. Typical resistances range from about 200 to 1000 Ω/V. The higher resistance values minimize self-heating and do not load the source as heavily. Errors due to leakage currents, which could flow through the insulators supporting the resistors, are minimized by using a guard circuit (see Chap. 15.4).

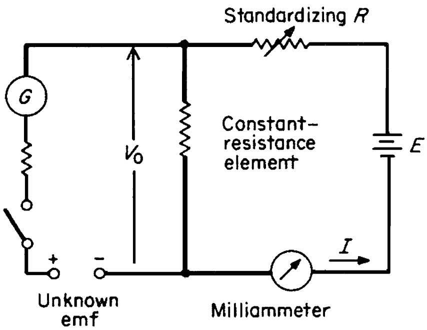


FIGURE 15.2.3 Constant-resistance potentiometer.

Decade voltage dividers provide a wide range of precisely defined, and very accurate voltage ratios. The Kelvin-Varley vernier decade circuit is shown in Fig. 15.2.7. The slide arms in the first three decades are arranged so that they always span two contacts. The shunting effect of the second gang resistance across the slide arms of the first decade is equal to $2R$, thereby giving a net resistance of R between

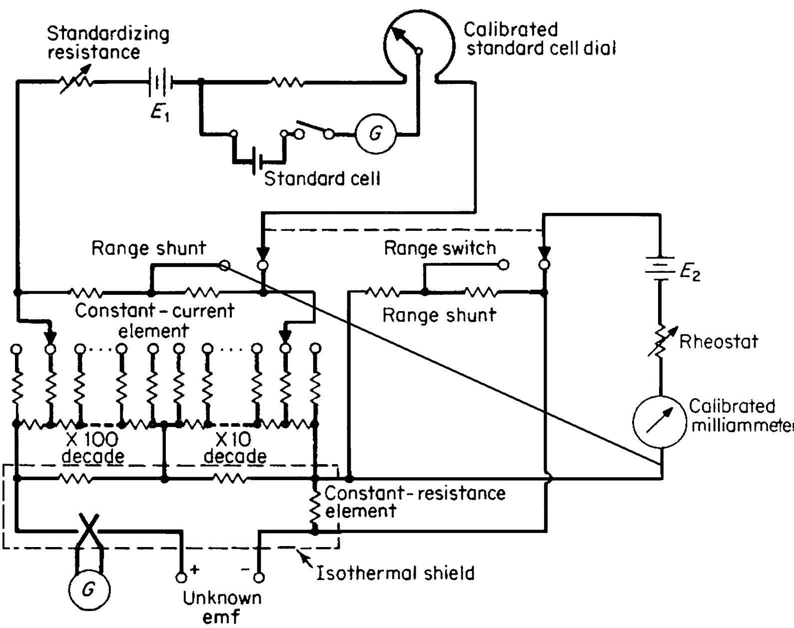


FIGURE 15.2.4 Microvolt potentiometer.

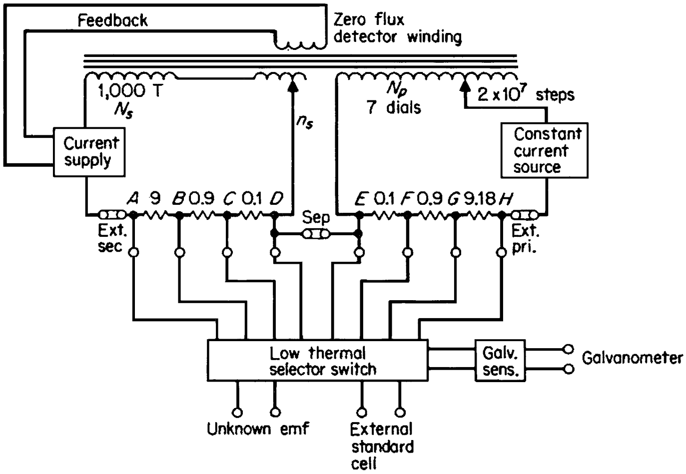


FIGURE 15.2.5 Direct-current comparator potentiometer.

the slide-arm contacts. With no current drawn from the output, the resistance loading on the input is equal to $10R$ and is independent of the slide-arm settings. In each of the first three decades, 11 resistors are used, while only 10 resistors are used in the final decade, which has a single sliding contact. Potentiometers with six decades have been constructed using the Kelvin-Varley circuit.

DECADE BOXES

Decade resistor boxes contain an assembly of resistances and switches, as shown in Fig. 15.2.8. The power rating of each resistance step is approximately constant; therefore, each decade has a different maximum current rating, which should not be exceeded. Boxes having four to seven decades are available with accuracies of 0.02 percent. Two typical seven-decade boxes provide resistance values from 0 to 1,111,111 Ω in 0.1-Ω steps and values from 0 to 11,111,110 Ω in 1-Ω steps. The accuracy at higher frequencies is affected by skin effect, series inductance, and shunt capacitance. The equivalent circuit of a resistance decade is shown in Fig. 15.2.9, where ΔL is the undesired incremental inductance added with each resistance step ΔR. Silver contacts are used to obtain a zero resistance R_o, as low as 1 mΩ/decade at dc. Zero inductance values L_o as low as 0.1 μH/decade are obtainable. The shunt capacitance for the configuration of Fig. 15.2.8 is a function of the highest decade in use, i.e., not set at zero. The shunt capacitance with the low terminal connected to the shield is typically 10 to 15 pF for the highest decade in use plus an equal value for each higher decade not in use.

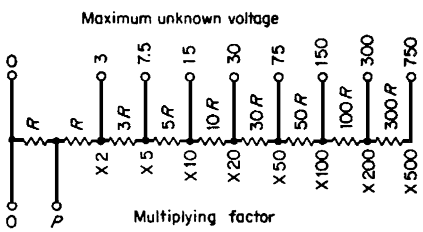


FIGURE 15.2.6 Volt-box circuit.

Some applications, e.g., the determination of small inductances at audio frequency and the determination of resistance at radio frequency by the substitution method, require that the equivalent series inductance of the resistance box remain constant, independent of the resistance setting. In the inductively compensated decade resistance box small copper-wound coils each having an inductance equal to the inductance of an individual resistance unit are selected by the decade switch so as to maintain a constant total inductance.

Decade capacitor units generally consist of four capacitances which are selectively connected in parallel by a four-gang 11-position switch (Fig. 15.2.10). The individual capacitors and their associated switch are shielded to

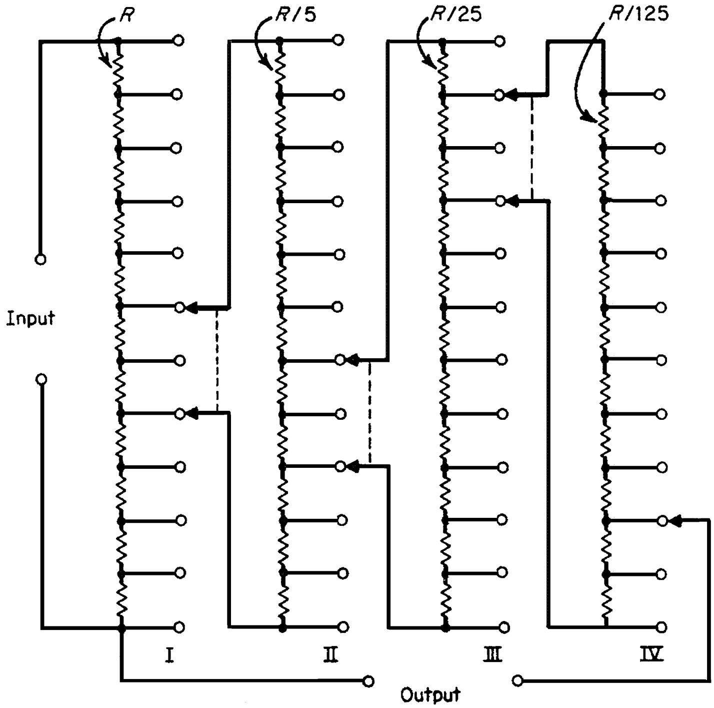


FIGURE 15.2.7 Decade voltage divider.

ensure that the selected steps add properly. *Decade capacitor boxes* are available with six-decade resolution, which provides a range of 0 to 1.11111 μF in increments of 1 pF and with an accuracy of 0.05 percent. Air capacitors are used in the 1- and 10-pF decades, and silver-mica capacitors in the higher ranges. Polystyrene capacitors are used in some less-precise decade capacitors.

Decade inductance units can be constructed using four series-connected inductances of relative units 1, 2, 3, 4 or 1, 2, 2, 5. A four-gang 11-position switch is used to short-circuit the undesired inductances. Care must be

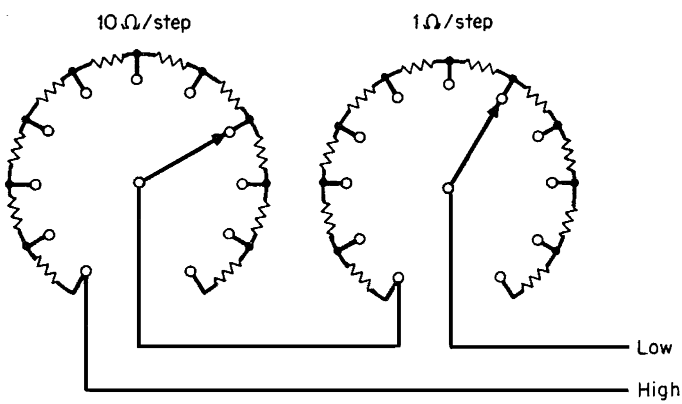


FIGURE 15.2.8 Decade resistance box.

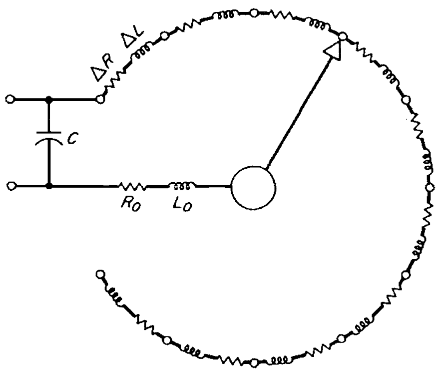


FIGURE 15.2.9 Equivalent circuit of a resistance decade. (*General Radio Co.*)

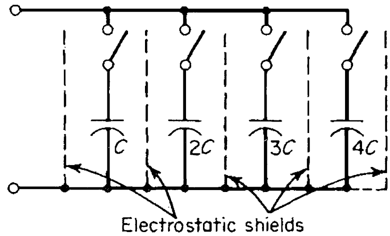


FIGURE 15.2.10 Capacitor decade.

taken to avoid mutual coupling between the inductances. *Decade inductance boxes* are available with individual decades ranging from 1 mH to 10 H total inductance. A commercial single-decade unit consists of an assembly of four inductors wound on molybdenum-Permalloy dust cores and a switch which enables consecutive values to be selected. Typical units have an accuracy of 1 percent at zero frequency. The effective series inductance of a typical decade unit increases with frequency. The inductance is also a function of the ac current and any dc bias current. The Q of the coils varies with frequency.

ANALOG MEASUREMENTS

Ohmmeter circuits provide a convenient means of obtaining an approximate measurement of resistance.

The basic series-type ohmmeter circuit of Fig. 15.2.11*a* consists of an emf source, series resistor R_1 and d'Arsonval milliammeter. Resistor R_2 is used to compensate for changes in battery emf and is adjusted to provide full-scale meter deflection (0-Ω indication) with terminals X_1 and X_2 short-circuited. No deflection (infinite resistance indication) is obtained with X_1 and X_2 open-circuited. When an unknown resistor R_x is connected

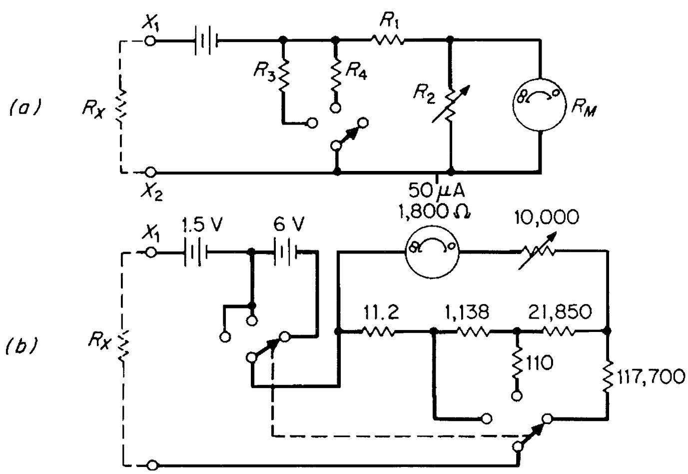


FIGURE 15.2.11 Series-type ohmmeters: (*a*) basic circuit; (*b*) commercial circuit (*Simpson Electric Co.*)

across the terminals, the meter deflection varies inversely with the unknown resistance. With the range switch in the position shown, half-scale deflection is obtained when the external resistance is equal to $R_1 + R_2R_M/(R_2 + R_M)$. A multirange meter can be obtained using current-shunting resistors R_3 and R_4. A typical commercial ohmmeter circuit (Fig. 15.2.11*b*) having midscale readings of 12 Ω, 1200 Ω, and 120 kΩ uses an Ayrton shunt for range selection and a higher battery voltage for the highest resistance range.

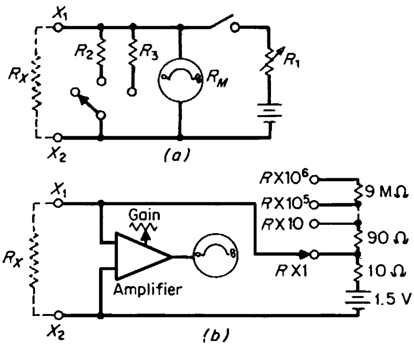


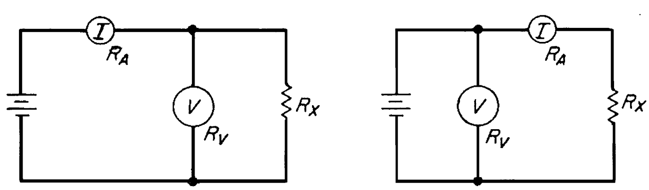
FIGURE 15.2.12 Shunt-type ohmmeter. (*Triplett Electrical Instrument Co.*)

In the shunt-type ohmmeter the unknown resistor R_x is connected across the d'Arsonval milliammeter, as shown in Fig. 15.2.12*a*. The variable resistance R_1 is adjusted for full-scale deflection (infinite-resistance indication) with terminals X_1 and X_2 open-circuited. The ohm scale, with 0 Ω corresponding to zero deflection, is the reverse of the series-type ohmmeter scale. The resistance range can be lowered by switching a shunt resistor across the meter. With the range switch selecting shunt resistor R_2, half-scale deflection occurs when R_x is equal to the parallel combination of R_1, R_M, and R_2. The shunt-type ohmmeter is therefore most suited to low-resistance measurements.

The use of a high-input impedance amplifier between the circuit and the d'Arsonval meter permits the shunt-type ohmmeter to be used for high- as well as low-resistance measurements. A commercial ohmmeter (Fig. 15.2.12*b*) uses a field-effect-amplifier input stage which draws negligible current. The amplifier gain is adjusted to provide full-scale deflection with terminals X_1 and X_2 open-circuited. Half-scale deflection occurs when R_x is equal to the total selected tap resistance.

Voltage-drop (or fall-of-potential) methods for determining resistance involve measuring the current flowing through the resistor with an ammeter, measuring the voltage drop across the resistor with a voltmeter, and calculating the resistance using Ohm's law. The circuit of Fig. 15.2.13*a* should be used for low-resistance measurements since the current drawn by the voltmeter V/R_v will be small with respect to the total current I. The circuit of Fig. 15.2.13*b* should be used for high-resistance measurements since the resistance of the ammeter R_A will be small with respect to the unknown resistance R_x. An accuracy of 1 percent or better can be obtained using 0.5 percent accurate instruments if the voltage source and instrument ranges are selected to provide readings near full scale.

Resonance methods can be used to measure the inductance, capacitance, and Q factor of components at radio frequencies. In Fig. 15.2.14, resistors R_1 and R_2 couple the oscillator voltage e to a series-connected known capacitance and an unknown inductance represented by effective inductance L' and effective series resistance r'. Resistor R_2 is chosen to be small with respect to resistance r', thereby minimizing the effect of source resistance of the injected voltage.



$$R_X = \frac{V}{I - V/R_V}$$

(*a*)

$$R_X = \frac{V}{I} - R_A$$

(*b*)

FIGURE 15.2.13 Fall-of-potential method: (*a*) for low resistances; (*b*) for high resistances.

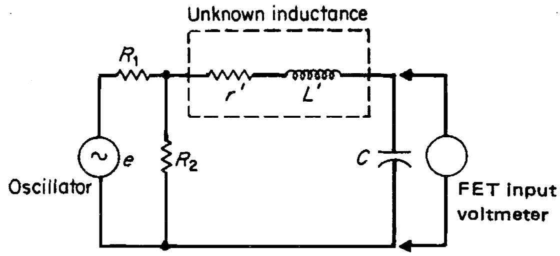


FIGURE 15.2.14 Inductance measurement.

A circuit containing reactive components is in resonance when the supply current is in phase with the applied voltage. The series circuit of Fig. 15.2.14 is in resonance when the inductive reactance $X_{L'}$ is equal to the capacitive X_C, which occurs when

$$\omega^2 = \omega_0^2 = 1/L'C$$

where $X_{L'} = \omega L'$, $X_C = 1/\omega C$, $\omega = 2\pi f$, $\omega_0 = 2\pi f_0$, f_0 = resonant frequency (Hz), L' = effective inductance (H), and C = capacitance (F).

If L', r', C, and e are constant and the oscillator frequency ω is adjusted until the voltage read across C by the FET input voltmeter is maximum, the frequency ω will be slightly less than the resonant frequency ω_0:

$$\omega^2 = \frac{1}{L'C} - \frac{(r' + R_s)^2}{2L'^2} = \omega_0^2\left(1 - \frac{1}{2Q^{*2}}\right)$$

where $R_s = R_1R_2/(R_1 + R_2)$ and $Q^* = \omega_0 L'/(r' + R_s)$. If $Q^* \geq 10$, ω and ω_0 will differ by less than 0.3 percent. The ratio m of the voltage across C to the voltage across R_2 can be measured while operating at ω. If $R_s \ll r'$, the value of the effective Q of the unknown inductance is related to m by

$$m = \frac{2Q'^2}{\sqrt{4Q'^2 - 1}} \quad Q' = \sqrt{\frac{m}{2(m - \sqrt{m_2 - 1})}}$$

where $Q' = \omega_0 L'/r'$ and $m = V_C/V_{R2}$ measured at ω. The values of m and Q' are nearly equal for high values of Q'; the difference is less than 0.4 percent for $Q' > 10$. If R_s is not small with respect to r', its value affects the determination of Q' only indirectly through its effect on ω. If $R_s = r$ and $Q' \geq 10$, the determination of Q' by the above equation is in error by less than 1 percent.

If ω, L', r', and e are constant and the capacitance C is adjusted until the voltage across it is maximum, the capacitance value C will be slightly less than the capacitance value C_R needed for resonance at the frequency ω:

$$C = \frac{1}{\omega^2 L'}\left(\frac{1}{1 + 1/Q^{*2}}\right) = C_R\frac{1}{1 + 1/Q^{*2}}$$

where $\omega = \omega_0$, $Q^* = \omega_0 L'/(r' + R_s)$, and $R_s = R_1R_2/(R_1 + R_2)$. For $Q^* \geq 10$, C differs from C_R by less than 1 percent. If $R_s \ll r'$, the value of the effective Q of the unknown inductance is related to m by

$$m = \sqrt{Q'^2 + 1} \quad Q' = \sqrt{m^2 - 1}$$

where $Q' = \omega L'/r'$, $m = V_C/V_{R2}$, $\omega = \omega_0 = 2\pi f_0$, and f_0 is the resonant frequency in hertz for L and C_R.

The circuit of Fig. 15.3.14 can be used to find the value of an unknown capacitance if a stable inductance of known value is available. Similar circuits are used in Q meters (see Chap. 15.4). Discussions of series resonance, parallel resonance, and Q factor are given in Refs. 3, 12, and 13.

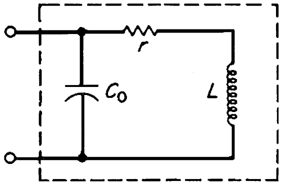


FIGURE 15.2.15 Equivalent circuit of inductance.

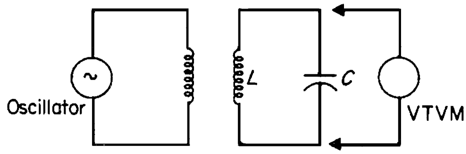


FIGURE 15.2.16 Loosely coupled tuned circuit.

Lead lengths should be kept short when making measurements at high frequencies, and the capacitance of the voltmeter must be added to capacitor C if it is not negligible.

The self-capacitance C_0 of an inductance is generally not negligible and can cause discrepancies in measurements. An inductance can be represented by the equivalent circuit of Fig. 15.2.15, where C_0 is the self-capacitance, L is the true inductance, and r is the true series resistance. The self-capacitance causes the effective inductance L' and the effective series resistance r' of the equivalent unknown inductance as measured in the circuit of Fig. 15.2.14 to differ somewhat from L and r, that is

$$r' = r\left(\frac{C + C_0}{C}\right)^2 \quad \text{and} \quad L' = L\frac{C + C_0}{C}$$

The value of C_0 can be measured with the aid of a Q meter (Chap. 15.4) having a calibrated variable capacitance.

The bandwidth of a tuned circuit can be determined by measuring the frequencies f_1 and f_2 at which the capacitor voltage in Fig. 15.2.16 is 0.707 times the maximum capacitor voltage. The oscillator, which maintains a constant voltage, is loosely coupled to the tuned circuit. The bandwidth Δf is equal to the difference $f_2 - f_1$ and is a function of the resonant frequency f_0 and the Q of the tuned circuit: $\Delta f = f_0/Q$.

The circuit Q of a tuned circuit for sinusoidal excitation is defined as 2π(maximum stored energy)/(energy dissipated per cycle). Circuit Q can be measured directly using the circuit of Fig.15.2.16. In this circuit the self-capacitance C_0 appears directly in parallel with the tuning capacitor C. The bandwidth is determined as explained above, and the Q is equal to $f_0/(f_2 - f_1)$. If the circuit of Fig. 15.2.14 is used, the self-capacitance no longer appears in parallel with the tuning capacitor C and the effective Q is measured. The real Q can be calculated if the self-capacitance C_0 is known; $Q = Q'(C + C_0)/C$, where Q' is the effective Q.

DIGITAL INSTRUMENTS

Digital multimeters provide a convenient means of making rapid and accurate measurements of voltage, current, and resistance. These instruments rely on the accuracy of A/D converters (see Chap. 15.3) and built-in precision voltage references. Many contain microprocessor-based intelligence and can be incorporated in automated-measurement and data-collection systems by means of the IEEE-488 bus (see Chap. 15.3). A typical block diagram is shown in Fig. 15.2.17. The performance is largely dependent on the A/D converter. Low-cost handheld units use a single-chip 3$^1/_2$-digit A/D converter, while high-accuracy 5$^1/_2$- and 6$^1/_2$-digit meters use microprocessor-controlled discrete A/D converters. Typical performance specifications are given in Table 15.2.1. Unlike analog meters that have accuracies specified in terms of full scale, accuracy figures for digital meters are given in percent of the reading plus the number of counts of the least significant digit. Therefore, a 5.00 V reading on a 3$^1/_2$-digit meter with a specified accuracy of ±(0.1 percent +1 digit) would be accurate to ±0.3 percent or ±0.015 V.

The input resistance R_{in} of the meter can introduce an error in the measurement of an unknown voltage V_x by loading the unknown as shown in Fig. 15.2.18. The voltage V_m indicated by the meter is equal to $V_x R_{in}/(R_{in} + R_x)$.

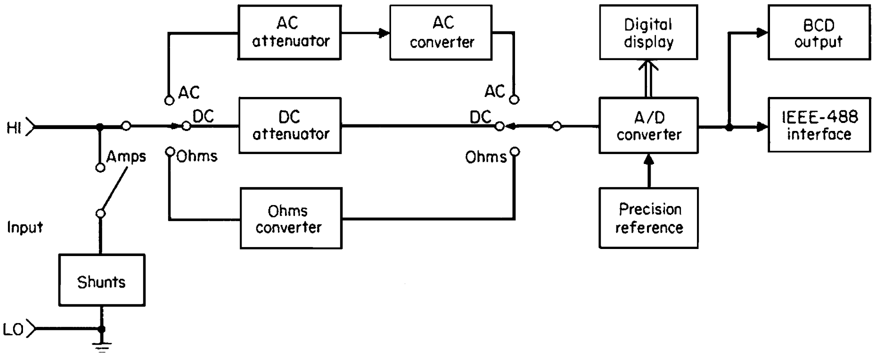


FIGURE 15.2.17 Digital multimeter. (*Keithley Instruments*)

TABLE 15.2.1 Typical Digital Multimeter Performance

	A/D digits		
	3½	4½	6½
Counts	1999	19,999	1,999,999
DC voltage			
Max. range	1000 V	1000 V	1000 V
Min. range	200 mV	200 mV	200 mV
Sensitivity	100 μV	10 μV	100 nV
Basic accuracy	± (0.1% + 1)	± (0.04% + 1)	±(0.0007% + 10)
Input resistance	10 MΩ	10 MΩ	≤ 20 V: 1 GΩ >20 V: 10 MΩ
CMRR (1 kΩ unbalance)	>100 dB	>120 dB	>120 dB
AC voltage			
Max. range	750 V	750 V	750 V
Min. range	200 mV	200 mV	2 V
Sensitivity	100 μV	10 μV	1 μV
Basic accuracy (<1 kHz)	±(1% + 2)	±(0.5% + 10)	±(0.25% + 1000)
Input resistance/cap	10 MΩ/100 pF	10 MΩ/100 pF	2 MΩ/50 pF
CMRR (1 kΩ unbalance)	>60 dB	>60 dB	>60 dB
Ohms			
Max. range	200 Ω	200 Ω	200 Ω
Min. range	20 mΩ	200 mΩ	200 mΩ
Sensitivity	100 mΩ	10 mΩ	100 μΩ
Basic accuracy	±(0.2% + 1)	±(0.05% + 2)	±(0.01% + 2)
DC current			
Max. range	2A	2A	2 A
Min. range	2 mA	200 μA	200 μA
Sensitivity	1 μA	10 nA	1 nA
Basic accuracy	±(0.75% + 1)	±(0.2% + 2)	±(0.09% + 10)
AC current			
Max. range	2 A	2 A	2 A
Min. range	2 mA	200 μA	200 μA
Sensitivity	1 μA	10 nA	1 nA
Basic accuracy	±(1.5% + 2)	±(1% + 20)	±(0.6% + 300)

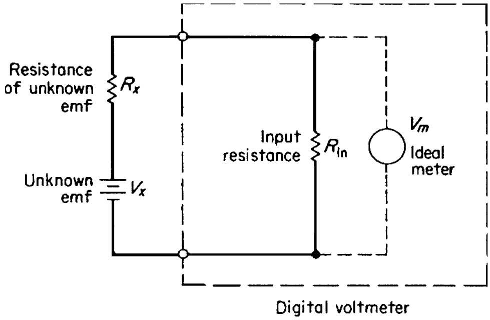


FIGURE 15.2.18 Effect of loading.

For example, if a meter having a 10-MΩ input resistance is used to measure an unknown emf having an equivalent resistance of 100 KΩ, the meter will read 99 percent of the correct value.

Resistance measurements rely on the accuracy of an internal current reference, shown as I_r in Fig.15.2.19. For an ideal meter with infinite input resistance R_{in}, the unknown resistance R_x is equal to V_m/I_r and the meter is calibrated to read the resistance value directly. For finite values of R_{in}, the indicated resistance value of R_x is equal to the parallel resistance of R_x and R_{in}, which can introduce a significant error in the measurement of high-value unknown resistances. The resistance of the leads connecting the unknown resistance to the meter of Fig. 15.2.19 can introduce an error in the measurement of low-value resistances. This problem is overcome in high-precision meters by using separate terminals and leads to connect the precision current reference to the unknown resistance. In this case the voltage measurement leads carry only the current drawn by the input resistance R_{in}.

Digital electrometers are highly refined dc multimeters which provide exceptionally high input resistance and sensitivity. The block diagram of a typical microcomputer-controlled digital electrometer is shown in Fig. 15.2.20. Performance depends on the high-input-resistance, low-offset-current JFET preamp and operational amplifier as well as the A/D converter. A state-of-the-art $4^1/_2$-digit autoranging meter* provides voltage ranges from 200 mV to 200 V, current ranges from 2 pA (2×10^{-12} A) to 20 mA, resistance ranges from 2 kΩ to 200 GΩ (200×10^9 Ω), and Coulomb ranges from 200 pC to 200 nC. Specified accuracies for the most-sensitive ranges are ±(0.05 percent + 4 counts) for voltage, ±(1.6 percent + 66 counts) for current, ±(0.20 percent + 4 counts) for resistance, and ±(0.4 percent + 4 counts) for charge. The input impedance is greater than 200×10^{12} Ω in parallel with 20 pF, and the input bias current is less than 5×10^{-15} A at 23°C. A bipolar 100-V, 0.2 percent-accuracy voltage source programmable in 50-mV steps and a 1-nA to 100-μA decade current source are built into the meter. Two techniques may be used to measure resistance or generate $I - V$ curves: The decade current source can be used to force a known current through the unknown impedance with the voltage measured by the high-input-impedance voltmeter, or the voltage source can be applied across the unknown with the resulting current measured by the current meter. The latter method is preferred for characterizing voltage-dependent materials.

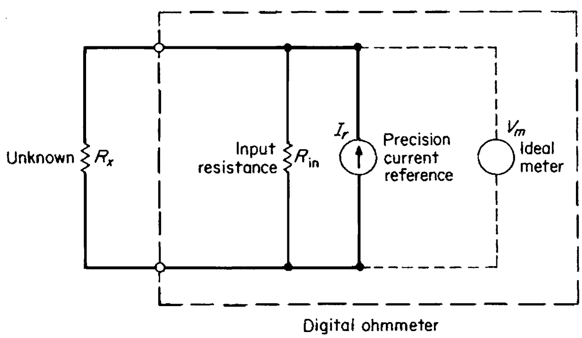


FIGURE 15.2.19 Effect of loading.

*Keithley Model 617 Electrometer.

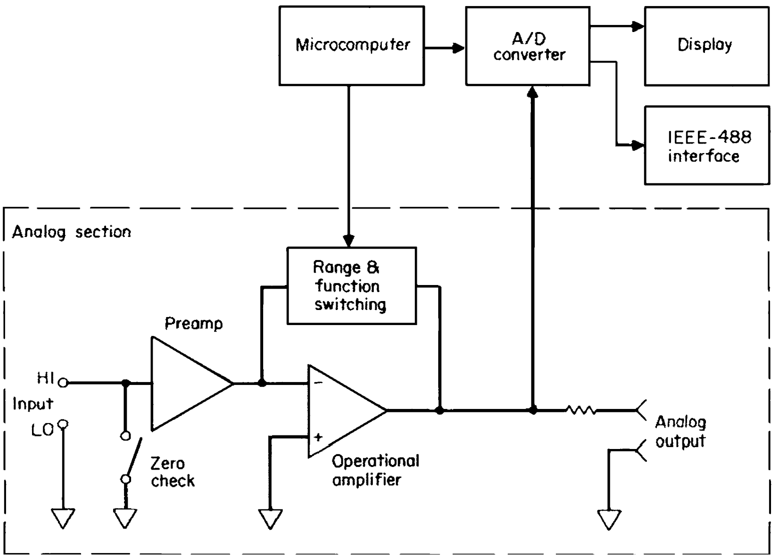


FIGURE 15.2.20 Typical digital electrometer.

A triaxial input cable, Fig. 15.2.21, is used with the electrometer to minimize input lead leakage and cable capacitance problems. The outer shield connected to the LO input can be grounded. The inner shield is driven by electrometer unity gain low-impedance output which maintains the guard voltage essentially equal to the high-impedance input signal HI. Insulator leakage through *r* becomes negligible since there is essentially no voltage difference between the center conductor and the guard.

Digital nanovoltmeters are digital voltmeters that are optimized for nanovolt measurements. A state-of-the-art meter uses a JFET input amplifier to obtain an input resistance of 10^9 Ω and a 5-nF input capacitance. Voltage ranges of 2 mV to 1000 V are available. Specified accuracy for a 24-h period is ± (0.006 percent + 5 counts) when properly zeroed. A special connector having a low thermoelectric voltage is used for the 200-mV and lower ranges.

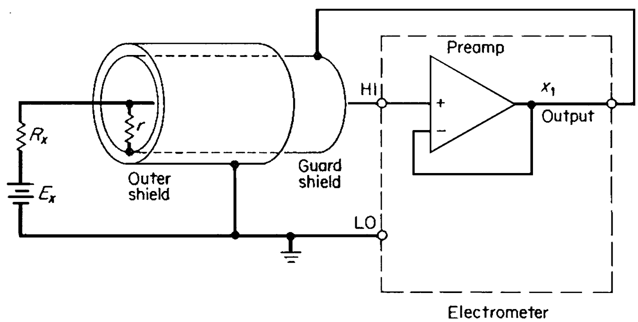


FIGURE 15.2.21 Guard shield minimizes input leakage.