In one sensor of this type the sharp melting line of Tempilaq paint on the surface of a tapered resistor provides an analog indication of the maximum power [42]. Thermooptical sensors on the other hand exhibit reversible colour changes. A number of cholesteric liquid crystals for thermooptical sensors are listed in [43]. Figure 30 [44] shows a thermooptical sensor with a tapered resistor. Recently this type of sensors has been incorporated into battery labels to measure a (small) discharge current [45] through a known resistance and to derive the battery voltage in order to assess the charge condition.

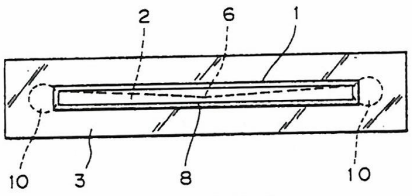


Figure 30 [44]

Voltage Measuring Sheet

The shape of the tapered resistor (2) with a narrow central portion (6) is indicated by the dashed line. A layer (1) of a decolouring agent that loses its colour at about 40 °C is in thermal contact with the conductor and is related to a scale pattern.

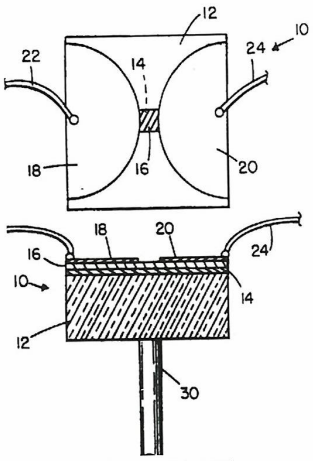


Figure 31 [46]

Thermooptical Current Sensor

The current to be measured is applied (18, 20, 22, 24) to a resistive or semiconducting element (16) to increase the temperature of this element and of the optical resonator (14). The change of the optical property of the resonator is detected through the modulation of the wavelength spectrum of light directed into and reflected from the resonator through the optical fiber (30) abutting the transparent substrate (12).

Optical resonators with temperature dependent optical properties can be used to sense a range of currents. In the device shown in Fig. 31 [46] a current is measured through the modulation of the wavelength spectrum of light by a temperature dependent optical resonance property of the sensing element. This particular sensor has been optimised for a short response time.

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1.2.5 Sensing the Energy of the Magnetic Field

A conductive ring that surrounds a current carrying conductor contains a magnetic flux. The energy stored in a ring with volume V (i.e. cross section a times circumference) is:

$$W = \frac{1}{2} B H V = \mu \mu_o \frac{a^2 I^2}{V} \quad (29)$$

When such a ring is interrupted by an air gap, the stored energy (given by a similar expression as above) leads to an attractive force between the pole pieces of the gap. Different methods can be employed to measure this force, which depends on the current I, e.g. a measurement of the strain on the ring, with a strain gauge or measurement of the movement of the pole piece with optical means [47]. With a variant of the first method current pulses between 1000 and 250 000 A have been measured in a resistive spot welding machine [48].

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1.2.6 Measuring the Lorentz Force on the Conductor

Charge carriers that move at an angle, α, with respect to magnetic field lines are deflected. The deflecting force acts in a direction perpendicular to the plane formed by the directions of the particle's velocity and the magnetic field. The force F on a charge q that moves with a velocity v in a magnetic field with a flux density B is given by the vector product:

$$\vec{F} = q\,(\vec{v} \times \vec{B}) \tag{30}$$

There is some confusion in the terminology: German texts refer to F as the Lorentz force, in other texts the electric field dependent (q E) term often is included in the expression for the Lorentz force. Rewriting Equation 30 in terms of the current I passing through a conductor of a given length ℓ, one obtains the force that acts on that conductor due to the magnetic flux density B:

$$\vec{F} = I\,(\vec{\ell} \times \vec{B}) \tag{31}$$

Since electric currents give rise to magnetic fields, a force acts between two parallel current carrying conductorss:

- an attractive force, if the currents flow in the same direction, and
- a repulsive force, if they flow in opposite directions.

The strength of the force F (in N) on each conductor can be derived from Equations 22 and 31:

$$F = \mu_o \frac{I_1 I_2 \ell}{2\pi r} = 2 \times 10^{-7} I_1 I_2 \frac{\ell}{r} \tag{32}$$

where I_1 and I_2 (in A) are the respective currents and r (in m) is the distance between the conductors. Different arrangements are used to exploit the Lorentz force on the current carrying conductor for current measurements.

They differ by the source of the magnetic field and how the Lorentz force is measured. Measurements according to Equation 31 rely on a known external magnetic field. When the force between two conductors is measured according to Equation 32 one may either work with one known current (I_1 or I_2) or feed the same current through both conductors ($I_1 = I_2$). In the later case the force is a quadratic function of the current.

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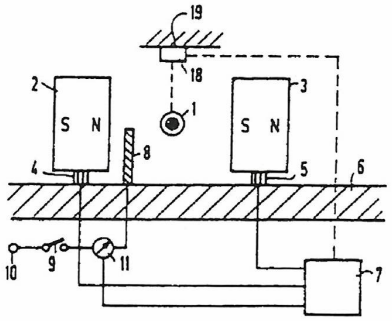


Figure 32 [49]

Measuring Electric Current in Conductor

A high current conductor (1) passes between two magnets (2, 3). The Lorentz force on the magnets is sensed with transducers (4, 5). In an alternative embodiment, the force on the conductor (1) is measured (18). Conductor (8) is used for calibration.

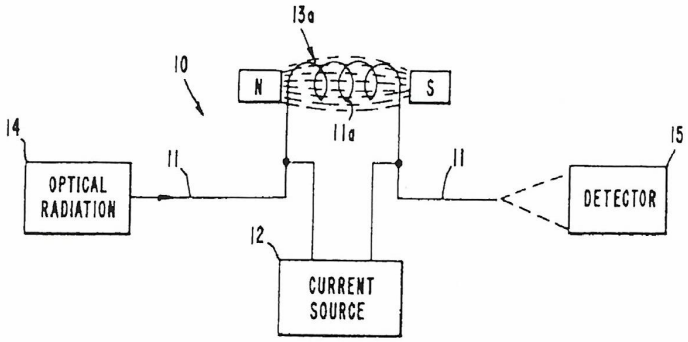


Figure 33 [50]

Fiber Optic Transducer for Measuring Current or Magnetic Field

A multi-mode optical fiber, which serves as the sensor, is composited with a metal capable of conducting electricity. Optical radiation is introduced into the fiber from a source which may be either coherent or incoherent. An electrical current is applied to a portion of the electrically conducting fiber and a magnetic field is applied to the current-carrying fiber. A known value of one permits a determination of the presence or absence of the other through electromotive forces on the metallic conductor in the magnetic field, which induce differential phase shifts (coherent optical radiation input) or losses (incoherent optical radiation input) between the fiber modes. These phase shifts or losses are detected by a suitable detector. For a magnetic flux density of 10^{-1} T good linearity is obtained in the 5 to 2,000 mA current range.

In all these setups the Lorentz force on at least one of the interacting components, i.e. one conductor or the source of the magnetic field, is measured. Various methods to measure the force may be used, e.g. spring loaded meters, piezoelectric sensors, resilient members or stress induced phase shifts in optical fibres.

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Moving coil instruments are the best known devices of this type (for moving iron instruments cf. Section 1.2.13.3). They consist of a rotatable coil mounted in the gap of a magnet. Direct current that flows through the coil gives rise to an angular force on the coil. With appropriate springs to bias the coil, a linear relation between current and angular deflection can be obtained.

The related figures illustrate
Fig. 32 [49]: an arrangment with either moving magnets or a moving conductor;
Fig. 33 [50]: a moving conductor with fixed magnets and a fibre sensor;
Fig. 34 [51]: two moving conductors with a piezoelectric transducer.

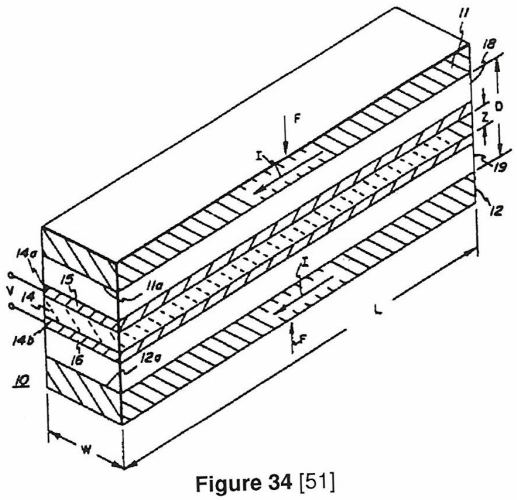


Figure 34 [51]

Current Sensor

A current sensor, integratable with power switching and control circuitry on a common substrate and insulatable from current-carrying conductors, utilizes a piezoelectric element (14) located between a pair of current-carrying conductors (11, 12). The conductors carry a current flowing in the same direction, causing an attractive force there between to act upon the piezoelectric element (14) and generate a voltage proportional to the square of the conductor currents. With a pair of sensors, each in a different portion of a half-bridge circuit, and associated electronics a sensor output proportional to the instantaneous value of the sensed current is obtained.

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1.2.7 Hall Elements

The Hall effect describes the action of an external magnetic field with a given flux density, B_z, on a current, I_x, in a conductive plate. The Lorentz force, F_y, acting on the charge carriers moving with a velocity v_x deflects them in the y direction. The magnitude of the resulting electric field, E_y, is such that the electrostatic force on the charge carriers q compensates the Lorentz force:

$$q\,E_y + (-\,q\,v_x\,B_z) = 0 \tag{33}$$

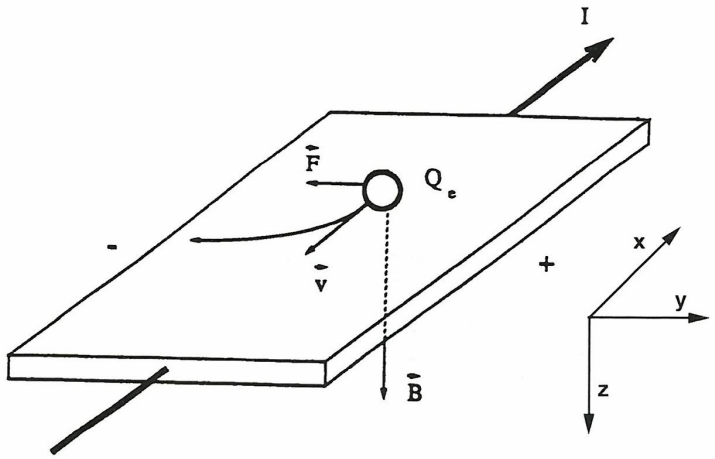


Figure 35 [303]

Hall Effect

The coordinate system has been added in this figure to illustrate the equation above:
Current I_x results from electrons ($q = -|e_o|$) moving with a velocity $v_x = -|v_x|$. A magnetic field $B_z = |B_z|$ gives rise to the Lorentz force $F_y(B_z) = -(-|e_o|)(-|v_x|)|B_z| = -|e_o\,v_x\,B_z|$. The resulting electric field $E_y = -|E_y|$ in turn gives rise to a force $F_y(E_y) = (-|e_o|)(-|E_y|) = |e_o\,E_y|$.

The electric field gives rise to a voltage U_H (Hall voltage) along the y direction. Its magnitude depends on the thickness on the conductor, d, and the Hall coefficient, R_H, of the material:

$$U_H = R_H \frac{I_x\,B_z}{d} \tag{34}$$

When the charge carriers are electrons, the Hall coefficient is negative, about $-5\cdot10^{-11}$ m³/C for copper and up to -10^{-4} m³/C for InAs-semiconductors [1].

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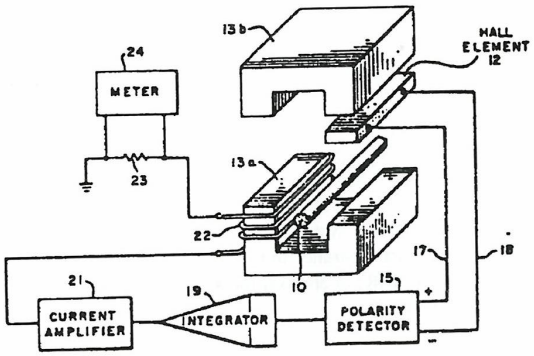


Figure 36 [54]

Sensing System for Measuring a Current

A current flowing through a conductor, is measured by initially passing an energizing magnetic flux through a Hall element (12) to produce a Hall voltage proportional to the sensed current. To eliminate temperature, gain and non-linear effects normally associated with a Hall element, the amplitude of the Hall voltage is effectively ignored and only is polarity is detected (15) and employed to control the operation of an integrator (19) to develop a cancelling current to establish an equal, but opposite cancelling flux in the Hall element. With a zero net flux, the integrator will hold the cancelling current at the level required to cancel the initial energizing flux. Since the cancelling current needed to null the energizing flux will be proportional to that flux, the current may be used to produce (23) a control voltage to represent the measured currrent.

Typical values for the thickness, d, the length, x, and width, y, of Hall elements are: $1 < d < 100$ µm, $0.5 < x,y < 10$ mm. The current, I_x, is chosen such that the resolution is highest but excessive heating of the Hall element is avoided. Usually values below 1 A are used. Solid state, GaAs ion implanted devices achieve a resolution of 1 V/T with $I_x = 5$ mA [52].

When direct currents are to be measured, the offset voltage of the Hall device has to be corrected for. A circuit suitable for this purpose is described in [53]. Two applications of Hall elements are shown in Figs. 36 [54] and 37 [55].

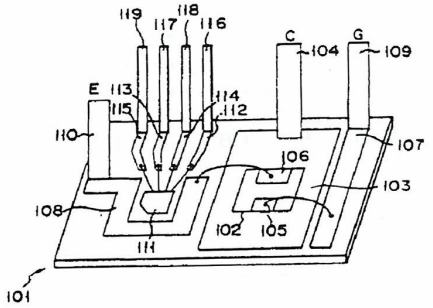


Figure 37 [55]

Composite Semiconductor Device with Overcurrent Detection

An insulating substrate (101) includes a conductive layer (103) on which a power semiconductor element (102) is mounted through its main electrode. The other main electrode of the semiconductor element is connected to one end of a U-shaped conductive layer (108) such that a load current through the semiconductor element also flows through the conductive layer (108) and produces a magnetic field which is detected by a current detector (111) such as a Hall element.

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1.2.8 Magnetoresistance Elements

Magnetoresistance is responsible for the change of resistance in metals and semiconductors when they are exposed to magnetic fields. Different materials and arrangements of materials exhibit different magnetoresistance effects. Magnetoresitance is measured by passing a sense current through the magnetoresistive material. Thus far the following effects have been studied, the following brief discussion is based on the references indicated:

- magnetoresitance in semiconductors like InSb, NiSb [56];
- anisotropic magnetoresistance (AMR) in ferromagnetic materials like Permalloy [57];
- giant magnetoristance (GMR) in systems in which the magnetic moments in different layers or clusters change between an antiparallel and a parallel orientation [58];
- giant magnetoresistance (GMR) in epitaxal films [59] (somtetimes also referred to as "colossal magnetoresistance").

For comparitive purposes these effects are best characterised by the ratio $\Delta R/R$ with the corresponding magnetic field range. Two definitions for the ratio $\Delta R/R$ are used concurrently:

$$\Delta R/R_H = (R_{H=0} - R_H)/R_H$$
$$\Delta R/R_{H=0} = (R_{H=0} - R_H)/R_{H=0} \qquad (35)$$

where $R_{H=0}$ and R_H are the resistance values in the absence of an external magnetic field or current to be measured and when the maximum field for the respective effect is applied respectively. If $\Delta R/R$ is small, both values are close to each other. Since the measurement of currents usually consists of the determination of the magnitude of the currrent (as compared to no current), we shall refer to $\Delta R/R_{H=0}$ in the following and convert the published values accordingly.

Semiconductor sensors are sensitive to external magnetic fields perpendicular to the direction of the sense current. The field dependence of a semiconductor magnetoristance can be described by:

$$R(B) = R_o + a\,B^2 \qquad (36)$$

To obtain a linear relationship between the resistance and the magnetic field one can bias the sensor with an external bias field $H_o >> H_x$, where H_x is the external field to be measured. If the H_x^2 term is negligible, the following linear relationship holds between the resistance R and the flux density B_x:

$$R(B) = (R_o + a\,B_o^2) + 2\,a\,B_o\,B_x \qquad (37)$$

In the corresponding reference [56] a linear $\Delta R/R_{H=0}$ response of $\approx \mp 1.1$ % for $\Delta B = \pm 2$ mT corresponding to $\Delta I = \pm 1$ A is reported. Figure 5 of this reference, however, suggests that the unlinearised $\Delta R/R_{H=0}$ value is considerably higher, i.e. because of the above sign convention for $\Delta R/R_{H=0}$ has a more negative value.

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The variation of the specific resistivity, ρ, of an anisotropic magnetoristance like a premagnetised Permalloy strip follows a $\cos^2\theta$ dependence, where θ is the angle between the direction of the sense current and the direction of magnetisation of the strip:

$$\Delta\rho = \rho - \rho_{\perp} = \Delta\rho_{max} \cos \Theta^2 = (\rho_{\parallel} - \rho_{\perp}) \cos \Theta^2 \quad (38)$$

where $\rho_{\perp}$ and $\rho_{\parallel}$ are the specific resistivities for orthogonal and parallel orientation of the directions of the sense current and the magnetisation of the strip, respectively. The strip typically consists of a 20 - 100 nm thin evaporated or sputtered layer of Permalloy which initially is magnetised along the strip parallel to the direction of the sense current. When an external magnetic field is applied in a plane orthogonal to the strip, the direction of magnetisation in the sensor plane is rotated whereby the angle θ increases and the resistance decreases. The variation of the resistance vs. the magnetic field follows a bell shaped curve. Typical value for the saturation field strengths at which the magnetisation aligns with the applied field are between ≈ 10^2 and 10^5 Am^{-1}. In the cited example [57] a curve with a nonlinear response of $\Delta R/R_{H=0} \approx 2$ % for $\Delta B = \pm 2$ mT ($1.6 \cdot 10^3$ Am^{-1}) is shown. A linearisation of the response over a limited field range can be achieved by arranging highly conductive equipotential strips (barber poles) at 45° with respect to the direction of original magnetisation on the sensor strip.

With zero field measurements, in which a feedback circuit responsive to the voltage drop across the sensor controls a compensation field (current), no linearisation of response of the sensor is required. Suitable bridge arrangements help to reduce the effects of the (considerable) temperature dependence of the magnetoresistance in semiconductors and the AMR in permalloy of about 0.06% $°K^{-1}$ and 0.25% $°K^{-1}$ respectively. One example each of a sensor based on semiconductor magnetoresistance and AMR is shown in Figures 37 [60] and 38 [61].

GMR on the other hand leads to considerably higher ΔR/R values. So far GMR sensors primarily have been used as magnetic field sensors. They are particularly useful for applications where a high signal-to-noise ratio is required like in reading heads for magnetic data carriers. Probably because of their still limited $\Delta R/(R \cdot \Delta H)$ resolution, which at room temperatures is not much better than that for AMR sensors, these sensors have not yet been employed on a larger scale for current measurements. Nevertheless some of their properties are summarized below.

The first type of giant magnetoresistances comprises layers or clusters of ferromagnetic materials separated by non-magnetic materials. In the absence of a magnetic field the magnetic moments of neighbouring domains are antiparallel. In this configuration the resistance is highest. Upon application of a magnetic field the relative orientation of the magnetic moments of the different changes until, at a specific saturation field strength, H_S, all are parallel. Resistance is lowest in this configuration. According to a simple model this change in resistance is due to the spin dependence of the electron scattering in the magnetized domains. Spin-dependent scattering on the interfaces between the different materials is also thought to contribute to GMR.

Two commonly used structures for this first type of giant magnetoresistances are the following:

- multilayer structures of up to about 50 ferromagnetic and non-magnetic layers in alternation (like Fe/Cr or Co/Cu). By an appropriate choice of the thickness of the non-magntic layers antiferromagnetic coupling between neighbouring magnetic layers, i.e. their antiparallel orientation in the absence of a field, can be achieved. Typical values for the thickness of the layers which commonly are produced by sputtering are in the order 10^{-9} m. With Co/Cu multilayers a nonlinear response of $\Delta R/R_{H=0} \approx 41$ % has been observed for $\Delta B \approx \pm 1$ T at room temperature [58].
- grannular allloys containing nanometer sized clusters of ferromagnetic material.

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The second type of giant magnetoresistances consists of expitaxial, single crystaline La-Ba-Mn or La-Ca-Mn oxide films. They are produced by laser deposition and are about 100 to 200 nm thick. The $\Delta R/R$ ratios of both types of giant magnetoresistances are highly temperature dependent. Because of thermal fluctuations the ratios are generally much higher at temperatures as low as 4.2 °K than at room temperatures. The properties and in particular the temperature dependence of the $\Delta R/R$ value of the second type of giant magnetoresistances can be significantly influenced by annealing, i.e. the particluar annealing conditions. With annealed La-Ca-Mn-O films a nearly isotropic, nonlinear $\Delta R/R_{H=0}$ response of ≈ 93 % has been observed for $\Delta B \approx \pm 6$ T at 260 °K [59].

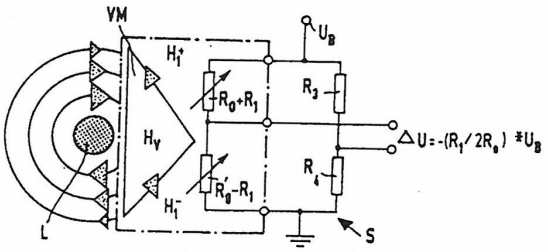


Figure 38 [60]

Test Arrangement for Electronic Circuits

The device to measure currents in bus-like conductors forms part of a small probe head. It comprises a serial arrangement of two biased magnetoresistances that sense one component of the magnetic field, H_1, only. When a bias field H_V parallel to H_1 is applied, the magnetoresistances sense the sum and the difference of H_V and H_1 respectively.

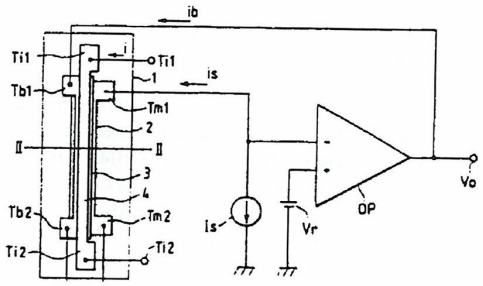


Figure 39 [61]

Current Sensor System

The sensing portion is composed of a ferromagnetic magnetoresistance element (2), a bias conductor (3) and a current conductor (4) all arranged on an insulating substrate (1). When a current flows through the conductor (4) the resistance of the magnetoresistance element changes accordingly. An amplifier (OP) provides a feedback current to cancel the change of resistance and an output signal (V_o).

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1.2.9 Magnetooptical Sensors

1.2.9.1 The Magnetooptical Effect

There are three effects that change the optical properties of materials in magnetic fields:
- the Faraday effect, i.e. the rotation of the polarisation of linearly polarised light
- the Cotton-Mouton effect, i.e. field induced birefringence
- the magnetooptic Kerr effect, i.e. the magnetisation dependent rotation of the polarisation of polarised light on reflection from magnetised materials

Nowadays mainly the Faraday effect is employed for current sensing. For the sake of completeness we will briefly describe the Cotton-Mouton effect as well. The magnetooptic Kerr effect, which is exploited rather for the determination of the magnetisation of ferrites than for the measurement of currents, will not be covered here.

Liquids with anisotropic molecules exhibit the Cotton-Mouton effect, i.e. birefringence in a transverse magnetic field. The phase difference induced between the light components, parallel and perpendicular to the magnetic field, is proportional to the square of the applied magnetic field. For a light wave that traverses a Cotton-Mouton cell of a given length length, ℓ, exposed to a magnetic field, H, the phase shift, δ, between the respective components amounts to:

$$\delta = 2 \pi \ell C H^2 \qquad (39)$$

C is the (temperature dependent) Cotton-Mouton constant ($C \approx 4 \cdot 10^{-14}$ m/A^2 for nitrobenzene [1]). A magnetic field of $\approx 4 \cdot 10^7$ A/m along a 10 mm cell causes a phase shift of about π. To measure the phase shift polariser/analyser arrangements like for Kerr cells (cf. Section 1.1.7.3) can be used. Most liquids that exhibit the Cotton-Mouton effect also exhibit the Kerr effect.

1.2.9.2 Faraday Sensors

Optically active materials (e.g. sugar solutions) rotate the polarisation of linearly polarised light. Under the influence of a magnetic field inactive isotropic materials rotate the polarisation of a light beam parallel to the field vector. The fact that only the field component parallel to the direction of propagation is rotated is expressed by the scalar vector product:

$$\alpha = V \int \vec{H} \, d\vec{\ell} \qquad (40)$$

where α is the angle of rotation (in degrees), H the magnetic field vector, $d\ell$ represents the path of the light in the sensor, V is the Verdet constant of the sensor material.

Two types of Faraday sensors are employed: optical (silicate) fibres and bulk material. In both cases light from a (laser) source is polarised before being sent through the sensor. To determine the magnetic field (current), one measures the rotation angle at the output of the sensor. Suitable materials for sensors preserve the polarisation of the light beam and do not rotate the plane of polarisation in the absence of a magnetic field. High signals can be obtained with fibres wound several times around the current carrying conductor. In order to preserve the polarisation, single mode fibres are used.

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Some bulk materials, however, have a much higher Verdet constant than fibres. Typical values for fibre sensors are $V = 2.5 \cdot 10^{-4}$ deg/A [62] and $\ell = 1$ m. Thus a magnetic field of $\approx 2 \cdot 10^5$ A/m is required to observe a phase shift of about $\pi/4$.

If the sensor forms one loop around a conductor that carries a current I, Equation 21 can be used to express the integral in Equation 40 in terms of I:

$$\alpha = V I \tag{41}$$

For a sensor placed inside a coil with N_C turns Equation 23 yields:

$$\alpha = V N_C I \tag{42}$$

Light that is reflected back at the end of the sensor and passes the sensor in the opposite direction undergoes a second Faraday rotation. This feature is referred to as "nonreciprocal" and can be derived from Equation 40. Most distortions are reciprocal and thus cancel when reflection is used.

We will shortly mention the various sources of error and the different methods employed to measure the Faraday rotation, a comprehensive treatment with the relevant references can be found in [62]. Impurities, bends and stress induce linear birefringence, torsion leads to circular birefringence. Vibration is another source of error. By twisting a bent fibre one can compensate the effects of linear birefringence through circular birefringence. The value of the Verdet constant varies with temperature and frequency, e.g. for optical fibres $(dV/dT) \approx V \cdot 10^{-4}\ ^\circ K^{-1}$. Paramagnetic and ferromagnetic bulk materials with a much higher Verdet constant than silica fibres exhibit a more pronounced temperature dependence. Diamagnetic glasses, however, offer higher Verdet constants at a reasonable temperature stability [62].

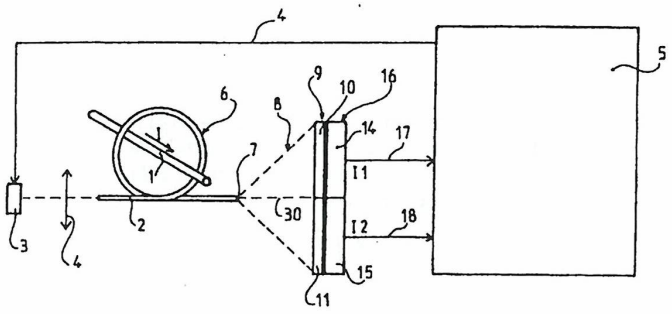


Figure 40 [63]

Device for Measuring Current with an Optical Fibre

The device measures, using the Faraday effect in a monomode optical fibre (2), the intensity (I) of the current in a conductor (1). It comprises a polariser-splitter (9) constituted, for example, by two coplanar half-discs (10,11) of polarising sheet, the preferred polarisation orientations of which make the desired angle 90° or 45° . Photodiodes (14, 15) constituted by a two-quadrant photodetector (16) in the form of a disc of the same diameter as that (9) formed by the polariser-splitter, are arranged downstream of the polariser-splitter. The phodiodes are connected to an electronic unit (5).

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The simplest way to measure the Faraday rotation consists in an arrangement with crossed or aligned polariser and analyser at the input and the output of the sensor. The intensities observed have a $\sin^2\alpha$ or $\cos^2\alpha$ dependence. The difference signal from two analysers at $\pm 45°$ relative to the polariser follows a $\sin(2\alpha)$ dependence. To achieve higher resolutions various interferometer designs can be used, e.g. Sagnac, Fabry Perot. With two light beams of different frequencies heterodyning techniques can be employed. This transforms the angle of the Faraday rotation into a phase modulation of a high frequency signal. Two applications of the Faraday effect are given in Figs. 40 [63] and 41 [64]. With the device shown in Figure 41 currents between 10 and 150 mA have been detected with a linear response in the frequency range from 25 to 125 Hz.

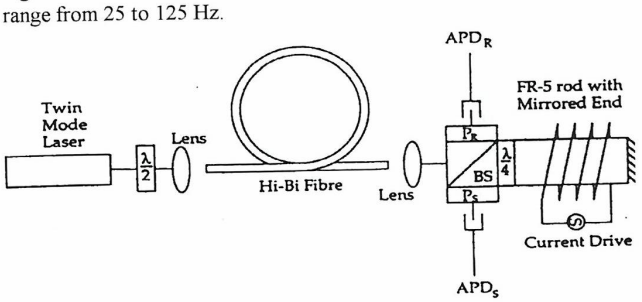


Figure 41 [64]

Current Sensing Using Faraday Rotation and a Common Path Optical Fiber Heterodyne Interferometer

Orthogonally polarised outputs of a two-frequency laser source are rotated with a $\lambda/2$ plate to the eigenaxis of a birefringent monomode fibre and transferred to the bulk sensor which is made of glass with a Verdet constant of $\approx 5 \cdot 10^{-3}$ deg/A. Part of the beam is directed into a reference detector (APD_R). The other part passes a $\lambda/4$ plate. The two emerging (orthogonal) circularly polarised modes pass the sensor, are reflected back, and pass the sensor again. They are detected after a back-transformation into linearly polarised states (through the $\lambda/4$ plate) at the signal detector (APD_S). The angle of the Faraday rotation is recovered electronically.

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1.2.10 Magnetostrictive Sensors

1.2.10.1 Magnetostriction

Ferromagnetic materials are elastically deformed when exposed to magnetic fields:

$$\frac{\Delta L}{L} = g\,H \tag{43}$$

For Ni g is in the order of 10^{-9} m/A [65]. For Fe, Co, Ni the deformation reaches values of $\Delta L/L \approx 10^{-5}$ at saturation. With magnetostrictive current sensors the current is derived from this deformation (length variation). All methods for length measurements with an appropriate resolution can be used. As with piezoelectric sensors (cf. Section 1.1.8) different approaches are employed to measure this deformation:

- through the change of resistance of resistive leads attached to the magnetostrictive material;
- through phase shifts (due to the length variation of the fibre or to strain induced birefringence in the fibre) in a light beam transmitted through an optical fibre bonded to the crystal;
- through the delay of a surface acoustic waves.

An example for the first approach in which, to increase the signal, the conductor has a meander like shape is shown in Fig. 42 [66].

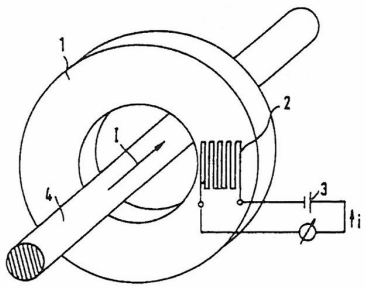


Figure 42 [66]

An annular body (1) which consists of magnetostrictive material surrounds the conductor (4). Conductor tracks (2) made of length dependent resistance are applied on the annular body (1). The resistance of the conductor tracks is a measure of the current (I) through conductor (4).

1.2.10.2 Magnetostrictive Materials with Optical Fibres

Measurements of length variations with optical fibres have been treated in detail in Section 1.1.8.2. Piezoelectric and magnetostrictive sensors with optical fibres look alike. However to avoid distortions due to the Faraday effect (cf. Section 1.2.9.2), the direction of propagation of the light must be perpendicular to the magnetic field vector.

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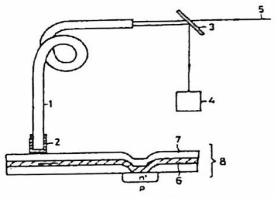


Figure 43 [67]

Measuring Device for Current on Wiring of Intergrated Circuit

When the tip of an optical fiber (1) covering a magnetostrictive material (2) is positioned on the wiring (6) of an integrated circuit (8), the material (2) is strained by a magnetic field generated by a current on the wiring, and thereby a stress is given to the fiber. A laser beam (5) in the fiber is modulated when it is reflected on the end surface and its propagation mode is changed. The beam is reflected by a half mirror (3) and falls on a photodetector (4) which delivers an output corresponding to the change in the propagation mode, and a current can be read from this output.

Certain materials like e.g. silica fibres exhibit the photo-elastic effect, i.e. stress induced birefringence. When such a fibre is wound around a magnetostrictive element, the length variation causes stress such that the magnetic field can be derived from the induced birefringence. With a nickel cylinder on which a monomode silica fibre of length ℓ was wound and through which a polarized light beam ($\lambda = 0.633$ µm) was propagated, the change of the state of polarisation Ψ has been measured. With a suitable detector for two polarisation states and an appropriate magnetic bias field ($\approx 5.6 \cdot 10^3$ A/m) a linear response $\Delta\Psi/(\ell \cdot H) \approx 2.2 \cdot 10^{-4}$ rad A^{-1} has been observed for H between 0.2 and 80 A/m [65]. In this field range a 200-mm long sensor would for example have of response of about $4 \cdot 10^{-5}$ rad A^{-1}. In a more qualitative measurement only the change of the propagation mode is observed (Fig. 43 [67]).

1.2.10.3 Magnetostrictive Materials and Surface Acoustic Waves

A surface acoustic wave sensor for current measurements is a resonance circuit, whose resonance frequency depends on the length of a magnetostrictive delay line. It consists of a transmitter and receiver with a feedback loop and a magnetostrictive delay line. Typically the delay line is made of Yttrium-Iron-Garnet (YIG) and Gallium-Gadolinium-Garnet (GGG), its length and thickness are several mm and µm respectively. Since the dependence of the resonance frequency on the magnetic field is linear only for a limited range of magnetic field strengths, a compensation coil is required to limit the field to be sensed to this range. In the arrangement shown in Fig. 44 [68] the sensor is mounted in the gap of a magnetic core with two coils, one for the current to be measured and the other one for a compensating current.

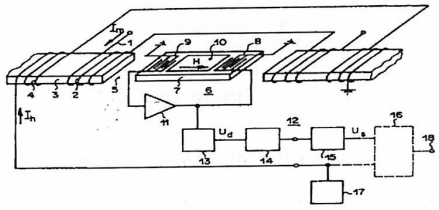


Figure 44 [68]

Non-Contact Current Measurement

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1.2.11 Magnetic Resonance

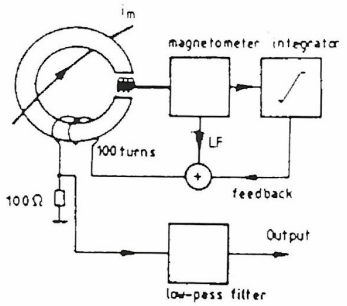


Figure 45 [70]

ESR Current Sensor

The ESR probe is made of AcTCNQ$_2$, a material that is rather cheap and easy to handle. With this sensor a current resolution of about 10 mA Hz$^{-1/2}$ has been achieved.

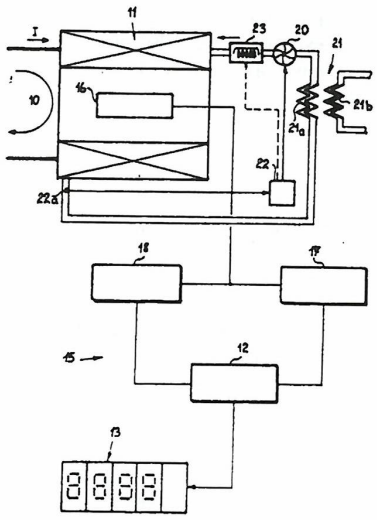


Figure 46 [71]

Device for Measuirng a Direct Current Intensity

A coiled magnet (11) is mounted in series in the circuit (10) wherein the current to be measured circulates, the temperature of the magnet is stabilized and the magnetic field is measured at the centre of the magnet by means of a NMR probe (16).

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Magnetic resonance is a well-known technique in material sciences and for chemical analysis (electron spin resonance, ESR, and nuclear magnetic resonance, NMR). When an electron/nucleus is exposed to a magnetic field, its magnetic energy levels split according to the orientation of its magnetic moment (parallel or antiparallel to the magnetic field). An external RF field perpendicular to the magnetic field stimulates transitions between the energy levels. The energy difference ΔE is given by:

$$\Delta E = g \mu B = h f_o \tag{44}$$

For electrons/protons the value for the g factor is about 2/5.58 and μ corresponds to the Bohr magneton, $\mu_B = 9.27 \cdot 10^{-24}$ J/T or the nuclear magneton, $\mu_N = 5.05 \cdot 10^{-27}$ J/T [1], h is the Planck constant, f_o is the Larmor frequency, i.e. the frequency of the precession of the spin. Its value for electrons and protons, f_e and f_p, can be derived from the expression for the energy difference:

$$f_e = 2.8\ 10^{10}\ Hz/T; \quad f_p = 4.3\ 10^7\ Hz/T \tag{45}$$

In spin resonance magnetometers (current metres) a test probe is exposed to the magnetic field and the field is determined from the Larmor frequency. When the frequency of an inductively coupled RF circuit is swept, absorption is observed at the Larmor frequency. The frequency is in the GHz/100 MHz range for ESR/NMR. In general NMR magnetometers are used for flux densities down to several mT [69]. Modern ESR systems give a resolution down to the nT range. NMR is observed in nuclei with a nonzero nuclear spin (e.g. hydrogen s=1/2) only. To achieve accurate results, both temperature stability and field homogeneity over the whole probe have to be ensured. One example each for an ESR and a NMR current sensor is shown in Figures 45 [70] and 46 [71].

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1.2.12 SQUIDS

Superconducting **QU**antum **I**nterference **D**evice**s** (SQUIDS) have been used for the measurement of magnetic fluxes and related quantities since the mid-1960s. SQUIDS consist of a superconducting ring interrupted by one or two Josephson junctions respectively (RF and DC SQUIDS). They act as magnetic flux sensors with a resolution of about 10^{-19} Wb. We will briefly summarize some of their basic properties on the basis of two review articles [72], [73] where further details can be found. We start with RF SQUIDS because they are simpler to understand. Commercially RF SQUIDS were available long before DC SQUIDS. Today, however, because of their higher accuracy (see below) DC SQUIDS are more common. The discovery of high temperature superconductors helped to promote SQUIDS.

When a moving charge is exposed to a magnetic field, its wave function acquires a phase shift. For the wave function to be single valued, the phase shift accumulated over a closed loop must be a multiple of 2π. The pase shift θ acquired by a Cooper pair (two electrons) circulating in a superconducting ring is related to the magnetic flux Φ_i through the hole surrounded by the ring:

$$\theta = 2\pi \frac{2e}{h} \Phi_i = 2\pi n \tag{46}$$

(h is the Planck constant and e the charge of the electron).As a consequence the magnetic flux through the hole is quantized as well - in units of Φ_o
($\Phi_o = 2 \cdot 10^{-15}$ Wb):

$$\Phi_i = n \frac{h}{2e} = n \Phi_o \tag{47}$$

The application of an external flux Φ_e causes a supercurrent, i_s, to maintain the flux quantisation. The magnitude of the current is determined by the flux, $i_s \cdot L$ (L is the selfinductance of the ring), required:

$$n \Phi_o = \Phi_e + i_s L \tag{48}$$

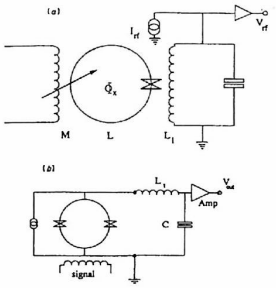


Figure 47 [73]

SQUIDS, the Josephson Effects and Measurement
(a) Schematic diagram of RF SQUID with L-C circuit readout
(b) Schematic diagram of DC SQUID readout system

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In a superconducting ring with one Josephson junction (RF SQUID) the situation changes due to the Josephson effect: the junction acts as tunnelling barrier and introduces an additional phase shift $\delta\theta$ in the wave function according to

$$i_s = i_c \sin \delta\theta \tag{49}$$

where i_c is the critical current of the junction. When a modulating RF flux $\Phi_e(t)$

$$\Phi_e(t) = \Phi_{RF} \sin \omega t + \Phi_x \tag{50}$$

is applied (with $\Phi_{RF} < i_c \cdot L$; f is typically about 20 MHz), new features appear. Both, the supercurrent i_s and the enclosed flux Φ_i become multi-valued functions of the external flux Φ_e. For $\beta = (2\pi\, i_c L/\Phi_o) > 1$ the internal flux Φ_i passes through a hysteresis loop.

While this loop is performed, a flux quantum Φ_o is absorbed and consecutively emitted. The necessary energy ($\Phi_o^2/2L$) is absorbed from the RF field. This absorption can be detected through a change in the RF voltage U_{RF} at a RF circuit inductively coupled to the SQUID. For $n\,\Phi_o < \Phi_x < (n+1/2)\,\Phi_o$, U_{RF} depends linearly on Φ_x. The dependence is given by:

$$\Delta V_{RF} \sim \omega_{RF} \sqrt{\frac{Q\, L_{RF}}{L}}\, \Delta \Phi_x \tag{51}$$

where Q and L_{RF} are the quality factor and the inductance of the RF circuit.The DC SQUID consists of a superconducting ring with two Josephson junctions that are resistively shunted through resistors (R). The device is biased with a direct current that exceeds the critical current of the junctions. The direct voltage (U) that appears at the junctions leads to an oscillation (frequency f) of the phase shift, θ, introduced at the junction (AC Josephson effect):

$$\frac{d\,\theta}{dt} = \frac{2\pi V}{\Phi_o} \tag{52}$$

When an external flux is applied, the circulating current and therefore the direct voltage becomes a periodic function of this flux. The modulation depth is $R \cdot \Phi_o/I$ for a flux variation of $\Phi_o/2$.

As SQUIDS act as very accurate flux to voltage converters in a very limited flux range ($\Phi_o/2$) only, they must be operated with some kind of compensation and have to be shielded against external fluxes, cf. Fig. 48 [74].

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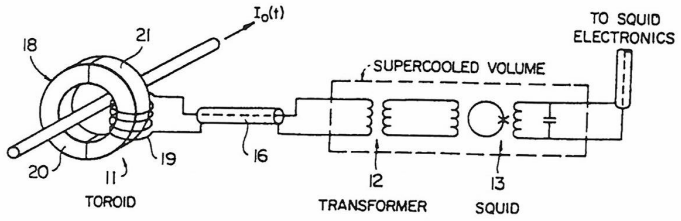


Figure 48 [74]

Method and Apparatus for Measuring Magnetic Fields and Electrical Currents in Biological and Other Systems

The measurement probe is at room temperature from where the signal is transferred into the supercooled housing of the SQUID which also contains the inductive coupling elements. Superconducting impedance matching transformers increase the resolution.

Sensitivities of several 10^{-11} T for magnetic fields (10^{-13} A for currents and 10^{-20} V for DC voltages) have been achieved. The flux resolution is about $10^{-6}\ \Phi_o\ Hz^{-1/2}$ for DC SQUIDS as compared to about $10^{-4}\ \Phi_o\ Hz^{-1/2}$ for RF SQUIDS. The resolution is worse for RF SQUIDS because the absorbed energy has to be bigger than the thermal fluctuations in the order of kT. This leads to an upper limit of about 10 nH for L at T = 10 °K.

The good voltage resolution of SQUIDS results from the fact that the voltage measurement is performed as a current measurement with a small measuring resistance. A circuit with a flux-locked loop and current compensation with a dynamic range from 10^{-13} to 10^{-9} V is described in [75].

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1.2.13 Other Magnetic Field Sensors

1.2.13.1 Transistors

Magnetic field responsive transistors operate similarly to Hall elements. They are constructed in a planar configuration and have two collectors that share the emitter current equally. When a magnetic field is applied perpendicular to the emitter/collector plane, the current is deflected by the Lorentz force. As a result the current distribution between the two collectors becomes asymmetric. A field resolution of ≈ 0.1 A/m has been achieved with this kind of device [76].

Figure 49 [77] shows a device where the measurement range can be extended by attenuation or intensification of the magnetic field.

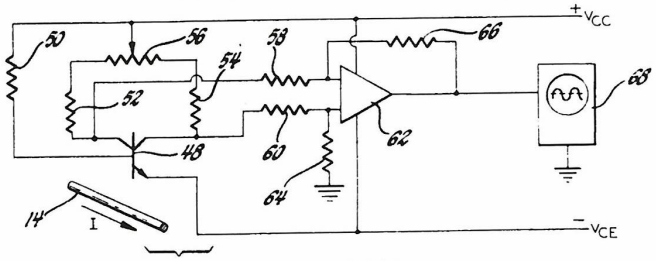


Figure 49 [77]

System for Calibrated High Level Current Measurement

Transistor (48) provides a differential output voltage proportionate to the magnetic field created by the current in the conductor (14). Calibration of the current sensor is effected by the intensifying or shunting the magnetic field to permit measurement of low and high current levels respectively. The field incident on the transistor is intensified by use of a magnetic material to direct the magnetic flux on the transistor to sense low current levels. High current levels are sensed by interposing magnetic material between the transistor and the conductor to provide a shunt path for shunting the magnetic field in a calibrated fashion such that the field at the transistor is maintained within the operational limits of the device.

1.2.13.2 Reeds

Reed switches (two blades of resilient ferromagnetic material, facing each other, sealed in a glass tube) are actuated by magnetic fields. One type opens when there is no field and closes when the field strength exceeds a specified value (e.g. $H = 10^2$ A/m). When such a reed switch is equipped with a compensating coil, it operates as a measuring instrument for (DC) currents. The reed switches whenever the superimposed fields from the current to be measured and the compensating coil exceed a predetermined value (Fig. 50 [78]).

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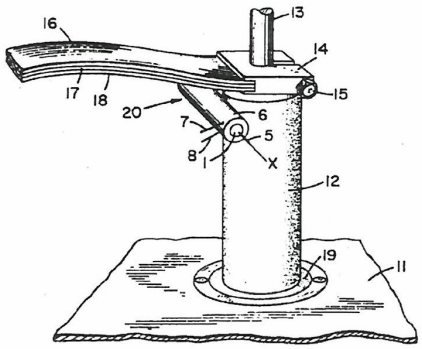


Figure 50 [78]

Current Measuring Apparatus Employing Magnetic Switch

Reed relay (20) consists of a sealed glas tube (1) which contains two reeds and is surrounded by an actuacting coil. To measure the current through the conductors (16, 17, 18), the magnitude of the current through the actuating coil is increased at a constant rate and the instant at which the reed switch changes its state is recorded. The magnitude of the current through the actuating coil when the switch changes its state is a measure of the current through the conductors.

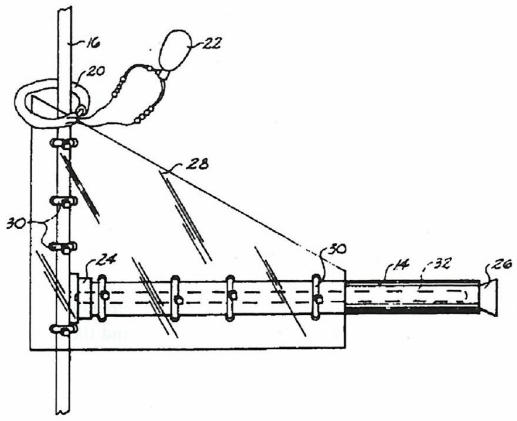


Figure 51 [79]

Lighting Current Detector

The apparatus for measuring the intensity of current produced in an elongated electrical conductive member (16) by a lightning strike includes an elongated strip of magnetic material that is carried within an elongated tubular housing (14). Initially a predetermined electrical signal is recorded along the length of the strip of magnetic material. One end of the magnetic material is positioned closely adjacent the conductive member so that the magnetic field produced by current flowing through the conductive member disturbs a portion of the recorded signal directly proportional to the intensity of the lightning strike.

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1.2.13.3 Different Magnetisation Sensors

When ferromagnetic materials are exposed to magnetic fields, their magnetisation changes (cf. Equation 24). Various properties related to the magnetisation change as well. The sensors described here either exploit just the change in magnetisation or respond to related parameters.

Moving iron amperemeters are the oldest type of magnetisation responsive devices. They consist two soft iron plates placed in the field of a fixed coil, one of which is movable and connected to an indicator. The magnetic field from a direct or an alternating current passing through the coil magnetises both plates in the same way. Their mutual repulsion is transformed into a deflection of the indicator. Correction irons, e.g. ferromagnets, mounted close to the moving iron are used to adjust the indication characteristic of such an instrument.

When a premagnetised sensor is exposed to a magnetic field, part of the premagnetisation gets lost. Placed next to a conductor they thus can serve as peak current detectors. The current is derived from the loss of the premagnetisation. Such a device may be a magnetic data carrier where the loss of data is measured (Fig. 51 [79]).

Some magnetic properties of anisotropic magnetic materials change linearly with the applied magnetic field. Magnetic fields move the direction of the saturation magnetisation (e.g. in ferromagnetic magnetoresistances -Section 1.2.8). In the device shown in Figure 52 [80] an AC coil is placed over the probe to sense the change of its (magnetisation dependant) self-inductance L.

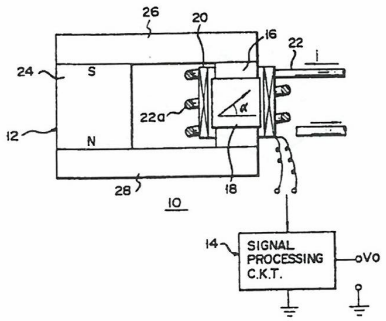


Figure 52 [80]

Current Sensor Having an Element Made of Amorphous Magnetic Metal

The current sensor comprises an element (18) located near a conductor (22) and made of amorphous magnetic metal whose magnetic property varies in accordance with the intensity of a magnetic field generated from a current flowing through the conductor. It further comprises magnetic field-applying means (24) applying a DC-bias magnetic field to the element, a coil (20) exhibiting electrical property corresponding to the magnetic property of the element (18), and output means (14) for outputting a signal corresponding to the electrical property of the coil. The coil has a specific positional relationship with the element, such that at least part of the magnetic flux generated by said coil passes through the element.

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In materials with unbalanced magnetic moments (ferrites) oscillations and waves of the magnetic moments can be excited (magnetostatic spin waves). Microwave radiation in the GHz range coupled in a slab of yttrium iron garnet (YIG) produces such an oscillation. Because of this interaction, the microwave signal is delayed on its passage through the probe. The delay time depends on geometrical factors of the probe, the frequency, and the magnetic field. The sensor shown in Figure 53 [81] provides an intensity signal, V_m, which reflects the field dependent phase shift, $\delta\phi$:

$$V_m = 1 + \cos(2\,\delta\phi - \theta) \qquad (53)$$

where θ is a phaseshift introduced by an additional delay line which is chosen to achieve maximum sensitivity, e.g. $\theta = -\pi/2$ or $-\pi$. The reported sensitivity is 0.1 mA over a dynamic range of six orders of magnitude using suitable alternators.

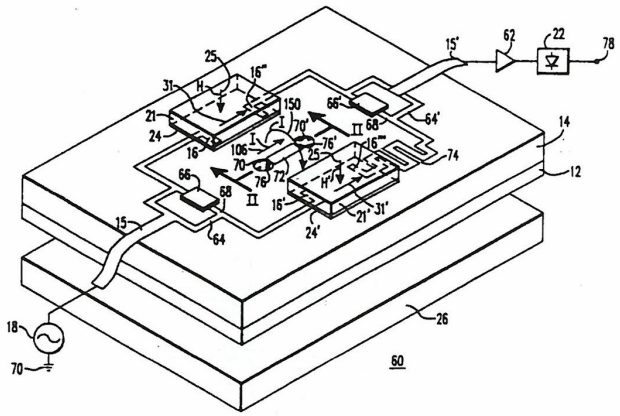


Figure 53 [81]

Wide Band Large Dynamic Current Sensor

Two YIG delay lines in parallel (24,24') are exposed to the same bias field (H, H'). A current (I) that passes through a conductor (72) produces a field (150) that adds with an opposite sign to the bias field in the probes. The intensity of the sum of the transmitted signals can be expressed in terms of the different phase shifts acquired, i.e. $+\delta\phi$ and $\theta - \delta\phi$ where θ is a delay caused by the delay line (74) on the right-hand side.

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1.2.14 Inductive Measurements

1.2.14.1 Simple Coils

Alternating currents give rise to an alternating magnetic flux, Φ, in the vicinity of the conductor. An alternating flux that passes through an opening of a conductive loop induces a voltage U_{ind}. This fact is expressed by Faraday's law of induction (equivalent to the integral form of Maxwell's second equation):

$$U_{ind} = \oint \vec{E}\, d\vec{s} = -\frac{d}{dt}\left(\int_s \vec{B}\, d\vec{a}\right) = -\frac{d}{dt}\,\Phi \qquad (54)$$

A simple inductive sensor consists of a coil with one or more windings. Its both ends are connected through a measuring resistor. The voltage drop over the resistor is proportional to the alternating current. For a circular coil (radius r) with a single winding Equation 22 can be used to determine the induced voltage:

$$|U_{ind}| = \mu_o \frac{a}{2\pi\, r}\,\frac{d\,I}{dt} = 10^{-7}\, 2\pi\, r\,\frac{d\,I}{dt} \qquad (55)$$

where a is the area of the winding, $a = \pi\, r^2$. Figure 54 [82] shows the principle of an inductive sensor. Arrangements to measure also direct currents inductively are discussed in Section 1.2.14.3.

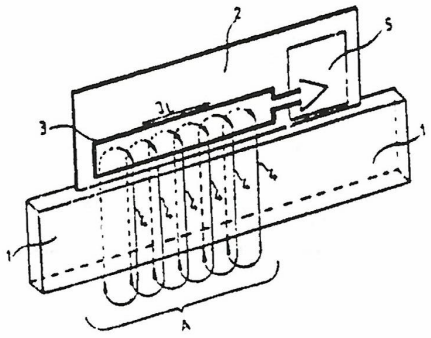


Figure 54 [82]

Device for the Potential-Isolated Measurement of Alternating Currents
Primary conductor (1) and inductive loop (3) are inductively coupled by field lines (4).

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To avoid capacitively coupled distortions, the whole coil can be mounted in a nonferromagnetic housing. The influence of magnetic stray fields can be reduced by using a pair of antiparallel coils in series (astatic coils), cf. Figure 55 [83].

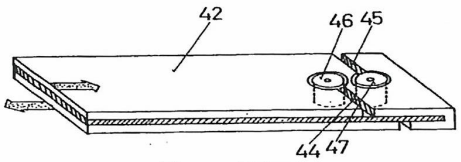


Figure 55 [83]

Current Transformer for a Static Electricity Counter

The current transformer on the secondary side is provided with a secondary winding consisting of two astatically designed coils (46, 47) laying next to one another and connected in series. As a result external parasitic magnetic fields have practically no effect on the transmission of the measured value. A large output signal is achieved by a large magnetic coupling of the primary conductor (42) and the secondary coil without the use of ferromagnetic materials.

1.2.14.2 Rogowski Coils

Coils wound on a nonmagnetic core that surround a conductor are referred to as Rogowski coils. Such a coil consists of many inductive loops perpendicular to the magnetic field H. Integration of Equation 55 around the circumference s of the core yields:

$$|U_{ind}| = \mu_o \, n \, a \, \frac{dI}{dt} = 1.26 \cdot 10^{-6} \, n \, a \, \frac{dI}{dt} \qquad (56)$$

a being the cross section of the individual windings and n the number of windings per unit length. In the derivation of the above expression it has been assumed that the magnetic field is constant over the cross section of the windings. For a straightforward analysis of the results the radial dimension of such a coil should be rather small. The advantage of Rogowski coils over a simple winding is twofold:

- extraneous fields cancel in the integral around the closed loop (provided they are uniform over this range)
- the signal is higher than for a single coil only.

The voltage U_{ind} gives rise to a current I_c that is related to the current to be measured:

$$|U_{ind}| = I_c \, R + L \, \frac{dI_c}{dt} = \mu_o \, n \, a \, \frac{dI}{dt} \qquad (57)$$

whereL is the inductance of the coil and R the combined resistance of the coil and measuring circuit.

Two simple solutions arise when one of the terms of Equation 57 is negligible:

- when $I_c R >> L \, dI_c/dt$ one has to integrate the voltage over a measuring resistor to obtain I; with $a = 10^{-4}$ m^2 and $n = 500$ m^{-1} a 10 MA/s current pulse yields a signal of about 0.6 V;
- when $I_c R << L \, dI/dt$ (short circuited coil,) the device acts as current transformer with $I_c = I/N$, where N is the number of windings.

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A shielded Rogowski coil is shown in Figure 56 [84]. In order to facilitate the measurements, Rogowski coils can also be made in shape of a belt that can easily be mounted around a conductor.

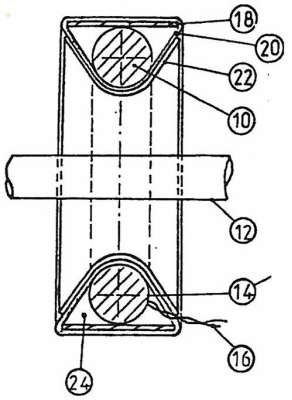


Figure 56 [84]

Current Transformer with Non-Magnetic Core

The Rogowski coil (14) is provided with a mangetic shielding (18) against extraneous fields, an inner electrostatic shield (22) to prevent capacitive coupling between the coil and the conductor (12) and an insulating layer (20).

1.2.14.3 Coils with Magnetic Cores

Alternating current transformers consist of a magnetic core with a primary and a secondary coil with N_1 and N_2 windings. Ideally input and output current I_1 and I_2 are related through:

$$I_2 = \frac{N_1}{N_2} I_1 \qquad (58)$$

With a sufficiently large resistor R in the secondary circuit, I_1 can be derived from the voltage drop $U = I_2 R$. Low and high frequency components can be isolated with appropriate circuitry, cf. Figs. 57 [85] and 58 [86]. The response of a transducer can further be modified by introducing nonlinear portions into the flux path, cf. Fig. 59 [87].

Soft magnetic materials with high permeability, low coercive force and linear B-H characteristic are chosen as core materials for alternating current transducers. For high frequency applications the frequency dependence of the permeability has to be taken into account. In some metallic amorphous alloys it is constant up to about 100 kHz [88].

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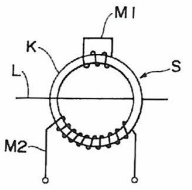


Figure 57 [85]

Apparatus for Assessing Insulation Conditions

The short-circuited tertiary winding (M1) has a low resistance for low frequency components. It therefore cancels the low frequency flux and thus acts as a high-pass for the secondary coil (M2).

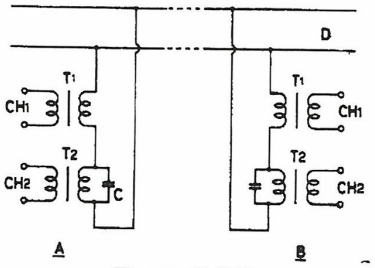


Figure 58 [86]

Branching and Inserting Circuit for Data Transmission Line

The capacitance (C) connected in parallel to the primary winding of the second transformer (T2) coil acts as a low-pass for the secondary winding (CH2).

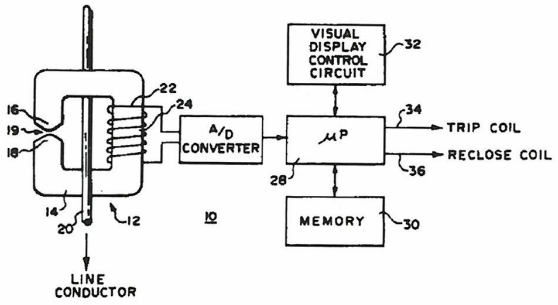


Figure 59 [87]

Non-Linear Alternating Current Transducer

Yoke (14) has an air gap (19) to exhibit a nonlinear magnetic response to current changes in the line conductor (20) and to provide a logarithmic output across the secondary winding (22). Because of the logarithmic response the relative accuracy of the transformer output is uniform over a wide range of primary currents.

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As soon as a secondary current flows in a current transformer, power is transferred to the secondary side, and the primary current is affected. When the primary current is purely AC, zero flux transformers offer a suitable alternative. In the circuit shown in Fig. 60 [89] there is no feed back into the primary circuit and Equation 58 holds exactly.

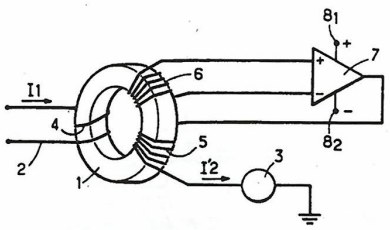


Figure 60 [89]

Device for Generating a Secondary Alternating Current whose Strenght is Proportional to that of a Primary Current

The voltage induced in the sense coil (6) controls an amplifier (7) that drives a current through the compensation coil (5). When the amplifier has a negligible offset and an infinite amplification the flux induced has the same magnitude and opposite sign as the flux from the primary coil (4).

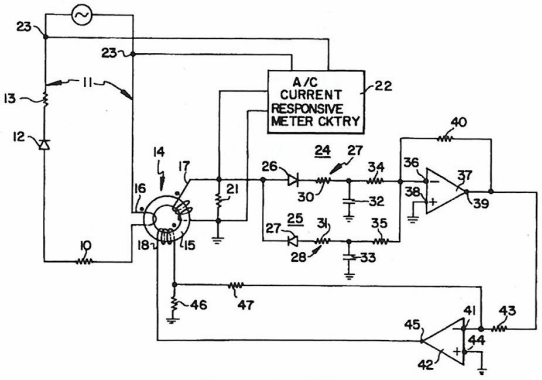


Figure 61 [90]

Current Measuring and DC Magnetic Flux Compensating Apparatus

DC magnetic flux bias imposed on a magnetic core (15) of a current transformer (14) is compensated in response to an indication of the amplitude and duration of opposite polarity secondary winding current components. In a second embodiment, the magnetic bias is compensated by responding to the integral of the secondary current. The integral of the secondary current is a measure of the magnetic core flux variations. The durations of the positive and negative flux variations are compared.

To measure direct currents one exploits the fact that they give rise to DC bias flux in the magnetic core. This flux offset alters the operation characteristics of the transformer such that amplitude and duration of the positive and negative half waves differ. In the circuit shown in Fig. 61 [90] a compensating DC flux is derived from this asymmetry.

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In the arrangement shown in Figure 62 [91] the DC flux is detected through the difference of the amplitudes of the positive and negative half waves required to drive a core alternatingly into opposite saturations. Saturation is detected through the vanishing self-inductance of the AC coil, which acts as "flux gate". Induction of the AC signal onto the primary conductor is prevented by driving two identical AC coils in opposite directions. For this kind of converters core materials with a steep linear magnetisation curve that flattens off rapidly at saturation are required. The field to be measured must be well below the hard saturation field and the AC field should be considerably higher.

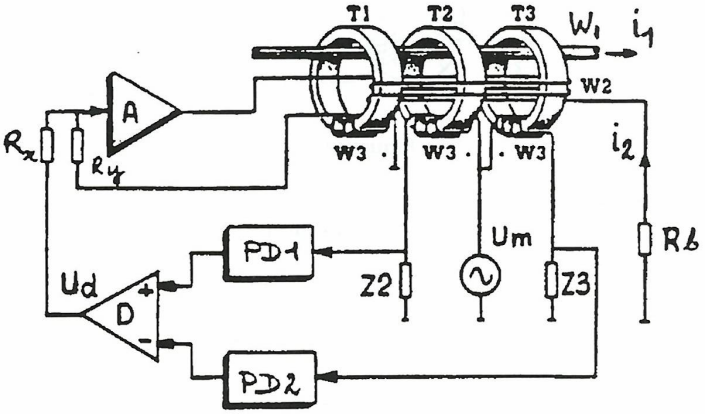


Figure 62 [91]

Circuit for the Detection of an Asymmetry in the Magnetization Current of a Magnetic Modulator

Two tertiary coils (W3) are wound in opposite directions on two cores (T2, T3) and fed by an AC reference voltage (Um) to drive the cores periodically into saturation in both directions. Upon saturation the voltages across the impedances (Z2, Z3) sharply rise. Any DC magnetic flux Φ_{DC} in the two cores due to a direct current in the primary coil (W1) gives rise to an asymmetry in the voltages over the two impedances which appear at the outputs of two peak detectors (PD1, PD2). An amplifier (A) provides a direct current to the secondary coil (W2) which compensates the magnetic flux Φ_{DC}. Any alternating current (accompanied by a magnetic flux Φ_{AC}) in the primary coil induces a signal in the tertiary coil on an additional core (T1) which gives rise to an alternating current through the secondary coil which compensates the magnetic flux Φ_{AC}. The secondary current (i_2) is an image of a primary direct or alternatiung current (i_1).

1.2.14.4 Planar-Type Detector

When an alternating primary current passes through a first ferromagnetic plate which faces a similar second plate, eddy currents are induced in the second plate. If the frequency of the current is sufficiently high (MHz) and the geometry of the plates is chosen accordingly , the skin effect is observed, i.e. the current densities of both, the primary and the induced eddy current, peak on the surfaces of the respective plates. Figure 63 [92] shows a device with 5 µm thick plates in which this effect is exploited to measure currents in the 10 to 100 MHz range. The sensitivity of the sensor crucially depends on the matching of the output impedance of the power source and the impedance of the transmission line between the power source and the detector. Its frequency range can be increased by decreasing the capacitance between the plates.

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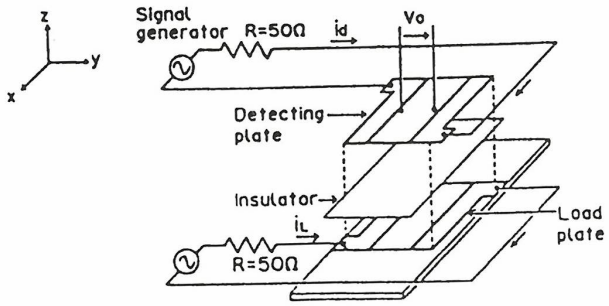


Figure 63 [92]

Planar-Type Current Detector

Two ultra-thin permallow plates with sputtered electrodes are mounted facing each other. The load current (i_L) is sent through the load plate and a detecting current (i_d) in the opposite direction through the detecting plate. Due to the skin effect the applied currents and the induced eddy currents flow in opposite directions on the outer surfaces of the plates such that the current density on the outer surface of the plates is minimised when both currents are equal. To measure the amplitude and phase of the load current of a known frequency one adjusts the amplitude and the phase of the measuring current such as to minimise the voltage drop (Vo) across the outer side of the upper plate.

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CHAPTER 4

MEASURING POWER

4.1 INTRODUCTION

By "power" we mean the product of current and voltage, measured in Watts. The word is often used when referring to energy (power x time, Ws) in conjunction with electricity consumption and utility meters. It is also used to express the proportion of signal power that falls within a certain frequency band, "spectral power" (e.g. in %), of a complex signal.

A Watt hour energy meter is a power meter with an integrating component added. Therefore energy meters are also treated in this chapter, when measuring power is the essential characteristic.

Most common power meters are ac instruments, or work both for ac and dc. Some dedicated dc instruments are treated below.

Power can be sensed directly, e.g. with Hall or thermal devices, or indirectly via combined current and voltage measurement.

4.1.1 Definitions

This chapter treats mainly currents at utility power frequencies (50-60HZ). Let us recall some definitions (sinusoidal voltage and current in a 2 phase current).

Instantaneous Power	$p(t) = u(t)\, i(t)$		**(61)**
Apparent Power	$S = U\, I$	[W]	**(62)**
Active Power	$P = U\, I \cos\alpha$	[W]	**(63)**
Reactive Power	$Q = U\, I \sin\alpha$	[VAr]	**(64)**
Power Factor	$\varphi = \cos\alpha = P/S$		**(65)**
Average Power	$p_{av} = 1/T \int_0^T u i dt$	[W]	**(66)**

where

U = effective voltage [V]
I = effective current [A]
α = phase angle between voltage and current
u = instantaneous voltage [V]
i = instantaneous current [A]
φ = power factor

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Power Vector $S = P + j\,Q + kD$ (j,k are imaginary unity vectors) **(67)**

|S| = Apparent Power [VA]
P = Active Power [W]
Q = Reactive Power [VAr]
D = Distortion Power [VA] (contribution from distorted wave forms).
If voltage and current have the same waveform D = 0. Then $P^2 + Q^2 = S^2$

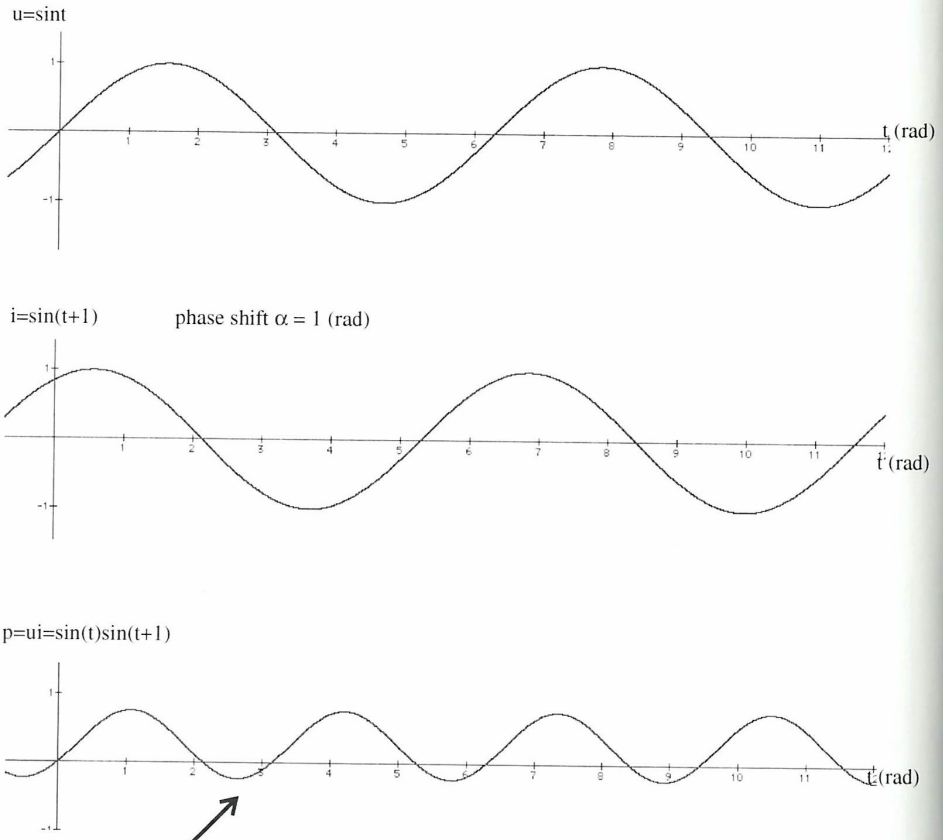


Figure 144
Voltage, current and power wave forms

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The definition of time-dependent electric power consumed by a load is: $p(t) = i(t)\,u(t)$. One can measure p(t) directly, or measure i(t) and u(t) and multiply them.

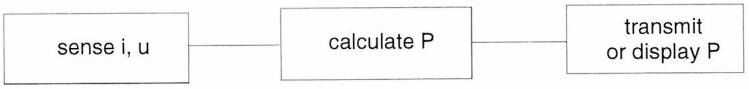


Figure 145

Measuring i and u and multiplying in real time is theoretically straightforward, but turns out to be difficult to do.

Digital technology performs this by sampling, more or less simultaneously, the current (i) and the voltage (u), and then multiplying the samples pair wise (Fig. 146) [196]. The illustrated device additionally averages the power-value.

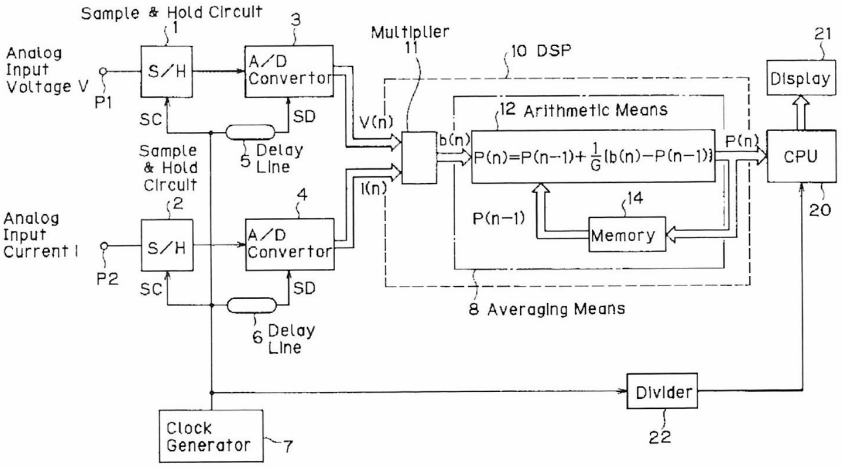


Figure 146 [196]

Sampling Type Power Measuring Device

A sampling type power measuring device measures an analog input signal by performing sampling on the analog input signal converting each value picked up by sampling into digital form and processing the digital values. Exponential averaging calculation is performed directly on the digital values; therefore no provision nor time is needed for regulating a sampling period this shortening the time of measurement.

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4.2 MEASURING PRINCIPLES

4.2.1 Overview

The standard methods for measuring power are:

- Measuring a non electric effect of power absorption.
- Measuring an electric effect of power absorption.
- Using the exponential or square-law characteristics of semiconductors, and applying Ohm's law.
- Measuring current or voltage, knowing the impedance.
- Measuring current and voltage separately, and multiplying the values.

A digital DC-power meter works like its AC-cousin but with DC-sensors: Measure i and u and multiply.

In order to achieve a high accuracy in measuring on AC signals, high speed sampling and large range analog to digital converters (ADC's) are required. Such a meter, although simple in principle, becomes expensive. Even at lower precision, an electronic meter is costly due to components such power supply, protection elements, sensors and shielding.

This is one of the reasons for inventing other, cheaper designs. Others being linked to aspects like maintenance and compatibility with existing systems, components and standards

Compensation for non-linearity of the transducers is implemented in precise instruments.

Using the data on t, p, i and u, combination instruments, like those just cited in [196], elaborate information such as average power (Pave $=1/T\int I(t)\,u(t)\,dt = I\,V\cos\alpha$ [VA]), reactive power (IU sin α [VAR]),maximum power and consumed energy.

An elementary analog device can function as in Figure 147 [198].

High frequencies are usually measured via diode detectors or by thermal methods.

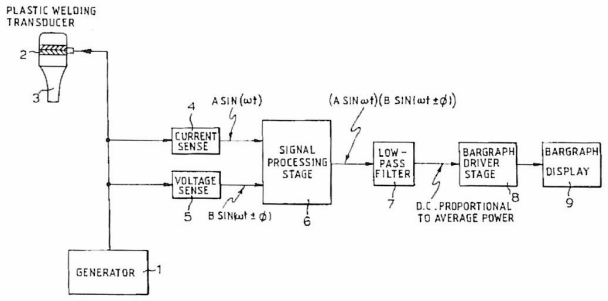


Figure 147 [198]

Indicating Power Developed by Ultrasonic Transducer

Analog device for measuring and displaying power supplied by a generator to a plastic welding transducer. Note the similarity with the digital power meter of fig. 146.

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4.2.2 Direct Sensing Of Power

4.2.2.1 Measuring a Non-Electric Effect of Power Absorption

Non-Thermal Methods

Most fluids have variable states of their constituent atoms or molecules such that when energy is applied, visible spectra may result. These could be, for example, emission, absorption or fluorescent spectra and may be discrete or continuous. By using certain characteristics of the spectra an indication of power levels may be given. A spectrum may be produced by electrons changing state and emitting or absorbing photon, or also, in the case of molecules, by interactions of rotational and vibrational states with incident energy [207].

Figure 148 [207] exploits the phenomena of power dependent light emission in an absorbing gas, and detects the light with a photo detector.

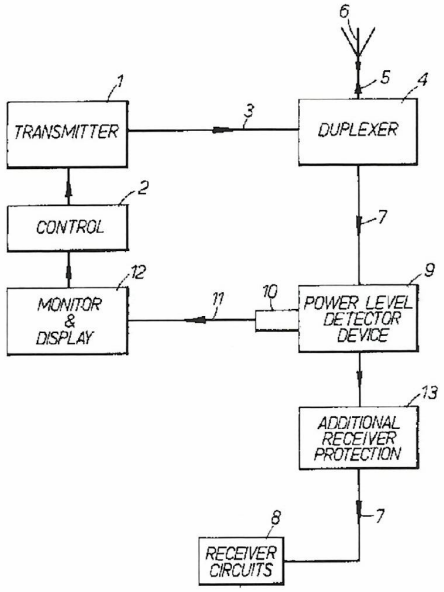


Figure 148 [207]

Apparatus for Monitoring Power

A power level detector device (9) acts to protect receiver circuits 8 from too high a level of incident power. When power at the device (9) is above a certain threshold level the gas contained by the device (9) generates an emission spectrum of electro- magnetic energy. This energy is detected by a photodetector (10) which produces an electrical signal to control the power of a transmitter (1) and hence the power transmitted to the receiver circuits (8).

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Thermal Methods

Electric power absorbed by a solid or a fluid induces physical changes. Usually the temperature raise following resistive heat production in the load (Joule's law) is measured $\delta Q = RI^2 dt$. **(68)**

Calorimetric measurements of electric power can be very precise, especially at high power levels. They are therefore being used for standards [201] and for calibration. Their simplicity and transparency make them popular in high voltage, radio and microwave applications. Radio engineers use them for measuring output power of radio transmitters.

Figure 149 [199] shows a calorimetric device for measuring power of an incident RF wave. The device determines the temperature difference between in- and outgoing cooling water. Dissipated power = P (Kw) = const x f x ΔT, where f = massflow (kg/s)and ΔT = temperature change. **(68 bis)**

In this apparatus the electric radio frequent power is absorbed by the water, but the same principle could of course be used with current producing heat in a resistive load, cooled by liquid.

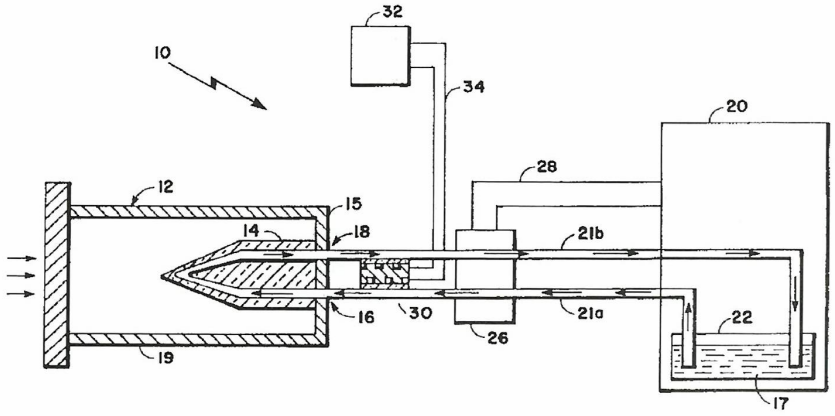


Figure 149 [199]

Power Measurement Calorimeter

One type of measurement device widely used for measuring high RF power signals is the waterload. In using a waterload, the high level RF power is absorbed in a matched-impedance section of a transmission line that is wholly or partially filled with a flowing stream of liquid, usually water. It is generally desired that the liquid itself dissipate all of the incident power, so that the temperature difference between the liquid entering the load and the liquid leaving the load can be measured by calibrated thermocouples, thermistors, or thermometers. One of several techniques can then be used for relating the desired absolute RF power to the corresponding rise in temperature.In one type of waterload, a water filled chamber is disposed at an end portion of a housing and input and output ports are provided to the chamber such that, a pressurized water supply is provided to the chamber. In this way cool water is provided to the load through the input port and the heated water is removed from the chamber from the output port.

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Instead of measuring ΔT directly, one can measure heat dependent expansion (Fig.150) [200].

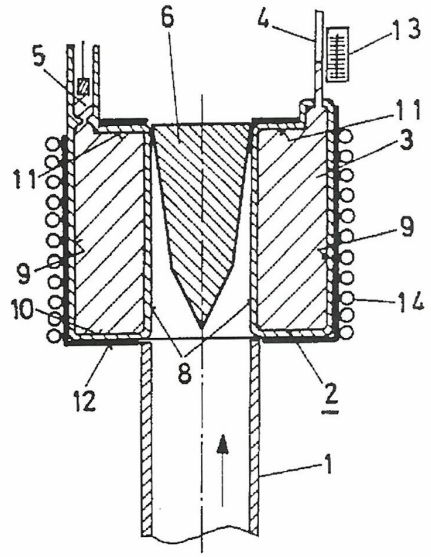


Figure 150 [200]

Microwave Pulse Energy Measurement by Absorbent Liquid Heating

A vessel (2) placed coaxially across the waveguide end (1) contains a quantity of absorbent liq. (3) in the space between concentric cylindrical walls (8, 9). A capillary tube (4) with a graduated scale (13) and a compensating tube (5) extend upward from the cover (11) of the vessel (2). A conical reflecting wedge (6) blocking the inner cylinder scatters the incident microwave energy uniformly around the inner wall (8), through which microwaves penetrate into the absorber (3). The outer wall (9), base (10) and cover (11) are coated with reflecting material (12) to prevent radiative loss. The absorbed energy is evaluated from the rise of liq. in the capillary (4).

There are other ways of using the thermal transformation: In [203] current and voltage proportional currents can be used to heat resistors that in their turn influence a bimetallic spring which rotates a pointer. It is used as a maximum demand meter in combination with a normal Wh-meter [203].

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Optical thermometers, as recently developed, are being used - see Figure 151 [205].

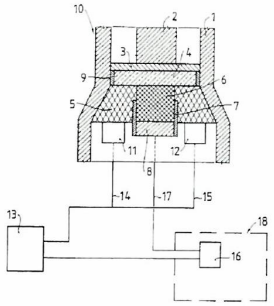


Figure 151 [205]
Sensor and Device for the Measurement of Radiant Energy,in Particular the Energy Associated with Radio-Frequency, Microwave and Light Radiation Signals

The sensor provides a termination for the guide along which the signal is propagated which termination is capable of completely absorbing the incident energy and converting it into heat. The termination is in thermal contact with a monocrystal or with a layer of crystals for example of ruby (oxide of aluminium) which is capable of being illuminated by one or more sources of light with a predetermined and sinusoidally modulated wavelength in order to emit fluorescent radiation at a second wavelength the phase displacement of which with regard to the modulating phase is a function of the temperature of the termination and thus a function of the incident energy. For radio-frequency and microwave measurements the termination terminates the coaxial cable or the wave-guide with the characteristic impedance and comprises a layer of metal deposited on the monocrystal or on the layer of ruby crystals. In the case of measurements of optical power the termination is constituted by a black body which the input optical fiber abuts the black body being in a thermal exchange relationship with the monocrystal or with the layer of ruby crystals.

A similar principle is used in the now well known **battery tester**, where the power dissipated in a thin film resistor is sensed by a layer (e.g. a liquid crystal) that changes colour with temperature (thermochromic transformation) [206] (Fig. 152).

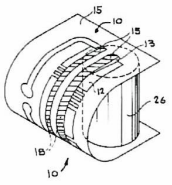


Figure 152 [206]
Battery Tester and Method for Making the Tester

A device (14) for testing a battery particularly a small portable battery (25 or 26) comprising a flexible transparent substrate (15) on which is deposited a narrow band of a black absorbant material (18). A conductive material (10) which tapers outwardly in opposite directions from a central point (11) to a pair of outer terminals (13 or 17) is then deposited atop the substrate (15) on the same side of the substrate (15) as absorber layer (18). A layer of a cholesteric liquid crystal material (15') is then deposited on the substrate (15) on the opposite side from the black absorber layer (18). The conductive material (10) is an epoxy cement-based conductor preferably silver printed or painted directly on to the substrate (15). An indicator scale (19-22) is located along sections of the tapered conductive material (10). To test a dry cell battery (25 or 26) the terminal ends (13 or 17) of the conductive material (10) are placed in contact with the battery terminals causing a current to flow which heats the conductive material (10) the heat being most intense at the central point (11) and radiating outwardly. The heat is transferred through the thin substrate (15) to the liquid cystal layer (15) which results in a color change in the liquid crystal. The traverse of the color change along the length of the indicator scale (19-22) outwardly from the central point (11) is proportional to the current or voltage output condition read on the indicator scale (19-22) which is calibrated accordingly.

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4.2.2.2 Using Thermo-electric Conversion or Resistivity Change

We have several useful transformations from heat to an electrical variable.

- Thermoelement give a voltage as an output signal.
- Resistors, e.g. thermistors, that change resistance with temperature. When used for measuring the power of radio frequent or microwave radiation they are called Bolometers.
- The characteristic curves of semiconductors change with temperature.

Whereas calorimetric methods are absolute in character, the transformations just mentioned imply relative measurements, and thus rely on calibrations. The principle is this:

1) The current to be measured heats up a resistor in a manner proportional to power. The resistor looses some unknown amount of heat to the surrounding. For each input power level, the thermal equilibrium temperature is different.

2) The temperature, or temperature change, of the resistor is sensed by a temperature sensitive element with an electrical output which is proportional to the power to be measured.

In the case of the thermoelement, it is possible to lead the signal to be measured directly through it, and use it as a measurement resistor (Fig. 153) [202] below.

4.2.2.2.1 Thermocouples

The heat difference between the hot and cold junctions, whether induced by Joule's effect in the element or by heat transfer, creates a measurable voltage by the Seebeck and Thomson effects [335].

Thermocouples can also be used as multipliers (i x u) in power measuring instruments: See ch. 4.2.5, "Thermal Converters".

Thermocouple sensors have been used for many years, especially for measuring at RF and microwave frequencies. They can however be used at lower frequencies or DC as well. An instrument based on AC/DC conversion using thermocouples, designed for power frequencies, i.e. 45-65 Hz is described in [201]. This instrument was developed as a standard.

Thermocouple sensors are faster than calorimetric devices.

The meter of [202] (Fig. 153) shows how radio frequent power creates a low level (e.g. 160 nV for 1 mW of applied RF power) DC-signal when applied to a thermocouple pair (106). The low level output of the thermocouple is chopped at 220 Hz to facilitate processing.

The low chopping frequency reduces the interfering effect of "spiking" caused by the chopper.

The meter perform synchronous detection and a microprocessor in a feed back loop is used to zero and calibrate it automatically.

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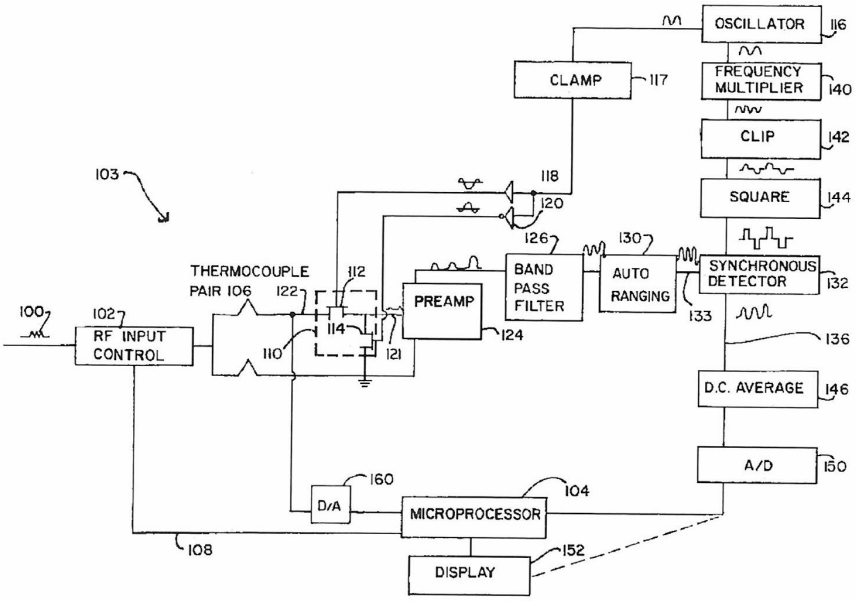


Figure 153 [202]
Thermocouple Power Meter

An RF input 100 is shown entering an RF input control element 102 of a thermocouple power meter 103. Also entering the RF input control element 102 is an input line from a microprocessor 104. The RF input control element 102 selectively permits any one of several RF inputs to be applied to a thermocouple pair 106—including the RF input 100 and a zero input or a reference input such as one milliwatt selected by the microprocessor 104 via line 108.

The thermocouple pair 106 converts the RF input, if any, exitting the control element 102 into heat and thereafter into a d.c. signal in conventional fashion. The amplitude of the d.c. signal generated by the thermocouple pair 106 is low and difficult to measure with accuracy. Accordingly, the d.c. signal from the thermocouple pair 106 is passed through a chopper stage 110 that includes a series chopper 112 and a shunt chopper 114. The series chopper 112 and the shunt chopper 114 are driven to operate complementarily, one being ON (or conductive) when the other is OFF (or non-conductive). For simplification, capacitors and resistors and the like that would be included in a detailed circuit diagram are omitted, such elements being within the knowledge of one of ordinary skill in the art to include.

The low end sensitivity of thermocouples is not very good.

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Figure 154 [319] illustrates how a thermoelectric measurement converter can be integrated in a single chip. The circuit measures the product of two instantaneous values.

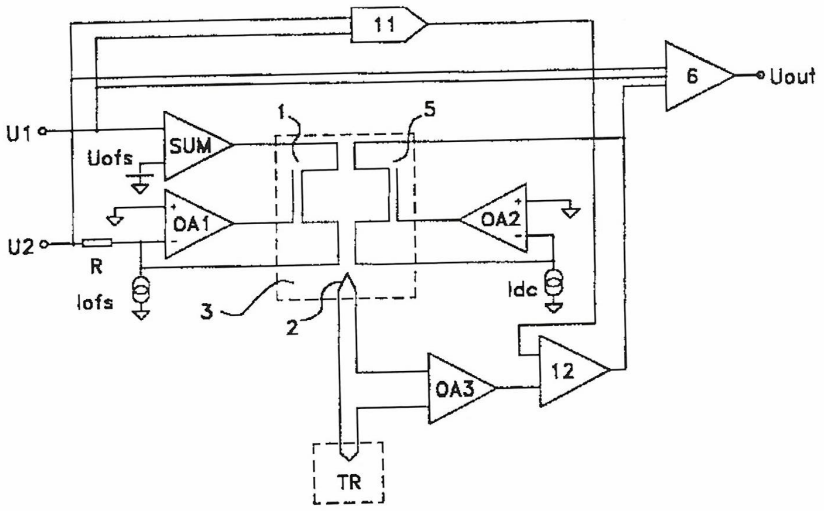


Figure 154 [319]
Integrated Thermoelectric Measurement Converter

FET transistors (1) and (5) act as controlled resistors and are seated on a heat insulating layer (3). FET (1) is coupled to the input signals. FET (5) performs an automatic compensation for the thermal influence of power change on FET (1). Temperature sensor (2) measures the temperature difference between the layer (3) and the reference body (TR). A feed-back system drives the voltage, and thus the power on transistor (5) so as to establish temperature balance between the layer (3) and the reference body (TR).

The integrated device, which shape is illustrated in Figure 155 is manufactured by epitaxial growth. Here the substrate (20) is mounted on multiple Peltier-elements (29) which cool the device so as to lower the noise level.

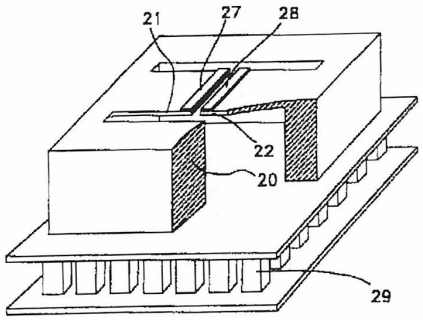


Figure 155 [319]
The Device of Fig. 154

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4.2.2.2.2 Semiconductors as Temperature Sensors in Power Measuring devices

The **characteristic curves** of semiconducting elements change with temperature. A transistor which is heated will give a higher collector-emitter current for a fixed base-emitter voltage.

Figure 156 [336] shows a circuit arrangement exploiting this principle. (The dashed lines indicate thermally isolated monolithic semiconductor islands 96 and 98):

The input is at T1-T2, and resistor 42 heats up transistor 52, which collector-emitter current thus varies in a way proportional to power.

This operational mode is relatively insensitive to noise and operates linearly at very small signal levels.

At higher amplitude levels the diode 60 conducts, and the resistor 44 is effective in heating island 98, and establishing thermal equilibrium.

The output signal 92 over T7-T8 is in this case proportional to the square of the input power.

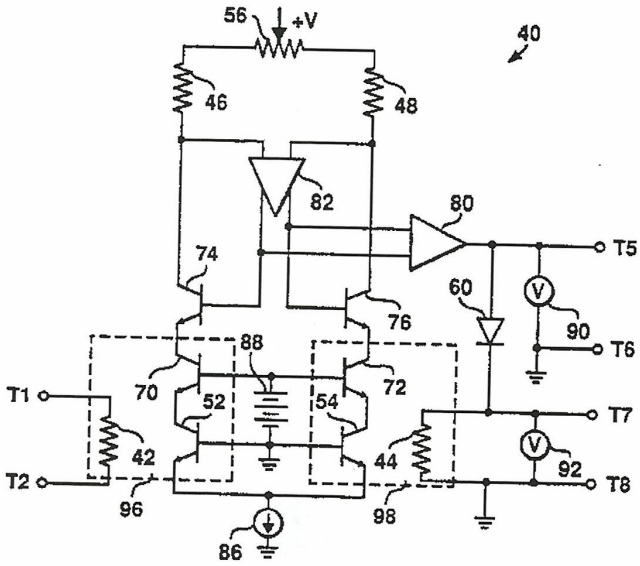


Figure 156 [336]
Power Measuring Device Comprising Transistor Heat Sensors

The circuit has a linear feed back part with an output T5-T6. For small signal amplitudes the feed back through amplifier 82 brings the islands 96 and 98 into thermal equilibrium (power dissipated in transistor 72 heats island 98). It has to be noted that in this mode of operation the diode 60 is not conducting, because the forward bias over it is to small.

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Combing the two outputs, T5-T6 and T7-T8 in an additional stage is illustrated in Figure 157 [336].

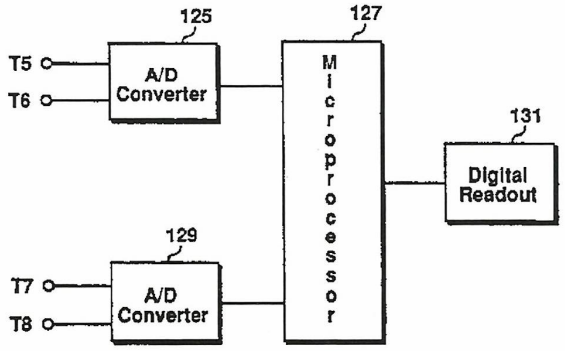


Figure 157 [336]
Automatic means for combining the indications of the meters 90 and 92 in "fig. 156"

This instrument has a good linearity, sensitivity and accuracy over a wide dynamic range of measured signals.

4.2.2.2.3 Resistive Sensors

Bolometer elements (resistance change with microwave power absorption) such as **Baretter** (positive temperature coefficient) elements and **thermistors** (negative temperature coefficient) are being used in microwave applications. Resistive sensing of microwave power is faster than using thermocouples. Baretters are faster than thermistors [331]. Traditionally they are inserted in Wheatstone bridges. Figure 158 [208] shows an example of the use of a Baretter element.

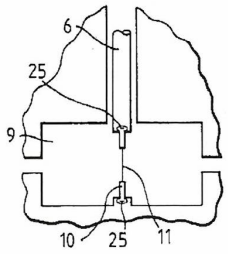


Figure 158 [208]
Baretter Elements

A known method of measuring microwave power is to suspend a very thin wire in a waveguide section with a moulded D.C. lead out. When illuminated by microwave the resistance of the wire changes considerably. It is normal to operate barretters at constant resistance by means of a D.C. self balancing bridge. The change in D.C. bias power when microwave power is absorbed by the element is known as the D.C. substitute power and the ratio of this to the incident power is the effective efficiency. This is less than unity because of dissipation in the structure which does not result in a change in the element resistance, microwave leakage and differing power co-efficients of resistance for D.C. and microwave heating.

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A method of measuring **microwave** power is to suspend a very thin wire in a waveguide section with a grounded DC lead out. When illuminated by microwaves the resistance of the wire changes considerably. It is normal to operate barretters at constant resistance by means of a DC self -balancing bridge. The change in DC bias power when microwave power is absorbed by the element is known as the DC. substitute power and the ratio of this to the incident power is the effective efficiency. This is less than unity because of dissipation in the structure which does not result in a change in the element resistance, microwave leakage and differing power coefficients of resistance for DC and microwave heating.

A self-balancing system with bolometer element without a Wheatstone bridge is shown in Figure 159 [209]

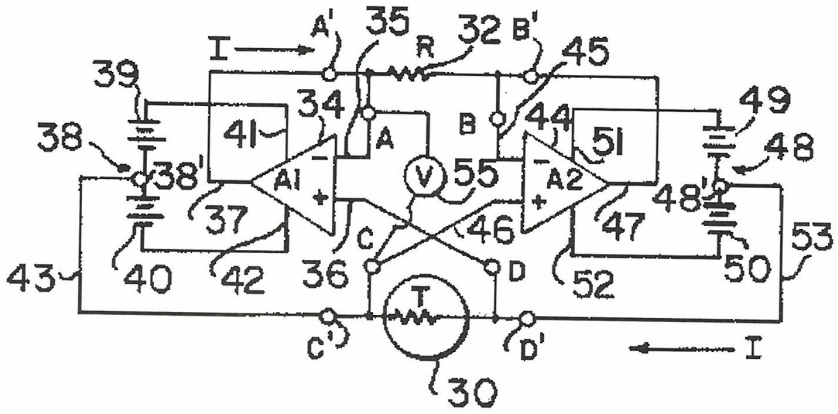


Figure 159 [209]
Self-Balancing D.C.-Substitution Measuring System

A self-balancing D.C.-substitution R.F. power measuring system includes first and second high gain differential operational amplifiers, a bolometer element, and a reference resistor element. The amplifiers and the two elements are connected in a current loop with one of the elements connected between the output terminals from the differential amplifiers and the other of the elements connected between center points of isolated dual power supplies associated with each of the amplifiers. The inputs to one amplifier are connected from an adjacent end of one of the elements and the far end of the other element, while the inputs to the second amplifier are connected to the far end of the one element and the adjacent end of the other element. Current flows out of one amplifier and into the other. The current is driven to a value which maintains the potential between the input terminals of the first amplifier essentially equal to zero and the potential between the input terminals of the second amplifier essentially equal to zero. Thus, the current drives the value of the bolometer element to a resistance which is equal to the resistance of the reference element. An output connection to a voltmeter may be taken between corresponding ends of the elements. The bolometer element may be a thermistor or a barretter. The system may also be used in a hot-wire anemometer. In another version of the system, the current loop is established with one of the elements connected between the output of one amplifier and the center point of the power supply of the other amplifier and the other of the elements connected between the output of the other amplifier and the center point of the power supply of the one amplifier; the input connections to the amplifiers are taken from the same points, but with the input leads to the second amplifier interchanged.

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Digital technology eliminates the need for bridges. See Figure 160 [210].

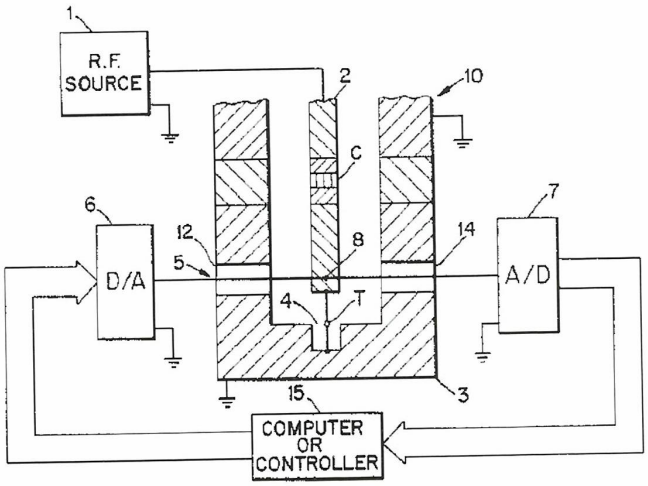


Figure 160 [210]
High Frequency Power Sensing Device

The device uses a single thermistor, directly converts a digital signal to the current through the thermistor, and converts the voltage across the thermistor to a digital signal. The change of the digitized dc signals from no RF power to power to be measured provides the measure of RF power by the substituted dc method. The above approach eliminates the need for bridges and voltmeters between the power sensor and computer circuits. This approach permits the computer to set the dc resistance of the thermistor to the proper value to provide optimum matching at the desired frequency and to calculate the dc power dissipated in the thermistor. The present invention using only a single thermistor, eliminates the need for bypass capacitors and the nonlinearity error produced by thermistor pairs. The dc blocking capacitor in this invention, not being an integral part of the power sensor, is interchangeable permitting a wide range of operating frequencies with a single basic power sensor.

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