Section C
Control

11 Electrical Measurement

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11.1 Introduction

With increased interest in quality in manufacturing industry, the judicious selection of measuring instruments and the ability to demonstrate the acceptability of the measurements made are more important than ever. With this in mind, a range of instruments is described and instrument specification and accuracy are considered.

11.2 Terminology

The vocabulary of this subject is often a source of confusion and so the internationally agreed definitions[1] of some important terms are given.

Measurement: The set of operations having the object of determining the value of a quantity.

Measurand: A quantity subject to measurement.

Metrology: The field of knowledge concerned with measurement. This term covers all aspects both theoretical and practical with reference to measurement, whatever their level of accuracy, and in whatever field of science or technology they occur.

Accuracy: The closeness of agreement between the result of a measurement and the true value of the measurand.

Systematic error: A component of the error of measurement which, in the course of a number of measurements of the same measurand, remains constant or varies in a predictable way.

Correction: The value which, added algebraically to the uncorrected result of a measurement, compensates for an assumed systematic error.

Random error: A component of the error of measurement which, in the course of a number of measurements of the same measurand, varies in an unpredictable way.

Uncertainty of measurement: An estimate characterising the range of values within which the true value of a measurand lies.

Discrimination: The ability of a measuring instrument to respond to small changes in the value of the stimulus.

Traceability: The property of the result of a measurement whereby it can be related to appropriate standards, generally international or national standards, through an unbroken chain of comparisons.

Calibration: The set of operations which establish, under specified conditions, the relationship between values indicated by a measuring instrument or measuring system and the corresponding known value of a measurand.

Adjustment: The operation intended to bring a measuring instrument into a state of performance and freedom from bias suitable for its use.

Influence quantity: A quantity which is not the subject of the measurement but which influences the value of a measurand or the indication of the measuring instrument.

11.3 The role of measurement traceability in product quality

Many organisations are adopting quality systems such as ISO 9000 (BS 5750). Such a quality system aims to cover all aspects of an organisation's activities from design through production to sales. A key aspect of such a quality system is a formal method for assuring that the measurements used in production and testing are acceptably accurate. This is achieved by requiring the organisation to demonstrate the traceability of measurements made. This involves relating the readings of instruments used to national standards, in the UK usually at the National Physical Laboratory, by means of a small number of steps. These steps would usually be the organisation's Calibration or Quality Department and a specialist calibration laboratory. In the UK, the suitability of such laboratories is carefully monitored by NAMAS. A calibrating laboratory can be given NAMAS accreditation in some fields of measurement and at some levels of accuracy if it can demonstrate acceptable performance. The traceability chain from production to National Measurement Laboratory can, therefore, be established and can be documented. This allows products to be purchased and used or built into further systems from organisations accredited to ISO 9000 with the confidence that the various products are compatible since all measurements used in manufacture and testing are traceable to the same national and international measurement standards.

11.4 National and international measurement standards

At the top of the traceability chain are the SI base units. Definitions of these base units are given in Chapter 1. The SI base units are metre, kilogram, second, ampere, kelvin, candela and mole. Traceability for these quantities involves comparison with the unit through the chain of comparisons. An example for voltage is shown in *Figure 11.1*.

The measurement standards for all other quantities are, in principle, derived from the base units. Traceability is, therefore, to the maintained measurement standard in a country. Since establishing the measurement standards of the base and derived quantities is a major undertaking at the level of uncertainty demanded by end users, this is achieved by international cooperation and the International Conference of Weights and Measures (CIPM) and the International Bureau of Weights and Measures (BIPM) have an important role to play. European cooperation between national measurement institutes is facilitated by Euromet.

11.4.1 Establishment of the major standards

11.4.1.1 Kilogram

The kilogram is the mass of a particular object, the International Prototype Kilogram kept in France. All other masses are related to the kilogram by comparison.

11.4.1.2 Metre

The definition of the metre incorporates the agreed value for the speed of light ($c = 299\,792\,458$ m/s). This speed does not have to be measured. The metre is not in practice established by the time-of-flight experiment implied in the definition, but by using given wavelengths of suitable laser radiation stated in the small print of the definition.

11.4.1.3 Second

The second is established by setting up a caesium clock apparatus. This apparatus enables a man-made oscillator to be locked onto a frequency inherent in the nature of the caesium atom, thus eliminating the inperfections of the man-made oscillator, such as temperature sensitivity and ageing.

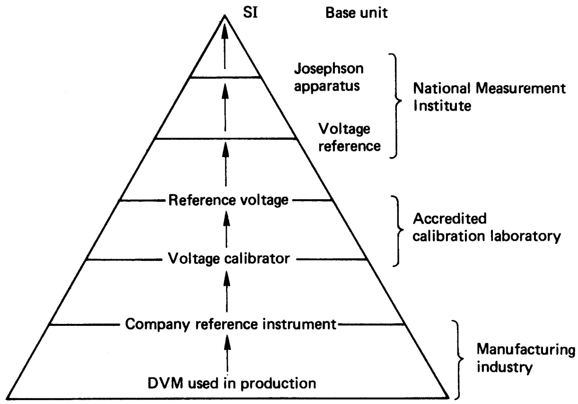

Figure 11.1 The traceability chain. DVM, digital voltmeter

Using this apparatus, the second can be established extremely close in size to that implied in the definition. Time and frequency can be propagated by radio signals and so calibration of, for example, a frequency counter consists in displaying the frequency of a stable frequency transmission such as BBC Radio 4 on 198 kHz. Commercial off-air frequency standards are available to facilitate such calibration.

11.4.1.4 Voltage

Work by Kibble[2] and others has led to the value of 483 597.9 GHz/V being ascribed to the height of the voltage step which is seen when a Josephson junction is irradiated with electromagnetic energy. A Josephson junction is a device formed by the separation of two superconductors by a very small gap. The voltage produced by such a junction is independent of temperature, age, materials used and apparently any other influencing quantity. It is therefore an excellent, if somewhat inconvenient, reference voltage. It is possible to cascade thousands of devices on one substrate to obtain volt level outputs at convenient frequencies. Industrial voltage measurements can be related to the Josephson voltage by the traceability hierachy.

11.4.1.5 Impedance

Since 1990 measurements of electrical resistance have been traceable to the quantised Hall effect at NPL.[3] Capacitance and inductance standards are realised from DC standards using a chain of AC and DC bridges. Impedance measurements in industry can be related through the traceability hierarchy to these maintained measurement standards. Research is in progress at NPL and elsewhere aimed at using the AC quantum hall effect for the direct realisation of impedance standards.

11.4.1.6 Kelvin

The fundamental fixed point is the triple point of water, to which is assigned the thermodynamic temperature of 273.16 K exactly; on this scale the temperature of the ice point is 273.15 K (or 0°C). Platinum resistance and thermocouple thermometers are the primary instruments up to 1336 K (gold point), above which the standard optical pyrometer is used to about 4000 K.

11.5 Direct-acting analogue measuring instruments

The principal specification for the characteristics and testing of low-frequency instruments is IEC 51 (BS 89). The approach of this standard is to specify intrinsic errors under closely defined reference conditions. In addition, variations are stated which are errors occurring when an influencing quantity such as temperature is changed from the reference condition. The operating errors can, therefore, be much greater than the intrinsic error which is likely to be a best-case value.

Instruments are described by accuracy class. The class indices are 0.05, 0.1, 0.2, 0.3, 0.5, 1, 1.5, 2, 2.5, 3 and 5 for ammeters and voltmeters. Accuracy class is the limit of the percentage intrinsic error relative to a stated value, usually the full-scale deflection. For example, for a class index of 0.05, the limits of the intrinsic error are ±0.05% of the full-scale value. It must be stressed that such performance is only achieved if the instrument is at the reference conditions and almost always extra errors in the form of variations must be taken into account.

11.5.1 Direct-acting indicators

These can be described in terms of the dominating torque-production effects:

(1) *Electromagnetic torque*: moving-coil, moving-iron, induction and electrodynamic (dynamometer).
(2) *Electric torque*: electrostatic voltmeter.
(3) *Electrothermal torque*: maximum demand.

A comparison is made in *Table 11.1*, which lists types and applications.

11.5.1.1 Torque effects

Instrument dynamics is discussed later. The relevant instantaneous torques are: *driving* torque, produced by means of energy drawn from the network being monitored; *acceleration* torque; *damping and friction* torque, by air dashpot or eddy current reaction; *restoring* torque, due usually to spring action, but occasionally to gravity or opposing magnetic field.

The higher the driving torque the better, in general, are the design and sensitivity; but high driving torque is usually associated with movements having large mass and inertia. The torque/mass ratio is one indication of relative performance, if associated with low power demand. For small instruments the torque is 0.001–0.005 mN-m for full-scale deflection; for 10–20 cm scales it is 0.05–0.1 mN-m, the higher figures being for induction and the lower for electrostatic instruments.

Friction torque, always a source of error, is due to imperfections in pivots and jewel bearings. Increasing use is now made of taut-band suspensions (*Figure 11.2*); this eliminates pivot friction and also replaces control springs. High-sensitivity moving-coil movements require only 0.005 mN-m for a 15 cm scale length: for moving-iron movements the torques are similar to those for the conventional pivoted instruments.

11.5.1.2 Scale shapes

The moving-coil instrument has a linear scale owing to the constant energy of the permanent magnet providing one of the two ‘force’ elements. All other classes of indicators are

Table 11.1 Comparison of instruments

Measurement parameter	*Type*	*Advantages*	*Disadvantages*
Direct voltage and current	Moving coil	Accurate: wide range	Power loss 50–100 mW
	Induced moving-magnet	Cheap	Inaccurate; high power loss
Alternating voltage and current	Moving iron	Cheap; reasonable r.m.s. accuracy	No low range; high power loss
	Electrodynamic	Adequate a.c./d.c. comparator; more accurate r.m.s.	Expensive; high power loss; square-law scale
	Induction	Long scale	Single frequency; high power loss; inaccurate
	Moving-coil rectifier	Low power loss (a few mW); wide range; usable up to audio frequency	Non-sinusoidal waveform error
	Moving-coil thermocouple	True r.m.s.; good a.c./d.c. comparator; usable up to radiofrequency	No overload safety factor; slow response; power loss (e.g. 1 W)
Direct and alternating voltage	Electrostatic	Negligible power; good a.c./d.c. comparator; accurate r.m.s.; wide frequency range	High capacitance; only for medium and high voltage; poor damping
A.c. power (active)	Electrodynamic	Accurate; linear scale	Expensive; high power loss
	Thermocouple electronic	High frequency; pulse power; true r.m.s.	Expensive
	Induction	Cheap; long scale	Single frequency; high power loss
A.c. power (reactive)	Electrodynamic	Accurate; linear scale	Power loss; needs phase-rotation network
A.c. maximum current	Thermal maximum demand	Long adjustable time constant; thermal integration; cheap	Relatively inaccurate

inherently of the double-energy type, giving a square-law scale of the linear property (voltage or current) being measured; for a wattmeter, the scale is linear for the average scalar product of voltage and current. The rectifier instrument, although an a.c. instrument, has a scale which is usually linear, as it depends on the moving-coil characteristic and the rectifier effect is usually negligible. (In low-range voltmeters, the rectifier has some effect and the scale shape is between linear and square law.) Thermal ammeters and electrodynamic voltmeters and ammeters usually have a true square-law scale, as any scale shaping requires extra torque and it is already low in these types. Moving-iron instruments are easily designed with high torque, and scale shaping is almost always carried out in order to approach a linear scale. In some cases the scale is actually contracted at the top in order to give an indication of overloads that would otherwise be off-scale. The best moving-iron scale shape is contracted only for about the initial 10% and is then nearly linear. Logarithmic scales have the advantage of equal percentage accuracy over all the scale, but they are difficult to read, owing to sudden changes in values of adjacent scale divisions. Logarithmic scales are unusual in switchboard instruments, but they are sometimes found in portable instruments, such as self-contained ohmmeters.

11.5.2 Direct voltage and current

11.5.2.1 Moving-coil indicator

This instrument comprises a coil, usually wound on a conducting former to provide eddy-current damping, with taut-band or pivot/control-spring suspension. In each case the coil rotates in the short air gap of a permanent magnet. The direction of the deflection depends on the polarity, so that unmodified indicators are usable only on d.c., and may have a centre zero if required. The error may be as low as $\pm 0.1\%$ of full-scale deflection; the range, from a few microamperes or millivolts up to 600 A and 750 V, which makes possible multirange d.c. test sets. The scale is generally linear (equispaced) and easily read. Non-linear scales can be obtained by shaping the magnet poles or the core to give a non-uniform air gap.

11.5.2.2 Corrections

The total power taken by a normal-range voltmeter can be 50 μW/V or more. For an ammeter the *total* series loss is 1–50 mW/A. Such powers may be a significant fraction of the total delivered to some electronic networks: in such

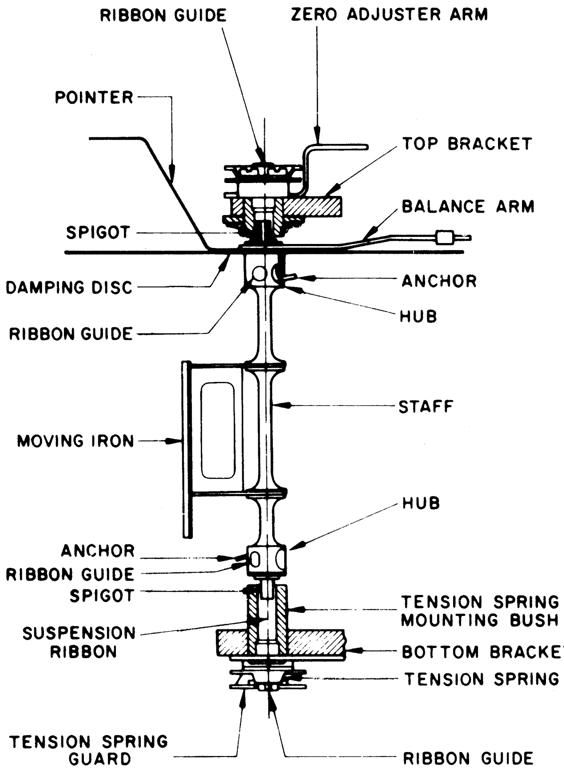

Figure 11.2 Taut-ribbon suspension assembly (Crompton Parkinson Ltd)

cases electronic and digital voltmeters with trivial power loss should be used instead.

11.5.2.3 Induced moving-magnet indicator

The instrument is polarised by a fixed permanent magnet and so is only suitable for d.c. A small pivoted iron vane is arranged to lie across the magnet poles, the field of which applies the restoring torque. One or more turns of wire carry the current to be measured around the vane, producing a magnetic field at right angles to that of the magnet, and this drive torque deflects the movement to the equilibrium position. It is a cheap, low-accuracy instrument which can have a nearly linear centre-zero scale, well suited to monitoring battery charge conditions.

11.5.3 Alternating voltage and current

11.5.3.1 Moving-iron indicator

This is the most common instrument for a.c. at industrial frequency. Some (but not all) instruments may also be used on d.c. The current to be measured passes through a magnetising coil enclosing the movement. The latter is formed either from a nickel–iron alloy vane that is *attracted* into the field, or from two magnetic alloy members which, becoming similarly polarised, mutually *repel*. The latter is more usual but its deflection is limited to about 90°; however, when combined with attraction vanes, a 240° deflection is obtainable but can be used only on a.c. on account of the use of material of high hysteresis.

The inherent torque deflection is square law and so, in principle, the instrument measures true r.m.s. values. By careful shaping of vanes when made from modern magnetic alloys, a good linear scale has been achieved except for the bottom 10%, which readily distinguishes the instrument from the moving-coil linear scale. Damping torque is produced by eddy-current reaction in an aluminium disc moving through a permanent-magnet field.

11.5.3.2 Precautions and corrections

These instruments work at low flux densities (e.g. 7 mT), and care is necessary when they are used near strong ambient magnetic fields despite the magnetic shield often provided. As a result of the shape of the B/H curve, a.c./d.c. indicators used on a.c. read high at the lower end and low at the higher end of the scale. The effect of hysteresis with d.c. measurements is to make the 'descending' reading higher than the 'ascending'.

Moving-iron instruments are available with class index values of 0.3–5.0, and many have similar accuracies for a.c. or d.c. applications. They tend to be single-frequency power instruments with a higher frequency behaviour that enables good r.m.s. readings to be obtained from supplies with distorted current waveforms, although excessive distortion will cause errors. Nevertheless, instruments can be designed for use in the lower audio range, provided that the power loss (a few watts) can be accepted. Ranges are from a few milliamperes and several volts upward. The power loss is usually significant and may have to be allowed for.

11.5.3.3 Electrodynamic indicator

In this the permanent magnet of the moving-coil instrument is replaced by an electromagnet (field coil). For ammeters the moving coil is connected in series with two fixed field coils; for voltmeters the moving and the fixed coils each have series-connected resistors to give the same time-constant, and the combination is connected in parallel across the voltage to be measured. With ammeters the current is limited by the suspension to a fraction of 1 A, and for higher currents it is necessary to employ non-reactive shunts or current transformers.

11.5.3.4 Precautions and corrections

The flux density is of the same order as that in moving-iron instruments, and similar precautions are necessary. The torque has a square-law characteristic and the scale is cramped at the lower end. By using the mean of reversed readings these instruments can be d.c. calibrated and used as adequate d.c./a.c. transfer devices for calibrating other a.c. instruments such as moving-iron indicators. The power loss (a few watts) may have to be allowed for in calculations derived from the readings.

11.5.3.5 Induction indicator

This single-frequency instrument is robust, but limited to class index (CI) number from 1.0 to 5.0. The current or voltage produces a proportional alternating magnetic field normal to an aluminium disc arranged to rotate. A second, phase-displaced field is necessary to develop torque. Originally it was obtained by pole 'shading', but this method is obsolete; modern methods include separate magnets one of which is shunted (Ockenden), a cylindrical aluminium movement (Lipman) or two electromagnets coupled by loops (Banner). The induction principle is most generally employed for the measurement of energy.

11.5.3.6 Moving-coil rectifier indicator

The place of the former copper oxide and selenium rectifiers has been taken by silicon diodes. Used with taut-band suspension, full-wave instrument rectifier units provide improved versions of the useful rectifier indicator.

11.5.3.7 Precautions and corrections

Owing to its inertia, the polarised moving-coil instrument gives a mean deflection proportional to the mean rectified current. The scale is marked in r.m.s. values on the assumption that the waveform of the current to be measured is sinusoidal with a form factor of 1.11. On non-sinusoidal waveforms the true r.m.s. value cannot be inferred from the reading: only the true *average* can be known (from the r.m.s. scale reading divided by 1.11). When true r.m.s. values are required, it is necessary to use a thermal, electrodynamic or square-law electronic instrument.

11.5.3.8 Multirange indicator

The rectifier diodes have capacitance, limiting the upper frequency of multirange instruments to about 20 kHz. The lower limit is 20–30 Hz, depending on the inertia of the movement. Indicators are available with CI numbers from 1.0 to 5.0, the multirange versions catering for a wide range of direct and alternating voltages and currents: a.c. 2.5 V–2.5 kV and 200 mA–10 A; d.c. 100 μV–2.5 kV and 50 μA–10 A. Non-linear resistance scales are normally included; values are derived from *internal* battery-driven out-of-balance bridge networks.

11.5.3.9 Moving-coil thermocouple indicator

The current to be measured (or a known proportion of it) is used as a heater current for the thermocouple, and the equivalent r.m.s. output voltage is steady because of thermal inertia—except for very low frequency (5 Hz and below). With heater current of about 5 mA in a 100 Ω resistor as typical, the possible voltage and current ranges are dictated by the usual series and shunt resistance, under the restriction that only the upper two-thirds of the square-law scale is within the effective range. The instrument has a frequency range from zero to 100 MHz. Normal radiofrequency (r.f.) low-range self-contained ammeters can be used up to 5 MHz with CI from 1.0 to 5.0. Instruments of this kind could be used in the secondary circuit of a r.f. current transformer for measurement of aerial current. The minimum measurable voltage is about 1 V.

The thermocouple, giving a true r.m.s. indication, is a primary a.c./d.c. transfer device. Recent evidence[4] indicates that the transfer uncertainty is only a few parts in 10^6. Such devices provide an essential link in the traceability chain between microwave power, r.f. voltage and the primary standards of direct voltage and resistance.

11.5.3.10 Precautions and corrections

The response is slow, e.g. 5–10 s from zero to full-scale deflection, and the overload capacity is negligible: the heater may be destroyed by a switching surge. Thermocouple voltmeters are low-impedance devices (taking, e.g., 500 mW at 100 V) and may be unsuitable for electronic circuit measurements.

11.5.4 Medium and high direct and alternating voltage

11.5.4.1 Electrostatic voltmeter

This is in effect a variable capacitor with fixed and moving vanes. The power taken (theoretically zero) is in fact sufficient to provide the small dielectric loss. The basically square-law characteristic can be modified by shaping the vanes to give a reasonably linear upper scale. The minimum useful range is 50–150 V in a small instrument, up to some hundreds of kilovolts for large fixed instruments employing capacitor multipliers. Medium-voltage direct-connected voltmeters for ranges up to 15 kV have CI ratings from 1.0 to 5.0. The electrostatic indicator is a true primary alternating/direct voltage (a.v./d.v.) transfer device, but has now been superseded by the thermocouple instrument. The effective isolation property of the electrostatic voltmeter is attractive on grounds of safety for high-voltage measurements.

11.5.5 Power

11.5.5.1 Electrodynamic indicator

The instrument is essentially similar to the electrodynamic voltmeter and ammeter. It is usually 'air-cored', but 'iron-cored' wattmeters with high-permeability material to give a better torque/mass ratio are made with little sacrifice in accuracy. The current (fixed) coils are connected in series with the load, and the voltage (moving) coil with its series range resistor is connected either (a) across the load side of the current coil, or (b) across the supply side. In (a) the instrument reading must be *reduced* by the power loss in the volt coil circuit (typically 5 W), and in (b) by the current coil loss (typically 1 W).

The volt coil circuit *power* corrections are avoided by use of a *compensated wattmeter*, in which an additional winding in series with the volt circuit is wound, turn-for-turn, with the current coils, and the connection is (a) above. The m.m.f. due to the volt-coil circuit current in the compensating coil will cancel the m.m.f. due to the volt circuit current in the current coils.

The volt circuit terminal marked ±⇐is immediately adjacent to the voltage coil; it should be connected to the current terminal similarly marked, to ensure that the wattmeter reads positive power and that negligible p.d. exists between fixed and moving coils, so safeguarding insulation and eliminating error due to electric torque.

When a wattmeter is used on d.c., the power should be taken from the mean of reversed readings; wattmeters read also the active (average) a.c. power. For the measurement of power at very low power factor, special wattmeters are made with weak restoring torque to give f.s.d. at, e.g., a power factor of 0.4. Range extension for all wattmeters on a.c. can be obtained by internal or external current transformers, internal resistive volt-range selectors or external voltage transformers. Typical self-contained ranges are between 0.5 and 20 A, 75 and 300 V. The usable range of frequency is about 30–150 Hz, with a best CI of 0.05.

Three-phase power can be measured by single dual-element instruments.

11.5.5.2 Thermocouple indicator

The modern versions of this instrument are more correctly termed electronic wattmeters. The outputs from current- and voltage-sensing thermocouples, multiplied together and amplified, can be displayed as average power. Typical d.c. and a.c. ranges are 300 mV–300 V and 10 mA–10 A, from pulsed and other non-sinusoidal sources at frequencies up to a few hundred kilohertz. The interaction with the network is low; e.g. the voltage network has typically an impedance of 10 kΩ/V.

11.5.5.3 Induction indicator

The power-reading instrument includes both current and voltage coils. The energy meter is referred to later. The instruments are frequency sensitive and are used only for fixed-frequency switchboard applications. The CI number is 1.0–5.0, usually the latter.

11.5.5.4 Electrodynamic reactive-power indicator

The active-power instrument can read reactive power if the phase of the volt circuit supply is changed by 90° to give $Q = VI \sin \phi$ instead of $P = VI \cos \phi$ (for sinusoidal conditions).

11.5.6 Maximum alternating current

11.5.6.1 Maximum-demand instrument

The use of an auxiliary pointer carried forward by the main pointer and remaining in position makes possible the indication of maximum current values, but the method is not satisfactory, because it demands large torques, a condition that reduces effective damping and gives rise to overswing. A truer maximum demand indication, and one that is insensitive to momentary peaks, is obtainable with the aid of a thermal bimetal. Passing the current to be indicated through the bimetal gives a thermal lag of a few minutes. For long-period indication (1 h or more) a separate heater is required.

Such instruments have been adapted for three-phase operation using separate heaters. A recent development permits a similar device to be used for two phases of a three-phase supply. The currents are summed *linearly*, and the maximum of the combined unbalanced currents maintained for 45 min is recorded on a kV-A scale marked in terms of nominal voltage.

11.5.7 Power factor

Power factor indicators have both voltage and current circuits, and can be interconnected to form basic electrodynamic or moving-iron direct-acting indicators. The current coils are fixed; the movement is free, the combined voltage- and current-excited field producing both deflecting and controlling torque. The electrodynamic form has the greater accuracy, but is restricted to a 90° deflection, as compared with 360° (90° lag and lead, motoring and generating) of the moving-iron type.

One electrodynamic form comprises two fixed coils carrying the line current, and a pair of moving volt coils set with their axes mutually at almost 90° (*Figure 11.3*). For one-phase operation the volt-coil currents are nearly in quadrature, being, respectively, connected in series with an inductor L and a resistor R. For three-phase working, L is

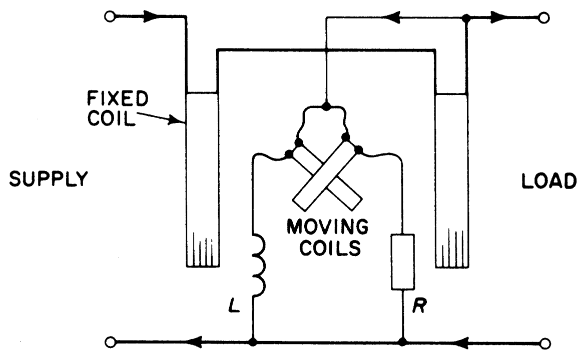

Figure 11.3 Single-phase electrodynamic power factor indicator

replaced by a resistor r and the ends of the volt-coil circuits taken, respectively, to the second and third phases. A three-phase moving-iron instrument for balanced loads has one current coil and three volt coils (or three current and one voltage); if the load is unbalanced, three current and three volt coils are required. All these power factor indicators are industrial frequency devices.

The one-phase electrodynamic wattmeter can be used as a phase-sensing instrument. When constant r.m.s. voltages and currents from two, identical-frequency, *sinusoidal* power supplies are applied to the wattmeter, then the phase change in one system produces power scale changes which are proportional to cos ϕ.

Many electronic phase-meters of the analogue and digital type are available which, although not direct-acting, have high impedance inputs, and they measure the phase displacement between any two voltages (of the same frequency) over a wide range of frequencies. Either the conventional solid state digital instruments generate a pulse when the two voltages pass zero, and measure and display the time between these pulses as phase difference, or the pulses trigger a multivibrator for the same period, to generate an output current proportional to phase (time) difference. These instruments provide good discrimination and, with modern high-speed logic switching, the instant for switch-on (i.e. pulse generation) is less uncertain: conversely, the presence of small harmonic voltages around zero could lead to some ambiguity.

11.5.8 Phase sequence and synchronism

A small portable form of *phase sequence* indicator is essentially a primitive three-phase induction motor with a disc rotor and stator phases connected by clip-on leads to the three-phase supply. The disc rotates in the marked forward direction if the sequence is correct. These instruments have an intermittent rating and must not be left in circuit.

The *synchroscope* is a power factor instrument with rotor slip-rings to allow continuous rotation. The moving-iron type is robust and cheap. The direction of rotation of the rotor indicates whether the incoming three-phase system is 'fast' or 'slow', and the speed of rotation measures the difference in frequency.

11.5.9 Frequency

Power frequency monitoring frequency indicators are based on the response of reactive networks. Both electrodynamic and induction instruments are available, the former having the better accuracy, the latter having a long scale of 300° or more.

One type of indicator consists of two parallel fixed coils tuned to slightly different frequencies, and their combined currents return to the supply through a moving coil which lies within the resultant field of the fixed coils. The position of the moving coil will be unique for each frequency, as the currents (i.e. fields) of the fixed coils are different unique functions of frequency. The indications are within a restricted range of a few hertz around nominal frequency. A ratiometer instrument, which can be used up to a few kilohertz, has a permanent-magnet field in which lie two moving coils set with their planes at right angles. Each coil is driven by rectified a.c., through resistive and capacitive impedances, respectively, and the deflection is proportional to frequency.

Conventional solid state counters are versatile time and frequency instruments. The counter is based on a stable crystal reference oscillator (e.g. at 1 MHz) with an error between 1 and 10 parts in 10^8 per day. The displays have up to eight digits, with a top frequency of 100 MHz (or 600 MHz with heterodyning). The resolution can be 10 ns, permitting pulse widths of 1000 ns to be assessed to 1%: but a 50 Hz reading would be near to the bottom end of the display, giving poor discrimination. All counters provide for measurements giving the period of a waveform to a good discrimination.

11.6 Integrating (energy) metering

Integrating meters record the time integral of active, reactive and apparent power as a continuous summation. The integration may be limited by a specified total energy (e.g. prepayment meters) or by time (e.g. maximum demand). Meters for a.c. supplies are all of the induction type, with measuring elements in accordance with the connection of the load (one-phase or three- or four-wire three-phase). Manufacturing and testing specifications are given in BS 37. Instrument transformers for use with meters are listed in BS 3938 and BS 3941.

11.6.1 Single-phase meter

The rotor is a light aluminium disc on a vertical spindle, supported in low-friction bearings. The lower bearing is a sapphire cup, carrying the highly polished hemispherical hardened steel end of the spindle. The rotor is actuated by an induction driving element and its speed is controlled by an eddy current brake. The case is usually a high-quality black phenolic moulding with integral terminal block. The frame is a rigid high-stability iron casting which serves as the mounting, as part of a magnetic flux path, and as a shield against ambient magnetic fields.

The driving element has the basic form shown in *Figure 11.4*. It has two electromagnets, one voltage-excited and the other current-excited. The volt magnet, roughly of E-shape, has a nearly complete magnetic circuit with a volt coil on the central limb. Most of the flux divides between the outer limbs, but the *working* flux from the central limb penetrates the disc and enters the core of the current magnet. The latter, of approximate U-shape, is energised by a coil carrying the load current. With a condition of *zero* load current, the working flux from the volt magnet divides equally between the two limbs of the current magnet and returns to the volt magnet core through the frame, or through an iron path provided for this purpose. If the volt magnet flux is symmetrically disposed, the eddy current induced in the disc does not exert any net driving torque and the disc remains

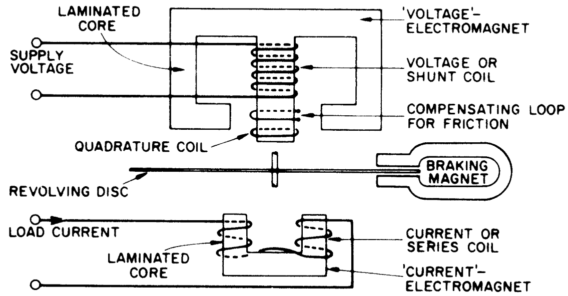

Figure 11.4 Essential parts of an induction meter

stationary. The volt magnet flux is approximately in phase quadrature with the applied voltage.

When a load current flows in the coil of the current magnet, it develops a co-phasal flux, interacting with the volt magnet flux to produce a resultant that 'shifts' from one current magnet pole to the other. Force is developed by interaction of the eddy currents in the disc and the flux in which it lies, and the net torque is proportional to the voltage and to that component of the load current in phase with the voltage—i.e. to the active power. The disc therefore rotates in the direction of the 'shift'.

The disc rotates through the field of a suitably located permanent isotropic brake magnet, and induced currents provide a reaction proportional to speed. Full-load adjustment is effected by means of a micrometer screw which can set the radial position of the brake magnet.

The volt magnet working flux lags the voltage by a phase angle rather less than 90° (e.g. by 85°). The phase angle is made 90° by providing a closed *quadrature coil* on the central limb, with its position (or resistance) capable of adjustment.

The accuracy of the meter is affected on low loads by pivot friction. A *low-load adjustment*, consisting of some device that introduces a slight asymmetry in the volt magnet working flux, is fitted to mitigate frictional error. There are several methods of producing the required asymmetry: one is the insertion of a small magnetic vane into the path of the working flux. The asymmetry results in the development at zero load of a small forward torque, just sufficient to balance friction torque without causing the disc to rotate.

11.6.1.1 Performance of single-phase meters

The limits of permissible error are defined in BS 37. In general, the error should not exceed ±2% over most of the working range for credit meters. For prepayment meters, +2 and −3% for loads above 1/30 full load at unity power factor, at any price setting to be used, are the specified limits. Tests for compliance with this requirement at and below 1/10 marked current must be made with the coin condition that not less than one nor more than three coins of the highest denomination acceptable by the meter are prepaid.

It is usual for manufacturers to adjust their meters to less than one-half of the permissible error over much of the working range. The mean error of an individual meter is, of course, less than the maximum observable error, and the mean collective error of a large number of meters is likely to be much less than ±1% at rated voltage and frequency. If no attempt is made during production testing to bias the error in one direction, the weighted mean error of many meters taken collectively is probably within 0.5%, and is likely to be positive (see *Figure 11.5*(*a*)).

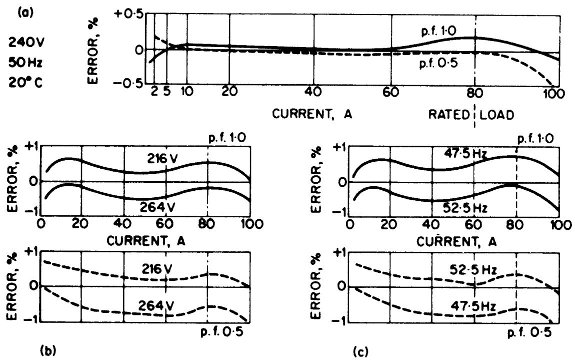

Figure 11.5 Typical performance of 240 V, 50 Hz, 20–80 A house-service meter (Ferranti Ltd)

11.6.1.2 Temperature

The temperature error of a.c. meters with no temperature compensation is negligible under the conditions normally existing in the UK. In North America it is common practice to fix meters on the outside of buildings, where they are subjected to wide temperature ranges (e.g. 80°C), which makes compensation desirable. However, even in the UK it is common to provide compensation (when required) in the form of a strip of nickel-rich alloy encircling the stator magnet. Its use is advantageous for short-duration testing, as there is an observable difference in error of a meter when cold and after a 30 min run.

11.6.1.3 Voltage and frequency

Within the usual limits of ±4% in voltage and ±0.2% in frequency, the consequent errors in a meter are negligible. Load-shedding, however, may mean substantial reductions of voltage and frequency. A significant reduction in voltage usually causes the meter to read high on all loads, as shown in *Figure 11.5*(*b*). Reduction of the supply frequency makes the meter run faster on high-power-factor loads and slower on lagging reactive loads: an increase of frequency has the opposite effect (*Figure 11.5*)(*c*). The errors are cumulative if both voltage and frequency variations occur together.

11.7 Electronic instrumentation

The rapid development of large-scale integrated (l.s.i.) circuits, as applied to analogue and especially digital circuitry, has allowed both dramatic changes in the format of measurement instrumentation by designers and in the measurement test systems available for production or control applications.

A modern instrument is often a multipurpose assembly, including a microprocessor for controlling both the sequential measurement functions and the subsequent programmed assessment of the data retained in its store; hence, these instruments can be economical in running and labour costs by avoiding skilled attention in the recording and statistical computation of the data. Instruments are often designed to be compatible with standardised interfaces, such as the IEC bus or IEEE-488 bus, to form automatic measurement systems for interactive design use of feedback digital-control applications; typically a system would include a computer-controlled multiple data logger with 30 inputs from different parameter sensors, statistical and system analysers, displays and storage. The input analogue parameters would use high-speed analogue-to-digital (A/D) conversion, so minimising the front-end analogue conditioning networks while maximising the digital functions (e.g. filtering and waveform analyses by digital rather than analogue techniques). The readout data can be assembled digitally and printed out in full, or presented through D/A converters in graphical form on *X–Y* plotters using the most appropriate functional co-ordinates to give an economical presentation of the data.

The production testing of the component subsystems in the above instruments requires complex instrumentation and logical measurements, since there are many permutations of these minute logic elements resulting in millions of high-speed digital logic operations—which precludes a complete testing cycle, owing to time and cost. Inspection testing of l.s.i. analogue or digital devices and subassemblies demands systematically programmed measurement testing sequences with varied predetermined signals; similar automatic test equipments (ATE) are becoming more general in any industrial production line which employs these modern instruments in process or quality-control operations. Some companies are specialising in computer-orientated measurement techniques to develop this area of ATE; economic projections indicate that the present high growth rate in ATE-type measurement systems will continue or increase during the next few years.

The complementary nature of automated design, production and final testing types of ATE leads logically to the interlinking of these separate functions for optimising and improving designs within the spectrum of production, materials handling, quality control and costing to provide an overall economic and technical surveillance of the product.

Some of the more important instruments have been included in the survey of electronic instrumentation in the following pages, but, owing to the extensive variety of instruments at present available and the rapid rate of development in this field, the selection must be limited to some of the more important devices.

Digital instruments usually provide an output in visual decimal or coded digit form as a discontinuous function of a smoothly changing input quantity. In practice the precision of a digital instrument can be extended by adding digits to make it better than an analogue instrument, although the final (least significant) digit must always be imprecise to ±0.5.

11.7.1 Digital voltmeters

These provide a digital display of d.c. and/or a.c. inputs, together with coded signals of the visible quantity, enabling the instrument to be coupled to recording or control systems. Depending on the measurement principle adopted, the signals are sampled at intervals over the range 2–500 ms. The basic principles are: (a) linear ramp; (b) charge balancing; (c) successive-approximation/potentiometric; (d) voltage to frequency integration; (e) dual slope; (f) multislope; and (g) some combination of the foregoing.

Modern digital voltmeters often include a multiway socket connection (e.g. BCD; IEC bus; IEEE-488 bus, etc.) for data/interactive control applications.

11.7.1.1 Linear ramp

This is a voltage/time conversion in which a linear time-base is used to determine the time taken for the internally generated voltage v_s to change by the value of the unknown voltage V. The block diagram (*Figure 11.6*(*b*)) shows the use of comparison networks to compare V with the rising

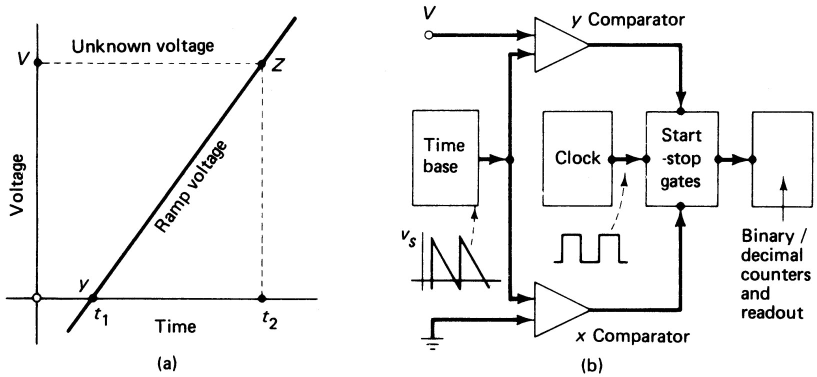

Figure 11.6 Linear ramp digital voltmeter. (a) Ramp voltage; (b) block diagram

(or falling) v_s; these networks open and close the path between the continuously running oscillator, which provides the counting pulses at a fixed clock rate, and the counter. Counting is performed by one of the binary-coded sequences; the translation networks give the visual decimal output. In addition, a binary-coded decimal output may be provided for monitoring or control purposes.

Limitations are imposed by small non-linearities in the ramp, the instability of the ramp and oscillator, imprecision of the coincidence networks at instants y and z, and the inherent lack of noise rejection. The overall uncertainty is about $\pm 0.05\%$ and the measurement cycle would be repeated every 200 ms for a typical four-digit display.

Linear 'staircase' ramp instruments are available in which V is measured by counting the number of equal voltage 'steps' required to reach it. The staircase is generated by a solid state diode pump network, and linearities and accuracies achievable are similar to those with the linear ramp.

11.7.1.2 Charge balancing

This principle[5] employs a pair of differential-input transistors used to charge a capacitor by a current proportional to the unknown d.c. voltage, and the capacitor is then discharged by a large number of small $+\delta q$ and $-\delta q$ quantities, the elemental discharges being sensed and directed by fast comparator/flip-flop circuits and the total numbers being stored. Zero is in the *middle* of the measurement range and the displayed result is proportional to the difference between the number of $+\delta q$, $-\delta q$ events.

The technique which, unlike dual-ramp methods, is inherently bipolar, claims some other advantages, such as improved linearity, more rapid recovery from overloads, higher sensitivity and reduced noise due to the averaging of thousands of zero crossings during the measurement, as well as a true *digital* auto-zero subtraction from the next measurement, as compared with the more usual capacitive-stored analogue offset p.d. being subtracted from the measured p.d.

Applications of this digital voltmeter system (which is available on a monolithic integrated circuit, Ferranti ZN 450) apart from d.c. and a.c. multimeter uses, include interfacing it *directly* with a wide range of conventional transducers such as thermocouples, strain gauges, resistance thermometers, etc.

11.7.1.3 Successive approximation

As it is based on the potentiometer principle, this class produces very high accuracy. The arrows in the block diagram of *Figure 11.7* show the signal-flow path for one version; the resistors are selected in sequence so that, with a constant current supply, the test voltage is created within the voltmeter.

Each decade of the unknown voltage is assessed in terms of a sequence of accurate stable voltages, graded in descending magnitude in accordance with a binary (or similar) counting scale. After each voltage approximation of the final result has been made and stored, the residual voltage is then automatically re-assessed against smaller standard voltages, and so on to the smallest voltage discrimination required in the result. Probably four logic decisions are needed to select the major decade value of the unknown voltage, and this process will be repeated for each lower decade in decimal sequence until, after a few milliseconds, the required voltage is stored in a coded form. This voltage is then translated for decimal display. A binary-coded sequence could be as shown in *Table 11.2*, where the numerals in **bold** type represent a logical *rejection* of that number and a progress to the next lower value. The actual sequence of logical decisions is more complicated than is suggested by the example. It is necessary to sense the initial polarity of the unknown signal, and to select the range and decimal marker for the read-out; the time for the logic networks to settle must be longer for the earlier (higher voltage) choices than for the later ones, because they must be of the highest possible accuracy; offset voltages may be added to the earlier logic choices, to be withdrawn later in the sequence; and so forth.

The total measurement and display takes about 5 ms. When noise is present in the input, the necessary insertion of filters may extend the time to about 1 s. As noise is more troublesome for the smaller residuals in the process, it is sometimes convenient to use some different techniques for the latter part. One such is the voltage–frequency principle (see below), which has a high noise rejection ratio. The

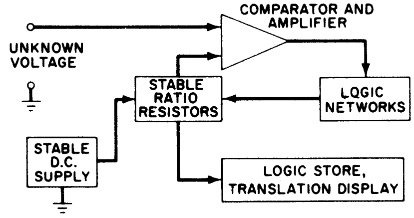

Figure 11.7 Successive-approximation digital voltmeter

Table 11.2 Voltages obtained from residual (difference) currents across high-stability resistors

Unknown analogue voltage	3	9	.	2	0	6
Logic decisions in	**8**	8		**8**	**8**	**8**
vertical binary	**4**	**4**		**4**	**4**	4
sequences and in	2	**2**		2	**2**	2
descending 'order'	1	1		**1**	**1**	**1**
Decoded decimal display	3	9	.	2	0	6

reduced accuracy of this technique can be tolerated as it applies only to the least significant figures.

11.7.1.4 Voltage–frequency

The converter (*Figure 11.8*) provides output pulses at a frequency proportional to the instantaneous unknown input voltage, and the non-uniform pulse spacing represents the variable frequency output. The decade counter accumulates the pulses for a predetermined time T and in effect measures the average frequency during this period. When T is selected to coincide with the longest period time of interfering noise (e.g. mains supply frequency), such noise averages out to zero.

Instruments must operate at high conversion frequencies if adequate discrimination is required in the final result. If a six-digit display were required within 5 ms for a range from zero to 1.000 00 V to $\pm 0.01\%$, then 10^5 counts during 5 ms are called for, i.e. a 20 MHz change from zero frequency with a 0.01% voltage–frequency linearity. To reduce the frequency range, the measuring time is increased to 200 ms or higher. Even at the more practical frequency of 0.5 MHz that results, the inaccuracy of the instrument is still determined largely by the non-linearity of the voltage–frequency conversion process.

In many instruments the input network consists of an integrating operational amplifier in which the average input voltage is 'accumulated' in terms of charge on a capacitor in a given time.

11.7.1.5 Dual-slope

This instrument uses a composite technique consisting of an integration (as mentioned above) and an accurate measuring network based on the ramp technique. During the integration (*Figure 11.9*) the unknown voltage V_x is switched at time zero to the integration amplifier, and the initially uncharged capacitor C is charged to V_x in a known time T_1, which may be chosen so as to reduce noise interference. The ramp part of the process consists in replacing V_x by a *reversed* biased direct reference voltage V_r which produces a constant-current discharge of C; hence, a known linear voltage/time change occurs across C. The total voltage change to zero can be measured by the method used in the linear-ramp instrument, except that the slope is negative and that counting begins at maximum voltage. From the diagram in *Figure 11.9(a)* it follows that $\tan\theta_x/\tan\theta_r = (T_2 - T_1)/T_1 = V_x/V_r$ so that V_x is directly proportional to $T_2 - T_1$. The dual-slope method is seen to depend ultimately on time-base linearity and on the measurement of time difference, and is subject to the same limitations as the linear-ramp method, but with the important and fundamental quality of inherent noise rejection capability.

Noise interference (of series-mode type) is principally due to e.m.f.s at power supply frequency being induced into the d.c. source of V_x. When T_1 equals the periodic time of the interference (20 ms for 50 Hz), the charging p.d. v_c across C would follow the dotted path (*Figure 11.9(a)*) without changing the final required value of v_c, thus eliminating this interference.

11.7.1.6 Multislope

At time T_1 (above) the maximum dielectric absorption in C will coincide with v_c maximum; this effect degrades (a) the linearity of the run-down slope during $T_2 - T_1$ and (b) the identification of zero p.d. One improved 'multislope' technique reduces dielectric absorption in C by inserting various reference voltages during the run-up period and, after subtraction, leaves a lower p.d., v'_c, prior to run-down while storing the most significant digit of V_x during the run-up period. During run-down, C is discharged rapidly to measure the next smaller digit; the residue of v'_c, including overshoot, is assessed to give the remaining three digits in sequence using three v_r/t functions of slope +1/10, −1/100, +1/1000 compared with the initial rapid discharge. Measurement time is reduced, owing to the successive digits being accumulated during the measurement process.[6]

11.7.1.7 Mixed techniques

Several techniques can be combined in the one instrument in order to exploit to advantage the best features of each. One accurate, precise, digital voltmeter is based upon precision inductive potentiometers, successive approximation, and the dual-slope technique for the least significant figures. An uncertainty of 10 parts in 10^6 for a 3-month period is claimed, with short-term stability and a *precision* of about 2 parts in 10^6.

11.7.1.8 Digital multimeters

Any digital voltmeter can be scaled to read d.c. or a.c. voltage, current, immittance or any other physical property, provided that an appropriate transducer is inserted. Instruments scaled for alternating voltage and current

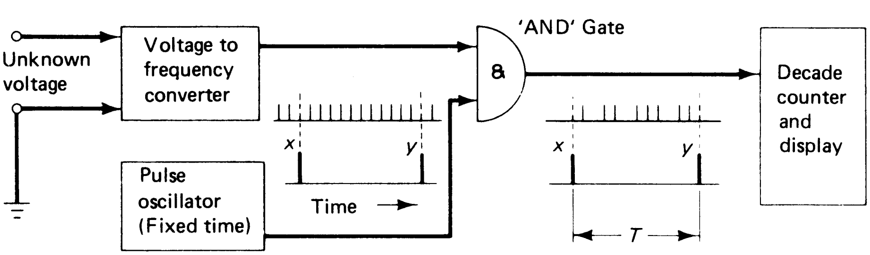

Figure 11.8 Voltage-to-frequency digital converter

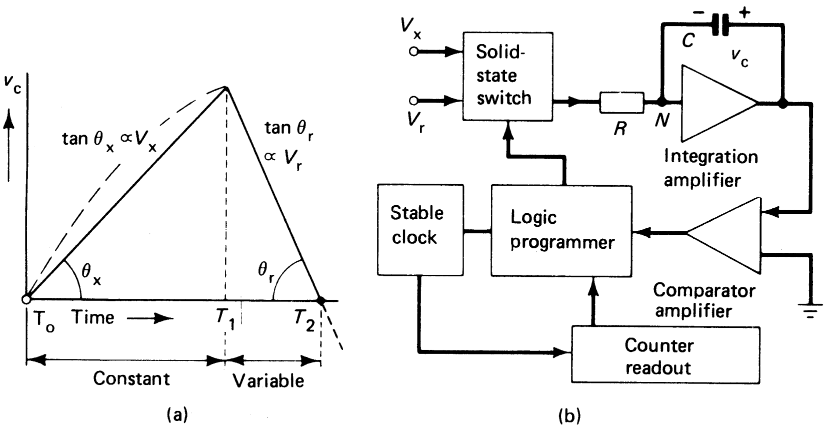

Figure 11.9 Dual-slope digital voltmeter. (a) Ramp voltage; (b) block diagram

normally incorporate one of the a.c./d.c. converter units listed in a previous paragraph, and the quality of the result is limited by the characteristics inherent in such converters. The digital part of the measurement is more accurate and precise than the analogue counterpart, but may be more expensive.

For systems application, programmed signals can be inserted into, and binary-coded or analogue measurements received from, the instrument through multiway socket connections, enabling the instrument to form an active element in a control system (e.g. IEC bus, IEEE-488 bus).

Resistance, capacitance and inductance measurements depend to some extent on the adaptability of the basic voltage-measuring process. The dual-slope technique can be easily adapted for two-, three- or four-terminal ratio measurements of resistance by using the positive and negative ramps in sequence; with other techniques separate impedance units are necessary. (See *Table 11.3*.)

11.7.1.9 Input and dynamic impedance

The high precision and small uncertainty of digital voltmeters make it essential that they have a high input impedance if these qualities are to be exploited. Low test voltages are often associated with source impedances of several hundred kilo-ohms; for example, to measure a voltage with source resistance 50 kΩ to an uncertainty of ±0.005% demands an instrument of input resistance 1 GΩ, and for a practical instrument this must be 10 GΩ if the loading error is limited to one-tenth of the total uncertainty.

The dynamic impedance will vary considerably during the measuring period, and it will always be lower than the quoted null, passive, input impedance. These changes in dynamic impedance are coincident with voltage 'spikes' which appear at the terminals owing to normal logic functions; this noise can adversely affect components connected to the terminals, e.g. Weston standard cells.

Input resistances of the order of 1–10 GΩ represent the conventional range of good-quality insulators. To these must be added the stray parallel reactance paths through unwanted capacitive coupling to various conducting and earth planes, frames, chassis, common rails, etc.

11.7.1.10 Noise limitation

The information signal exists as the p.d. between the two input leads; but each lead can have unique voltage and impedance conditions superimposed on it with respect to the basic *reference* or *ground* potential of the system, as well as another and different set of values with respect to a local *earth* reference plane.

An elementary electronic instrumentation system will have at least one ground potential and several earth connections—possibly through the (earthed) neutral of the main supply, the signal source, a read-out recorder or a cathode ray oscilloscope. Most true earth connections are at different electrical potentials with respect to each other, owing to circulation of currents (d.c. to u.h.f.) from other apparatus, through a finite earth resistance path. When multiple earth connections are made to various parts of a high-gain amplifier system, it is possible that a significant frequency spectrum of these signals will be introduced as electrical noise. It is this interference which has to be rejected by the input networks of the instrumentation, quite apart from the concomitant removal of any electrostatic/electromagnetic noise introduced by direct coupling into the signal paths. The total contamination voltage can be many times larger (say 100) than the useful information voltage level.

Electrostatic interference in input cables can be greatly reduced by 'screened' cables (which may be 80% effective as screens), and electromagnetic effects minimised by transposition of the input wires and reduction in the 'aerial loop' area of the various conductors. Any residual effects, together with the introduction of 'ground and earth-loop' currents into the system, are collectively referred to as *series* and/or *common-mode* signals.

11.7.1.11 Series and common-mode signals

Series-mode (normal) interference signals, V_{sm}, occur in series with the required information signal. Common-mode interference signals, V_{cm}, are present in both input leads with respect to the reference potential plane: the required information signal is the difference voltage between these leads. The results are expressed as *rejection ratio* (in decibels) with respect to the *effective* input error signal, V_e, that the interference signals produce, i.e.

$$K_{sm} = 20\log(V_{sm}/V_e) \text{ and } K_{cm} = 20\log(V_{cm}/V_e)$$

where K is the rejection ratio. The series-mode rejection networks are *within* the amplifier, so V_e is measured with *zero* normal input signal as V_e = output interference

Table 11.3 Typical characteristics of digital multimeters

(a) Type *(b) ±Uncertainty of maximum reading* *(c) Resolution*	*Voltage*	*Current*	*Resistance*	*Voltage and frequency min ≃ 40 Hz*	*Current and frequency min ≃ 40 Hz*	*Parallel input R and C*	*Recalibration period* (months)*
(a) $3\frac{1}{2}$ digits general-purpose (mean sensing)	0.1 V–1 kV	0.1 A–1 A	1.1 kΩ–11 MΩ	0.1 V–1.0 kV	0.1 A–1.0 A	10 MΩ	3
(b)	0.3%	0.8%	0.5%	1.5% → 8% (2 kHz → 10 kHz)	2% → 3.5% (2 kHz → 5 kHz)	30 pF	
(c)	100 μV	100 μA	1 Ω!	100 μV	100 μA		
(a) $4\frac{1}{2}$ digits general purpose (mean sensing)	20 mV–1 kV	200 μA → 2 A	200 Ω → 20 MΩ	200 mV – 100 V	200 μA – 100 mA	1 MΩ	
(b)	0.03%	0.1% → 0.6%	0.02% → 0.1%	0.15% → 0.5% (10 kHz → 20 kHz)	0.3% → 0.7% (10 kHz → 20 kHz)	100 pF	12
(c)	1 μV	10 nA	10 mΩ!	10 μV	10 nA		
(a) $5\frac{1}{2}$ digits (r.m.s.)	0.1 V–1 kV	—	0.1 kΩ → 15 MΩ	*r.m.s.* 1 V–1 kV	—	*d.c.* 10^{10} Ω → 10^{7} Ω	3
(b)	0.01%	—	50 → 1000 ppm	0.1% → 9% (20 kHz → 1 MHz)	—	2 MΩ	
(a) $5\frac{1}{2}$ digits (mean)				*mean* 1 V–1 kV		100 pF	
(b)				0.2% → 0.8% (100 Hz → 250 kHz)			
(c)	1 μV	—	1 mΩ!	10 μV	—		
(a) $6\frac{1}{2}$ digits precision (r.m.s. up to 200 V)	10 mV–1 kV	—	14 Ω–14 MΩ	0.1 V–1 kV	—	1 MΩ 150 pF	6
(b)	20 ppm	—	50 ppm	0.15% → 0.9% (10 kHz → 100 kHz)	—		
(c)	1 μV	—	1 mΩ!	1 μV	—		
(a) $8\frac{1}{2}$ digits	0.1 V–1 kV	—	0.1 kΩ → 1.4 GΩ	0.1 V–1 kV r.m.s.	—	*a.c.* 1 MΩ 150 pF	3
(b)	15 ppm	—	17–46 ppm (0.1% 1 GΩ range)	0.02% → 2% 40 Hz → 1 MHz	—	*d.c.* 10 GΩ to 10 MΩ 10 MΩ (0.1 → 1 kV)	
(c)	10 nV	—	10 μΩ!	1 μV	—		

* Specifications often include 12-, 6-, and 3-month periods each with progressively smaller uncertainties of measurement down to 24-h statements for high precision instruments.

voltage ÷ gain of the amplifier appropriate to the bandwidth of V_e.

Consider the elementary case in *Figure 11.10*, where the input-lead resistances are unequal, as would occur with a transducer input. Let r be the difference resistance, C the cable capacitance, with common-mode signal and error voltages V_{cm} and V_e, respectively. Then the common-mode numerical ratio (c.m.r.) is

$$V_{cm}/V_e = V_{cm}/ri = 1/2\pi fCr$$

assuming the cable insulation to be ideal, and $X_C \gg r$. Clearly, for a common-mode statement to be complete, it must have a stated frequency range and include the resistive imbalance of the source. (It is often assumed in c.m.r. statements that $r = 1\,\text{k}\Omega$.)

The c.m.r. for a digital voltmeter could be typically 140 dB (corresponding to a ratio of $10^7/1$) at 50 Hz with a 1 kΩ line imbalance, and leading consequently to $C = 0.3$ pF. As the normal input cable capacitance is of the order of 100 pF/m, the situation is not feasible. The solution is to inhibit the return path of the current i by the introduction of a guard network. Typical guard and shield parameters are shown in *Figure 11.11* for a six-figure digital display on a voltmeter with ±0.005% uncertainty. Consider the magnitude of the common-mode error signal due to a 5 V, 50 Hz common-mode voltage between the shield earth E_1 and the signal earth E_2:

N–G not connected. The a.c. *common*-mode voltage drives current through the guard network and causes a change of 1.5 mV to appear across r as a *series*-mode signal; for $V_s = 1$ V this represents an error V_e of 0.15% for an instrument whose quality is ±0.005%.

N–G connected. The *common*-mode current is now limited by the shield impedance, and the resultant *series*-mode signal is 3.1 μV, an acceptably low value that will be further reduced by the noise rejection property of the measuring circuits.

11.7.1.12 Floating-voltage measurement

If the d.c. voltage difference to be measured has a p.d. to E_2 of 100 V, as shown, then with N–G open the change in p.d. across r will be 50 μV, as a *series*-mode error of 0.005% for

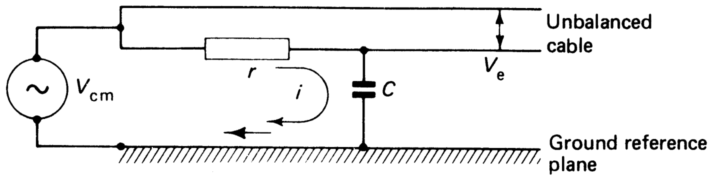

Figure 11.10 Common-mode effect in an unbalanced network

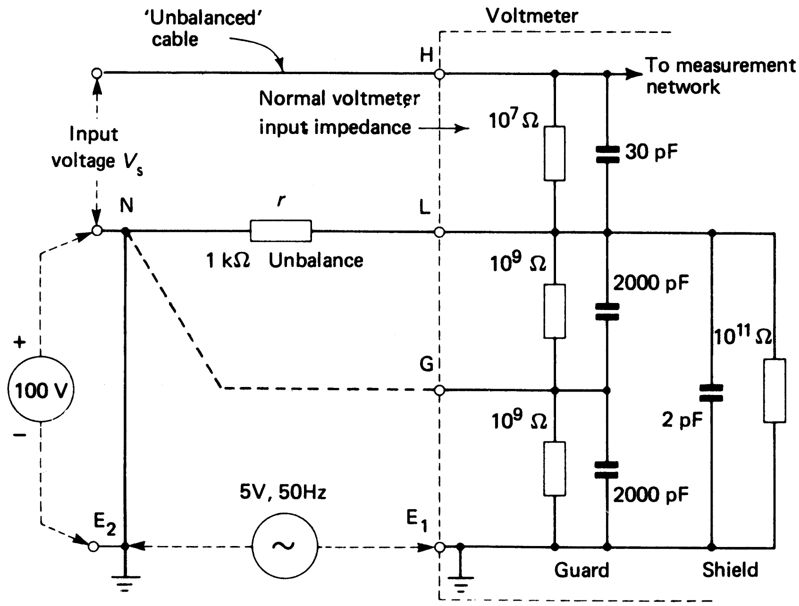

Figure 11.11 Typical guard and shield network for a digital voltmeter

a 1 V measurement. With N–G connected the change will be 1 μV, which is negligible.

The interconnection of electronic apparatus must be carefully made to avoid systematic measurement errors (and short circuits) arising from incorrect screen, ground or earth potentials.

In general, it is preferable, wherever possible, to use a *single* common reference node, which should be at zero signal reference potential to avoid leakage current through *r*. Indiscriminate interconnection of the shields and screens of adjacent components can increase *noise* currents by short-circuiting the high-impedance stray path between the screens.

11.7.1.13 Instrument selection

A precise seven-digit voltmeter, when used for a 10 V measurement, has a discrimination of ±1 part in 10^6 (i.e. ±10 μV), but has an uncertainty ('accuracy') of about ±10 parts in 10^6 (i.e. ±100 μV). The distinction is important with digital read-out devices, lest a higher quality be accorded to the number indicated than is in fact justified. The quality of any reading must be based upon the time stability of the total instrument since it was last calibrated against *external* standards, and the cumulative evidence of previous calibrations of a like kind.

Selection of a digital voltmeter from the list of types given in *Table 11.3* is based on the following considerations:

(1) No more digits than necessary, as the cost per extra digit is high.
(2) High input impedance, and the effect on likely sources of the *dynamic* impedance.
(3) Electrical noise rejection, assessed and compared with (a) the common-mode rejection ratio based on the input and guard terminal networks and (b) the actual inherent noise rejection property of the measuring principle employed.
(4) Requirements for binary-coded decimal, IEEE-488, storage and computational facilities.
(5) Versatility and suitability for use with a.c. or impedance converter units.
(6) Use with transducers (in which case (3) is the most important single factor).

11.7.1.14 Calibration

It will be seen from *Table 11.3* that digital voltmeters should be recalibrated at intervals between 3 and 12 months. Built-in self-checking facilities are normally provided to confirm satisfactory operational behaviour, but the 'accuracy' check cannot be better than that of the included standard cell or Zener diode reference voltage and will apply only to *one* range. If the user has available some low-noise stable voltage supplies (preferably batteries) and some high-resistance helical voltage dividers, it is easy to check logic sequences, resolution, and the approximate ratio between ranges. In order to demonstrate traceability to national standards, a voltage source whose voltage is traceable to national standards should be used. This can either be done by the user or, more commonly, by sending the instrument to an accredited calibration laboratory.

11.7.2 Digital wattmeters[7–9]

Digital wattmeters employ various techniques for sampling the voltage and current waveforms of the power source to produce a sequence of instantaneous power representations from which the average power is obtained. One new NPL method uses high-precision successive-approximation A/D integrated circuits with the multiplication and averaging of the digitised outputs being completed by computer.[10,11] A different 'feedback time-division multiplier' principle is employed by Yokogawa Electric Limited.[1]

The NPL digital wattmeter uses 'sample and hold' circuits to capture the two instantaneous voltage (v_x) and current (i_x) values, then it uses a novel double dual-slope multiplication technique to measure the average power from the mean of numerous instantaneous ($v_x i_x$) power measurements captured at precise intervals during repetitive waveforms. Referring to the single dual-ramp (*Figure 11.9*(*a*)) the a.c. voltage v_x (captured at T_0) is measured as $v_x T_1 = V_r(T_2 - T_1)$. If a second voltage (also captured at T_0) is proportional to i_x and integrated for $T_2 - T_1$, then, if it is reduced to zero (by V_r) during the time $T_3 - T_2$ it follows that $i_x(T_2 - T_1) = V_r(T_3 - T_2)$.

The instantaneous power (at T_0) is $v_x i_x = K(T_3 - T_2)$ and, by scaling, the mean summation of all counts such as $T_3 - T_2$ equals the average power. The prototype instrument measures power with an uncertainty of about ±0.03% f.s.d. (and ±0.01%, between 50 and 400 Hz, should be possible after further development).

The 'feedback time division multiplier' technique develops rectangular waveforms with the pulse height and width being proportional, respectively, to the instantaneous voltage (v_x) and current (i_x); the average 'area' of all such instantaneous powers ($v_x i_x$) is given (after l.p. filtering) as a d.c. voltage measured by a DVM scaled in watts; such precision digital wattmeters, operating from 50 to 500 Hz, have ±0.05% to ±0.08% uncertainty for measurements up to about 6 kW.

11.7.3 Energy meters

A *solid state* single-phase meter is claimed to fulfil the specifications of its traditional electromechanical counterpart while providing a reduced uncertainty of measurement coupled with improved reliability at a similar cost. The instrument should provide greater flexibility for reading, tariff and data communication of all kinds, e.g. instantaneous power information, totalised energy, credit limiting and multiple tariff control. The instrument is less prone to fraudulent abuse.

11.7.4 Signal generators

Signal generators provide sinusoidal, ramp and square-wave output from fractions of 1 Hz to a few megaherz, although the more accurate and stable instruments generally cover a more restricted range. As voltage sources, the output power is low (100 mW–2 W). The total harmonic distortion (i.e. r.m.s. ratio between harmonic voltage and fundamental voltage) must be low, particularly for testing-quality amplifiers, e.g. 90 dB rejection is desirable and 60 dB (10^3:1) is normal for conventional oscillators. Frequency selection by dials may introduce ±2% uncertainty, reduced to 0.2% with digital dial selection which also provides 0.01% reproducibility. Frequency-synthesiser function generators should be used when a precise, repeatable frequency source is required. Amplitude stability and constant value over the whole frequency range is often only 1–2% for normal oscillators.

11.7.5 Electronic analysers

Analysers include an important group of instruments which operate in the frequency domain to provide specialised analyses, in terms of energy, voltage, ratio, etc., (a) to characterise networks, (b) to decompose complex signals which include analogue or digital intelligence, and (c) to identify interference effects and non-linear cross-modulations throughout complex systems.

The frequency components of a complex waveform (Fourier analysis) are selected by the sequential, or parallel, use of analogue to digital filter networks. A frequency-domain display consists of the simultaneous presentation of each separate frequency component, plotted in the *x*-direction to a base of frequency, with the magnitude of the *y*-component representing the parameter of interest (often energy or voltage scaled in decibels).

Instruments described as network, signal, spectrum or Fourier analysers as well as digital oscilloscopes have each become so versatile, with various built-in memory and computational facilities, that the distinction between their separate objectives is now imprecise. Some of the essential generic properties are discussed below.

11.7.5.1 Network analysers

Used in the design and synthesis of complex cascaded networks and systems. The instrument normally contains a swept-frequency sinusoidal source which energises the test system and, using suitable connectors or transducers to the input and output ports of the system, determines the frequency-domain characteristic in 'lumped' parameters and the transfer function equations in both magnitude and phase: high-frequency analysers often characterise networks in 'distributed' *s*-parameters for measurements of insertion loss, complex reflection coefficients, group delay, etc. Three or four separate instruments may be required for tests up to 40 GHz. Each instrument would have a wide dynamic range (e.g. 100 dB) for frequency and amplitude, coupled with high resolution (e.g. 0.1 Hz and 0.01 dB) and good accuracy, often derived from comparative readings or ratios made from a built-in reference measurement channel.

Such instruments have wide applications in research and development laboratories, quality control and production testing—and they are particularly versatile as an automatic measurement testing facility, particularly when provided with built-in programmable information and a statistical storage capability for assessment of the product. Data display is by $X-Y$ recorder and/or cathode-ray oscilloscope (c.r.o.) and the time-base is locked to the swept frequency.

11.7.5.2 Spectrum analysers

Spectrum analysers normally do not act as sources; they employ a swept-frequency technique which could include a very selective, narrow-bandwidth filter with intermediate frequency (i.f.) signals compared, in sequence, with those selected from the swept frequency; hence, close control of frequency stability is needed to permit a few per cent resolution within, say, a 10 Hz bandwidth filter. Instruments can be narrow range, 0.1 Hz–30 kHz, up to wide range, 1 MHz–40 GHz, according to the precision required in the results; typically, an 80 dB dynamic range can have a few hertz bandwidth and 0.5 dB amplitude accuracy. Phase information is not directly available from this technique. The instrument is useful for tests involving the output signals from networks used for frequency mixing, amplitude and frequency modulation or pulsed power generation. A steady c.r.o. display is obtained from the sequential measurements owing to digital storage of the separate results when coupled with a conventional variable-persistence cathode-ray tube (a special cathode-ray tube is not required for a digital oscilloscope): the digitised results can, in some instruments, be processed by internal programs for (a) time averaging of the input parameters, (b) probability density and cumulative distribution, and (c) a comprehensive range of statistical functions such as average power or r.m.s. spectrum, coherence, autocorrelation and cross-correlation functions, etc.

11.7.5.3 Fourier analysers

Fourier analysers use digital signalling techniques to provide facilities similar to the spectrum analysers but with more flexibility. These 'Fourier' techniques are based upon the calculation of the discrete Fourier transform using the fast Fourier transform algorithm (which calculates the magnitude and phase of each frequency component using a group of time-domain signals from the input signal variation of the sampling rate); it enables the long measurement time needed for very-low-frequency ($\ll 1$ Hz) assessments to be completed in a shorter time than that for a conventional swept measurement—together with good resolution (by digital translation) at high frequencies. Such specialised instruments are usable only up to 100 kHz but they are particularly suitable for examining low-frequency phenomena such as vibration, noise transmission through media and the precise measurement of random signals obscured by noise, etc.

11.7.6 Data loggers

Data loggers consist of an assembly of conditioning amplifiers which accept signals from a wide range of transducers or other sources in analogue or digital form; possibly a small computer program will provide linearisation and other corrections to signals prior to computational assessments (comparable to some features of a spectrum or other class of analyser) before presenting the result in the most economical manner (such as a graphical display). The data logger, being modular in construction, is very flexible in application; it is not limited to particular control applications and by the nature of the instrument can appear in various guises.

11.8 Oscilloscopes

The cathode-ray oscilloscope (c.r.o.) is one of the most versatile instruments in engineering. It is used for diagnostic testing and monitoring of electrical and electronic systems, and (with suitable transducers) for the display of time-varying phenomena in many branches of physics and engineering. Two-dimensional functions of any pair of repetitive, transient or pulse signals can be developed on the fluorescent screen of a c.r.o.

These instruments may be classified by the manner in which the *analogue* test signals are conditioned for display as (a) the 'analogue c.r.o.', using analogue amplifiers, and (b) the 'digital storage oscilloscope', using A/D converters at the input with 'all-digit' processing.

11.8.1 Cathode-ray tube

The cathode-ray tube takes its name from Pluecker's discovery in 1859 of 'cathode rays' (a better name would be 'electron beam', for J. J. Thomson in 1897 showed that the 'rays' consist of high-speed electrons). The developments by Braun, Dufour and several other investigators enabled Zworykin in 1929 to construct the essential features of the modern cathode-ray tube, namely a small thermionic cathode, an electron lens and a means for modulating the beam intensity. The block diagram (*Figure 11.12*) shows the principal components of a modern cathode-ray oscilloscope. The tube is designed to focus and deflect the beam by means of structured electric field patterns. Electrons are accelerated to high speed as they travel from the cathode through a potential rise of perhaps 15 kV. After being focused, the beam may be deflected by two mutually perpendicular X and Y electric fields, locating the position of the fluorescing 'spot' on the screen.

11.8.1.1 Electron gun and beam deflection

The section of the tube from which the beam emerges includes the heater, cathode, grid and the first accelerating anode; these collectively form the *electron gun*. A simplified diagram of the connections is given in *Figure 11.13*. The electric field of the relatively positive anode partly penetrates through the aperture of the grid electrode and determines on the cathode a disc-shaped area (bounded by the -15 kV equipotential) within which electrons are drawn away from the cathode: outside this area they are driven back. With a sufficiently negative grid potential the area shrinks to a point and the beam current is zero. With rising grid potential, both the emitting area and the current density within it grow rapidly.

The diagram also shows the electron paths that originate at the emitting area of the cathode. It can be seen that those which leave the cathode at right angles cross the axis at a point not far from the cathode. This is a consequence of the powerful lens action of the strongly curved equipotential surfaces in the grid aperture. But, as electrons are emitted with random thermal velocities in all directions to the normal, the 'cross-over' is not a point but a small *patch*, the image of which appears on the screen when the spot is adjusted to maximum sharpness.

The conically shaped beam of divergent electrons from the electron gun have been accelerated as they rise up the steep potential gradient. The electrical potentials of the remaining nickel anodes, together with the post-deflection accelerating anode formed by the graphite coating, are selected to provide

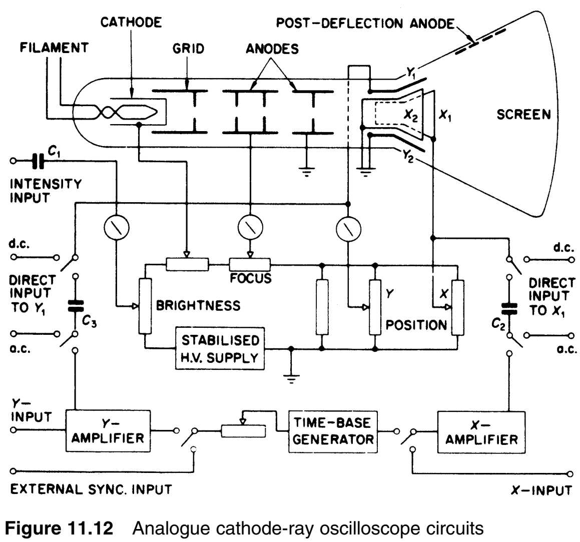

Figure 11.12 Analogue cathode-ray oscilloscope circuits

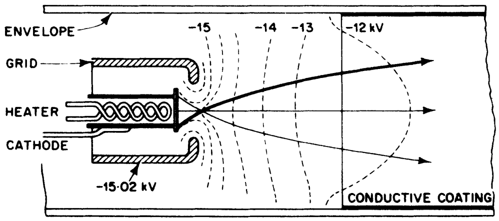

Figure 11.13 Electron beam in an electron gun

additional acceleration and, by field shaping, to refocus the beam to a spot on the screen. To achieve sharply defined traces it is essential that the focused spot should be as small as possible; the area of this image is, in part, dictated by the location and size of the 'point source' formed in front of the cathode. The rate at which electrons emerge from the cathode region is most directly controlled by adjustment of the negative potential of the grid; this is effected by the external *brightness control*. A secondary result of changes in the brightness control may be to modify, slightly, the shape of the point source. The slight de-focusing effect is corrected by the *focus control*, which adjusts an anode potential in the electron lens assembly.

11.8.1.2 Performance (electrostatic deflection)

If an electron with zero initial velocity rises through a potential difference V, acquiring a speed u, the change in its kinetic energy is $Ve = \frac{1}{2}mu^2$, so that the speed is

$$u = \sqrt{(2Ve/m)} \simeq 6 \times 10^5 \sqrt{V}\ \mathrm{m/s}$$

using accepted values of electron charge, e, and rest-mass, m. The performance is dependent to a considerable extent on the beam velocity, u.

The deflection sensitivity of the tube is increased by shaping the deflection plates so that, at maximum deflection, the beam grazes their surface (Y_1Y_2; *Figure 11.12*). The sensitivity and performance limits may be assessed from the *idealised* diagrams (and notation) in *Figure 11.14*. Experience shows that a good cathode-ray tube produces a beam-current density J_s at the centre of the luminous spot in rough agreement with that obtained on theoretical grounds by Langmuir, i.e.

$$J_s = kJ_c\theta^2 V/T \qquad (11.1)$$

where J_c is the current density at the cathode, V is the total accelerating voltage, T is the absolute temperature of the cathode emitting surface, θ is one-half of the beam convergence angle, and k is a constant. If i_b is the beam current, then the apparent diameter of the spot is given approximately by $\sqrt{(i_b/J_s)}$.

The voltage V_d applied to the deflection plates produces an angular deflection δ given by

$$V_d = \tfrac{1}{2}\delta V d/L_d \qquad (11.2)$$

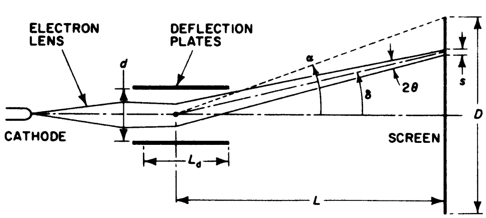

Figure 11.14 Deflection parameters

Deflection is possible, however, only up to a maximum angle α, such that

$$\alpha L_d + \theta L = \tfrac{1}{2}d \qquad (11.3)$$

From these three expressions, certain quantities fundamental to the rating of a cathode-ray tube with electric field deflection can be derived.

11.8.1.3 Definition

This is the maximum sweep of the spot on the screen, $D = 2\alpha L$, divided by the spot diameter, i.e. $N = D/s$. In television N is prescribed by the number of lines in the picture transmission. In oscillography N is between 200 and 300. Better definition can be obtained only at the expense of deflection sensitivity.

11.8.1.4 Specific deflection sensitivity

This is the inverse of the voltage that deflects the spot by a distance equal to its own diameter, i.e. the voltage which produces the smallest perceptible detail. The corresponding deflection angle is $\delta_0 = s/L$. Use of equations (11.1)–(11.3) gives for the specific deflection sensitivity

$$S = k_1 L/[(L/L_d + \alpha/\theta)\sqrt{(Vi_b)}] \qquad (11.4)$$

where k_1 is a constant. Thus, the larger the tube dimensions the better the sensitivity at the smallest beam power consistent with adequate brightness of trace.

11.8.1.5 Maximum recording speed

The trace remains visible so long as its brightness exceeds a certain minimum. The brightness is proportional to the beam power density VJ, to the luminous efficiency σ of the screen, and to the time s/U during which the beam sweeps at speed U over one element of the screen. Hence,

$$U = k_2 sJV\sigma \qquad (11.5)$$

Recording speeds up to 40 000 km/s have been achieved with sealed-off tubes, permitting single-stroke photography of the trace.

11.8.1.6 Maximum specified recording speed

Unless compared with the spot size s, the value of U does not itself give adequate information. The specific speed $U_s = U/s$ is more useful: it is the inverse measure of the shortest time-interval in which detail can be recorded. Combining equations (11.5) and (11.1)

$$U_s = k_2 JV\sigma = k_3\sigma(V\theta)^2 \qquad (11.6)$$

The product $\sigma(V\theta)^2$ is the essential figure in the speed rating of a cathode-ray tube.

11.8.1.7 Limiting frequency

A high recording speed is useful only if the test phenomena are recorded faithfully. The cathode-ray tube fails in this respect at frequencies such that the time of passage of an electron through the deflection field is more than about one-quarter of the period of an alternating quantity. This gives the condition of maximum frequency f_m:

$$f_m = 1.5 \times 10^5 \sqrt{V}/L_d \qquad (11.7)$$

which means that the smallest tube at the highest acceleration voltage has the highest frequency limit. This conflicts with the condition for deflection sensitivity. Combining the two expressions gives the product of the two:

$$(Sf_{\mathrm{m}}) = k_4/[1 + (L_{\mathrm{d}}/L)(\alpha/\theta)]\sqrt{i_{\mathrm{b}}} \qquad (11.8)$$

Compromise must be made according to the specific requirement. It may be observed that the optimum current i_{b} is not zero as might appear, because for such a condition the convergence angle θ is also zero. In general, the most advantageous current will not differ from that for which the electron beam just fills the aperture.

Limitation of cathode-ray-tube performance by space charge is an effect important only at low accelerating voltages and relatively large beam currents. For these conditions the expressions above do not give practical ratings.

11.8.1.8 Fluorescent screen

The kinetic energy of each incident electron is converted into light owing to the exchange of energy which occurs within the phosphors coated on the inside of the screen. The result, *luminescence*, is usually described as either fluorescence or phosphorescence. *Fluorescence* is the almost instantaneous energy conversion into light, whereas *phosphorescence* is restricted to the phosphors which 'store' the converted energy and release it, as light, after a short delay. The *afterglow* can last many seconds. The respective terms are applied to the phosphors in accordance with their dominant light-emitting behaviour; however, each property is always present in a phosphor. In most cathode-ray tubes fluorescent phosphors are used with a very short afterglow (40 μs); a longer afterglow would blur a quickly changing display.

The most common fluorescent substances are silicates and sulphides, with calcium tungstate for photographic work. An excellent zinc silicate occurs naturally in the mineral willemite, but all other substances are prepared artificially. After purification, the substances are activated by the addition of very small quantities of suitable metals, and then ground to a powder of grain size 5–10 μm (for sulphides 10–50 μm).

In applying the powder to the glass several methods are used. Very finely ground silicates can be applied without a binder, by setting the powder from water, or an electrolyte. Sulphides are usually compounded with the glass wall by means of a binder such as sodium silicate; good results have been obtained also by baking them directly into the glass. In choosing the method two conflicting considerations must be taken into account. Baking or compounding gives good thermal contact, and prevents overheating of the powder particles by the electron beam. But at the same time good optical contact is established with the glass, with the result that light from the powder enters the glass also at angles exceeding the total reflection angle. This part of the light cannot escape at the viewing side, but is reflected and forms a luminous 'halo' which blurs the picture.

While silicates fluoresce usually green or blue-green, white fluorescence can be obtained by mixing cadmium sulphide and zinc cadmium sulphide in suitable proportions. The luminous efficiency of modern fluorescent powders is very high. Zinc silicate emits up to 2.5 cd/W of beam power, and some sulphides are even more effective. The luminous efficacy increases with voltage up to about 10 kV, beyond which it decreases, partly because fast electrons are not stopped by the crystal grains.

The range of phosphors is classified by 'P' numbers, as in *Table 11.4*.

The light persistence of the most widely used phosphor (P31) is 40 μs; however, there are tubes available, using this phosphor, which have both variable persistence and 'mesh storage' facilities. Adjustable persistence enables very-low-frequency signals, lasting for a minute, to be viewed as a complete trace while retaining the normal 'writing' properties of the tube. Storage, at reduced brightness, is possible for up to 8 h.

Phosphor materials are insulators. The current circuit for all the incident beam electrons is completed by an equal number of low-energy electrons being attracted along various paths through the vacuum to the positive conducting-graphite coating. The result of secondary emission of electrons is to leave the screen with a positive charge only a little lower in potential than that of the graphite coating; thus, the surplus secondary emission electrons return to the phosphor-coated screen surface. This mechanism works perfectly at the voltages usually employed in oscillography and television (3–20 kV), also in small projection-tube sealed-off high-speed oscillographs (10–25 kV). Above 25 kV the mechanism is unsatisfactory and the dispersion of charge has to be supplemented by other means.

The power concentration in the luminous spot of a cathode-ray tube is very high. In direct-vision television tubes the momentary luminance may exceed 10 cd/mm^2. A 3 mA 80 kV beam of a large-screen projection tube may be concentrated in an area of 0.1 mm^2, representing a power density of 2 kW/mm^2. (By comparison, melting tungsten radiates about 10 W/mm^2.) The width of the trace in a conventional 10 kV cathode-ray tube with a beam current of 10 μA is only 0.35 mm.

11.8.2 Deflection amplifiers (analogue c.r.o.)

Amplifiers are built into cathode-ray oscilloscope (c.r.o.) equipment to provide for time base generation and the amplification of small signals. The signals applied to the vertical (Y) deflecting plates are often repetitive time functions. To display them on a time base, the vertical displacements must be moved horizontally across the screen at uniform speed repeatedly at intervals related to the signal frequency, and adjusted to coincidence by means of a synchronising signal fed to the time base from the Y amplifier.

Table 11.4 Classification of phosphors

Phosphor	*Persistence*	*Colour of trace*	*Relative brightness*	*Application*
P2	Medium/short	Yellow/green	6.5	Low repetition rates (general oscillography)
P4	Medium	White	5.0	High-contrast displays (monochrome television)
P7	Long	White or yellow/green	4.5	Long-persistence low-speed (radar)
P11	Medium/short	Blue	2.5	High writing speeds (photography)
P31	Medium/short	Green	10	High-brightness, general use

An additional constraint may be the requirement to start the time base from the left side of the screen at the instant the Y-plate signal is zero: a synchronising trigger signal is fed to the time base circuit to initiate each sweep. Finally, the spot must return (flyback) to the origin at the end of each sweep time. Ideally, flyback should be instantaneous: in practice it takes a few score nanoseconds. During flyback the beam is normally cut off to avoid return-trace effects, and this is the practice for television tubes.

Linear time bases are described in terms either of the sweep time per centimetre measured on the horizontal centre line of the graticule, or of the frequency. The repetition frequency range is from about 0.2 Hz to 2.5 MHz.

11.8.2.1 Ramp voltage (saw-tooth) generator

If a constant current I is fed to a capacitor C, its terminal voltage increases linearly with time. An amplified version of the voltage is applied to the X plates. Various sweep times are obtained by selecting different capacitors or other constant current values.

The method shown in *Figure 11.15* employs a d.c. amplifier of high gain $-G$ with a very large negative feedback coupled through an ideal capacitor C_f. The amplifier behaves in a manner dictated by the feedback signal to node B rather than by the small direct input voltage E_s. The resultant signal v_i is small enough for node B to be considered as virtually at earth potential. Analysing the network based on the 'virtual earth' principle, then $v_1 = 0$ and $i_i = 0$, the input resistance R_i being typically 1 MΩ. Then $i_s + i_f = 0$, whence

$$(E_s/R_s) = C_f(\mathrm{d}v_o/\mathrm{d}t) = 0 \text{ giving } v_o = -(1/C_f R_s)\int E_s \cdot \mathrm{d}t$$

The output voltage v_o increases linearly with time to provide the ramp function input to the X amplifier.

11.8.2.2 Variable delay

When two similar X amplifiers, A and B, are provided, B can be triggered by A; if B is set for, say, five times the frequency of A and this had a variable delay facility, then any one-fifth of the complete A waveform display can be extended to fill the screen. Both complete and expanded waveforms can be viewed simultaneously by automatic switching of A and B to the X plates during alternate sweeps.

11.8.2.3 Signal amplifiers

The X- and Y-plate voltages for maximum deflection in a modern 10 cm cathode-ray tube may lie between 30 and 200 V. As many voltage phenomena of interest have amplitudes ranging between 500 V and a few millivolts, both impedance matched attenuators and voltage amplifiers are called for. The amplifiers must be linear and should not introduce phase distortion in the working frequency range; the pulse response must be fast (e.g. 2 ns); and the drift and noise must be small (e.g. 3 μV).

Amplifiers have d.c. and a.c. inputs. The d.c. coupling is conventional, to minimise the use of capacitors (which tend to produce low-frequency phase distortion). The d.c. inputs also permit the use of a reference line. The a.c. input switch introduces a decoupling capacitor.

Two separate Y amplifiers are often provided; in conjunction either with one switched or with two separate X amplifiers, they permit the examination of two voltages simultaneously. It is important to distinguish the dual-trace oscilloscope (in which a single beam is shared by separate signal amplifiers) and the true double-beam device (with a gun producing two distinct beams).

11.8.3 Instrument selection

A general purpose c.r.o. must compromise between a wide frequency range and a high amplification sensitivity, because high gain is associated with restricted bandwidth. An economically priced c.r.o. with solid state circuitry may have a deflection sensitivity of 5 mV/cm, bandwidth 10 MHz and a 3 kV accelerating voltage for a 10 cm useful display. For greater flexibility, many modern c.r.o.s comprise a basic frame unit with plug-in amplifiers; the frame carries the tube, all necessary power outlets and basic controls; the plug-in units may have linear or special characteristics. The tube must be a high-performance unit compatible with any plug-in unit. The cost may be higher than that of a single function c.r.o., but it may nevertheless be economically advantageous.

Many classes of measurement are catered for by specially designed oscilloscopes. In the following list, the main characteristic is given first, followed by the maximum frequency, deflection sensitivity and accelerating voltage, with general application notes thereafter.

Low-frequency (30 MHz, or 2 MHz at high gain; 0.1 mV/cm–10 V/cm; 3–10 kV). General purpose oscillography, for low-frequency system applications. Usually, high X-sweep speeds are needed for transient displays and complex transducer signals. Two traces available with two-beam or chopped single-beam tube. Single-shot facility. Y-amplifier signal delay useful for 'leading-edge' display. Rise times 10–30 ns.

Medium-frequency (100 MHz; 0.1 mV/cm; 8–15 kV). General-purpose, including wide-band precision types with a rise time of 3–10 ns.

High-frequency (non-sampling) (275 MHz; 10 mV/cm; 20 kV). May include helical transmission-line deflecting

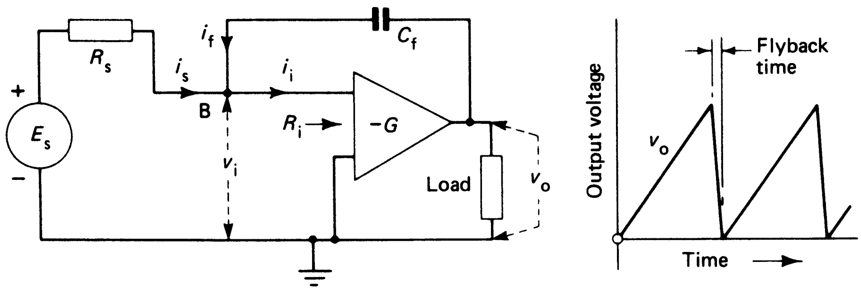

Figure 11.15 Time-base operational amplifier

plates. Rise time 1.5 ns for testing digital system and computer displays.

Variable persistence storage cathode-ray tube for h.f. transients.

Very-high frequency (sampling) (12 000 MHz; 5 mV/cm; 20 kV). The v.h.f. waveforms are 'instantaneously' measured and optically stored, n times at n successively retarded instants during the period of each of n successive waveforms (with n typically 1000). The n spots accumulate in the correct time sequence as a single waveform derived from the n waveforms sampled.

Large-screen (20 MHz; 60 mV/cm; 28 kV). Screen area up to $0.25\,m^2$. Useful for industrial monitoring, classroom display and computer graphics. (Magnetic deflection gives a shorter overall length of tube, probably with improved focusing towards the edges but with a lower maximum frequency.)

11.8.3.1 Other features

With all c.r.o.s, the X–Y cartesian coordinate display can be augmented by Z-modulation. This superimposes a beam-intensity control on the $X-Y$ signals to produce the illusion of three dimensions. The beam control may brighten one spot only, or gradually 'shape' the picture.

Oscilloscopes are supplemented by such accessories as special probes for matching and/or attenuating the signal to the signal/cable interface. Anti-microphony cables are available which minimise cable-generated noise and capacitance change. Many types of high-speed camera can be used, including single-shot cameras that can employ ultraviolet techniques to 'brighten' a record from a trace of low visual intensity.

11.8.4 Operational use

To avoid degradation of the signal source, a c.r.o. has a high input impedance (1 MΩ, 20 pF) and a good common-mode rejection (50 dB). The quality can be considerably impaired by impedance mismatch and incorrect signal/earth (or ground) connections.

11.8.4.1 Impedance matching

Voltage signals are often presented to a c.r.o. through wires or coaxial cables terminated on crocodile clips. Although this may be adequate at low frequencies, it can behave like a badly matched transmission line, causing attenuation distortion and phase distortion. The earth screen of a coaxial cable is not fully effective, and in consequence the cable can pick up local radiated interference. In a normal cable the central conductor is not bonded to the insulation, and flexure of the cable may distort the signal waveform because of changes in the capacitance distribution, or generated noise. The cable capacitance of about 50 pF/m is additive to the c.r.o. input capacitance, and will distort the signal, particularly if the frequency and source impedance are high.

These adverse effects can be mitigated by use of fully screened probes and cables of low and constant capacitance associated with low loss. The probes may be *active* (including an impedance matching unity-gain amplifier) or *passive* (including resistance with adjustable capacitance compensation); attenuators are incorporated in each to reduce both the signal level and the capacitive loading of the source, the compensation being most effective for high-impedance sources. The active probe can be used to match the source to the standard 50 Ω input impedance of a high-frequency c.r.o. (100 MHz and above), an important matter in that a quarter wavelength at 100 MHz is about 75 cm, and severe standing-wave effects may otherwise occur to invalidate the measurements.

11.8.4.2 Earths and grounds

Ground planes (e.g. extended screens) and earth planes (e.g. trunking) constitute low-impedance paths through which interfering signals can circulate, of a wide variety of frequencies and waveforms. The arbitrary connection of screens to earth and ground terminals can have the unwelcome result of increasing noise, due to earth-loop currents. The latter must not enter the input signal lines of either the system or the c.r.o.; and the interference can be minimised by using a single ground/earth connection to serve the whole equipment, if this is possible. The common connection should be made at the zero-potential input signal lead.

11.8.5 Calibration

Operational specifications will include performance statements, typically: voltage accuracy, ±3%; time-base frequency ±3%; Y-amplifier linearity, ±3%, phase shift, 1°; pulse response, $\not>$5% after 2 ns; sweep delay accuracy, ±2%. The calibration facility provided usually consists of a 1 or 2 kHz square waveform with a 3 μs rise time, with both the frequency and voltage amplitude within ±1%.

The practical application of the c.r.o. as a test instrument includes (a) the general diagnostic display of waveforms, and (b) the testing of a product during development or manufacture in relation to its accuracy and quality control. For (a) the built-in calibration facility is adequate; but for (b) it is prudent (or essential) to make regular checks of the actual parameters of the self-calibration facility and of the basic characteristics of the instrument. Such tests would include assessments of the accuracy on alternating voltages, and the ratio and phase angle of the amplifiers and attenuators over the frequency range. The most important and critical test for a c.r.o., which will inevitably be used with modern digital networks, is its response to high frequency square waveforms. The rise time is quite easily measured, but the initial resonant frequency of the 'flat' part of the waveform and its rate of attenuation are important in assessing the actual duration of this part.

The quality and usefulness of these tests is ultimately related to the readability of the display, which would be no better than ±0.3% in a 10 cm trace. The quality of the standard reference devices would, for example, be ±0.1% for a variable-frequency alternating voltage source; hence, there would be a total measurement uncertainty of about 0.4% in the ±1% waveform. It follows that the actual c.r.o. voltage source must be correct to within 0.6% if it is to be used with confidence as a ±1% source. It is a fundamental requirement that the quality of all the reference standards be known from recent linearity and traceability tests.

11.8.6 Applications

The principal application is for diagnostic testing. In addition, the c.r.o. is used to monitor the transfer characteristics of devices and systems, as well as for numerical measurements. Among special uses are to be found the television camera, electron diffraction camera and electron microscope.

Single-deflection measurements The tube functions with a single pair of deflecting plates as a voltmeter or ammeter, with the advantages of being free from damping, unaffected by change of frequency or temperature or over-deflection, and imposing only a minute load on the test circuit. Examples are the monitoring of signals in broadcast and recording studios, modulation checks, indication of motor peak starting currents, thickness measurement, null detection and the measurement of current and voltage in networks of very low power.

Differential measurements Similar or related phenomena are compared by use of both pairs of deflecting plates, the resulting Lissajous figure being observed. In this way phase difference can be measured, frequencies accurately checked against a standard, armature windings tested, modulation indicated by a stationary waveform, and the distortion measured in receivers, amplifiers and electro-acoustic equipment. *Figure 11.16* is a Lissajous figure comparing two frequencies by counting the horizontal and vertical loops, giving the ratio 4/5. The diagram in *Figure 11.17* shows the connections for testing the distortion of an amplifier: the oscillator is set to a known frequency, the two pairs of deflectors connected, respectively, to the input and output (with suitable attenuators). A straightline display indicates absence of distortion, while curvature or looping indicates distortion.

Repetitive time base Using a linear time base enables sustained waveforms to be displayed for machines, transformers and rectifiers, rapid operations to be timed, and surge phenomena to be shown with the aid of a recurrent-surge generator.

Single sweep Non-repetitive transients, such as those arising by lightning or switching, are most readily traced by use of the continuously evacuated tube, but sensitive sealed-off tubes can also be employed. In either case a non-repetitive time base is used consisting of the voltage of a capacitor discharged through a resistor and triggered by a gap. The transient discharge is normally obtained by use of a surge generator for power system plant testing, but other single transients, such as the small voltages of cardiac origin, may also require to be recorded.

Independent bases The c.r.o. is well suited to the display of quantities related by some variable other than time. The current due to an impressed voltage can be displayed on a frequency base as a resonance curve, radio receivers aligned, and B/H loops taken for steel samples. Transistor curve tracers are useful applications capable of giving a complete family of characteristics in a single display.

Modified time bases There are several modified forms of time base. If two voltages of the same amplitude and frequency are displaced in time phase by a quarter-period and applied, respectively, to the X and Y plates as in *Figure 11.18*, they will produce a circular trace. One application, in decade digital logic circuits, involves a Z modulation of the c.r.o. grid. To check that a frequency divider section is working correctly, the period of the circular time base is made 10 times that of the next higher count rate section. With the latter used to modulate, 10 equal bright and dark peripheral sectors should appear on the trace.

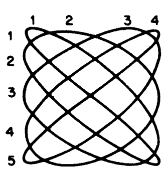

Figure 11.16 Lissajous figure

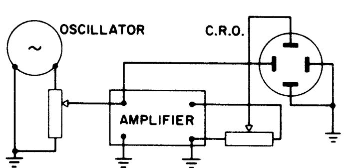

Figure 11.17 Distortion test on amplifier

11.8.7 Digital storage oscilloscopes

Digital storage oscilloscopes (d.s.o.) (*Figure 11.19*) offer advantages over the analogue c.r.o. in some circumstances:

(1) they allow long-term display of a transient quantity;
(2) it is easy to produce a hard copy of the display;
(3) they allow computation and signal processing within the d.s.o.;
(4) they allow easy transfer of data to a computer; and
(5) they use cheap cathode-ray-tube construction because the deflection is not at signal frequency but much lower than the maximum signal frequency.

The d.s.o. has become available for laboratory and industrial use because of the development of relatively cheap, accurate and fast A/D converters.

11.8.8 Digital oscilloscope characteristics

11.8.8.1 Sampling rate

Clearly the sampling rate must be high enough to give a faithful representation of the applied signal. Nyquist's theorem states that a periodic signal must be sampled at more than twice the highest frequency component of the signal. In practice, because of the finite time available, a sample rate somewhat higher than this is necessary. A sample rate of 4 per cycle at oscilloscope bandwidth would be typical. Therefore, for single shot, a digital oscilloscope specification will give a bandwidth of 100 MHz or a sample rate of 400 megasamples/s. If the display is a dot display, then something like 20 samples per cycle is necessary if the eye is to interpret the display easily. Usually some form of interpolation is provided to aid interpretation. The simplest method is to draw straight lines between successive sample parts. Other interpolation methods are used such as sine interpolation, designed for fairly pure sine waves. This interpolation will give erroneous displays for non-sinusoidal inputs.

All this means that great care must be taken in interpreting a display when the input approaches the oscilloscope bandwidth, particularly if it has significant high frequency components such as a square wave. This is also true of course for an analogue c.r.o.

The d.s.o. in repetitive mode can build up the display over a number of sections of the displayed waveform, and so the sampling-rate problems are not so severe. Specifications will give bandwidth, single-shot bandwidth and sampling rate. If single-shot mode is to be used, then care must be taken to

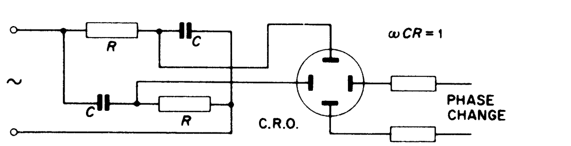

Figure 11.18 Connections for circular time base

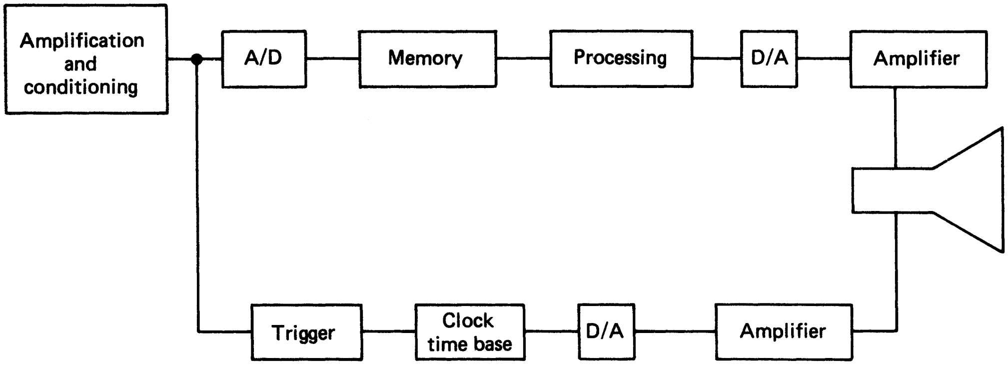

Figure 11.19 The digital storage oscilloscope

look for the adequacy of the single-shot bandwidth or the sampling rate.

For an analogue c.r.o., a rule of thumb relates the rise time of the c.r.o. to its bandwidth rise time (s) × bandwidth (Hz) = 0.35. So, a 100 MHz oscilloscope has a rise time of 3.5 ns. A signal rise time approaching this cannot be faithfully displayed, as the fastest displayed rise time is that of the instrument itself. Usually the fastest rise time displayable is 8 times the c.r.o. rise time. So the fastest signal rise time = 2.8 ÷ bandwidth. For a 100 MHz c.r.o. this would be 28 ns.

A similar restriction occurs for the d.s.o. A fast signal pulse on an expanded time scale is shown in *Figure 11.20*. The worst case giving the slowest rise time is when one of the samples is in the middle of the edge. The time from 10% to 90% of the rise (the rise time) is given by (80/100) × 2 sample periods (= 1.6 sample periods).

So for a 100 megasample/s d.s.o., the sampling interval is 10 ns and the displayed edge, regardless of the rise time of the applied signal, is at worst 16 ns. Again, in order to faithfully display the signal, and not the characteristics of the d.s.o., the fastest signal rise time should be about 8 times this limit, say 10 sample intervals. So, as a rule of thumb:

$$\text{Fastest signal rise time} = 10 \text{ sample intervals} = \frac{10}{\text{d.s.o. sampling rate}}$$

e.g. for 100 megasamples/s d.s.o. the fastest signal rise time is 100 ns. For a signal rise time faster than this it is likely that the performance of the d.s.o. will show itself in the display of the signal.

11.8.8.2 Vertical resolution

Vertical resolution is another important parameter of the d.s.o. An important contribution to this is the number of bits of the A/D converter. The more bits, the greater the resolution. The minimum discernible level is

$$\frac{\text{Input voltage}}{2^n - 1}$$

where n is the number of bits of the A/D.

For example, for an input voltage of 1 V and an 8-bit A/D, the minimum discernible level is

$$\frac{1}{2^8 - 1} = \frac{1}{255}\ \text{V}\ (\simeq 4\ \text{mV})$$

This value is sometimes given as a percentage and in this case would be ±0.2%.

So on a 1-V signal, noise of less than 4 mV will not be displayed. Increasing the number of bits in the A/D converter will help, but internally generated noise limits the advantage achieved. This leads to the concept of effective bits of the d.s.o., which is performance of a hypothetical idealised A/D converter which gives the same vertical resolution of the actual A/D converter with its noise limitation.

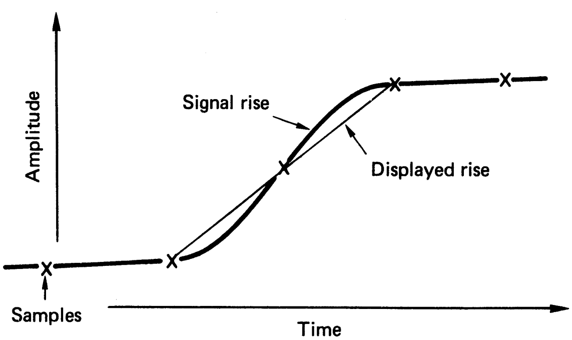

Figure 11.20 The effect of sampling rate on displayed rise time

11.9 Potentiometers and bridges

11.9.1 D.c. potentiometers

D.c. potentiometers may be used to measure a direct voltage by comparison with the known ratio of the e.m.f. of a Weston standard cell or electronic solid state reference device. The methods of achieving the linear ratio between the unknown p.d. and the standard-cell e.m.f. are by successive approximation to the balance condition (a) using groups of matched resistors, (b) using opposed m.m.f.s between matched windings and current-ratio resistors (comparator potentiometer), and (c) pulse-width modulation (ratio of time interval). The d.c. potentiometer is a ratio device of high precision with the uncertainty of the standard voltage source limiting the ultimate accuracy.

11.9.1.1 Weston standard cell

The highest quality saturated mercury/cadmium-mercury cell has an e.m.f. of about 1.018 620 V at 20°C. The net e.m.f./temperature coefficient is about $-40\,\mu V/°C$, arising from the two different coefficients of the limbs, each of the order of $350\,\mu V/°C$. It is, therefore, important to avoid a temperature differential across the cell by mounting it in oil or air within a constant temperature enclosure (e.g. for 20°C ±40 m°C, the uncertainty can be only ±0.4 μV).

Standard cells are stable, the e.m.f. falling by, perhaps, 1 μV per year. For accurate work, *regular* confirmation of the e.m.f. should be made at an accredited calibration laboratory.

The internal resistance is 600–800 Ω. The current drawn from a cell should ideally be zero: in practice it should be limited to a few nanoamperes for a few seconds.

Should a cell be compared with an electronic solid state voltage reference (below), the cell must be presented and removed while the electronic source is switched on.

11.9.1.2 Electronic solid state e.m.f. reference

These devices use the constant voltage property of Zener diodes to deliver highly stable d.c. voltages, usually at 1 V and 10 V levels for direct voltage meter calibrations, and 1.018 61 V to simulate a Weston cell for d.c. potentiometric standardisation. Up to four diodes (each stabilising at about 6.3 V) are used, having been carefully selected for stability, low noise and low temperature/voltage coefficient properties. After further ageing, these temperature-stabilised monolithic devices may be used either separately with a carefully designed operational amplifier, or connected in cascade in specialised networks to achieve the required output p.d. by potential division using highly stable resistors.

It would appear that these devices are excellent transportable standards of voltage (which should be reliable to within 2 μV/year) and they are much more convenient than Weston cells in this respect. Except for the most extreme requirements of accuracy, they could be used as laboratory standards, provided that at least three separate units are available for intercomparison. With more experience and/or minor refinement to these devices, they could shortly replace the Weston cell completely, since at present the Weston cell can be used as a transportable standard to less than 0.5 μV, but only if very great care is exercised.

Typical electronic devices can deliver, say, 5 mA, 10 V from 2 mΩ and 1.0186 V from 2kΩ.

11.9.1.3 Potentiometer principle (using groups of matched resistors)

In *Figure 11.21* AB is a manganin wire of length a little greater than 1 m, through which the current I from a 2 V secondary cell may be varied by means of R. The current is adjusted so that the p.d. between points F and D, 1.0186 m apart, balances the e.m.f. of the Weston standard cell, the balance condition being obtained by use of the switch S and galvanometer G. The volt drop along the wire is now 1 mV per mm length. The switch S may now be set to apply the unknown voltage V_x, and a new balance point H found. Then the length DH in mm represents the unknown voltage in millivolts.

This very simple arrangement is inconvenient in practice. It suffers also from the disadvantage that the precision of the voltage measurement depends upon the precision with which the point of contact of the slider is known. For example, if the length DH is known only to the nearest millimetre, the precision of the voltage measurement would, in general, be less than 1 in 1000. Much greater precision than this is often required, and to obtain it (and to secure a more compact form of apparatus) some elaboration and refinement in the circuit elements is required. Many practical forms of d.c. potentiometer have been devised.

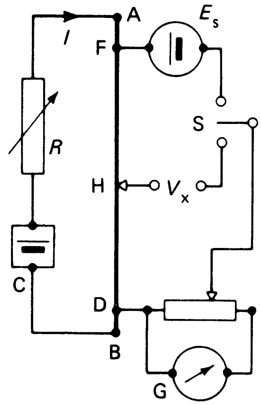

Figure 11.21 Simple potentiometer

11.9.1.4 Practical potentiometer

The potentiometer in *Figure 11.22* is supplied across AB from a 2 V secondary cell through a rheostat R. Resistors R_2, R_3 and R_4 are connected in series between A and B, but the conditions first considered are with R_4 short circuited by the switch at P_1. The resistor R_3 is divided into equal sections with 19 tappings brought out to studs, and there is an additional stud connected to a tapping on R_2.

One terminal of the standard cell is connected to a tapping on the parallel resistor R_1, and the other terminal may be connected (when the switch S is in the left-hand position

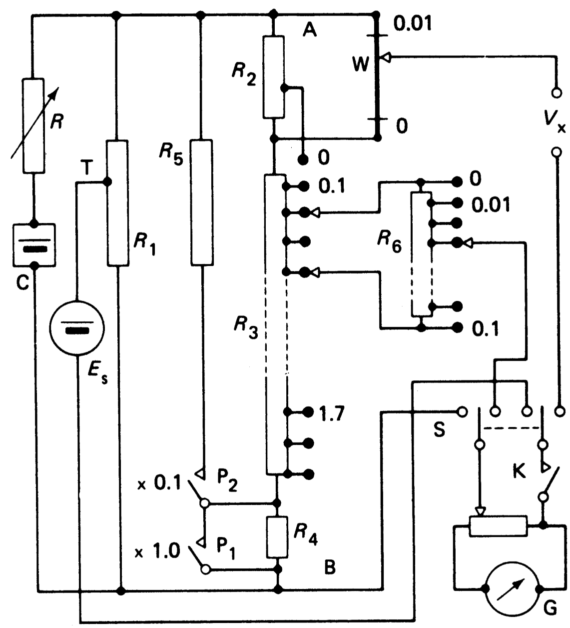

Figure 11.22 Practical potentiometer

and the key K is depressed) to the point B through galvanometer G. A balance may be obtained with the standard cell by adjusting the rheostat R, and the resistance values are such that the p.d. between successive studs on R_3 is then precisely 0.1 V, and the voltage between A and B a little over 1.8 V. In parallel with the shunting resistor R_2 is the slide wire W. The tapping on R_2 enables a true zero, and small negative readings, to be obtained on W. By suitable choice of the resistance of R_2 and W the voltage range covered by the whole slide wire is a little over 0.01 V.

R_6 is a tapped resistor with 10 equal sections and 11 studs. The total resistance of R_6 is made equal to the resistance of two sections of R_3, and R_6 always spans two sections by means of sliding contacts on the R_3 switch as shown. The total resistance of R_6 and two sections of R_3 in combination is therefore always equal to a single section of R_3 and the potential difference between successive tappings on R_6 is then 0.01 V.

The unknown voltage V_x is applied to terminals connected through the galvanometer switch S to the sliding contact on W and to the tapping on the R_6 switch. If the slide-wire dial has 100 divisions, each division represents 0.0001 V. Thus, for the switches in the positions shown, V_x might be 0.2376 V. The maximum V_x readable, assuming that there are some graduations provided beyond the 0.01 position on the slider dial, is $1.7 + 0.1 + 0.0100 = 1.8100$ V.

The resistors R_4 and R_5 are introduced into the circuit by opening switch P_1 and closing P_2; this reduces the p.d.s in the potentiometer circuits to one-tenth of those marked. One division on the slider dial would then correspond with 0.000 01 V or 10 μV.

Medium- to high-precision potentiometers include a 'standard cell setter' so that the actual e.m.f. of the standard cell can be used for standardisation and, subsequently, for rechecking the standardisation during use *without* altering the measuring dials: this is more convenient, although less accurate than standardising against the dials. The contact T could be the slider of a 10-turn voltage divider with an indicated range, e.g. of 1.018 250 to 1.018 750 V in 1-μV steps.

D.c. potentiometers of very high precision have a ratio non-linearity of ±5 parts in 10^6. A stable current of 50 mA must be provided by an electronic controller, and a high-gain low-noise detector is necessary: this may be a photocell galvanometer amplifier with a display sensitivity of 5000 mm/μV in a 10-Ω circuit.

11.9.1.5 Comparator potentiometer (using m.m.f. balance)

The comparator potentiometer is one of a group of precision bridges employing mutual-coupled inductive ratio arms. The bridges exploit the high m.m.f. linearity and discrimination of a constant current in a variable number of turns.

A simplified network is shown in *Figure 11.23*. The measuring loop M is magnetically coupled to the balancing loop B by the turns N_m and N_b, respectively, on a magnetic core, which also carries a feedback winding with N_f turns. The feedback network includes an a.c. modulated sensor which actuates an a.c./d.c. electronic control network C_b to change the direct current I_b until a zero m.m.f. condition is achieved in the core.

11.9.1.6 Operation and standardisation

If I_m and N_b are constant, then for an automatically maintained m.m.f. balance $N_m I_m = N_b I_b$ it follows that linear changes in N_m result in corresponding linear changes in I_b. As N_m has in effect a discrimination and linearity of about 1 in 10^7, these properties are imposed also on I_b. To convert this linear current scale to voltage across R_b, the instrument has to be standardised against a standard cell. The cell, of e.m.f. E_r, is connected (through S1) in opposition to the p.d. across R_m, and balance achieved by use of the adjustable direct-current controller C_m, which is then left to maintain this current level.

The equivalent cell e.m.f. $I_m R_m$ is now connected (through S2) in opposition to $I_b R_b$, and N_m turns are selected to be numerically equal to E_r. The m.m.f is balanced automatically by the feedback control system and by a small adjustment to N_b so that $I_m R_m = I_b R_b = E_r$, represented numerically by N_m. With switch S set to S_0 the unknown p.d. V_x is presented through S_x and galvanometer G_1 in opposition to $I_b R_b$ and balance achieved by manual adjustment of N_m. Then N_m is numerically the unknown voltage.

Instruments of this kind are probably the most linear d.c. potentiometers available. They incorporate additional windings that enable the linearity to be checked by the operator. While the actual values of R_m and R_b are not of first importance, it is vital that they should be constant. Resistors are incorporated for subdivision of I_m for lower decades of N_m in order to simulate the 10^7 turns implied by a discrimination of 1 in 10^7, but they do not contribute unduly to the overall non-linearity, which is about $\pm 0.3 \times 10^{-6}$ of full scale. The system uncertainty must include that of the standard cell, at least ±1–2 μV.

11.9.1.7 Pulse width modulation potentiometer[12]

This method subdivides a known stable d.c. voltage in terms of a time-period ratio. If a standard voltage source E_r is repeatedly connected to a low-pass filter through a very-high-speed semiconductor switch for a time t_1, and then disconnected for an additional time t_2, the high-frequency rectangular, chopped waveforms will, after smoothing, give an output $E_0 = E_r \cdot t_1/(t_1 + t_2)$ with $t_1 + t_2 = T$, a constant period. The time *ratio* can be precisely set using a variable time-interval counter for t_1 and a fixed-period counter for T, with each counter being driven from a common highly stable crystal-controlled oscillator. An unknown voltage V_x, in opposition to E_0 through a galvanometer, can be measured by variation of E_0 (through the voltage-scaled setting of the t_1 counter) until galvanometer zero balance is achieved. The e.m.f. E_r of the solid state source of steady voltage must be known accurately and the source must be capable of delivering small current surges into the low-pass filter without any significant change in voltage. Conversely, V_x and E_r could be interchanged when $V_x > E_r$.

One d.c. variable-voltage source uses the accurately determined voltage as input to constant gain d.c. amplifiers (designed for decade steps) to give a wide-range d.c. voltage standard source with ±50 p.p.m. uncertainty (3-month stability) up to 1.2 kV with a 25 → 10 mA (at 1 kV) current capability and 0.5 → 3.0 s settling times.

11.9.1.8 Applications

The d.c. potentiometer may be used to determine temperature by measuring the thermo-e.m.f.s in calibrated thermocouples. In conjunction with standard four-terminal resistors and shunts it is applied to the calibration of voltmeters, ammeters and wattmeters, and for the comparison of resistors. In *Figure 11.24* the values V_x, I_x, R_x and P_x are unknowns, and E_p is measured by the potentiometer. Resistors $R_1, \ldots, R_7$ have known values. At (*a*) a voltage divider is used to measure V_x, and at (*b*) a shunt is employed to measure I_x. At (*c*) R_x is compared with R_4 by means of

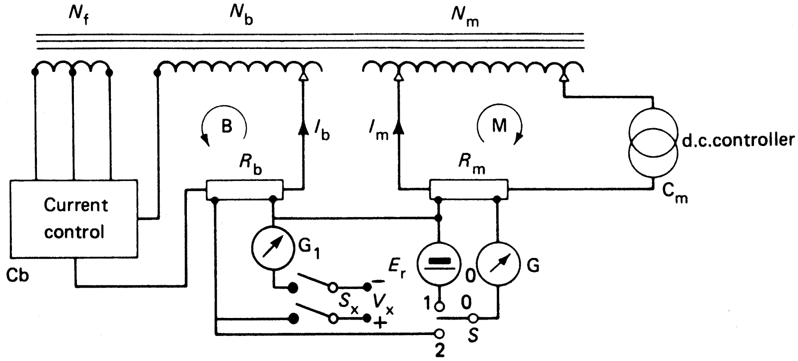

Figure 11.23 D.c. comparator potentiometer

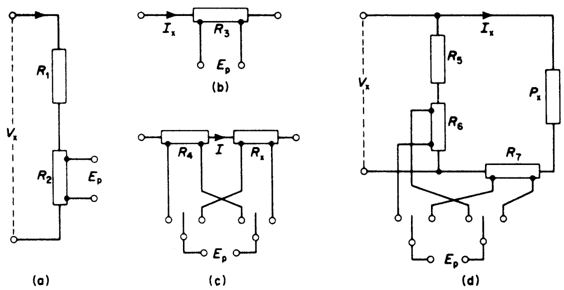

Figure 11.24 D.c. potentiometer measurement of voltage current, resistance and power

their respective p.d.s. The network (*d*) enables the power $P_x = V_x I_x$ to be determined from V_x and I_x. The resistance of four-terminal resistors is known between the potential (inner) terminals, and *not* the current (outer) terminals, which minimises errors due to non-uniform current densities and contact resistances.

The National Physical Laboratory (NPL) designed 'Wilkins' resistor is a high-stability, four-terminal, a.c./d.c. resistor available in various decade and other values (Tinsley Co., London) which has an exceptionally low time constant for accurate a.c./d.c. comparator networks, etc.

11.9.2 A.c. potentiometers

The a.c. potentiometer may be used to measure the magnitude of an alternating voltage and its phase relative to a datum waveform. The measurements are normally restricted to sinusoidal waveforms, as the presence of harmonics makes a null balance impossible; and operation is usually confined to a single frequency. One recent instrument (Yorke) is capable of measurements up to audiofrequency, with an uncertainty approaching ±0.05%. The general principles of the Larsen and Gall potentiometers are set out below.

11.9.2.1 Larsen potentiometer

The primary winding of a variable mutual inductor is connected in series with a tapped resistor to a source of voltage V. The unknown voltage V_x is normally derived from the same source, because identity of frequency is essential. The supply current I is controlled by resistor R_1 and indicated on an ammeter. The inductor secondary e.m.f. in series with a tapped fraction of the volt drop across R is balanced against the unknown V_x (*Figure 11.25*) by means of the vibration galvanometer G, the sensitivity of which can be varied by resistor R_2. If M and R are the mutual inductance and resistance values at balance, then

$$V_x = I\sqrt{|R^2 + \omega^2 M^2|} \text{ and } \phi = \arctan\,(\omega M/R)$$

where ϕ is the phase angle between V_x and I. In practice it may be necessary to reverse either or both of the component voltages to secure balance, and reversing switches are incorporated for this purpose. The mutual inductor should be astatic, to avoid pick-up errors.

11.9.2.2 Gall potentiometer

The Gall potentiometer (*Figure 11.26*) provides two quadrature voltage components, summed and balanced against the unknown V_x. The supply voltage V is applied to the primaries of isolating transformers T_1 and T_2. The secondary of T_1 supplies a current, approximately in phase with V, to R_1 which consists of a tapped resistor and slide wire in series. The current is adjusted by resistor R_3 and read on an ammeter A, and passes through the primary of a fixed mutual inductor M. Any desired value of in-phase voltage (within the available range) is obtained from tappings on R_1. The primary of transformer T_2 has a series resistor R_5 and a variable capacitor C, and may be so adjusted that the current in R_2 is in quadrature with that in R_1. The exact phase angle, and the calibration of the quadrature circuit, are achieved by balancing the secondary e.m.f. of the mutual inductor M against a fraction of the p.d. across R_2. Thus, variable voltages, phase and quadrature, can be obtained from R_1 and R_2, respectively. These, in series, are balanced against the unknown V_x through the vibration

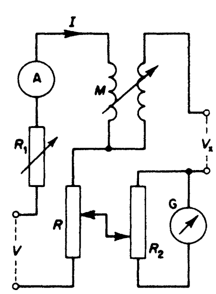

Figure 11.25 Larsen a.c. potentiometer

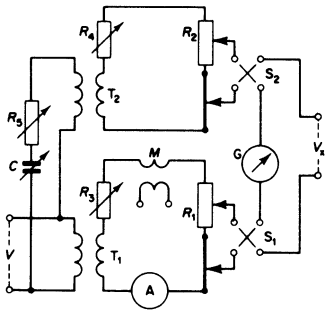

Figure 11.26 Gall coordinate a.c. potentiometer

galvanometer G. The reversing switches S_1 and S_2 facilitate balance and provide for a phase angle over a full 360°.

11.9.3 D.c. bridge networks

A bridge network has as its distinctive feature a 'bridge' connection between two nodes, with a null detector to sense the balance condition of zero p.d. between the nodes.

11.9.3.1 Wheatstone bridge

The Wheatstone bridge is an arrangement in very wide use for the determination of one unknown resistance in terms of three known resistances. The network is shown in *Figure 11.27*, where R_1, R_2, R_3 and R_4 are resistors connected at the nodes a and b through a reversing switch S to a d.c. supply. The galvanometer G, with a shunting resistor to control its sensitivity, and the key K are connected to the nodes c and d as shown. R_1 and R_3 may be set at known fixed values. R_2 is a variable resistor and R_4 is the unknown resistance to be measured. The bridge is balanced by adjusting the value of R_2 until the deflection of the galvanometer, set for maximum sensitivity, is brought to zero. The condition for balance is easily seen to be $R_4/R_3 = R_2/R_1$ and is independent of the voltage applied to the bridge: whence the unknown is $R_4 = (R_3/R_1)R_2$. The bridge may be built up as a single unit with, for example, R_1 and R_3 each able to be set at one of the resistance values 1, 10, 100 or 1000 Ω. The bridge ratio R_3/R_1 may therefore have a range of values from 1000/1 to 1/1000. In this case R_2 might conveniently be adjustable, by means of dial switches, in steps of 1 Ω from 1 to 11 110 Ω.

The bridge resistors are wound from manganin wire, since manganin is an alloy with a low temperature coefficient of resistance and a low thermoelectric e.m.f. to copper. Any thermoelectric e.m.f.s in the bridge connections which did not on average cancel themselves out could give a false result. Errors from this cause are eliminated by taking a balance for each position of the changeover switch S and using the mean of two observed values of R_2.

High-precision Wheatstone bridges are capable of measuring resistances between 1 Ω and 100 MΩ with uncertainties which vary between 5 and 100 in 10^6 of the reading over this range of values. Modern resistors are very stable devices and they should not change in value by more than a few parts in 10^6 per year; however, it is prudent to check the linearity of resistive bridges regularly, particularly if they are being used to the limit of the original specification. Some bridges have trimming facilities available with the principal resistors, although corrections can always be applied which although tedious, can avoid imposing new resistance drifts on the bridge as a consequence of changing *stable* resistors.

Various forms of high-quality bridges are used for measuring the changes which occur in resistive transducers. The Smith and Mueller bridges, used for measurements with precision platinum resistance thermometers, can yield 0.001°C discrimination in readings. For strain gauge transducers there are portable bridges that can detect changes of 0.05% or less.

11.9.3.2 Kelvin double bridge

The Kelvin double bridge is an adaptation of the Wheatstone bridge which may be used for the accurate measurement of very low resistances, such as four-terminal shunts or short lengths of cable. In the network (*Figure 11.28*) R_x is the unknown and R_s is a standard of approximately the same ohmic value. The two are connected in series in a heavy-current circuit (e.g. 50 A for $R_s = 10$ mΩ, giving 0.5 V drop and adequate sensitivity). Suitable values for R_1, R_2, R_3 and R_4 are in each case a few hundred ohms. To use the bridge, R_1 and R_2 are made equal; then R_3 and R_4 are adjusted (keeping $R_3 = R_4$) until balance is achieved. Then the balance condition is $R_x/R_s = R_4/R_2 = R_3/R_1$. Provided that it is short and of adequate cross-section, the connection between R_x and R_s does not affect the measurement.

Good-quality commercial bridges claim an uncertainty of about ±0.05% for unknowns between 1 Ω and a few micro-ohms. Near the latter level a small constant resistance becomes a limitation. As $R_1, \ldots, R_4$ constitute a Wheatstone bridge network, a Kelvin bridge is often extended to measure two-terminal resistances up to 1 MΩ with an uncertainty of ±0.02%.

11.9.3.3 Digital d.c. low resistance instruments

Digital milliohmmeter/micro-ohmmeter instruments have built-in d.c. supplies; they are auto-ranging and make four-terminal measurements by internal comparator-ratio techniques to give, e.g. $4\frac{1}{2}$-digit displays in the case of the Tinsley 5878 micro-ohmmeter with 0.1 μΩ best resolution and uncertainty ±(0.1% → 0.3%) of reading. Also, a IEEE 488 bus attachment is available for use in automated test systems. Thermal e.m.f. balance is incorporated as well as lead compensation.

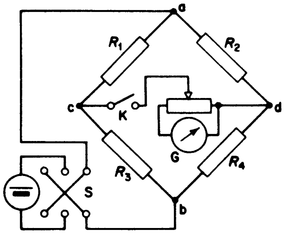

Figure 11.27 Wheatstone bridge

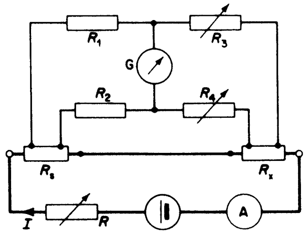

Figure 11.28 Kelvin double bridge

11.9.3.4 Comparator resistance bridge

The comparator resistance bridge is a variant of the comparator potentiometer (*Figure 11.23*). In this case the network connections and adjustments are such that, at balance,

$$I_m R_m = I_b R_b \text{ and } I_m N_m = I_b N_b, \text{ whence } R_m = N_m(R_b/N_b)$$

where R_b is the known standard four-terminal resistor and N_b has been adjusted to make R_b/N_b an exact decade value. The unknown four-terminal resistance R_m is found in terms of a number of turns N_m and a decade factor.

The two resistors can be compared at different current levels, so that, for example, the rated current can be used with the test resistor, while the standard resistor can be used at a current well below the level at which self-heating effects would change its resistance value. The actual comparison uncertainty with this class of bridge is less than 1 in 10^6, to which must be added the basic uncertainty in R_b, probably 2 in 10^6 if it is a class 'S' (or 'Wilkins') standard four-terminal resistor *known* to be stable.

11.9.4 A.c. bridge networks

A.c. network parameters (impedance, admittance, phase angle, loss angle, etc.) can be measured in the following ways:

(1) Single-purpose or multipurpose a.c. bridge networks, with four or six arms or with inductive ratio arms.
(2) Analogue and digital networks developed for special applications and consisting of frequency-selective or resonant or filter networks; they are associated usually with measurements at radiofrequencies and high audiofrequencies.
(3) Digital multimeter instruments for measuring the modulus of the unknown parameter (but not other qualities).

The treatment in this subsection is confined largely to (1).

11.9.4.1 Balance conditions

In the four-arm bridge of *Figure 11.29*, the arms comprise *impedance operators* Z_1, Z_2, Z_s and Z_x. The network is supplied at nodes a and b from a signal generator, and a sensitive null detector is connected between nodes c and d. If Z_1 and Z_2 (fixed standards) and Z_s (variable standard) are so adjusted that the p.d. between c and d is zero, then the unknown impedance Z_x depends on whether Z_1 and Z_2 are adjacent as in *Figure 11.29(a)*, or in opposite arms as in *Figure 11.29(b)*; in either case the balance condition is that the product of pairs of opposite arms is equal, giving

(*a*) *ratio bridge*: $Z_x = Z_s(Z_1/Z_2)$;
(*b*) *product bridge*: $Z_x = Y_s(Z_1/Z_2)$

where $Y_s = 1/Z_s$. These are *complex* expressions, so that the equality involves both magnitude and phase angle; alternatively, both phase ('real') and quadrature ('imaginary') components. Thus, equation (*a*) above can be written in terms of *magnitudes* in either of the following ways:

$$Z_x \angle \phi_x = Z_s(Z_1/Z_2)\angle(\phi_s + \phi_1 - \phi_2)$$
$$R_x + jX_x = (R_s + jX_s)(R_1 + jX_1)/(R_2 + jX_2)$$

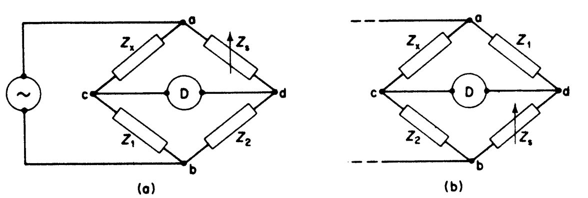

Figure 11.29 Basic four-arm a.c. bridge. (a) Ratio bridge; (b) product bridge

The bridge arms will include intentional or stray inductance and capacitance, and the corresponding reactances are functions of frequency. Most practical bridges are so chosen that only two or three final adjustments are necessary to obtain the optimum null condition of the detector.

11.9.4.2 Signal source

The bridge supply can be from a screened power-frequency instrument transformer, but more probably from a fixed- or variable-frequency signal generator (see Section 11.7.4). A voltage up to 30 V may be required, but a few volts is normally adequate, especially for radiofrequency bridges. The maximum power of a signal generator is of the order of 1000 mW, so that to avoid overload and distortion it is important that the appropriate voltage be set and a reasonable match be obtained between the generator and the effective input impedance of the bridge network.

11.9.4.3 Detector

A moving-coil vibration galvanometer may be used at discrete frequencies over the range of 10 Hz–1.2 kHz. The galvanometer has a taut suspension mechanically tuned to the operating frequency. The method is sensitive, and effectively filters harmonics, but it is often more convenient to employ the electronic detectors included in most commercial bridges. These include tuned-resonance amplifier networks or broad band high-gain electronic amplifiers; both give a rectified out-of-balance display on a d.c. instrument, with 'magic eye' or c.r.o. as possible alternatives. In some portable instruments it was the practice to use head-telephone sets, which are highly sensitive in the audio range of 0.8–1.0 kHz.

11.9.4.4 Bridge-arm components

Precision a.c. resistance boxes can be designed with very low time constants. They are available with switched or plugged groups of ten resistors in decade values between 0.1 Ω and 1 MΩ. Careful construction and screening enable some of these devices to be used at frequencies up to 1.2 MHz before the residual reactance becomes significant. An overall uncertainty of ±0.05% is possible, but 0.1–1.0% is more conventional. (See also Section 11.9.1.8.)

Multi-dial switched mica capacitors are available with good stability and low loss tangent. Single- and double-screen capacitors have respectively, three and four terminals, and care is needed to avoid unwanted stray capacitance, the screens being maintained at the proper potentials with respect to the equipment and the environment.

Continuously variable calibrated air-dielectric capacitors with ranges between 20 and 1200 pF are used for small adjustment, and ranges of fixed or variable mutual inductors are available. For high-voltage bridges the standard capacitors are of fixed value, and designed with air or compressed gas as dielectric for 300 kV and above.

11.9.4.5 Low-frequency and audiofrequency bridges

A selection is given of the more important bridges, many of which form the basis of commercial instruments; for example, the inductance bridges of Maxwell and Hay, the modified de Sauty bridges for capacitance, the Schering bridge for capacitance and loss tangent at low and high voltages, and the Wien bridge for frequency and a.c. resistance. These bridges are commonly made to give results with a ±1% uncertainty, but can be designed for ±0.1%.

All bridges should be checked occasionally to show that the reference standard components have not changed and that the range and ratio division has not drifted; also, if the bridge is frequency sensitive, the frequency of the oscillator and the drift rate should be investigated. Naturally, superior standard components must be available which have recent 'traceability' certification, in addition to accurately known decade-ratio measurement techniques. Where bridge measurements are implicit in contract specifications, it may be prudent or necessary to justify the claims by referring the instrument to a calibration service standards laboratory.

11.9.4.6 Mutual inductance

Figure 11.30(a) shows the simple Felici–Campbell bridge for the measurement of an unknown mutual inductance M_x in terms of a variable standard M. For balance on the null detector D, then $M_x = M$, on the assumption that the inductors are perfect, i.e. that the secondary e.m.f. is in phase quadrature with the primary current. Each, however, has a small in-phase component that can be represented by an equivalent resistance σ. The modification in *Figure 11.30(b)* shows Hartshorn's arrangement, which enables the impurities to be measured: at balance by adjustment of M and r, the conditions are $M_x = M$ and $\sigma_x = r + \sigma$.

11.9.4.7 Inductance

Inductance may be measured by Wien's modification of the Maxwell bridge (*Figure 11.31(a)*). If the unknown inductor has an inductance L_x and an equivalent series resistance R_x, balance gives

$$L_x = R_2 R_3 C_1; \quad R_x = R_2 R_3 / R_1; \quad Q_x = \omega L_x / R_x = \omega C_1 R_1$$

The advantage is that inductance is measured in terms of a high-quality and almost loss-free capacitor. The bridge can measure a wide range of inductance values with Q factors less than 10. High Q factors require excessively large values of R_1, and by creating the same effective phase characteristic by means of a low-valued resistor R_1 in series with C_1, the Hay bridge of *Figure 11.31(b)* is obtained. The balance conditions for it are

$$Q_x = 1/\omega C_1 R_1; \quad R_x = R_2 R_3 / R_1 (1 + Q_x^2);$$
$$L_x = R_2 R_3 C_1 / (1 + 1/Q_x^2)$$

A disadvantage is the need to measure the frequency.

A commercial instrument using the Maxwell and Hay bridges has built-in supplies at frequencies of 50 Hz, 1 kHz and 10 kHz, and an inductance range from 0.3 µH to 21 kH with an accuracy within ±1%. The loss resistance is known to about ±5%, and it is possible to introduce a d.c. bias.

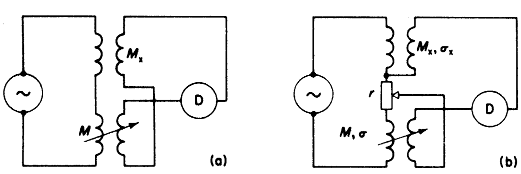

Figure 11.30 Mutual inductance bridges. (a) Felici–Campbell; (b) Hartshorn

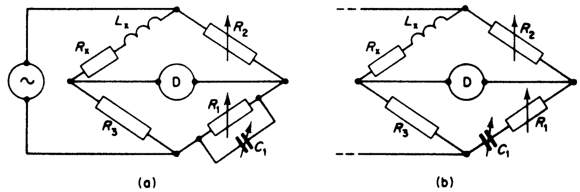

Figure 11.31 Inductance bridges. (a) Maxwell–Wein; (b) Hay

11.9.4.8 Capacitance at low voltage

Relatively pure capacitors can be measured by the de Sauty bridge (*Figure 11.32(a)*), the balance condition for which is $C_x = C_s(R_4/R_3)$. Imperfect capacitors can be compared by the arrangement shown in *Figure 11.22(b)*, in which the imperfection is represented by a series resistance r. To obtain balance, adjustment of all four resistors R is necessary:

$$C_x = C_s(R_4/R_3) = C_s[(R_2 + r_s)/(R_1 + r_x)]$$

For the loss tangent,

$$\tan \delta_x = \tan \delta_s + \omega C_s [R_2 - (R_1 R_4 / R_3)]$$

Portable commercial capacitance testers, with battery-driven 1 kHz oscillators and rectified d.c. or headphone detectors, use a modified de Sauty bridge network so that loss tangent can be estimated and the capacitance measured to an uncertainty of about ±0.25%.

11.9.4.9 Capacitance at high voltage

Dielectric tests at high voltages, particularly of the loss tangent, can be made with the Schering bridge (*Figure 11.33*). The bridge is in wide use for precision measurements on solid and liquid dielectrics, and for insulation testing of cables, high-voltage machine windings and capacitor bushings. The capacitance of the high-voltage standard capacitor C_s is calculable, and with an air or gas dielectric can be assumed to have zero loss. A typical standard capacitor for 150 kV is shown in *Figure 11.34*; for higher voltages the clearances necessary for avoiding corona and breakdown are large, but the dimensions can be reduced by a construction in which the dielectric gas is under pressure.

The test capacitor in *Figure 11.33* is represented by C_x and a series loss resistance r_x. R_3 is a variable resistor, and the fourth arm comprises a variable low-voltage capacitor C_4 in parallel with a resistor R_4 (which may be fixed). Then, for balance

$$C_x = C_s(R_4/R_3) \quad \text{and} \quad \tan \delta_x = \omega C_4 R_4$$

The loss tangent is a valuable and sensitive indication of the quality of the test insulation. If periodical tests reveal a gradual increase in the loss tangent, then deterioration is

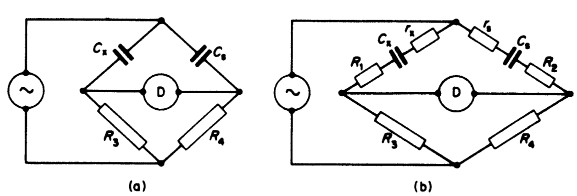

Figure 11.32 Capacitance bridges. (a) De Sauty; (b) modified de Sauty

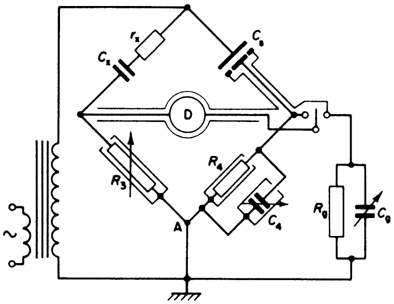

Figure 11.33 Schering bridge

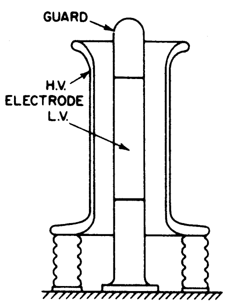

Figure 11.34 Air capacitor

occurring, the loss power is increasing and breakdown in service is probable.

To avoid electric field coupling between bridge components—in particular, between the high-voltage electrodes and the connecting leads—that would affect accuracy, components R_3, R_4 and C_4 are enclosed in metal screens connected to the earthed point A. Again, to avoid errors due to intercapacitive currents between the centre low-voltage electrode and the guard electrode of C_s, these electrodes should be brought to the same potential by aid of the auxiliary branches C_g and R_g. Balance is achieved with the detector switched to this combination, in addition to the main balance. If the test capacitor C_x is a cable with an earthed sheath, or a capacitor bushing on site and not readily insulated from earth, then the bridge node A must be isolated: careful screening is now even more necessary. *Figure 11.35* shows the arrangement for a portable Schering bridge equipment for testing an earthed bushing. It will be seen that an earthed screen isolates the bridge from the transformer primary, and a second screen connected to point A keeps capacitive currents from the high-voltage connections out of the bridge arms.

Discharge bridge This has been developed to detect the onset of void breakdown in cables and bushings. The network (*Figure 11.36*) can be used to test an unearthed component—for example, the bushing C_x; the arrangement resembles that of the Schering bridge. The stray capacitances of the guards and screening of C_s act as the capacitance C_4, which, however, at the discharge frequencies concerned, is greater than that required for balance. Hence, it is necessary to provide the variable capacitor C_3. The bridge output is fed to a rectifier milliammeter D through a filter and amplifier designed to pass, e.g. the band 10–20 kHz, so that the indication is a measure of the discharge current. On the supply side of the bridge is connected some

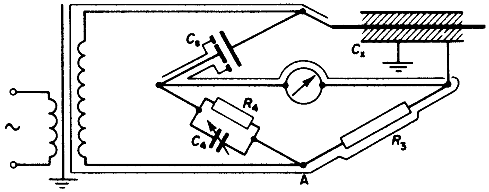

Figure 11.35 Schering bridge and earthed bushing

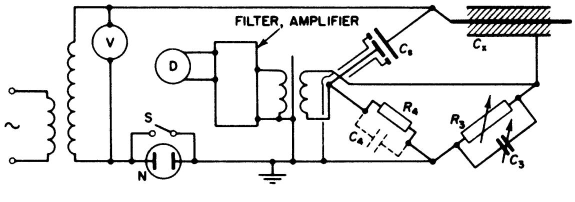

Figure 11.36 Discharge bridge

artificial source of discharge voltage, such as the neon tube N, which serves initially to balance the detector D.

11.9.4.10 Wagner earth

Capacitance between bridge arms and earth in the Schering bridge causes disturbing currents to flow so that false balance conditions may result. If, as in *Figure 11.37*, the bridge supply is not earthed but the nodes c and d are at earth potential, then the detector is also at earth potential and no stray capacitance currents can flow in it. Further, all branch capacitances to earth are stabilised. The Wagner earth method secures this condition by the use of the additional impedances Z_5 and Z_6, the junction of which is solidly earthed. Z_5 and Z_6 must be of the same type as either Z_1 and Z_2, or Z_3 and Z_4. If balance is achieved for both positions of the switch, then nodes c and d must have the same potential as the earthed junction, which is zero. Stray capacitance at a and b is innocuous as it merely 'loads' the supply.

The Wagner earth is usually applied to commercial Schering bridges, but the technique is applicable to any bridge and can be particularly helpful at higher frequencies.

11.9.4.11 Inductive ratio-arm bridge

The principle is explained by reference to the simplified diagram in *Figure 11.38(a)*, the essential features being the coupled windings of, respectively, N_x and N_s turns. If the voltage per turn is v, then the two windings have the voltages $V_x = N_x v$ and $V_s = N_s v$, respectively. Let $V_x = V_s$; then the current I_x through the unknown capacitive admittance Y_x can be made equal to the current I_s by variation of the parallel standards (conductance G_s and capacitance C_s),

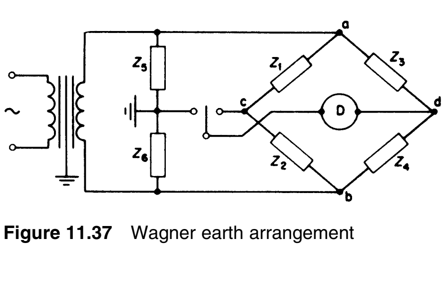

Figure 11.37 Wagner earth arrangement

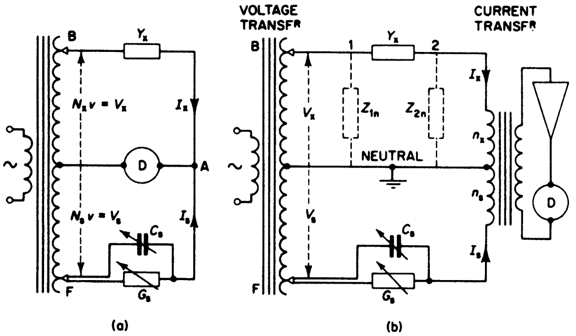

Figure 11.38 Inductive ratio-arm bridge. (a) Basic principle; (b) with isolated detector

until the detector shows a null reading. For this condition the standard admittances represent the *parallel* equivalent network values of Y_x. If Y_x is inductive, the C_s connection at F is transferred to B to avoid the use of standard inductors, which are very inferior to standard capacitors.

The voltage linearity of the input transformer is of fundamental importance to this class of bridge. With modern techniques and magnetic core materials, it is possible to wind toroidal cores to produce an extremely uniform reactive distribution (leakage inductive reactance and turn-to-turn capacitance) with a resultant non-linearity of about 1 in 10^7 in the magnitude of the mutually induced e.m.f. per turn.

In one well-known bridge (Wayne Kerr) the primary winding of a current transformer is inserted at A and the detector removed to the secondary winding as in *Figure 11.38*(*b*). The two parts of the primary winding (n_s and n_x turns) are wound in opposite senses and the detector can indicate the zero m.m.f. condition in the windings.

In general, N_s differs from N_x and n_s from n_x, so that $I_x n_x = I_s n_s$ for balance; further, the primary of the current transformer is at the ground (or earth) potential of the neutral if the trivial resistance drop of the windings is ignored, so that $I_x = (N_x v) Y_x$ and $I_s = (N_s v) Y_s$. Combining these results gives

$$Y_x = Y_s(N_s/N_x)(n_s/n_x)$$

The double turns ratio product can be used to permit *one* accurate standard resistor or capacitor to be switched to any N_s turns of value 1–10 to simulate *ten* accurate standard components by the precise selection of turns. Moreover, wide-range multiplication of the simulated standard is achieved by decade tappings on N_x and n_x.

An important advantage with this class of bridge is the possibility of connecting the neutral to earth; then small leakage currents from Y_x to earth are almost completely eliminated, since the current transformer is at earth potential and any *small* leakage currents from the high-voltage side of Y_x merely cause a slight *uniform* reduction in v but does not alter the turns ratio. Hence, determinations such as the following are possible:

(1) One-port low-capacitance measurement in the presence of a large shunt capacitance due to connecting leads.
(2) One-port *in situ* impedance measurement, the effect of the remainder of the network (shown dotted in *Figure 11.38*) is neutralised.
(3) Three-terminal impedance transfer functions for correctly terminated networks.

The results of measurements appear in *parallel admittance* form, and may require elementary computation to render them into *series impedance* form, with due regard to frequency. With medium to high values of impedance, resistance and capacitance are unaffected, but for inductance measurement (involving a 'resonance' balance) a reciprocal ω^2 correction is needed. For low-impedance networks, resistance and inductance are unchanged but capacitance readings require correction.

11.9.4.12 Characteristics

Universal bridges of the inductive ratio-arm type measure over a wide range of values, at an angular frequency of, e.g., 10 krad/s from a built-in oscillator, or at other frequencies up to 10 MHz using external signal generators and tuned electronic detectors. The measurement uncertainty can be within ±0.1–1.0%, provided that the standard component values are trimmed against superior standards and that the actual frequency is measured when frequency sensitive parameters are measured.

The Wayne Kerr precision bridge operates at 1591.55 Hz ($\omega = 10$ krad/s) and the six-figure display has a minimum uncertainty of 0.01%. The instrument uses an electronic null-seeking process to assist the *automatic* balancing procedure. With external source the bridge has a frequency range of 0.2–20 kHz with manual balancing.

Admittance bridges are commercially available to operate up to 100 MHz. At this frequency lumped parameter measurements are being displaced by distributed-parameter network characteristics.

11.9.4.13 Current comparator bridge

The a.c. current comparator instrument uses current transformers only in the load network. It is based on the principle, of m.m.f. (i.e. ampere-turn) balance on an ideal magnetic circuit, in which the current ratio is the inverse turns ratio.

The two basic current coils, P and S in *Figure 11.39*, are wound in opposing senses on a common magnetic core with a high initial permeability. The null m.m.f. condition is sensed by winding M and manual selection of the turns ratio N_p/N_s. The ideal balance condition can be closely approached in practice by shielding the detector from magnetic leakage by a laminated magnetic screen and by a compensation winding W_c. The shield must not form a short-circuited turn; but it does aid magnetic energy transfer between windings P and S. This is advantageous in current-transformer calibration, as it enables an external burden B to be supported under balance conditions. Any change in the capacitive error due to B can be neutralised by W_c connected on one side of the separate centre-tapped transformer W_b. By careful design the capacitance distribution is uniform.

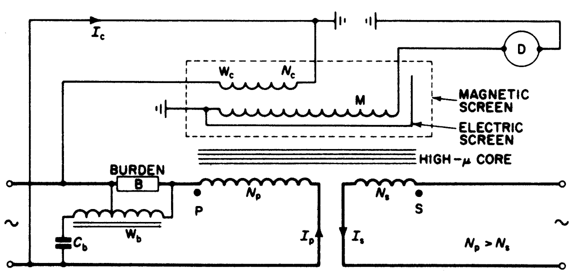

Figure 11.39 Current comparator with capacitive and burden compensation

The internal p.d. of the major winding can be neutralised by a parallel compensation winding W_c wound within the magnetic shield, so reducing capacitive currents, as it can be held at earth or reference ground potential.

The commercial range of comparators include d.c. potentiometers, direct voltage and current ratios, d.c. and a.c. four-terminal resistance comparison, voltage- and current-transformer errors, and high-voltage capacitor comparison. All the instruments derived from d.c. and a.c. comparators possess very linear properties, excellent discrimination and good stability. When certified they can be used as comparative reference standards.

11.9.4.14 H.V. capacitance comparator bridge

Standard capacitors used with high-voltage bridges (such as the Schering and comparator types) use compressed gas construction up to a maximum of 1000 pF, although 100 pF is more common. Measurement of such capacitors is based on the principle that a stable alternating voltage to two capacitors produces a current ratio equal to the capacitance ratio. The variable turns ratio of the comparator bridge can be used to balance an unknown capacitor up to 1000 times as large as the standard. By cascading, the range can be extended still further.

11.9.4.15 Substitution

In any active network the equivalent parameters of a component may be measured if variable reference standards can be either substituted for the component or connected in parallel with it—provided always that the standards can be adjusted so that no final change occurs in the voltage, current, phase or harmonic content of the waveforms in the network. The substitution principle can be applied to any type of bridge, or other class of network; resonant networks often provide the best discrimination between the two conditions. It is important that the stray capacitance, inductance and resistance paths, to earth and between components, should be unchanged by the act of substitution; hence, for the substitution test there should be (ideally) no change in the geometrical layout of leads and components. This can be easily achieved with low-loss changeover switches or coaxial connectors.

For the measurement of radiofrequency components, where shunt capacitance effects are prominent, the substitution principle is particularly valuable. Typical applications are discussed below.

11.9.4.16 Parallel-T network

Figure 11.40(*a*) shows a general parallel-T network, with an a.c. source as input and a detector across the output terminals. The balance condition (of zero output) in terms of the impedance operators is

$$Z_1 + Z_2 + (Z_1Z_2/Z_3) + Z_4 + Z_5 + (Z_4Z_5/Z_6) = 0$$

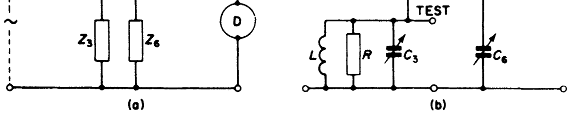

Figure 11.40 Parallel-T networks

Suppose that Z_1 and Z_2 are pure reactances of like sign, e.g. $-jX_1$ and $-jX_2$, and that Z_3 is purely resistive, then

$$Z_1Z_2/Z_3 = -(X_1X_2/R_3)$$⇐

which is equivalent to a negative resistance and capable of balancing out a positive resistance in Z_4 or Z_5. An example is shown in *Figure 11.40*(*b*). The source is a modulated signal generator and the detector is, in effect, a radio receiver tuned to the 'carrier' frequency. Balance is achieved by variation of C_3 and C_6 with the test terminals open-circuited. If an unknown admittance $G_x + jB_x$ is now connected across the test terminals and the disturbed balance restored by altering C_3 and C_6 to C_3'⇐and C_6', respectively, then

$$B_x = \omega(C_3 - C_3') \text{ and } G_x = \omega^2C^2(C_6' - C_6)P/C_4$$

Bridge screening is simplified because the common source and detector terminal can be earthed. The use of a variable resistor as a balance arm can often be avoided, a considerable simplification for radiofrequency measurement.

11.9.4.17 Bridged-T network

This can be used at frequencies up to about 50 MHz. *Figure 11.41*(*a*) shows the schematic arrangement, and *Figure 11.41*(*b*) illustrates a common application. If the unknown impedance $Z_x = R + j\omega L$ is such that R^2 is negligible compared with ω^2L^2, then for balance

$$1/\omega^2LC = 2 \quad \text{and} \quad R = S/4$$

whence R and L can be obtained if the angular frequency ως is known.

11.10 Measuring and protection transformers

Measuring (instrument) transformers are used primarily for changing currents and voltages in power networks to values more suited to the range of conventional indicators. Protection transformers are employed in systems of fault protection. Both types are dealt with in BS 3938 (*Current transformers*) (IEC 185) and BS 3941 (*Voltage transformers*) (IEC 186 and 186A).

11.10.1 Current transformers

Air-cooled current transformers (c.t.) are used in circuits of voltage up to 660 V and currents up to 75 kA. The *bar primary* form employs the actual cable or bus-bar of the main circuit as the primary winding. The core is built of high-grade steel laminations in rectangular or circular shape. The secondary has the appropriate number of complete turns uniformly around the annular core or on all four sides of

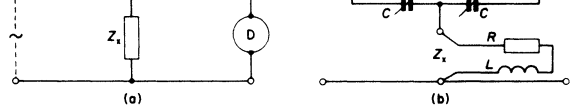

Figure 11.41 Bridged-T networks

the rectangular core. The assembly of core and secondary encloses the bar primary, which is equivalent to a single turn.

For lower primary currents it is necessary to provide a conventional primary winding. This gives a better primary/secondary coupling and a greater accuracy.

Current transformers must be insulated to withstand the service voltage, which may be up to 400 kV. Oil immersion may be necessary, in which case the core and coils are clamped to the top cover to facilitate removal as a complete unit.

11.10.1.1 Burden and output

Primary currents have preferred values between 1 A and 75 kA, with rated secondary currents of 5 A and 1 A (also 2 A). The *rated burden* is the ohmic impedance Z_T of the secondary circuit when carrying the rated current I_s at a stated power factor. The *rated output* is the volt–ampere product $(I_sZ_T)I_s$. The rated output, the limiting value for which accuracy statements apply, appears in selected preferred values between 1.5 and 30 VA. The normal operating mode of a c.t. is as a low-impedance device; hence, a short-circuiting switch is provided for the secondary winding for use when a low-impedance load (e.g. an ammeter) is not connected to the secondary terminals.

11.10.1.2 Errors

In an ideal c.t. the primary/secondary current ratio is precisely the same as the secondary/primary turns ratio, and the currents produce equal m.m.f.s in exact antiphase. In a practical c.t. the current ratio diverges from the turns ratio, and the phase angle differs by a small defect from opposition. These *ratio* and *phase angle errors* arise from that component of the primary current required to magnetise the core and provide for core loss, and from the e.m.f. necessary to circulate the secondary current through its burden. In transformers intended for accurate measurement and for metering, these errors must be small.

11.10.1.3 Classification

The limits of error define the 'class' of transformer. BS 3938 defines six classes for measuring and three for protective c.t.

Measuring c.t. The classes in descending order of accuracy are designated AL, AM, BM, CM, C and D. The limits of ratio error vary from ± 0.1% for AL to ± 5% for D; and the phase error from ± 5 min in AL to ± 120 min (2°) in types CM and C, with no limits quoted for type D.

Protective c.t. The classes are termed S, T and X. The limits of ratio error for S and T are, respectively, ± 3 and 5%, and the corresponding phase limits are 3° and 6° approximately. Limits for type X are not stated explicitly. In testing S, T and X classes it is necessary to distinguish between low- and high-impedance types.

11.10.1.4 Measurement

Errors are normally measured by comparison with a c.t. of higher class. By convention, a phase error is positive when the secondary current phasor leads that of the primary current.

11.10.1.5 Arnold method

In *Figure 11.42* C is the c.t. under test and B is a standard of the same nominal ratio and having known, very small

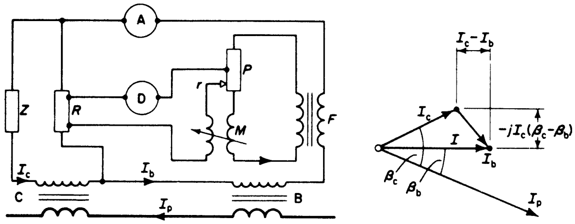

Figure 11.42 Arnold method for measurement of current-transformer errors

errors. The working burden of the test c.t. is Z. The network is such that the difference between the secondary currents I_b and I_c flows in the non-reactive resistor R and is measured by means of the mutual inductor M and slide wire r fed by the auxiliary c.t. F (conveniently of ratio 5/5 A). Phase balance is achieved by adjusting M (positive or negative), and ratio balance by selection of r around the centre-tap of P. From the phasor diagram, the components of the difference current I are approximately $I_b - I_c$ and $-jI_c(\beta_c - \beta_b)$ because the phase angles are actually very small. At balance the condition is given by $I_b(r - j\omega M) = RI$. Equating the 'real' parts gives $I_c/I_b = 1 - (r/R)$; whence in terms of the current ratios $k_c = I_p/I_c$ and $k_b = I_p/I_b$, where the primary current is common to both c.t.s,

$$k_c = k_b(I_c/I_b) = k_b/[1 - (r/R)] \simeq k_b[1 + (r/R)]$$

provided that $r \ll R$. Equating the 'imaginary' parts gives

$$\beta_c - \beta_b = (\omega M/R)(I_b/I_c) \simeq \omega M/R$$

The bridge readings give the errors directly by comparison with those of the reference transformer.

11.10.1.6 Kusters method

This is a comparator method (*Figure 11.43*) in which C is the c.t. under test with its rated burden Z, and B is an a.c. comparator with the same nominal ratio as C but possessing negligible errors. The errors of C can be deduced directly from the balance settings of the arms of the parallel $G - C_1$ network.

11.10.2 Voltage transformers

Magnetically coupled voltage transformers (v.t.) resemble power transformers in basic principle, and have recommended primary voltages up to $396/\sqrt{3}$ kV. Oil-immersion

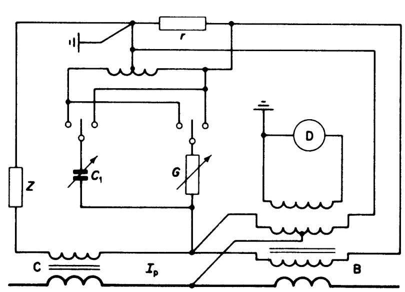

Figure 11.43 Kusters method for measurement of current-transformer errors

is necessary for these levels. The preferred secondary voltages lie between $110/\sqrt{3}$ and 220 V/ph.

11.10.2.1 Burden and output

The preferred rated output burdens lie between 10 and 200 VA/ph, and the rated burden is normally the limit for which stated accuracy limits apply.

11.10.2.2 Errors

Five classes are listed in descending order of accuracy, namely AL, A, B, C and D. The voltage ratio error limit varies between 0.25 and 5% and applies for *small* voltage changes ($\pm 10\%$) around the rated voltage; and the phase error limits are ± 10 min to ± 60 min from class AL to class C. The phase error of D is not defined.

For dual purpose v.t. (i.e. for both measuring and protective application) additional classes E and F are defined which quote the permitted limits of error for classes A, B and C when used at voltages between 0.05 and 1.9 times rated voltage. The v.t. is then denoted by both relevant class letters. The extended error limits are ± 3–5% and ± 2–5%.

11.10.2.3 Measurement

By convention, a phase error is positive when the secondary voltage phasor V_s leads the primary voltage phasor V_p. The ratio error is $(k_r V_s - V_p)/V_p$, where k_r is the nominal rated transformation ratio. The rated output is stated in volt-amperes at unity power factor for the specified class accuracy.

11.10.2.4 Dannatt method

This deduces the error in terms of network parameters (*Figure 11.44*). The v.t. primary is fed from a supply to which is also connected the standard air capacitor C_s in series with a low-valued resistor R_s. The v.t. secondary is loaded with the rated burden Z in parallel with a series-parallel network comprising high-valued resistors R_1 and R_2, a capacitor C_1 and the primary of a mutual inductor M. The secondary voltage of the mutual inductor is balanced against the volt drop across R_s. To ensure that the guard and guarded electrodes of C_s are held at the same potential, a variable resistance R_g is connected so that balance is obtained with either position of the switch S. The balance condition is such that the ratio and phase errors are given very closely by

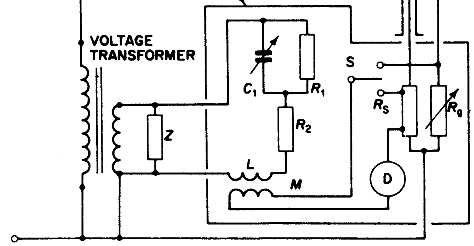

Figure 11.44 Dannatt method for measurement of voltage-transformer errors

$$M/C_s R_s(R_1 + R_2) \quad \text{and} \quad \omega(L - C_1 R_1^2)/(R_1 + R_2)$$

where L is the self-inductance of the mutual inductor primary.

11.10.2.5 Kusters method

The a.c. comparator bridge can be applied. The capacitance ratio of two gas-filled high-voltage capacitors supplied by a common secondary voltage is measured with the currents passing through the separate balancing windings. If one high-voltage capacitor and its series-connected balancing winding are now fed from the primary supply of the v.t., then the voltage ratio can be found by rebalancing the comparator when the two capacitors are connected, respectively, to the primary and secondary voltages.

11.10.2.6 Capacitor-divider voltage transformer

For high-voltage transformers for 100 kV and above, a more economical and satisfactory voltage division for measurement and protective purposes is given by use of the capacitor-divider system (*Figure 11.45*). The reduced intermediate voltage appears across the protecting gap G as the input to a tuned transformer. Changes in the secondary burden Z, which would adversely affect the errors, are minimised by adjusting the transformer leakage reactance and/or that of a series inductor L to resonate with the input capacitance $(C_1 + C_2)$ effectively across the primary. There is a consequent sensitivity to frequency.

The classification is the same as for magnetic transformers, except that AL does not apply and there is an additional limitation on frequency. The transient performance is naturally significantly different from that of the normal v.t.

11.11 Magnetic measurements

Because of the non-homogeneity of bulk ferromagnetic materials, magnetic metrology is relatively inaccurate and imprecise. The fields of interest lie in the basic physics of magnetism, the distribution of the magnetic field, the assessment of magnetic parameters, and the measurement of core loss.

Physical basis The origin of magnetic phenomena lies in the statistical quantum-mechanical behaviour of electrons and particles. Additional information can be obtained by photography of domain formation.

Field distribution Two-dimensional field distributions in air-gaps are readily traced by current-field analogues such as conducting paper or the electrolytic tank. The tank can also be employed for restricted three-dimensional fields, and tests around models or actual equipments can be

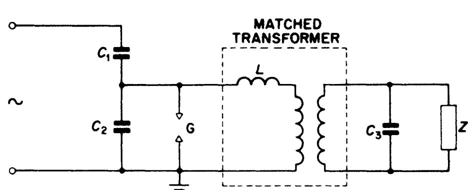

Figure 11.45 Capacitor-divider voltage transformer

based on various types of small sensor working into calibrated instruments (fluxmeter or ballistic galvanometer) or a Hall effect device.

Parameters The 'static' magnetisation characteristic, the relative permeability and the hysteresis loop are the principal parameters of interest.

Core loss The loss in a core is a matter of considerable technological importance. It is, however, very difficult to measure directly the losses in such a way that the results can be directly applied to machine constructions other than transformers.

11.11.1 Instruments

The most simple fundamental standard of magnetic flux density is based on the accurately known dimensions of a long, uniformly wound solenoid carrying a known current. Alternatively, a uniformly wound toroid can be used, although in this case the flux density will not be quite uniform across the core section. The measurement of flux and flux density can be carried out by means of the ballistic galvanometer or fluxmeter for the former, and the Hall effect instrument for the latter.

11.11.1.1 Ballistic galvanometer

The moving coil of a galvanometer with a long periodic time gives a first swing proportional to the time integral of the current through it (i.e. the charge) provided that the duration is very short. The charge results from a time integral of voltage (i.e. magnetic linkage) impressed on a known resistance. The moving coil is connected in series with a search coil and a resistor, and is immersed in the magnetic field due to a known current. When this current is *rapidly* reversed, the total change of linkage is presented to the galvanometer as an impulsive charge, and the first deflection of the instrument is used for calibration against a known (or calculable) linkage change, or for measuring an unknown linkage. The main limitations to accuracy are the observation of the scale and the uncertainty in the current measurement. A typical high sensitivity ballistic galvanometer has a period of 20 s, a moving-coil resistance of 850 Ω, and a sensitivity of 8.5 m/μC on a scale distant 1 m from the coil using a lamp and mirror technique.

11.11.1.2 Fluxmeter

The fluxmeter is a permanent-magnet instrument with a moving coil of low inertia and negligible control torque. Its damping is made high by use of a relatively thick aluminium former, so that the period is long. It is immaterial whether the time taken to change the linkage in a search coil connected to the moving coil is long or short, and in consequence the instrument is useful in iron testing, where the time taken for a flux to collapse or reverse may be several seconds. The deflection is read from the initial position of a pointer on a quadrant scale when the pointer reaches its maximum deflection; after this the pointer drifts slowly back to a zero position. A typical full-scale deflection would be given by a change of 10 μWb-t.

Strong air-gap flux densities can be measured by an alternative method in which a small coil is rotated at a high and known speed, the induced e.m.f. being proportional to the local flux density.

11.11.1.3 Hall effect instrument

The Hall effect applies to conductors generally, but its application is normally associated with semiconductor materials. These should be of low resistance so that, even for small signals, the thermal (Johnson) noise effects will be low. Bulk materials, indium arsenide and indium antimonide, have low resistances and good output voltage coefficients, with the InSb voltage larger than that of InAs; however, InAs is usually selected, as its temperature coefficient is one-tenth that of InSb. Thin-film InSb is used for switching applications where the higher output e.m.f. is the prime consideration. The Hall effect response is very fast, being usable up to the megaherz range. Bulk material InAs elements are more effective than thin-film elements, owing to the much lower resistances obtainable.

In commercial instruments the current I_c in the sensing element can be a direct current chopped at audio frequency; the resultant audiofrequency Hall voltage E_H, after linear amplification and demodulation, can be displayed on a taut-band d.c. moving coil indicator against a calibrated scale of flux density. Typical ranges have full-scale deflections between 0.1 mT and 5 T for steady fields. Pulsating fields can be measured with d.c. instruments at frequencies up to 500 Hz with a cathode-ray-oscilloscope detector. Instruments for alternating fields only are available for frequencies of about 30 kHz, with ranges similar to those of d.c. instruments.

An instrument and probe can be checked by use of stable reference magnets, which may have uncertainties of 0.5–1.0%. The overall uncertainty of a scale reading is typically ±3% of the full-scale deflection for the range.

The sensing elements are protected by epoxy glass-fibre or similar enclosures, and are available in a variety of forms for probing transverse, coaxial or tangential fields. The flat form is most common, typically with the dimensions 0.5 mm × 4 mm × 25 mm. Incremental measurements, using two matched probes and backing-off networks, enable perturbations as small as 0.1 μT to be observed either in the presence of, say, a given 0.1 T field or as a difference between two separate 0.1 T field systems. Hall effect instruments have many obvious applications where the magnitude and direction of the field distribution is required such as air gaps of machines and instruments, magnetron magnets, mass spectrometer fields, residual interference fields, etc.: in addition, the Hall-effect principle can be used in a variety of transducer applications by making I_c and B proportional to separate physical functions, and measuring the instantaneous or average results of this product. One example is a wide band, 50 kHz wattmeter in which I_c and B are made, respectively, proportional to the scalar values of the load current and voltage.

11.11.2 Magnetic parameters

The d.c. magnetisation curve, hysteresis loop and relative permeability can be obtained by use of a fluxmeter or calibrated ballistic galvanometer, using a method substantially the same as for calibration. The magnetic test material constitutes the core of the toroid, or is built in strips to form a hollow square. The a.c. cyclic magnetisation loop differs from the 'static' loop, and is still further modified when a d.c. bias m.m.f. is superposed. The a.c. loop can be viewed on a c.r.o. displaying the B and H functions on the Y and X axes, respectively. The H function can be obtained from the volt drop across a small resistor carrying the magnetising current, and the B function from the time integral of the search-coil voltage obtained with the aid of an operational amplifier.

11.11.2.1 Core loss

In a low-frequency test the loop area of the B/H relation is directly proportional to the core loss per cycle of magnetisation, which provides a simple comparative test for specimen material. A more convenient method employs low power factor wattmeters, but care is needed to exclude instrument and connection losses. Selected samples of sheet steel intended for the cores of inductors and power transformers are prepared for testing in the form of strips (e.g. 30 cm × 3 cm) assembled in bundles and butted or interleaved at the corners to form a hollow square. The sides are embraced by magnetising and search coils of known dimensions. Input power and current, and mean and r.m.s. voltages, are read at various frequencies and voltages to assess the overall power loss within the material and (if required) an approximate indication of the eddy and hysteresis loss components. The Lloyd–Fisher and Epstein squares are commonly used forms, the latter being referred to in BS 601 (*Sheet steel*).

The magnetic properties of bars, forgings and castings are derived from galvanometer or fluxmeter measurements (BS 2454). A.c. potentiometer and bridge methods are also used, especially for low-loss materials at frequencies in the audio range. With ferrite and powder cores, as with Q-value tests of a coil with and without a slug core, r.f. resonance methods are convenient.

Production and quality control test equipment for core-loss and coil-turns instruments include a 'coil-turns tester' (Tinsley 5812D) with a four-digit display of the number of turns in non-uniform and uniform windings, by using an inductive comparative measurement against a standard winding: the same range of 'magnetic type' instrumentation includes a 'shorted-turns tester' which will detect single and multiple shorted circuited turns in coils of any shape or number of windings; while a 'core tester' can check transformer core characteristics of EI stacks, C cores and toroids including hysteresis loop output for a cathode-ray-oscilloscope display.

11.11.3 Bridge methods

Two examples of bridges for audiofrequency tests are given. The non-linear magnetic characteristics of materials introduce harmonics at higher flux densities, imposing limitations because the bridge must be balanced at fundamental frequency.

11.11.3.1 Campbell method

The specimen, shown as a toroid, is uniformly wound with N_2 secondary turns, overlaid with N_1 primary turns (*Figure 11.46*(*a*)). The magnetising current I_1 passes through the primary of the mutual inductor M and the resistor r.

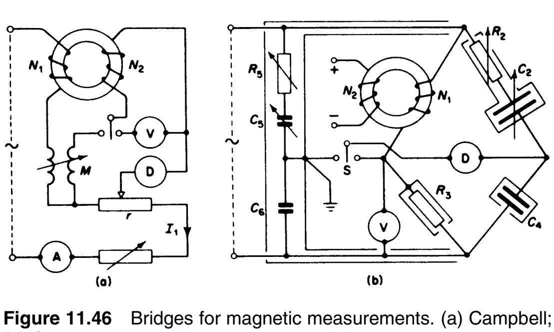

Figure 11.46 Bridges for magnetic measurements. (a) Campbell; (b) Owen

At balance indicated on detector D, the power factor and power loss are given by

$$\cos\phi = r/\sqrt{(r^2 + \omega^2 M^2)} \quad \text{and} \quad P = I_1^2 r(N_1/N_2)$$

The corresponding magnetising force and flux density are found from

$$H - (I_1 N_1/l)\sin\phi \quad \text{and} \quad B_m = V/4faN_2$$

where l is the mean length of the magnetic path in the toroid, a is its cross-sectional area, and V is the *mean* voltage as measured on the high-impedance *average* voltmeter V.

11.11.3.2 Modified Owen method

The toroidal specimen (*Figure 11.46*(*b*)), has N_1 a.c. magnetising turns, and an additional winding N_2 through which a steady polarising d.c. excitation can be superposed. The d.c. supply is taken from a battery through a high-valued inductor to minimise induced alternating current from N_1. The Wagner earth arrangement $R_5C_5C_6$ may be added to eliminate earth capacitance errors from the main bridge arms by balancing in both positions of the switch S. The high-impedance voltmeter V measures the r.m.s. voltage V across R_3. The a.c. magnetising force H_a due to the current V/R_3 in N_1, the d.c. magnetising force H_d due to I_d in N_2, the power loss in the specimen (including the I^2R loss in the winding N_1 which can be allowed for separately) and the peak flux density B_m are given by

$$H_a = VN_1/R_3 l; \quad H_d = I_d N_2/l; \quad P = V^2(C_4/C_3R_3)$$
$$B_m = \sqrt{2}V(C_4/\omega C_2 a N_1)\sqrt{(1 + \omega^2 C_2^2 R_2^2)}$$

where a is the cross-sectional area of the specimen, and it is assumed that all time-varying quantities have sinusoidal waveform and angular frequency ω.

11.12 Transducers

In instrumentation systems a transducer is a device that can sense changes of one physical kind and transpose them systematically into a different physical kind compatible with a signal processing system. The compatible signals considered here are generally electrical or magnetic. Transducers can sense most non-electrical quantities (e.g. humidity, pressure, temperature and force) and, so far as the electrical response is concerned, can be classified as 'active' or 'passive', as follows:

Passive. The output response produces proportional changes in a passive network parameter such as resistance, inductance and capacitance.

Active. Transducers act as generators, the class including piezoelectric, magnetoelectric and electrochemical devices.

The 'Code for Temperature Measurement' is included in BS 1041. The sensors and instruments for the purpose are classified according to temperature range:

(1) Specialised thermometers for measurements near absolute zero. One such, based on the magnetic susceptance of certain paramagnetic salts, is suitable for temperatures below 1.5 K; another, employing an acoustic resonant-cavity technique, can be used up to 50 K. The associated

instruments are a mutual inductance bridge for the former, and electronic measurement of length for the latter.
(2) Ge, Si and GaAs p–n junction diode and carbon resistor thermometers are used with d.c. potentiometers and Wheatstone bridges for temperatures up to 100 K.
(3) Vapour pressure thermometers are used for standard readings below 5 K and for general measurements up to 370 K.
(4) Thermistor (resistance) and quartz (resonance) methods, with Wheatstone bridge and electronic counter, respectively, cover the range 5–550 K.
(5) Electrical resistance sensors (usually of Pt) are used with d.c. bridges throughout the range 14–1337 K.
(6) Thermocouple e.m.f.s are measured by d.c. potentiometer. The method, widely employed in industry, has ranges between 100 and 3000 K with various combinations of metals.
(7) The expansion of mercury in glass or steel capillary tubes is applied to measurements from 230 to 750 K. The range is 80–300 K if the mercury is replaced by toluene.
(8) Radiation and optical thermometers employ thermopiles, photodiodes or photomultipliers for measurements up to 5000 K, using d.c. potentiometric or optical balance methods. These provide the only practical methods for very high temperatures.

Mercury-in-glass and optical thermometers are indicating instruments only, but the remainder can be made to furnish graphical records. The electrical resistance and thermoelectrical thermometers are especially suitable for multipoint recording as for heated solid surfaces, points in a mass, or inaccessible places in electrical machines.

11.12.1 Resistive transducers for temperature measurement

A resistive sensor based on a substantially linear resistance/temperature coefficient should have small thermal capacity for rapid response and avoidance of local temperature gradient. Hence, materials of high resistivity (giving small volume) and temperature coefficient are desirable. Thermistors are made to fulfil these requirements.

11.12.1.1 Pure-metal sensors

Pure conductors such as Pt, Ni and W have low resistance/temperature coefficients but these are stable and fairly linear. Pt is generally adopted for precision measurements and, in particular, for the definitive experiment within the range 100–903.5 K of the International Practical Temperature Scale. Platinum and other metals can be deposited on ceramics to produce metal-film resistors suitable (if individually calibrated) for temperature measurement.

11.12.1.2 Platinum resistance thermometer

To a reasonable approximation the resistance R of a platinum wire at temperature θ (in degrees Celsius) in terms of its resistance R_0 at 0°C is

$$R = R_0(1 + k\theta) \quad \text{where} \quad k = (R_{100} - R_0)/100R_0$$

The temperature θ_p for a given value R is

$$\theta_p = 100(R - R_0)/(R_{100} - R_0)$$

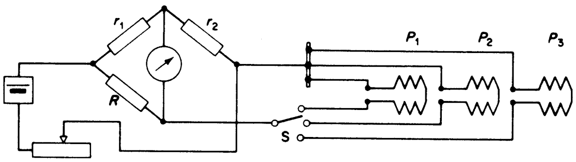

Figure 11.47 Temperature-sensor network

and is known as the *platinum temperature*. Conversion to true temperature is found from the difference formula

$$\theta - \theta_p = \delta[(\theta/100)^2 - (\theta/100)]$$

The value of δ depends upon the purity of the metal. It is obtained by measurement of the resistance of the sensor at 0, 100 and 444.67°C, the last being the boiling point of sulphur. The value of δ is typically 1.5.

A practical sensor usually consists of a coil of pure platinum wire wound on a mica or steatite frame, the coil being protected by a tube of steel or refractory material. The resistance measurement is carried out by connecting the coil in a Wheatstone bridge network or to a potentiometer. In the former case the bridge usually has equal ratio-arms and a pair of compensating leads is connected in the fourth arm. These leads run in parallel to the actual leads from the sensor coil and compensate for their resistance changes. If the initial resistance and coefficient k of the sensor are large, the compensating leads may not be required. *Figure 11.47* shows the method of connecting three such coils in turn in a Wheatstone bridge network, so as to measure the temperatures at three different locations. In this case the bridge is not balanced; the out-of-balance current through the galvanometer gives the temperature directly. Initial setting is done by adjustment of the battery current until some definite deflection is obtained, when a standard resistance replaces the sensors in the bridge circuit.

11.12.1.3 Thermocouple sensors

Thermocouple sensors are active transducers exploiting the Seebeck effect developed between two dissimilar metals when two junctions are at different temperatures. The International Practical Temperature Scale is defined in terms of a (0.9 Pt + 0.1 Rh)/Pt thermocouple over the range 903.5–1336 K. Readings between 100 and 3000 K are possible with Au/CoCu at the lower and W/Re at the higher end. Conventional materials for intermediate ranges include

copper/constantan (670 K)	iron/constantan (1030 K).
chrome/constantan (1270 K)	chromel/alumel (1640 K)

the figures giving the upper limits. The thermo-e.m.f. varies between 10 and 80 μV/K. To make an instrument direct-reading for the temperature at one junction, the second ('cold') junction must be kept at a reference temperature. The cold junction can be maintained at 0°C by immersion within chipped, melting ice in a thermos flask, or by using a commercial ice-point apparatus working upon the Peltier effect. When high temperatures are being measured, the terminals of the detector can sometimes be used as the 'cold' junction, but compensating leads should be connected between the thermocouple and the detector. The most accurate method for measuring the e.m.f. is with a d.c. potentiometer.

The e.m.f. e for a hot-junction temperature θ (in degrees Celsius) with the cold junction at 0°C is of the form $\log(e) = A\log(\theta) + B$, where the constants A and B have, for example, the values 1.14 and 1.36 for copper/constantan, with e in microvolts.

The small thermal capacity of thermocouples makes them suitable for the measurement of rapidly changing temperatures and for the temperatures at particular points in a piece of apparatus. One useful application is the measurement of surface temperatures, in which case the thermocouple consists of the two metals in the form of a flat flexible strip with a welded junction at the centre, this strip being applied to the surface under test.

11.12.2 Thermistors

Thermistors are semiconductor sensors which are made from the sintered compounds of metallic oxides of Cu, Mn, Ni and Co formed into beads, rings and discs. A high resistivity is achieved with a large resistance/temperature coefficient. The units can have resistances at 20°C from 1 Ω to several mega-ohms. The thermal-inertia time-constant is not more than 1 s for small beads, but rather more for discs and coated specimens. The outstanding property is the negative resistance/temperature coefficient of several parts in 10^2 per degree Celsius. Thermistors can be used over the range 200–550 K. One consequence of the current/voltage characteristic is self-heating if the thermistor carries appreciable current.

Wheatstone bridge networks are widely used, with the thermistor in one arm and the out-of-balance detector current as an indication of temperature. Although the characteristics are strongly non-linear, it is possible to obtain thermistors in matched pairs, which can be applied to differential temperature measurement. Another application, not directly a temperature measurement, is to the indication of air flow in pipe, or as anemometers: one thermistor is embedded in a metal block, acting as a thermal reservoir, the other senses the air speed as a cooling effect.

11.12.3 p–n Junctions

Certain semiconductor diodes are suitable for temperature measurement over a wide range; Ge, for example, is useful below 35 K. One differential thermometer has two matched p–n diodes which, when linearly amplified, give a discrimination better than 0.0001°C. Such instruments have several indirect applications, such as sensing thermal gradients in 'constant-temperature' enclosures or vats, monitoring load changes, and displaying the input–output liquid temperature differences in fluid pumps.

11.12.4 Pyrometers

Radiation thermometers respond to the total radiation (heat and light) of a hot body, while optical thermometers make use only of the visible radiation. Both forms of *pyrometer* are specially suited to the measurement of very high temperatures, because they do not involve contact with the source of heat.

Radiation thermometers Both the Fery variable-focus and the Foster fixed-focus types comprise a tube containing at the closed end a concave mirror which focuses the radiation on to a sensitive thermocouple. In use the tube is 'sighted' on to the hot body. The sighting distance is not critical, provided that the image formed is large enough to cover the thermocouple surface. Calibration is by direct sighting on to a body of known surface temperature.

Optical thermometers The commonest form is the disappearing-filament type, which consists of a telescope containing a lamp, the filament brightness of which can be so adjusted by circuit resistance that, when viewed against the background of the hot body, the filament vanishes. The lamp current passes through an ammeter scaled in temperature. The telescope eyepiece contains a monochromatic glass filter to utilise the phenomenon that the light of any one wavelength emitted by the hot body depends on its temperature. Although calibration is based on the assumption that the hot body is a uniform radiator, departure from this condition involves less error than in the radiation pyrometer.

11.12.5 Pressure

The piezoelectric effect is the separation of electronic charge within a material when applied pressure deforms the crystal structure. Conversely, the application of charge to the crystal will cause changes in the dimension of the crystal. Quartz is the only *natural* crystal in general use as a sensor. It is important that the crystal be prepared by slicing along the plane most sensitive for the particular application. Quartz is used in oscillators as the stable frequency-reference element. The self-resonant frequency is temperature-dependent, so constant temperature (e.g. 35°C) ovens are required for very stable oscillators: this application of quartz is not as a sensor, but the same temperature dependence of resonant frequency is applied for very-low-temperature measurement (below 35 K) where the temperature change can be interpreted by changes in frequency. Numerous ceramic crystals have been developed based upon barium titanate with controlled added impurities; the piezoelectric properties have to be applied by a special polarising treatment during the cooling period following the sintering process in a kiln.

Active four-arm strain gauge bridges, diffused into a single crystal silicon diaphragm, form the basic sensor for a wide range of sensitive pressure elements which are encapsulated in solid state transducers/transmitters (e.g. Druck Limited, Groby, Leicestershire, England). Minimum to maximum pressures can be measured, from venous or arterial values in physiology, to aerospace applications in engines and satellites, and to high-pressure industrial processes. The sensors have 10 V d.c. or a.c. input, with 15 mV $\rightarrow$ 10 V outputs, 0.1% non-linearity working into a $4\frac{1}{2}$-digit multichannel indicator/BCD-recorder which, when coupled with a $5\frac{1}{2}$-digit, 0.04% calibrator, provides traceability through a low-cost primary standard.

11.12.6 Acceleration

Accelerometers can be used for velocity measurement by integration of the output signal. The acceleration force of a mass is made to increase or decrease the spring pressure on a ceramic or quartz crystal to produce proportional piezoelectric effect. The device is insulated and hermetically sealed, with care taken to avoid loss of signal through parallel insulation paths of resistance comparable with that of the crystal itself (e.g. 1000 MΩ). Conditioning of the signal involves matching to the much lower input impedance of conventional amplifiers by use of an emitter–follower network. 'Integrated circuit' forms of the network can be located actually within the transducer unit.

The signal amplifier can be a voltage amplifier or a charge amplifier (i.e. an operational amplifier with capacitive feedback). The latter is preferred because the equivalent feedback capacitive effect is large and dominates the input capacitance associated with varying lengths of coaxial cable.

Other accelerometers are based either on changes in the reluctance of differential transformers or on variations in the resistance of a strain gauge. Slow changes in acceleration can be sensed by connecting the seismic mass to a wiper on a resistance voltage divider.

The upper frequency of the output is limited by the natural frequency of the transducer, while the minimum is almost zero. If the lowest useful frequency of an a.c. electronic amplifier is 20 Hz, the analysis of very-low-frequency signals may be achieved by first recording them on a precision frequency-modulated tape-recorder, which is then replayed at high speed for amplification and analysis.

Constant bandwidth frequency analysers are convenient for stable periodic complex waveforms. Constant per cent bandwidth types suit cases such as machines in which there is some small fluctuation in the nominal periodic behaviour. Truly random vibrations of a stochastic nature may more profitably be analysed in terms of the power spectral density function. Analysers are available for measuring many properties of random signals—e.g. 'probability density function' devices based on instantaneous signal amplitudes (see Section 11.7.5).

11.12.7 Strain gauges

Electrical strain gauges are devices employed primarily for detecting and measuring small dimensional variations in the surfaces to which they are attached, particularly where direct measurement is difficult. The gauge essentially converts mechanical displacement into a change in some electrical quantity (usually resistance). The essential feature is that strain shall be communicated to the gauge without fatigue or inertia effects.

11.12.7.1 Resistance-wire, p–n junction and carbon gauges

A grid of resistance wire is cemented between two insulating films. The grid is usually smaller than 30 mm × 15 mm, and has a resistance in the range 60–2000 Ω. When the gauge is cemented to the surface under test, the change ΔR in total resistance R due to a displacement ΔL in a length L is converted into strain ϵ by means of the gauge factor

$$k_g = (\Delta R/R)(L/\Delta L) = (\Delta R/R)/\epsilon$$

If F represents the force and E the elastic modulus (i.e. the stress/strain ratio), then

$$F = k_g EaR/\Delta R$$

where a is the area normal to the direction of the applied force. Resistance-wire strain gauges have been in use for many years. Recent developments include metal-foil strain gauges based upon printed circuit and photoetching techniques: these can be made smaller. Thin-film techniques have been developed to apply the gauges directly to very small surface areas.

Semiconductor *silicon junction* strain gauges are perhaps 30 times more sensitive than the resistive gauge; however, their stress and temperature ranges are more restricted, and they are more difficult to 'cement' in position. The inherent non-linearity of the response by sensors can be offset by special bridge techniques using compensation networks.

Gauges consisting of a layer of *carbon*, which responds to strain in a manner similar to that of a carbon microphone, have a gauge factor typically of 20, considerably higher than for metals but much less than for p–n junctions. The inherent disadvantage is sensitivity to temperature and humidity.

11.12.7.2 Strain-gauge measurement techniques

Strain An 'active' gauge is cemented to the workpiece, and a similar 'compensating' gauge is left free. The gauges form two arms of a bridge (*Figure 11.48*(*a*)), which is inherently self-compensating for temperature. Balance is achieved initially by adjusting R_1, the load is applied and the bridge rebalanced. For multiple-gauge arrangements, 100-channel instruments with zero-balancing facilities are available, giving a printed read-out. Alternatively, after initial balance, the residual bridge voltage is measured, or applied to a cathode-ray-oscilloscope for recording. Dynamic strain may be measured by the circuit of (*Figure 11.48*(*b*)), usable with a flat-response amplifier over the range 0.5–1000 Hz. As in any structural stress analysis it is important to know the time relation as well as the amplitude of stress, the amplifier must have either a negligble phase shift or one directly proportional to frequency over the required working range.

Torque For a circular-section shaft in torsion, the principal stresses lie at 45° to the axis. If four gauges are applied as in *Figure 11.49*, and connected to a bridge through suitable slip-ring gear (usually silver ring surfaces and silver graphite brushes), the bridge unbalance is four times that of a single gauge. The arrangement is inherently temperature compensated. For engine testing it is normal to make up a special length of shaft with gauges and slip-rings, to be fitted between the engine and the brake or driven unit. With automatic recorders it is possible to record simultaneous data such as torque, pressure variation, temperature, etc., to give all phenomena on a time base display.

Thrust Various other measurements of mechanical power transmission can be measured. The arrangement in *Figure 11.50* is for thrust. Four axial (A) and four circumferential (C) gauges are disposed symmetrically and connected as indicated in a bridge network, with r as balancing adjustment at zero load. The monitoring system (which can be used with other strain-gauge applications) consists of a self-balancing potentiometer with chart (Ch), digital (Dg) and indicator (Id) readouts. The speed of a self-balancing potentiometer is restricted to about one reading per second.

Principal stresses The direction and magnitude of principal stresses can be determined from a *rosette* of three or

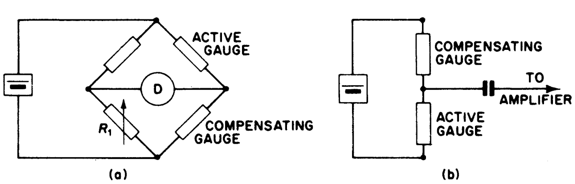

Figure 11.48 Strain-gauge networks

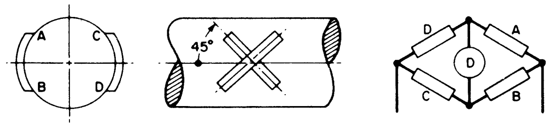

Figure 11.49 Strain-gauge torque measurement

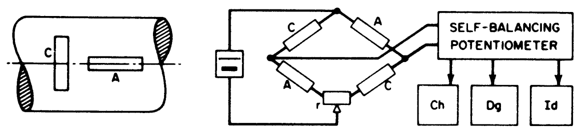

Figure 11.50 Strain-gauge thrust measurement

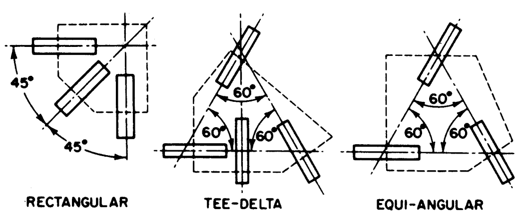

Figure 11.51 Typical strain-gauge rosettes

more strain gauges arranged with suitable orientations (*Figure 11.51*). Commercial rosettes are available in several geometrical formations. As it is unlikely that the principal strain will be on the axis of any one gauge, note must be taken of the sensitivity of the gauges in a direction at right angles to the axis if accurate results are required. The correction is not likely to exceed 3%.

11.12.8 Magnetostrictive transducers

Magnetostriction is the dimensional change that occurs in a ferromagnetic material when subjected to magnetisation, and an inverse effect of change in magnetisation when the physical dimensions are altered by the application of external force. The dimensional changes are very small (e.g. 30 parts in 10^6), but if the magnetic field alternates at a frequency corresponding to one of the natural frequencies of vibration of the material in some form, such as a bar, resonance increases the vibration amplitude considerably to perhaps 1 part in 10^3.

Materials vary widely in the magnitude and sign of the magnetostrictive effect, which also depends on the magnetic field intensity. Pure iron may have positive or negative magnetostriction: the addition of nickel makes the effect positive at all frequencies. With 30% Ni the longitudinal magnetostriction falls to zero, a property utilised in invar–nickel alloys to give very low coefficients of thermal expansion, because the latter is neutralised by the magnetostrictive contraction. In nickel alloys there is a peak dimensional change when the Ni content is about 45% and at about 63% the most sensitive magnetostrictive condition is reached.

11.12.8.1 Sensors

For the purpose of detecting surface and submarine ships, magnetostriction transducers have been employed using under-water supersonic vibrations. The compression stresses generated by the transmitter are reflected by the submerged object and are picked up by a detector utilising the inverse effect—i.e. change of magnetic properties under mechanical stress.

The presence of flaws, air pockets or impurities can similarly be detected in some opaque substances (e.g. rubber) which should normally be homogeneous. The phase displacement between the incident and reflected waves can be interpreted to indicate the position of the obstruction.

Liquid level can be controlled by a sensor probe just above the required level and connected to an oscillator network. As the level of the liquid reaches the probe, the oscillation ceases because of the greatly increased damping. For the measurement of mechanical stress (*Figure 11.52*) the bridge is balanced when the sensor and compensating elements are equally stressed, the compensating element eliminating thermal errors. If the stress in the sensor element is increased, its permeability is reduced, and the out-of-balance current is indicated on the detector D, which can be calibrated in stress units. If the stress is rapidly varying, D can be replaced by a cathode-ray oscilloscope.

11.12.9 Reactance sensors

Many transducers use the properties of inductive and capacitive reactance. The inductance of an iron-cored inductor varies very rapidly with the length of an included air gap. The basic network in *Figure 11.53* employs a bridge with inductors 1 and 2, one with a fixed gap and the other with a gap variable in accordance with some physical quantity such as pressure, strain, thickness, acceleration, etc. The detector D is calibrated appropriately for direct indication. Reliable readings of displacement down to about 3×10^{-5} mm are readily secured with a properly designed instrument. If the displacement fluctuates, it will modulate the supply waveform, and the modulation waveform can be filtered out from the carrier.

Capacitive reactance can also be used, but as the magnitude of its change with displacement is very small, the capacitor sensor is made to be part of the capacitance in an

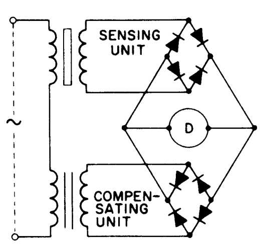

Figure 11.52 Magnetorestriction stress sensor circuit

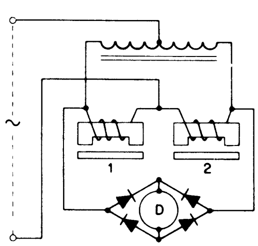

Figure 11.53 Reactance bridge electric guage

oscillatory circuit, the frequency of which can be related to the displacement. One type of capacitor pressure sensor consists of two flexible parallel plates with a 'vacuum' dielectric. A cylindrical capacitor suitable for sensing high pressures has as electrodes an inner deforming cylinder and an outer ring section: dimensional changes in the radial gap occur when the axial force on the inner cylinder causes a change in its diameter. The change of capacitance is well within the range of commercial straingauge bridges.

11.12.9.1 Angular velocity sensors

Measurements can be read from the output of a permanent-magnet d.c. tachometer for speeds up to 3000 rev/min. The instrument is direct-reading, but a proportional direct output voltage can be used as an electrical transducer. For low speeds the commutator ripple can be excessive and an a.c. tachometer with full-wave rectification will provide a more uniform output, particularly if the instrument is of the multi-tooth variable-reluctance type. These instruments act as loads on the source and they are unable to assess the fine detail of the angular velocity such as superimposed small amplitude oscillations, or non-uniform accelerations and retardations due to changes in load. Simple optical solutions to the problem involve the use of a stroboscope or photosensor coupled with digital counters, and each method avoids any loading of the machine.

11.12.10 Stroboscope

The stroboscope generates high-intensity impulsive flashes of light at controllable repetition frequency. If the repetition frequency corresponds to the time of one revolution of a rotating object or to one complete excursion of a reciprocating object (or any multiple thereof), the object appears to be at rest. If the period is a submultiple $1/n$, then the object appears stationary in n different positions. It is desirable for the repetition frequency to exceed 30 Hz in order to avoid visual flicker.

From the flash frequency and observation, the character of velocity perturbations can be viewed and counted and dynamic distortion effects seen. Electrical stroboscopes may involve special neon lamps connected to the secondary side of an induction coil, the primary of which is interrupted by a driven tuning fork. In modern instruments the lamp is a xenon gas discharge tube emitting white light. The flashing frequency is adjusted by means of a transistorised multivibrator-shaper circuit producing pulses of constant energy and fast rise time. Each trigger pulse initiates the discharge of a capacitor to create the flash. Capacitor recharging limits the upper frequency to about 1.5 kHz with a 200 lux illumination level. High frequencies up to 10 kHz can be obtained from a Z-modulated cathode-ray tube as a low-level light source.

11.12.11 Photosensors

In this context, 'light' usually means radiation at wavelengths covering the visible spectrum from the near infrared to the near ultraviolet. Light transducers can be identified by three basic forms of physical behaviour: photovoltaic, photoresistive/photojunction and photoemissive.

11.12.11.1 Photovoltaic

Photovoltaic sensors are p–n junction devices with the normal barrier layer potential present between the two materials. When illuminated, the higher-energy protons raise electrons from the valence into the conduction band to create electron–hole pairs. The consequent continual charge separation is enough to drive current through a resistive load without any external source.

Selenium cell This has a linear current/illumination characteristic for load resistances up to 100 Ω, but the relation is progressively more non-linear for higher resistance, and reaches an open-circuit e.m.f. of 0.6 V. The peak spectral response is in the centre of the visible band (0.57 μm), but the energy efficiency is only 0.5%.

Silicon p–n junction cell The energy conversion is 10–15%, which is adequate for solar cell use in space vehicles. As the time response (a few microseconds) is very fast, silicon photocell arrays are used for reading punched-card and tape, and for optical tracking. The output of a single sensor is typically 70 μA for 5000 lux, the current being constant up to 200 mV. When they are reverse-biased, silicon photosensors behave as photoresistive devices.

11.12.11.2 Photoresistive

A photoresistive sensor consists of a single homogeneous semiconductor material such as n-type cadmium sulphide or cadmium selenide. The material is doped to permit of a large electron charge amplification by the selective absorption of holes, and a spectral response that can simulate that of the eye. This type of sensor is well known as a photographic exposure meter. The CdS sensor is not quite linear over wide illumination ranges, and its response time (100 ms), is slow compared with that (10 ms) of the CdSe type.

The sensors can be used in series with a source and a relay, the latter operating when illumination reduces the sensor resistance to about 1 kΩ. The power dissipation by the sensor can be minimised if the slow resistance changes are used to alter the state of a conventional Schmitt trigger circuit (*Figure 11.54*). The required illumination level is set by R_1; when the illumination is reduced, the cell resistance rises to switch on T_1, which causes T_2 to switch off and T_3 then saturates. The relay operates when the Zener diode conducts at about 12 V. The sensor is useful for alarm circuits, street-lighting control and low-rate counting.

11.12.11.3 Photojunction

Normal diodes and transistors are light-shielded. In the photojunction devices controlled light is admitted to enhance the light effect.

Photodiodes These are normally of silicon unless the greater infra-red response of germanium is needed. Time constants of 10–100 ns are normal. Silicon planar p-type/intrinsic/n-type photodiodes can respond to laser pulses of 1 ns, and are usable at very low light level equivalent to a

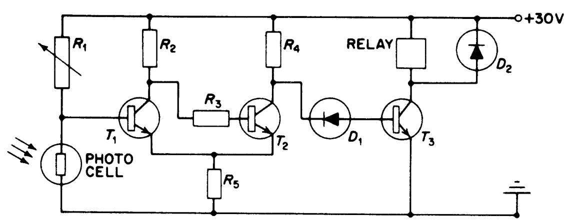

Figure 11.54 Photoresistive cell in Schmitt trigger circuit

current of about 100 pA. Special low-noise transistor amplifiers are required.

Phototransistors The normal form is the n–p–n silicon planar transistor, with conventional bias except for the reverse-biased photodiode between base and collector. It behaves as a common-collector transistor amplifier with a photocurrent generator between base and collector. The maximum cut-off frequency is 50 kHz. A derivative of the field effect transistor, the 'photofet', is more sensitive and the gain–bandwidth product is higher.

11.12.11.4 Photoemissive

Photoemissive sensors are vacuum or gas-filled phototubes. The emission of electrons from the cathode occurs when light falls on it, to be collected by a positive anode. Vacuum phototubes have been superseded for light measurement by photodiodes. In the *photomultiplier* tube the vacuum emission is enhanced by secondary emission at a succession of anodes to give an electron multiplication of 10^8 or 10^9, providing significant output for flashes of very short duration. The photomultiplier is used as a *scintillation* sensor by viewing the weak light emissions that occur in selected phosphors exposed to α, β- or γ-radiation. These scintillation 'counters' have wide application in spectrophotometry, flying-spot scanning, photon counting and whenever a very weak light signal is to be amplified. An alternative device is the Bendix magnetic multiplier. This consists of a single layer of resistive film with the emissive cathode at one end; by suitable accelerating p.d.s applied along the film, together with shaped magnetic fields, the electrons are multiplied by secondary emission and bounced along the strip to accumulate on an anode to give an overall electron gain of 10^9. With different emissive materials, ultraviolet rays and X-rays can be amplified for subsequent analysis.

Gas filled phototubes are low-frequency devices but have a high signal/noise ratio. They are used for film sound-track sensing, and have been applied to monitor natural gas flames (which emit considerable ultraviolet radiation) to avoid danger from unburnt gas.

11.12.11.5 Selection

Photovoltaic cells do not require an external supply, but they are generally less sensitive than photoresistive types, which do. Photodiodes can be selected for a range of optimum spectral sensitivities and are fast-operating, particularly so for the silicon planar p–i–n type. Phototransistors provide amplification at the cost of a relatively slow response (minimum 20 μs). Photomultiplier tubes provide high amplification, even for single photons.

Integrated-circuit chips about 4 mm^2 in area, produced with progressively more complex networks, can include scores of photodiode arrays, with amplifiers, logic counting networks, scanning circuits, etc., to give such outputs as digital pulse rates proportional to illumination, logic for character recognition and punched-card reading, and as replacements for photomultipliers. These devices find a wide range of application in transducer instrumentation for control and communication.

11.12.12 Nuclear radiation sensors

Radiation sensors are required in particle physics research, for monitoring reactions in nuclear power stations, as gauges for industrial processes, and for measurements employing radioactive isotopes. The emanations used for industrial processes include α, β- and γ-rays with X-rays for special applications such as the examination of welds; the sensor characteristics largely determine the radiation to be processed. The devices include the following.

Low-pressure gas ionisation chambers to measure α-particles by accumulating the electrons released by their collisions with gas molecules, and without subsequent multiplication.

Proportional counters, similar to the foregoing but with a higher accelerating voltage (500–800 V) to give cumulative gas amplification (Townsend avalanche) and an electron gain of 10^5–10^6. The use is for counting α-particles in the presence of β-particles and γ-rays.

Geiger counters operating at 800–1500 V, or at low voltage, for β- and γ-ray detection. The complete ionisation must be quenched between counts, and the dead time restricts the counting rate to about 1000/s.

Scintillation sensors (or 'counters') for α, β- or γ-detection. There are also semiconductor sensors for the same application and for X-rays.

α-, β- and γ-Ray sensors require to be followed by electronic pulse amplifiers and counters. A counter may accumulate a total/fractional scaled count, or it may give a rate-meter display. These outputs, together with pulse height analysers, are used to assess the energy spectra of the received signals. Coincidence counters have multiple input sensors so arranged that only the required type of emanation is measured during any selected period.

The detectors can form the bases for industrial gauges and monitors when used with radioisotope sources. The selective absorption of materials enables β-radiations through paper to indicate the uniformity of the product; liquid-level gauges give correcting signals when the rise of level absorbs the radiation; radioisotope tracers with short half-lives can be introduced into liquid and gas channels to indicate flow rate, uniformity of mixing and the detection of leakages.

11.13 Data recording

Chart recorders responsive to signals at frequencies up to a maximum of 25 kHz include: (1) a.c. ultraviolet, (2) $X-Y$, and (3) analogue and digital strip-chart forms. All provide records of related phenomena that need to be preserved in graphical form.

The visual discrimination of the best-quality trace, when stated as a fraction of the chart width, should not be significantly different from the accuracy and linearity of the display; this condition dictates the width of the chart paper. Many different types of treated chart paper are needed in order to be compatible with the various writing systems mentioned below. Paper is provided in rolls or overlapping folds, the latter being very convenient for quick retrieval of data.

Conventional pens for ink-feed methods use felt, ball-point or nylon disposable tips. The ink cartridges are easily replaced, and the supply is drawn to the tip by capillary action. The nylon pen is suitable for higher writing speeds; to retain a uniform trace, a variable pressure is automatically applied to the ink cartridge. In multiple-display applications, the possibility of easy trace separation by colour is a unique convenience of ink writing.

Many alternative writing methods have been developed in an attempt to overcome the inertia (response time) limitation of conventional ink writing, as well as to improve the quality of the trace. These include: (a) a heated stylus and sensitive plastic-coated paper; (2) mechanical pressure on chemically treated paper; (3) ultraviolet light beams directed on to photographic paper, the trace appearing and becoming fixed within a few seconds of exposure to natural light; (4) the passage of current through treated metallised paper causing the chemical reduction process to develop a black imprint; and (5) electrostatic copying, in which a flat capacitor is formed from conductive paper with a dielectric coating—a very small area can be charged, and particles are attracted to the charged area to provide the visible trace, which may require treatment to make it permanent.

References

1 BRITISH STANDARDS INSTITUTION, *BS 5233 Glossary of terms used in metrology*, Milton Keynes (1986)

2 KIBBLE, B. P., ROBINSON, I. A. and BELISS, J. H., *Metrologia* 27, 173–192 (1990)

3 http://www.npl.co.uk/npl/cem/dclf/index.html

4 GALAKHOVA, O. P., HARKNESS, S., HERMACH, F. L., HIRAYAMA, H., MARTIN, P., ROZDESTVENSKA, T. H. and WILLIAMS, E. S., *IEEE Transactions on Instrumentation and Measurement*, **IM-29**, 396–399 (1980)

5 EMMENS, T., *IEE Electronics and Power*, 166 (February 1981)

6 JONES, L. T., RESSMEYER, J. J. and CLARK, C. A., *Hewlett-Packard Journal*, **32**(4), 23 (1981)

7 IKEDA, Y., *NPL DES Memorandum No. 21* (1976)

8 TURGEL, R. S., *NBS Technical Note No. 870* (June 1975)

9 TURGEL, R. S., *IEEE Transactions on Instrumentation and Measurement*, **IM-23**, 337–341 (1974)

10 STOCKTON J. R., *IEE Electronics Letters*, **13**(14) 406–407 (1977)

11 KNIGHT, R. B. D. and STOCKTON, J. R., *NPL Reports DES60* (1981)

12 STOCKTON, J. R. and CLARKE, F. J. J., *DES71* (August 1981)