

PRINCIPLES AND APPLICATIONS IN ENGINEERING SERIES

ELECTRICAL MEASUREMENT, SIGNAL PROCESSING, and DISPLAYS

Edited by
JOHN G. WEBSTER



CRC PRESS

Petitioner has added romanettes to pages i-vi. Otherwise, it leaves the original page numbering.

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CRC PRESS

Boca Raton London New York Washington, D.C.

Library of Congress Cataloging-in-Publication Data

Electrical measurements, signal processing, and displays / John G. Webster, editor-in-chief.

p. cm. -- (Principles and applications in engineering)

Includes bibliographical references and index.

ISBN 0-8493-1733-9 (alk. paper)

1. Electronic measurements. 2. Electric measurements. 3. Signal processing. I. Webster, John G., 1932- II. Series.

TK7878.E435 2003

621.3815'48—dc21

2003048530

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International Standard Book Number 0-8493-1733-9

Library of Congress Card Number 2003048530

Printed in the United States of America 1 2 3 4 5 6 7 8 9 0

Printed on acid-free paper

Contents

SECTION I Electromagnetic Variables Measurement

1	Voltage Measurement	<i>Alessandro Ferrero, Jerry Murphy, Cipriano Bartoletti, Luca Podestà, and Giancarlo Sacerdoti</i>	1-1
2	Current Measurement	<i>Douglas P. McNutt</i>	2-1
3	Power Measurement	<i>Pasquale Arpaia, Francesco Avallone, Aldo Baccigalupi, Claudio De Capua, and Carmine Landi</i>	3-1
4	Power Factor Measurement	<i>Michael Z. Lowenstein</i>	4-1
5	Phase Measurement	<i>Peter O'Shea</i>	5-1
6	Energy Measurement	<i>Arnaldo Brandolini and Alessandro Gandelli</i>	6-1
7	Electrical Conductivity and Resistivity	<i>Michael B. Heaney</i>	7-1
8	Charge Measurement	<i>Saps Buchman, John Mester, and Timothy J. Sumner</i>	8-1
9	Capacitance and Capacitance Measurements	<i>Halit Eren and James Goh</i>	9-1
10	Permittivity Measurement	<i>Devendra K. Misra</i>	10-1
11	Electric Field Strength	<i>David A. Hill and Motohisa Kanda</i>	11-1
12	Magnetic Field Measurement	<i>Steven A. Macintyre</i>	12-1
13	Permeability and Hysteresis Measurement	<i>Jeff P. Anderson and Richard J. Blotzer</i>	13-1
14	Inductance Measurement	<i>Michał Szyper</i>	14-1

15	Immittance Measurement	<i>Achim Dreher</i>	15-1
16	Q Factor Measurement	<i>Albert D. Helfrick</i>	16-1
17	Distortion Measurement	<i>Michael F. Toner and Gordon W. Roberts</i>	17-1
18	Noise Measurement	<i>W. Marshall Leach, Jr.</i>	18-1
19	Microwave Measurement	<i>A. Dehé, K. Beilenhoff, K. Fricke, H. Klingbeil, V. Krozer, and H. L. Hartnagel</i>	19-1

SECTION II Signal Processing

20	Amplifiers and Signal Conditioners	<i>Ramón Pallás-Areny</i>	20-1
21	Modulation	<i>David M. Beams</i>	21-1
22	Filters	<i>Rahman Jamal and Robert Steer</i>	22-1
23	Spectrum Analysis and Correlation	<i>Ronney B. Panerai, A. Ambrosini, C. Bortolotti, N. D'Amico, G. Grueff, S. Mariotti, S. Montebugnoli, A. Orfei, and G. Tomassetti</i>	23-1
24	Applied Intelligence Processing	<i>Peter H. Sydenham and Rodney Pratt</i>	24-1
25	Analog-to-Digital Converters	<i>E.B. Loewenstein</i>	25-1
26	Computers	<i>A.M. MacLeod, P.F. Martin, and W.A. Gillespie</i>	26-1
27	Telemetry	<i>Albert Lozano-Nieto</i>	27-1
28	Sensor Networks and Communication	<i>Robert M. Crovella</i>	28-1
29	Electromagnetic Compatibility	<i>Daryl Gerke, William Kimmel, and Jeffrey P. Mills</i>	29-1

SECTION III Displays

30	Human Factors in Displays	<i>Steven A. Murray and Barrett S. Caldwell</i>	30-1
31	Cathode Ray Tube Displays	<i>Christopher J. Sherman</i>	31-1

.....	15-1
.....	16-1
.....	17-1
.....	18-1
eil, V. Krozer,	 19-1
.....	20-1
.....	21-1
.....	22-1
C. Bortolotti,	 23-1
tti	 24-1
ratt	 25-1
.....	26-1
.....	27-1
.....	28-1
i	 29-1
ell	 30-1
.....	31-1

32	Liquid Crystal Displays	Kalluri R. Sarma	32-1
33	Plasma-Driven Flat Panel Displays	Robert T. McGrath, Ramanapathy Veerasingam, William C. Moffatt, and Robert B. Campbell	33-1
34	Electroluminescent Displays	William A. Barrow.....	34-1
35	Light-Emitting Diode Displays	Mohammad A. Karim.....	35-1
36	Reading/Recording Devices	Herman Vermariën, Edward McConnell, and Yufeng Li.....	36-1
Index.....			I-1

Current Measurement

2.1	Definition of the Ampere	2-2
2.2	Magnetics	2-3
2.3	Shunts.....	2-5
2.4	The Moving Magnet Meter	2-5
2.5	The D'Arsonval Meter.....	2-6
2.6	The Electrodynamometer	2-6
2.7	The RF Ammeter and True rms.....	2-6
2.8	The Current Transformer	2-7
	Safety Note	
2.9	Gapped Inductive Sensors	2-8
2.10	Hall Effect Sensor	2-9
2.11	Clamp-On Sensors	2-10
2.12	Magnetoresistive Sensors	2-10
2.13	The Magnetic Amplifier	2-10
2.14	Fluxgates	2-11
2.15	Optical Sensors.....	2-12
2.16	Fault Indicators	2-12
2.17	Other Schemes.....	2-12
2.18	Some Generalities and Warnings	2-12
2.19	Current Actuated Switches and Indicators.....	2-13
2.20	Where to Get Current Sensors	2-13

Douglas P. McNutt

The MacNaughtan Laboratory

Current measuring devices are selected for:

- Precision
- Stability
- Frequency response, including dc
- Galvanic isolation
- Presentation of the data
- Effect on measured circuit

A few of the most common sensors will be introduced in this chapter. Details and some less common sensors will follow after definitions and a bit of magnetic theory, which can be skipped. Any magnetic field sensor can be used as a current sensor and there are some exotic examples, such as quantum effects in low-temperature superconductors used to measure currents in neurons within the brain. This discussion is limited to measurement of currents in wires with commercially practical devices.

An isolated current sensor is free of any metallic connection to the circuit being measured. It is also essentially free of capacitive coupling so that it is safe to use with grounded amplifiers and other equipment. The quality of the isolation is measured in volts and is usually the breakdown potential of an insulator. 5 kV is typical for personal safety around light industrial power.

By far, the simplest current-to-voltage converter is the resistor. In current measuring service, it is called a shunt although it is typically placed in series with the load. That is because shunts are sometimes used to increase the range of another current-measuring device using a connection that bypasses part of the current around a meter. Frequency response of a shunt is good and includes dc. Shunts produce a voltage output that can be presented by a variety of secondary meters, including analog meters, digital meters, oscilloscopes, and 4 to 20 mA converters. Shunts provide no isolation and have a potentially unacceptable effect on the circuit being measured. Shunts used for dc are as accurate as the resistance and the associated voltmeter.

The common moving-coil meter, the D'Arsonval movement [1, 2], probably with an internal shunt and/or rectifier, is an easily used device and still available. Its isolation is by means of the human eye since that is the only way to read out the result. It is useful for power panels where an operator needs quick data. Accuracy is no better than 2%.

For power frequency, 25 to 400 Hz service, the current transformer, called a "donut" transformer, or CT in the trade, is commonly employed. The current-carrying conductor is passed through the hole in a toroid of magnetic material. A shorted secondary winding of n turns carries current, which is $1/n$ times the measured current, and is typically passed to another ammeter or used as the current input to a power measuring device. Isolation is as good as the insulation on the primary conductor; frequency response is fair but does not include dc; there is minimal effect on the measured circuit, and the cost is low. Operational safety is an issue; see the note below.

A variety of noncontact sensors is available for dc sensing. Most depend on the Hall effect and all require a source of operating power. Frequency response from dc to 200 kHz is advertised. Because operating power is available, output to later processing can be voltage, current, or digital. Accuracy is whatever one wants to pay for. Long-term stability depends on dc-coupled operational amplifiers and can exhibit zero drift. Externally, these look much like CTs.

CTs, Hall devices, and other similar noncontact schemes are available in wrap-around form so they can be installed without disconnecting power. The wrapping process always involves breaking a magnetic path, and consistency of reassembly becomes a limit on precision. Everything from current-sensing probes for oscilloscopes to CTs for 10000 A circuits can be found as wrap-arounds.

2.1 Definition of the Ampere

There is perpetual argument about the number of "fundamental" quantities required to describe our environment. Is there a fundamental unit of electricity or are the electrical units derived from more basic things such as mass, length, and time? Standards laboratories such as the U.S. National Institute of Standards and Technology, NIST, measure the force between current-carrying wires to compare the ampere to the meter, the kilogram, and the second, and provide a standard stated as "that constant current which, if maintained in two straight parallel conductors of infinite length, of negligible circular cross-section, and placed 1 m apart in vacuum, would produce between these conductors a force equal to 2×10^{-7} newton per meter of length" [3]. For the rest of this discussion, it is assumed that instrumentation is "traceable to the NIST," meaning that one way or another, an electrical unit as measured is compared to a standard ampere maintained by NIST. Absolute calibrations using the quantum properties of the Hall effect and the Josephson junction are not practical in the field.

For practical reasons, it is often easier to distribute a voltage standard than a current standard. Chemical cells and, more recently, semiconductor voltage references are quite stable and do not depend on the length of the wires used to connect them together during calibration and use. As a result, current measuring is usually a matter of conversion of current to an equivalent voltage.

An exception is current measurement by comparison to magnetic forces provided by a permanent magnet and that might be why the older unrationalized centimeter-gram-second, the cgs units, with no separate electrical unit is still found in specifications of magnetic quantities. The gauss and the oersted are particular examples of these now deprecated units. It is this confusion of units that frightens many who would measure current away from even attempting calculations involving magnetic devices.

2.2 Magnetics

Magnetic current sensors have advantages over shunts. To understand them, one should delve into the interaction between currents and magnetic fields. Following Maxwell by way of Sommerfeld [4], it is convenient to describe magnetic effects in terms of two vector fields, $\mathbf{B}$ and $\mathbf{H}$. $\mathbf{H}$ is the field created by an electric current, and $\mathbf{B}$ is the field that acts on a moving charge or a current-carrying wire. $\mathbf{B}$ and $\mathbf{H}$ are related by characteristics of the material in which they coexist. Strictly speaking, $\mathbf{B}$ is the flux density and $\mathbf{H}$ is the field, but they are both called the field in less than precise usage.

The SI units of $\mathbf{B}$ and $\mathbf{H}$ are the tesla (T) and the ampere per meter (A m^{-1}). They are called rationalized because a ubiquitous 4π has been suppressed in the underlying equations.

For practical engineering, it is still necessary to understand the unrationalized equivalents, the gauss and the oersted, because they are universally used in specifications for magnetic materials. To convert from gauss to tesla, divide by 10^4 . To convert from oersted to amperes per meter, multiply by $1000/(4\pi)$, a number that is commonly approximated as simply 80; but strictly speaking, the units of $\mathbf{H}$ in the two systems are dimensionally different and cannot be converted.

The relationship between $\mathbf{B}$ and $\mathbf{H}$ is most generally a tensor that reflects spatial anisotropy in the material, but for common magnetic materials used in current sensing, a scalar value μ applies. In SI units, $\mathbf{B}$ and $\mathbf{H}$ have different physical dimensions and are not equal in vacuum. The "permeability of free space" μ_0 is defined so that the μ for a material is a dimensionless constant. Thus,

$$\mathbf{B} = \mu\mu_0\mathbf{H} \quad (2.1)$$

where μ_0 is exactly $4\pi \times 10^{-7} \text{ H m}^{-1}$. (H here is the henry, the SI unit of inductance, base units $\text{m}^2 \text{ kg s}^{-2} \text{ A}^{-2}$). For many problems, μ will be a constant of the magnetic material, but when magnetic saturation needs to be considered it will be a variable. For engineering calculations, μ is often treated as though it depends on frequency. Values of μ range from less than 100 for high-frequency ferrite to 5000 for transformer iron, to 10^5 for carefully annealed magnetic alloys.

The field due to a long straight wire is shown in Figure 2.1 and Equation 2.2. The $\mathbf{H}$ vector obeys the right-hand rule and is everywhere perpendicular to the wire. The amplitude of $\mathbf{H}$ is proportional to the current and falls off linearly with the radial distance from the wire. This is the field that makes noncontact sensing of current possible.

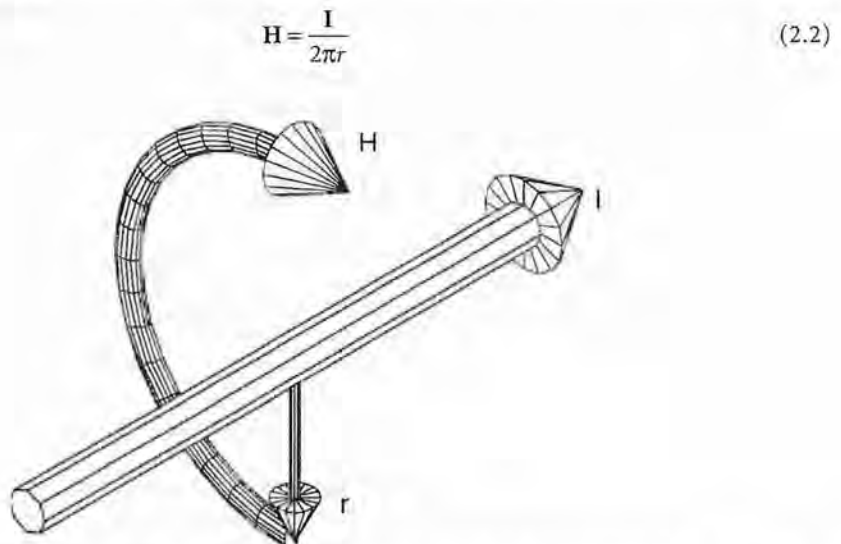


FIGURE 2.1 The magnetic field associated with a current-carrying long wire. It is this field that makes contactless sensing of current possible.

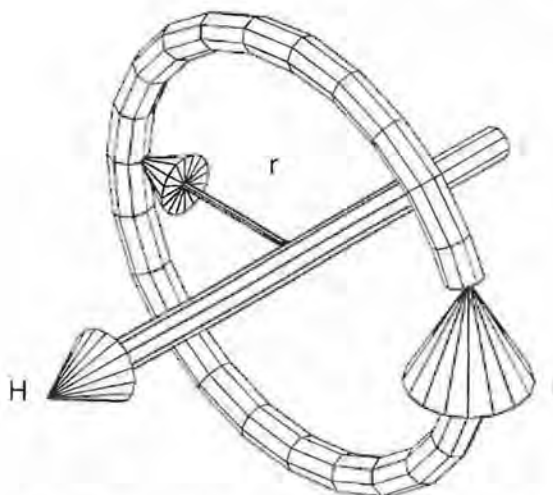


FIGURE 2.2 The magnetic field at the center of a loop of wire. This is the starting point for many calculations involving inductors and transformers.

The field at the center of a circular loop of wire is shown in Figure 2.2 and Equation 2.3. It obeys the right-hand rule, is proportional to the current, and inversely proportional to the radius of the loop.

$$H = \frac{I}{r} \quad (2.3)$$

A magnetic field can often be calculated by direct application of Maxwell's relations in integral form [4]. The line integral over any closed path of $\mathbf{H} \cdot d\mathbf{l}$ is equal to the current passing through a surface delimited by the closed path.

Magnetic flux passing through a surface, usually denoted by ϕ , is the integral of $\mathbf{B} \cdot \mathbf{n} dA$ over the surface with normal vector $\mathbf{n}$. The SI unit of ϕ is the weber. The unrationalized unit, the line of force, is best relegated to history except that the lines, which form continuous loops, dramatize the fact that it does not matter how a surface is drawn to close a bounding line. The flux is the same. It is convenient to think of magnetic lines, forming closed loops, that can be bent but not interrupted or removed.

A carrier of charge, an electron in a wire, a particle in vacuum, or a hole in a semiconductor, moving in a magnetic field is acted on by a force that is perpendicular to the field and the velocity. For positive charge, the force obeys the right-hand rule. The magnitude of the force is proportional to the magnitude of $\mathbf{B}$, the velocity $\mathbf{V}$, and the sine of the angle between them.

$$\mathbf{F} = \mathbf{V} \times \mathbf{B} \quad (2.4)$$

A carrier, moving or not, is affected in a similar way by a changing magnetic field. The result is an electromotive force, *EMF*, in a loop of wire through which a changing magnetic flux passes. The *EMF* is equal to the rate of change of the flux enclosed by the loop with a change of sign. That is Faraday's law of induction.

$$EMF = -\frac{d\phi}{dt} \quad (2.5)$$

Most magnetic materials exhibit hysteresis. That is, the relationship between $\mathbf{B}$ and $\mathbf{H}$ depends on the history of the applied $\mathbf{H}$. A plot of $\mathbf{B}$ vs. $\mathbf{H}$ is called a hysteresis loop and a sample is shown as Figure 2.3. The area of the loop represents an energy. If a magnetic core is repeatedly driven around its hysteresis loop, there is a power loss due to hysteresis that is proportional to the area of the loop and the frequency. A material with very large hysteresis is a permanent magnet.

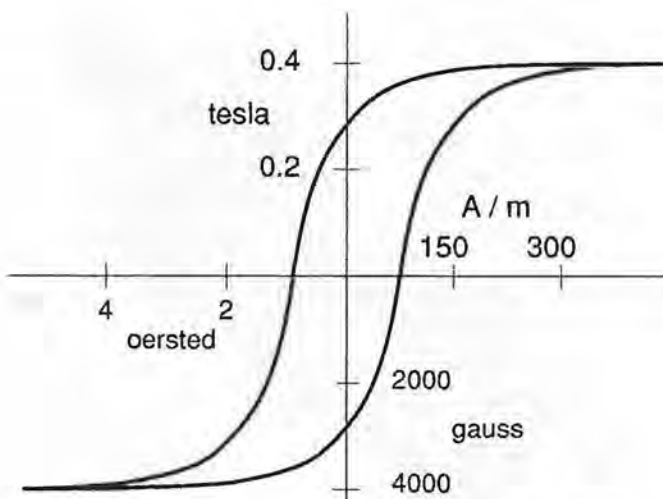


FIGURE 2.3 A hysteresis curve for some rather poor transformer iron. Unrationalized units are shown below and left of the origin because that remains standard practice in the industry.

Magnetic materials which are also electric conductors have free carriers which are affected by alternating magnetic fields. As they move back and forth they encounter electric resistance and dissipate energy in the form of heat. These eddy currents can be minimized by use of high-resistance ferrites, powdered iron, by laminating cores, or by keeping the flux small.

2.3 Shunts

Shunts were introduced above. They dissipate power as heat and the resistance changes in response to the rising temperature. The dissipation is proportional to the voltage across the shunt and a design compromise must be made because low voltage implies less accuracy in the voltmeter. A standard of 50 mV has evolved. Shunts do not provide galvanic isolation between the measured circuit and the measurement device.

Measurement of alternating current using shunts is also affected by skin effect and the inductance of the shunt. Skin effect can be minimized in the design of the shunt by the use of several parallel sheets of thin metal, Figure 2.4, a feature that also improves heat dissipation. There is not much to be done about inductance except to minimize size.

Safety is enhanced if shunts are placed in the ground leg of a circuit; that way, the output leads, usually only at 50 mV, are near ground. However, that introduces a resistance in the ground path and can interfere with common mode requirements of interdevice signal connections. If the shunt is placed in the high side, care must be taken to protect the wiring and the meter to which it is attached from accidental grounds.

Sometimes, sufficient accuracy can be obtained by measuring the voltage drop along a length of conductor that is otherwise required in the installation. In vehicular service, it is common to sense the voltage drop in a battery cable using a millivoltmeter. Such installations are always high side and should be protected with fuses installed near the points of measurement. It is also wise to provide a connection means that is independent of contact resistance where the current-carrying conductor is installed — a Kelvin connection. Including one lead of an in-line fuse holder in the terminals before they are crimped is one such technique.

2.4 The Moving Magnet Meter

The simplest current indicator balances the force on a permanent magnet created by current in a wire against a spring. A magnetic material is usually placed around the conductor to concentrate the field and

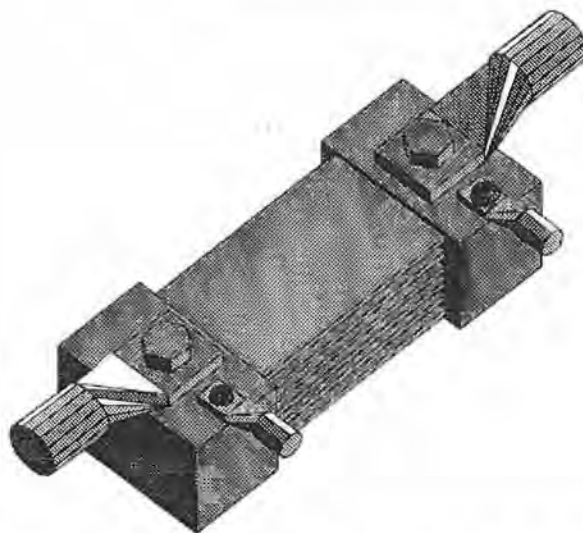


FIGURE 2.4 Multiple sheets of conductor are provided in this shunt to reduce skin effect and allow air cooling. Kelvin connections are provided for the voltmeter so that the voltage drop in the high-current connectors is not inadvertently included in the measurement.

reduce the effect of the magnetic field of the Earth. Use is limited to low-precision indicators such as a battery charging meter for a vehicle. It is a dc device.

2.5 The D'Arsonval Meter

This indicator balances the force on a current-carrying wire due to a permanent magnet against a spring. The measured current flows in a coil of wire supported in bearings. It is described in elementary texts [1, 2]. It is generally a dc instrument, but chart recorders have been built with frequency response in the kilohertz range using mirrors on the moving coil. They are then called galvanometers. For ac service, these meters are often equipped with internal copper oxide rectifiers and a nonlinear scale to correct for the diode drop. For current ranges above a few milliamperes, they will have an internal shunt.

The moving magnet meter and the D'Arsonval movement are the only current sensors that do not convert current to voltage and then depend on other devices to read out the voltage.

2.6 The Electrodynamometer

A variation of the D'Arsonval meter for ac service can be built by replacing the permanent magnet with an electromagnet. The force on the moving coil becomes proportional to both the current being measured and the voltage applied to the coil of the electromagnet. It is sensitive to the relative phase of the voltage and current in just the right way to be useful for measurement of power in a circuit with correction for power factor. An electrodynamometer in a power panel is often the load for a current transformer.

2.7 The RF Ammeter and True rms

Current to a radio transmitting antenna is commonly passed through a small resistor, a shunt, that is thermally connected to a thermocouple or other thermometer and mounted in a thermally insulating blanket. The rise in temperature is a measure of the current and is often sensed with a thermocouple.

This is an example of true rms indication. Root mean square current is the square root of the integral of the square of the instantaneous current over an unspecified time divided by that time. It is intended

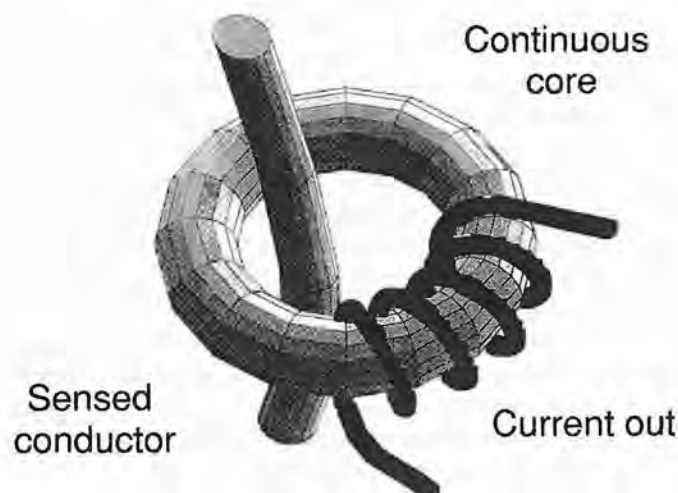


FIGURE 2.5 The ideal current transformer or CT is tightly coupled with no magnetic gap. Current in the secondary exactly balances current in the single-turn primary so that the magnetic flux in the core is zero.

to represent a stationary ac waveform by a single value that is equal to the direct current which would dissipate the same power in a resistive load. The RF ammeter does that precisely. The indication is not particularly linear, but it can easily be calibrated by applying dc to the input.

Other schemes for measuring rms current depend on analog multipliers and subsequent integration. They are limited by crest factor, the ratio of highest instantaneous peak to the rms over a measuring period. Inexpensive meters simply measure the peak, assume a sinusoidal waveform, and scale to rms.

2.8 The Current Transformer

Consider the magnetics of a toroidal core of high- μ material through which a current-carrying conductor passes. Include a secondary winding of n turns as shown in Figure 2.5. The secondary winding is connected to a low-resistance load.

In this current transformer, universally referred to as a CT, alternating current in the single-turn primary attempts to magnetize the core but in so doing, creates an emf and current in the secondary that tend to cancel the field. If the secondary truly has zero resistance, the current in it exactly cancels the field due to the primary. The result is a secondary current equal to the primary current divided by the number of secondary turns. The secondary current is in phase with the primary current. Because of the tightly closed magnetic loop, there is little effect from nearby conductors or position of the primary wire in the hole.

The secondary circuit can now be connected to a low-resistance current- or power-sensing device with assurance of calibration. But the secondary resistance is never really zero and the magnetic coupling is never perfect, so there are other considerations.

First, the concept of *secondary burden* is introduced. It is called that to avoid calling it a "load" because it behaves differently; the best burden is a short-circuit. Burden is sometimes expressed as a secondary resistance in ohms, but more often as an equivalent power in kVA for a defined current without consideration of phase. When the burden is not a perfect short-circuit, energy is dissipated and the magnetic field present in the core is no longer zero. The secondary current leads the primary current with a phase that depends on frequency.

Manufacturers of CTs have techniques to optimize accuracy of the CT when specified for a particular burden. The finished units might not have the number of secondary turns one would expect, but will nonetheless provide results accurate to a percent or so. They have laminations selected to minimize heating of the core. One will see ratings like 100:5, meaning 100 A in the primary will produce 5 A in

the secondary rather than "20 turns." They should be installed in a circuit that provides the burden for which they were calibrated. The voltage across a burden resistor is commonly amplified and passed to a data collection device. CTs for large currents need to be large to avoid magnetic saturation when burdened. Cores are prepared with laminates of silicon iron in the form of disks, concentric rings, or tape that is wound on a bobbin. Even with the best of materials, eddy current and hysteresis loss are present. When power dissipation is unacceptable, another choice of sensor might be preferable.

Most CTs are used for measurement of low-frequency power and energy. They are found at the front end of kilowatt-hour meters used by power providers. Radio frequency current in transmitting antennas can be measured with suitable core material. Ferrite cores with appropriate compensation are used for sensing pulse width modulated current in switching power supplies. Very large cores are used to sense pulsing beams of high-energy particles. Some oscilloscope probes are highly compensated CTs with a core that can be opened to allow a current-carrying wire to be introduced. With modern winding equipment for toroids, it is possible to put 2000 or more turns on a secondary [Coilcraft]. The CT then begins to look more like a current-to-voltage converter in its own right without need for very small values of the burden resistor and subsequent voltage amplification.

Safety Note

The secondary of a CT should always remain shorted while there is any possibility of current flow in the primary. With an open secondary, the core can be driven back and forth between saturation in opposite directions, resulting in high-speed changes in the internal B field. The result is very high, dangerous to life, voltages on the open secondary. Insulation in the secondary circuit can be damaged by arcing. Many CTs are made with a provision for shorting the secondary if the circuit must be opened. Use it.

2.9 Gapped Inductive Sensors

It is common practice in the design of transformers to introduce a small gap in the magnetic path. For even very small gaps, the magnetic properties of the magnetic loop become almost completely determined by the length of the gap, the rest of the material serving only to contain the lines of flux. Analysis of such a device begins with understanding that the B field is continuous in the core and through the gap. The H field is not, but it still satisfies the relation that the line integral of $\mathbf{H} \cdot d\mathbf{l}$ around the core is equal to the linked current. For a magnetic path of length s in material of permeability μ with a gap g , the ratio B/H , the effective permeability is given by:

$$\mu_{\text{eff}} = \frac{s}{g + \frac{s}{\mu}} \quad (2.6)$$

which applies for g much smaller than s . Note that when μ is sufficiently large, the effective permeability becomes independent of μ .

Introducing a gap into what would otherwise be a CT and drastically increasing the secondary turns count to 10,000 or more results in a current sensor that is intrinsically safe because the core cannot saturate: Figure 2.6 [SRT]. Because the B field is always small, the heating effect of eddy currents is less important than in the CT. When loaded with an appropriate resistor, the high inductance of the secondary causes the sensor to act like a current source that generates a voltage across the load proportional to the primary current with better than 1% linearity. Useful bandwidths of a sensor can be over 3 decades. The output impedance is high and requires the use of electronic voltmeters. Power dissipation is low, even for very high current models. Output voltage is high enough that simple diode rectifiers can be used to provide for dc output to further processing. In many cases, such a sensor can be used without any special electronics other than a voltmeter.

provides the burden for amplified and passed to a saturation when burdened. electric rings, or tape that is is loss are present. When e.

ey are found at the front in transmitting antennas compensation are used for e cores are used to sense compensated CTs with a . With modern winding [Coilcraft]. The CT then need for very small values

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(2.6)

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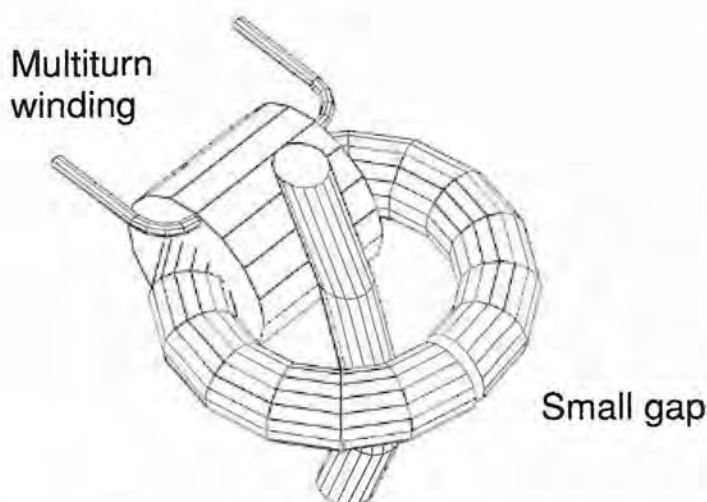


FIGURE 2.6 Placing a gap in the core and dramatically increasing the secondary turns count results in a current-to-voltage converter with high output voltage.

2.10 Hall Effect Sensor

The *Hall effect* as a sensor for magnetic fields is described in Chapter 12 of this volume and in [5]. It depends on a semiconductor crystal selected for its high carrier mobility and is placed in a magnetic field. A current is passed through the crystal along an axis perpendicular to the field. The carriers assume a mean velocity that causes them to be acted upon by the field and they move toward the other axis perpendicular to the field. The result is an emf at the faces of the crystal that can be measured. The emf is proportional to the field, the bias current, and the mobility.

In principle, such a field sensor could be placed near a current-carrying wire and oriented to sense the field created by the current, Figure 2.1, but the sensitivity is insufficient and there would always be interfering fields from currents in other nearby wires. A flux concentrator that looks like a CT with a large gap is always used. See Figure 2.7.

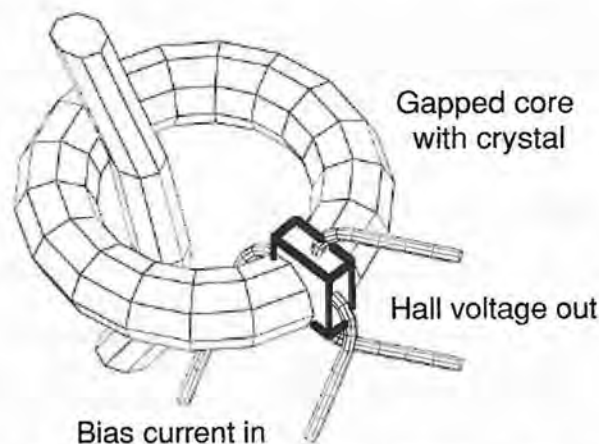


FIGURE 2.7 A semiconducting crystal is placed in the gap of a flux concentrating magnetic core. Bias current on one axis of the crystal produces a Hall voltage on the other.

The device is sensitive to direct current and the polarity is preserved. The Hall voltage is a few millivolts and amplification is always required. Zero drift in the amplifiers must be properly compensated although this is not so important for ac service. The bias current must be carefully controlled and it can be used to provide intrinsic analog multiplication and metering of power if it is made proportional to the circuit voltage.

Sensitivity is best with the smallest gap but there must be room for the crystal so gaps are larger than in the gapped inductive sensors. Fringing of the field in the larger gap reduces the natural shielding of the toroid from unwanted magnetic fields.

The accuracy and linearity of the Hall effect sensor can be improved in closed-loop mode. A feedback winding is added to the core and driven by a servo amplifier. The emf from the Hall device is used to drive the servo amplifier until the field is zero. The output is then the feedback current which is less than the sensed current by the number of turns in the feedback winding. The frequency response of the closed-loop system is surprisingly good, hundreds of kilohertz [F. W. Bell].

2.11 Clamp-On Sensors

It is often desirable to measure current in an existing system without removing power in order to install a device; thus, most of the magnetic devices are available in a clamp-on configuration. The variety ranges from tiny oscilloscope probes to clamps for 3000 A power buses.

Accuracy is always reduced in clamp-on mode because the clamp itself constitutes a gap that is uncontrollable and subject to wear. Some manufacturers provide highly polished surfaces that slide together. Others have iron fingers that interlace as the clamp is closed. Still others do not worry about it because the instrument is intended for field use where accuracy is not so critical.

Some units have handles for one-hand operation; others require a wrench or other tool. An interesting variation is the flexible core by [Flexcorp]. Installation is by bending the core and allowing it to spring back to shape.

2.12 Magnetoresistive Sensors

Most of the features of a Hall effect sensor are available if the Hall crystal is replaced by a device whose resistance changes with magnetic field. The discovery of giant magnetoresistive devices has recently made this idea attractive. [NVE]

Such devices still exhibit rather small resistance change and are sensitive to other effects such as temperature, so it is imperative that they be used in self-compensating bridge circuits in the manner of a strain gage. They are also insensitive to the polarity of the field.

Zetex has delivered magnetoresistive current sensors using thin-film permalloy in a variety of printed circuit packages. They are constructed in the form of a bridge and require bias and a differential amplifier. Measured current up to 20 A passes through the chip via its solder pins.

2.13 The Magnetic Amplifier

The efficiency of a transformer can be adjusted by a dc or low-frequency current that moves the operating point on a hysteresis curve. Excitation, a pump, is required at a higher frequency; it is passed through the transformer and then synchronously rectified and filtered into a higher power representation of the low-frequency signal. The magnetic configuration must be designed so that the pump does not put power into the signal circuit. Figure 2.8 shows one such arrangement. The pump coils are phased in series so that the associated flux cancels in the center leg of the E-E transformer core. The output coils are also in series and phased to add. When dc is applied to the center winding, it saturates both sides and reduces the signal on the output coils. More complicated arrangements can preserve polarity in the phase of the output.

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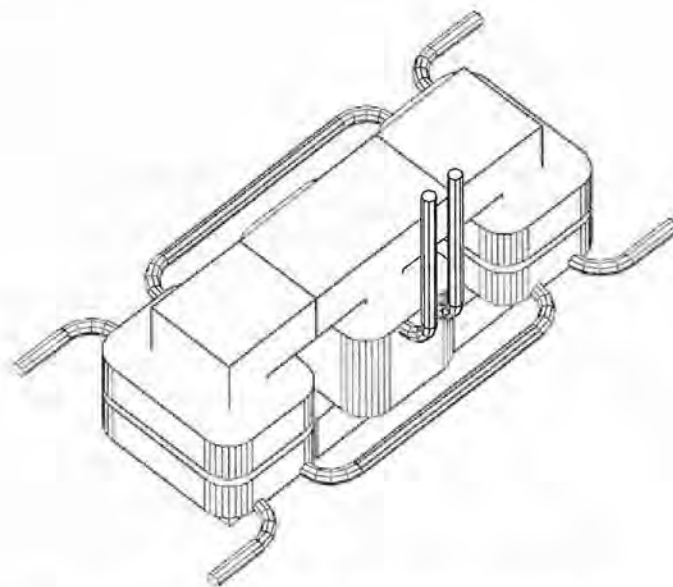


FIGURE 2.8 The windings of a magnetic amplifier. For simple current detection, the coils can be as little as two turns on ferrite toroids using an RF pump.

As a current measuring device the magnetic amplifier leaves much to be desired in linearity and frequency response, but it does provide isolation and is limited in sensitivity only by the number of turns placed on the center winding. The 20 mA dc off-hook current in a telephone is one such application.

2.14 Fluxgates

In its simplest form, a *fluxgate magnetometer* uses a driving coil to drive a high-permeability rod into saturation, first in one direction and then in the other. A second winding observes the rate of change of the B field inductively. In the absence of an external magnetic field, the observed signal is symmetric; but when an external field shifts the hysteresis loop to the right or left, the symmetry is lost. An amplifier tuned and phase-locked to the second harmonic of the driving frequency can be used to determine the amplitude of the external field.

In principle, such a fluxgate could be used to sense the field in the gap of a flux concentrator as with the Hall sensor, but it is too big and the linearity would be unacceptable. It is better to think of a way to drive the whole flux concentrator with a pumping frequency similar to that used in a magnetic amplifier.

Driving a toroid this way has the undesirable effect of coupling the pump energy into the circuit being measured, but one can pump two toroids in opposite directions. Now a current to be sensed passes through both cores and biases one in one direction and one in the other relative to the pump. A pickup winding senses the second harmonic, which is demodulated with a phase-locked sensor to preserve the direction of the sensed current.

Linearity is improved by adding one more winding through both toroids; it is driven by a servo amplifier that is satisfied only when the amplitude of the second harmonic is zero; the flux at the measured frequency is zero. This is now the ideal current transformer. At the frequency of the sensed current, which might be dc, there is zero average flux in the toroids and the output of the servo amplifier is a true representation of the current being measured, reduced by the turns count of the servo winding.

Unfortunately, complicated electronics and considerable power are required to accomplish all of this, but the resulting accuracy is significantly better than anything other than precision shunts. Two suppliers — GMW and Holec — provide equipment that uses similar principles. Many of the details are either patented or proprietary.

2.15 Optical Sensors

The Faraday effect is a rotation of the plane of polarization of light as it passes in a transparent medium parallel to a magnetic field. Optical fibers that do not randomize polarization are available with a small Faraday coefficient. Winding such a fiber on a form to be installed around a wire so that the light propagates parallel to the field produced by current in the wire, Figure 2.1, results in a sensor. A measurable rotation proportional to the current comes about because of the long optical path. When the myriad problems of preparation are solved, the result is a sensor that looks externally like a current transformer but has no wires. Using a reflecting ring made of yttrium-iron-garnet with a large Faraday coefficient, NIST reports sensitivity of 220 nA [6]. Matsushita Electric Industrial Company makes sensors using thin garnet films.

Analysis of the polarization requires a polarized light source and a polarization analyzer at the other. An advantage of such a sensor is that the fiber leading to and from can be long enough to allow isolation for very high voltages.

Winding and annealing of fiber sensors without destroying the polarization-preserving properties of the fiber and temperature sensitivity of the Faraday coefficient must be addressed. Further information on these sensors can be found in [7-9]. ABB, Sweden, reports a maximum detectable current of >23 kA, a sensitivity of about 2 A, and a relative error of $\pm 0.15\%$ [7].

2.16 Fault Indicators

Latching indicators which save an indication of high pulse current which was present sometime in the past are needed for power distribution systems subject to lightning strikes. Such indicators are often a bar magnet that moves into a latched position when a current pulse occurs. Readout is visual and they can be reset and installed on live wires using high-voltage tools.

2.17 Other Schemes

Cantor of Ford Aerospace has described a flux reset transformer scheme for measurement of direct current [10]. An oscillator and semiconductor switch periodically drive a core into reverse saturation. When the drive current is switched off, the flux rises due to dc in the sense winding and produces a voltage on the drive winding that is sensed synchronously with the oscillator. No commercial device based on this principle is available.

The double-balanced mixer, familiar to RF engineers, is also a sensor for low-frequency current. The IF port is usually dc-coupled through diodes in a ring configuration. dc applied there will modulate an RF pump applied to one of the RF ports and recovered on the other. There are no known products that use this principle.

2.18 Some Generalities and Warnings

Except for very simple loads, the waveform of the current drawn by a load does not resemble the voltage waveform. Besides the well-known phase shift and power factor, current flows in harmonics and sub-harmonics of the power frequency. Bridge rectifiers with capacitor input filters draw pulses of current near the peaks of the voltage waveform. Triacs cause a phase shift of the fundamental and introduce odd harmonics. Triacs in full-cycle mode with zero current switching can draw full current for a few cycles, followed by zero current for a few more introducing frequency components below the power frequency. Frequency changers are likely to draw current in pulse width modulated bursts.

A classic error is to measure true rms voltage and true rms current and multiply the two to get power. Even after correcting for phase, this is usually wrong. In short, accurate measurement demands some

knowledge of the characteristics of the load. Beware of sensors labeled "true rms" for they can be anything but.

2.19 Current Actuated Switches and Indicators

Another form taken by current sensors is the current actuated switch. There was a time when contacts were placed on the pointer of a D'Arsonval movement to establish upper and lower limits for a process variable. The modern way is to configure a sensor so that it operates a solid-state switch.

When operating power is available, any of the current sensors can be configured as a switch; but when the switch must be controlled solely by the power that can be extracted from the current being measured, the gapped toroid is superior. The high voltage available can be rectified to control MOSFETs or the base of an open collector Darlington transistor [SRT].

One company — CR Magnetics — markets a light-emitting diode, LED, indicator that shows the presence of alternating current without any connection or external source of power.

Circuit breakers can be implemented with a magnetic coil or with a bimetallic thermal element. For completeness, there is also the fuse.

Ground fault breakers use a toroidal CT through which the line current is passed twice: once on the way to the load and again on the way back to the neutral wire in such a way as to cancel. Any fault current to ground associated with the load causes the return current to be less than the source and induces voltage on the secondary of the toroid. This is amplified and used to trip a switch. This application demands absolute insensitivity to the position of a current-carrying wire in the hole of the toroid. Gapped sensors typically are not suitable because of magnetic leakage.

2.20 Where to Get Current Sensors

CTs, shunts, and some other sensors are commodity items available from a variety of manufacturers and distributors. Well-known manufacturers of switchgear — General Electric and Westinghouse — have their own CTs for use with their power meters. Test equipment manufacturers — Simpson, Tektronix, Fluke, Extech, and the like — offer hand-held current meters. Table 2.1 shows distributors and manufacturers of current sensing equipment. There is no way that a list such as this can be complete. It has selected itself from those companies willing to contribute to the author's library.

TABLE 2.1 Selected Manufacturers of Current Sensors

AEMC Instruments/Chauvin Arnoux, Inc.

200 Foxborough Blvd.

Foxborough, MA 02035

Tel: (800) 343-1391, (508) 698-2115, Fax: (508) 698-2118

<http://www.aemc.com>

Makes a selection of clamp-on and flexible-core current transformers which are suitable for extending the range of other alternating current meters.

ABB

SSAC division of ABB Inc has a line of current actuated switches and sensors.

ABB has its own line of current metering equipment for the power industry and has been involved with optical current sensing using the Faraday effect.

<http://www.ssac.com>

Sales are through:

Entrelec Inc. (USA headquarters)

1950 Hurd Drive

Irving, TX 75038-4312

Tel: (972) 550-9025, (800) 431 2308, Fax: (972) 550-9215, (800) 862-5066

entrelec.info@us.abb.com

<http://www.entrelec.com/ssacofclink>

TABLE 2.1 (continued) Selected Manufacturers of Current Sensors

Amecon, Inc.

1900 Chris Lane

Anaheim, CA 92805

Tel: (714) 634-2220, (800) 394-2234, Fax: (714) 634-0905

<http://www.ameconinc.com>sales@ameconinc.com

Offers a line of smaller CTs and a Hall effect device

American Aerospace Controls

570 Smith St.

Farmingdale, NY 11735

Tel: (516) 694-5100, Fax: (516) 694-6739

<http://www.a-a-c.com>

Has a line of ac and dc current sensors for mounting in remote locations with wiring to collect data.

Coilcraft

1102 Silver Lake Rd.

Cary IL, 60013

Tel: (800) 322-2645, Fax: (847) 639-6400, (708) 639-6400, (847) 639-1469

<http://www.coilcraft.com>

Makes smaller current transformers usable up to 100 amps for mounting on printed circuit boards.

CR Magnetics Inc.

544 Axminister Drive

Fenton (St. Louis)

MO 63026

Tel: (636) 343-8518, Fax: (636) 343-5119

<http://www.crmagnetics.com>sales@crmagnetics.com

Provides a full line of current transformers and current actuated switches. Their catalog contains a useful discussion of accuracy classes and selection criteria.

Dranetz-BMI (Headquarters)

1000 New Durham Rd.

Edison, NJ 08818-4019

Tel: (800) 372-6832, Fax: (732) 248-1834

<http://www.dranetz-bmi.com>

Offers a line of equipment for analysis of electrical power and is included here even though it does not make instruments dedicated to current sensing.

Extech Instruments Corporation

335 Bear Hill Road

Waltham, MA 02154-1020

Tel: (617) 890-7440, Fax: (617) 890 7864

<http://www.extech.com>

Builds a line of hand-held meters suitable for current measurement to 2000 A.

F.W. Bell, A Division of Bell Technologi (sic)

6120 Hanging Moss Rd.

Orlando FL 32807

Tel: (407) 677-6900, (800) 775-2550, Fax: (407) 677-5765

<http://www.fwbell.com>

Offers noncontact sensors mostly using the Hall effect with a variety of output connections.

Fluke

<http://www.fluke.com>

Numerous sales offices and distributors, has a line of sensors for their hand-held instruments.

TABLE 2.1 (continued) Selected Manufacturers of Current Sensors

Flex-Core division of Morlan and Associates

6625 McVey Blvd.

Columbus, OH 43235

Tel: (614) 889-6152, (614) 876-8308, Fax: (614) 876-8538

<http://www.flex-core.com>

mailto:flexcore ail.msn.com

Makes a line of sensors and current actuated switches. Of particular interest is their flexible iron core which can be installed over an existing conductor by bending it.

GMW Danfysik

GMW Associates

955 Industrial Road

San Carlos, CA 94070

Tel: (650) 802-8292, Fax: (650) 802-8298

<http://www.gmw.com/index.html>

sales@gmw.com

Offers a line of laboratory grade equipment for precision noncontact measurement. Precision in the ppm range and prices to match. Up to 10000 A, dc to more than 200 kHz. Large area CTs suitable for beam current in accelerators under the Bergoz tradename.

LEM USA, Inc. is the North American subsidiary of

LEM holding SA of Geneva, Switzerland

6643 W. Mill Road

Milwaukee, WI 53218

Tel: (414) 353-0711, Fax: (414) 353-0733

<http://www.lemusa.com>

Offers Hall effect sensors in both open- and closed-loop form for ac and dc currents up to 18000 A and up to 500 kHz.

Honeywell Inc. (Microswitch)

<http://www.honeywell.com>

<http://content.honeywell.com/sensing>

Offers current sensors mostly using the Hall effect along with their line of magnetic sensors and switches.

Magnetics Division of Spang, Inc.

P.O. Box 11422

Pittsburgh, PA 15238-0422

Tel: (412) 696-1333, (800) 245-3984, Fax: (412) 696-0333

<http://www.mag-inc.com>

<http://www.mag-inc.com/software/transformer.asp>

magnetics@spang.com

Offers core material suitable for current transformers with design data.

Metrosil unit of M&I Materials Ltd.

PO Box 136

Manchester, UK

M60 1AN

Tel: +44(0)161 875 4332, Fax: +44(0)161 875 2695

metrosilsales@metrosil.com

http://www.metrosil.com/protection_units.htm

Supplies protective devices for use on secondaries of current transformers.

Neilsen-Kuljian, Inc./NK Technologies

1400 Dell Avenue, Suite A

Campbell, CA 95008-6620

Tel: (800) 959-4014, (408) 871-7510, Fax: (408) 871-7515

<http://www.nktechnologies.com>

mailto:sales@nktechnologies.com

Current transformers, other current sensors.

TABLE 2.1 (continued) Selected Manufacturers of Current Sensors

NVE Corporation (Nonvolatile Electronics)

11409 Valley View Road

Eden Prairie, MN 55344-3617

Tel: (952) 829-9217, (800) 467-7141, Fax: (952) 996-1600

<http://www.nve.com>info@nve.com<http://www.nve.com>

Offers a giant magnetoresistive technology for current sensing.

NxtPhase Corporation

3040 East Broadway

Vancouver BC V5M 1Z4

Tel: (604) 215-9822, (604) 215-9833

<http://www.nxtphase.com>http://www.nxtphase.com/sub-products-optical_current-nxvct_voltage.htminfo@nxtphase.com

Offers high kilovolt isolated optical current and voltage measurement.

Ohio Semitronics, Inc.

4242 Reynolds Dr.

Hilliard, OH 43026

Tel: (800) 537-6732, (614) 777-1005, Fax: (614) 777-4511

sales@ohiosemitronics.com<http://www.ohiosemi.com>

Offers a wide range of current, voltage, and power transducers for ac and dc.

Pearson Electronics Inc.

4009 Transport Street

Palo Alto, California 94303

Tel: (650) 494-6444, (650) 494-6716

<http://www.pearsonelectronics.com>

Provides high-speed compensated current transformers for pulse work with oscilloscopes.

Smith Research & Technology, Inc. (SRT)

3109 S. Cascade Ave #201

Colorado Springs, CO, 80907-5190

Tel: (719) 634-2259, Fax: (719) 634-2601

Offers gapped inductive sensors with ac, dc, and 4-20 mA output up to 1000 A. Ac current actuated switches which control isolated 200 W loads using no external power for the switch are offered.

SSAC - see ABB

Tektronix

<http://www.tektronix.com>

Offers current sensors and hand-held instruments as well as current-sensing probes for oscilloscopes.

Zetex plc

Lansdowne Road

Chadderton

OL9 9TY

United Kingdom

Tel: +44 (0) 161 622 4444, Fax: +44 (0) 161 622 4720

Tel: (516) 543-7100, Fax: (516) 864-7630 in the USA

<http://www.zetex.com>This supplier of integrated circuits and transistors offers magnetoresistive current and magnetic field sensors (ZMZ20) for use on printed circuit boards.

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Power Measurement

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3.1	Power Measurements in Dc Circuits	3-1
3.2	Power Measurements in Ac Circuits	3-3
	Definitions • Low- and Medium-Frequency Power Measurements • Line-Frequency Power Measurements • High- Frequency Power Measurements	
3.3	Pulse Power Measurements	3-30

In this chapter, the concept of electric power is first introduced, and then the most popular power measurement methods and instruments in *dc*, *ac*, and *pulse* waveform circuits are illustrated.

Power is defined as the *work performed per unit time*. So, dimensionally, it is expressed as joules per second, $J s^{-1}$. According to this general definition, electric power is the electric work or energy dissipated per unit time and, dimensionally, it yields:

$$Js^{-1} = JC^{-1} \times Cs^{-1} = V \times A \quad (3.1)$$

where J = Joules
 s = Seconds
 C = Coulombs
 V = Volts
 A = Amperes

The product *voltage times current* gives an electrical quantity equivalent to *power*.

3.1 Power Measurements in Dc Circuits

Electric power (P) dissipated by a load (L) fed by a dc power supply (E) is the product of the voltage across the load (V_L) and the current flowing in it (I_L):

$$P = V_L \times I_L \quad (3.2)$$

Therefore, a power measurement in a dc circuit can be generally carried out using a voltmeter (V) and an ammeter (A) according to one of the arrangements shown in Figure 3.1. In the arrangement of Figure 3.1(a), the ammeter measures the current flowing into the voltmeter, as well as that into the load; whereas in the arrangement of Figure 3.1(b), this error is avoided, but the voltmeter measures the voltage drop across the ammeter in addition to that dropping across the load. Thus, both arrangements give a surplus of power measurement absorbed by the instruments. The corresponding measurement errors are generally referred to as insertion errors.

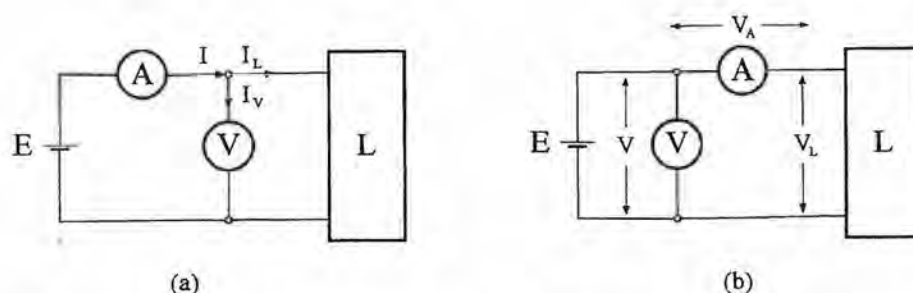


FIGURE 3.1 Two arrangements for dc power measurement circuits.

According to the notation:

- I Current measured by the ammeter (Figure 3.1(a))
- V Voltage measured by the voltmeter (Figure 3.1(b))
- R_V, R_A Internal resistance of the voltmeter and the ammeter, respectively
- R_L Load resistance
- I_V Current flowing into the voltmeter (Figure 3.1(a))
- V_A Voltage drop across the ammeter (Figure 3.1(b))

the following expressions between the measurand electric power P and the measured power $V \times I$ are derived by analyzing the circuits of Figures 3.1(a) and 3.1(b), respectively:

$$P = V_L \times I_L = V \times I \times \left(\frac{R_V - R_L}{R_V} \right) \quad (3.3)$$

$$P = V_L \times I_L = V \times I \times \left(\frac{R_L - R_A}{R_L} \right) \quad (3.4)$$

If:

- I_V , compared with I
- V_A , compared with V

are neglected for the arrangements of Figure 3.1(a) and 3.1(b), respectively, it approximately yields:

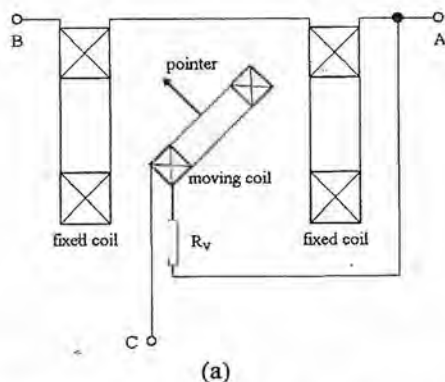
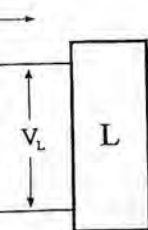
$$\frac{I_V}{I} = \frac{R_L}{R_V + R_L} \approx \frac{R_L}{R_V} \approx 0; \quad \frac{V_A}{V} = \frac{R_A}{R_A + R_L} \approx \frac{R_A}{R_L} \approx 0; \quad (3.5)$$

consequently, measured and measurand power will be coincident.

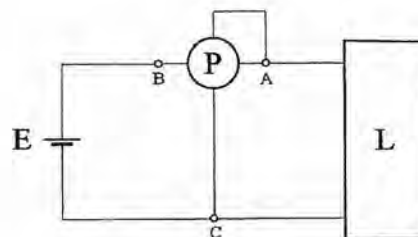
On this basis, from Equations 3.3, 3.4, and 3.5, analytical corrections of the insertion errors can be easily derived for the arrangement of Figures 3.1(a) and 3.1(b), respectively.

The instrument most commonly used for power measurement is the *dynamometer*. It is built by (1) two fixed coils, connected in series and positioned coaxially with space between them, and (2) a moving coil, placed between the fixed coils and equipped with a pointer (Figure 3.2(a)).

The torque produced in the dynamometer is proportional to the product of the current flowing into the fixed coils times that in the moving coil. The fixed coils, generally referred to as *current coils*, carry the load current while the moving coil, generally referred to as *voltage coil*, carries a current that is proportional, via the multiplier resistor R_V , to the voltage across the load resistor R_L . As a consequence, the deflection of the moving coil is proportional to the power dissipated into the load.



(a)



(b)

FIGURE 3.2 Power measurement with a dynamometer. (a) Working principle; (b) measurement circuit.

As for the case of Figure 3.1, insertion errors are also present in the dynamometer power measurement. In particular, by connecting the voltage coil between A and C (Figure 3.2(b)), the current coils carry the surplus current flowing into the voltage coil. Consequently, the power P_L dissipated in the load can be obtained by the dynamometer reading P as:

$$P_L = P - \frac{V^2}{R'_v} \quad (3.6)$$

where R'_v is the resistance of the voltage circuit ($R'_v = R_v + R_{vc}$, where R_{vc} is the resistance of the voltage coil). By connecting the moving coil between B and C, this current error can be avoided, but now the voltage coil measures the surplus voltage drop across the current coils. In this case, the corrected value is:

$$P_L = P - I^2 R_C \quad (3.7)$$

where R_C is the resistance of the current coil.

3.2 Power Measurements in Ac Circuits

Definitions

All the above considerations relate to *dc* power supplies. Now look at power dissipation in *ac* fed circuits. In this case, electric power, defined as voltage drop across the load times the current flowing through it, is the function:

$$p(t) = v(t) \times i(t) \quad (3.8)$$

referred to as the *instantaneous power*. In *ac* circuits, one is mainly interested in the mean value of instantaneous power for a defined time interval. In circuits fed by periodic *ac* voltages, it is relevant to define the mean power dissipated in one period T (*active power* P):

$$P = \frac{1}{T} \int_0^T p(t) dt \quad (3.9)$$

The simplest case is a sinusoidal power supply feeding a purely resistive load. In this case, $v(t)$ and $i(t)$ are in phase and $p(t)$ is given by:

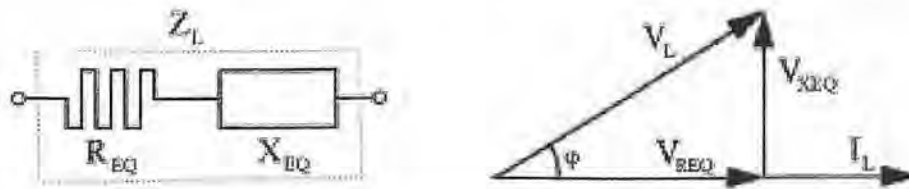


FIGURE 3.3 Voltage drop on the load and on its equivalent components.

$$p(t) = VI[1 - \cos(2\omega t)] \quad (3.10)$$

where V and I = rms values of $v(t)$ and $i(t)$, respectively

ω = power supply angular frequency

Therefore, the instantaneous power is given by a constant value VI plus the ac quantity oscillating with twice the angular frequency of the power supply; thus, the active power is simply the product VI . In this case, all the above considerations referring to active power for dc circuits are still correct, but voltages and currents must be replaced by the corresponding rms values.

The case of purely reactive loads is the opposite; the voltage drop across the load and current flowing through it are out of phase by 90° . Instantaneous power $p(t)$ is given by:

$$p(t) = VI \cos(2\omega t) \quad (3.11)$$

Thus, the active power dissipated by a reactive load is zero, owing to the phase introduced by the load itself between voltage and current.

The simplest cases of sinusoidal power sources supplying purely resistive and purely reactive loads have been discussed. In these cases, the load is expressed by a real or a pure imaginary number. In general, the load is represented by a complex quantity (the impedance value). In this case, load impedance can be represented by its equivalent circuit (e.g., a pure resistance and a pure reactance in series). With this representation in mind, the electric power dissipated in the load Z_L (Figure 3.3) can be expressed by the sum of power components separately dissipated by resistance R_{EQ} and reactance X_{EQ} of the equivalent circuit Z_L .

Considering that no active power is dissipated in the reactance X_{EQ} , it yields:

$$P = V_{REQ} I_L = V_L I_L \cos \phi \quad (3.12)$$

The term $\cos \phi$ appearing in Equation 3.12 is referred to as the *power factor*. It considers that only a fraction of voltage V_L contributes to the power; in fact, its component V_{REQ} (the drop across the resistance) does not produce any active power, as it is orthogonal to the current I_L flowing into the load.

Figure 3.4 plots the waveforms of instantaneous power $p(t)$, voltage $v(t)$, and current $i(t)$. The effect of the power factor is demonstrated by a dc component of $p(t)$ that varies from a null value (i.e., $v(t)$ and $i(t)$ displaced by 90°) toward the value VI (i.e., $v(t)$ and $i(t)$ in phase).

The term:

$$P_A = V_L I_L \quad (3.13)$$

is called the *apparent power*, while the term:

$$Q = V_{XEQ} I_L = V_L I_L \sin \phi \quad (3.14)$$



(3.10)

the ac quantity oscillating
is simply the product VI .
circuits are still correct, but

the load and current flowing

(3.11)

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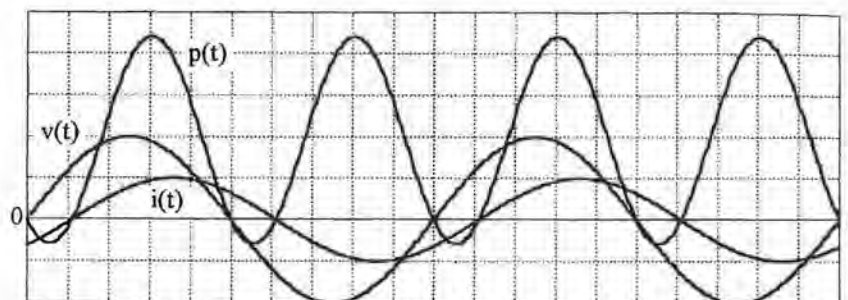


FIGURE 3.4 Waveforms of instantaneous power (p), voltage (v), and current (i).

is called the *reactive power* because it represents a quantity that is dimensionally equivalent to power. This is introduced as a consequence of the voltage drop across a pure reactance and, therefore, does not give any contribution to the active power. From Figure 3.3, the relationship existing between *apparent power*, *active power*, and *reactive power* is given by:

$$P_A = \sqrt{P^2 + Q^2} \quad (3.15)$$

Dynamometers working in ac circuits are designed to integrate instantaneous power according to Equation 3.9. Insertion errors can be derived by simple considerations analogous to the dc case. However, in ac, a phase uncertainty due to the not purely resistive characteristic of voltage circuit arises. In sinusoidal conditions, if ϵ_w (in radians) is the phase of the real coil impedance, and $\cos\phi$ is the load power factor, the relative uncertainty in active power measurements can be shown to be equal to $-\epsilon_w T_g \phi$. The phase uncertainty depends on the frequency. By using more complex circuits, the frequency range of the dynamometer can be extended to a few tens of kilohertz.

The above has presented the power definitions applied to ac circuits with the restrictions of sinusoidal quantities. In the most general case of distorted quantities, obviously symbolic representation can no longer be applied. In any case, active power is always defined as the mean power dissipated in one period.

As far as methods and instruments for ac power measurements are concerned, some circuit classification is required. In fact, the problems are different, arising in circuits as the frequency of power supply increases. Therefore, in the following, ac circuits will be classified into (1) line-frequency circuits, (2) low- and medium-frequency circuits (up to a few megahertz), and (3) high-frequency circuits (up to a few gigahertz). Line-frequency circuits will be discussed separately from low-frequency circuits, principally because of the existence of problems related specifically to the three-phase power supply of the main.

Low- and Medium-Frequency Power Measurements

In the following, the main methods and instruments for power measurements at low and medium frequencies are considered.

Three-Voltmeter Method

The power dissipation in the load L can be measured using a noninductive resistor R and measuring the three voltages shown in Figure 3.5 [1]. Although one of the voltages might appear redundant on a first analysis of the circuit, in actual fact, three independent data are needed in order to derive power from Equation 3.12. In particular, from voltage drops v_{AB} and v_{BC} , the load current and load voltage can be directly derived; instead, v_{AC} is used to retrieve information about their relative phase.

If currents derived by voltmeters are neglected and the current i_L flowing into the load L is assumed to be equal to that flowing into the resistor R , the statement can be demonstrated as follows:

TABLE 3.1 Operating Field of Main Radiation Power Sensors

Operating Field (Wavelength)	Microwave (1, 10^{-3} m)	Infrared (10^{-3} , 10^{-6} m)	Visible and Ultraviolet (10^{-6} , 10^{-9} m)	Nuclear Rays (10^{-8} , 10^{-15} m)
Sensors	Noncontacting displacement sensors	Pyroelectric, photoconductive, photovoltaic	Photoconductive, photovoltaic	Scintillation counters, plastic films, solid-state, thermolum

converts the incident radiation to pulses of light, and a light-electric pulse converter. The active material can be a crystal, a plastic fluorine, or a liquid. The scintillator size varies greatly according to the radiation energy, from thin solid films to large tanks of liquid to detect cosmic rays. A thin ($5\text{ }\mu\text{m}$) plastic polycarbonate film or a thermoluminescent material (e.g., LiF) can measure the radiation power falling on a surface. The film is mechanically damaged by the propagation of highly α -ionizing particles. Consequent etching of the film reveals tracks that can be observed and counted.

3.3 Pulse Power Measurements

Pulse waveforms are becoming more and more diffused in several fields such as telecommunications, power source applications, etc. The *pulse power* P_p is defined as the average power P_m in the pulse width:

$$P_p = \frac{P_m}{\tau_d} \quad (3.71)$$

where τ_d is the duty cycle of the pulse waveform (i.e., the pulse width divided by the waveform period). If the pulse width cannot be accurately defined (e.g., nonrectangular pulses in the presence of noise), the pulse power P_p becomes unmeaningful. In this case, the *peak envelope power* is introduced as the maximum of the instantaneous power detected on a time interval, including several periods of the pulse waveform (but negligible with respect to the modulation period, in the case of PWM waveforms).

Several techniques are used to measure pulse power [21]. In particular, they can be classified according to the *pulse frequency* and the necessity for *constraining real-time applications* (i.e., measuring times, including a few of the modulation periods). For real-time, low-frequency applications (up to 100 kHz), the algorithms mentioned in the above sections on wattmeters based on digital multipliers can be applied [9].

If constraining limits of real-time do not have to be satisfied, either digital or analog techniques can be utilized. As far as the digital techniques are concerned, for high-frequency applications, if the measurand pulse waveform is stationary over several modulation periods, digital wattmeters based on equivalent sampling can be applied, with accuracies increasing according to measuring times. As far as the analog techniques are concerned, three traditional methods are still valid: (1) average power per duty cycle, (2) integration-differentiation, and (3) dc/pulse power comparison.

A block diagram of an instrument measuring *average power per duty cycle* is illustrated in Figure 3.39. At first, the mean power of the measurand pulse signal, terminated on a suitable load, is measured by means of an average power meter; then, the pulse width and the pulse waveform period are measured by a digital counter. Finally, the power is obtained by means of Equation 3.71.

The *integration-differentiation* technique is based on a barretter sensor capable of integrating the measurand, and on a conditioning and differentiating circuit to obtain a voltage signal proportional to the measurand power. The signal is input to the barretter sensor, having a thermal constant such that the barretter resistance will correspond to the integral of the input. The barretter is mounted as an arm of a conditioning Wheatstone bridge; in this way, the barretter resistance variations are transformed into voltage variations, and an integrated voltage signal is obtained as an output of the bridge detecting arm. This signal, suitably integrated to reach a voltage signal proportional to the output, is detected by a peak

Nuclear Rays
(10^{-8} , 10^{-15} m)

scintillation counters,
photographic films, solid-state,
thermoluminescence

er. The active material
according to the radiation
A thin ($5\text{ }\mu\text{m}$) plastic
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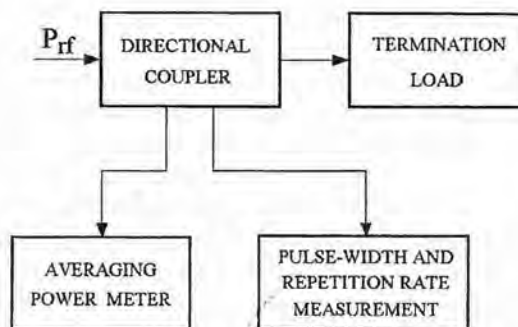


FIGURE 3.39 Block diagram of an instrument measuring average power per duty cycle.

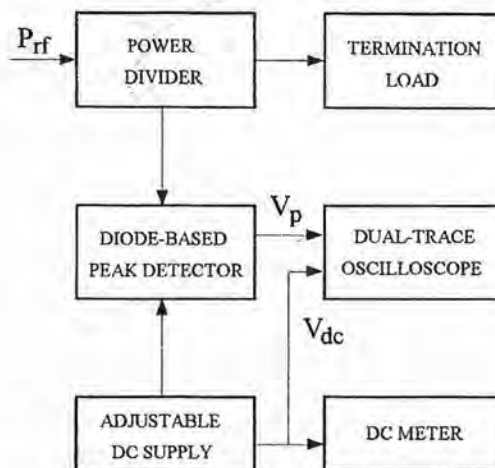


FIGURE 3.40 Block diagram of an instrument based on dc/pulse power comparison technique.

voltmeter calibrated in power. Analogously to the selection of the time constant of an RC integrating circuit, attention must be paid to the thermal constant selection of the barretter in order to attain the desired accuracy in the integration. With respect to the measurand pulse period, a very long thermal constant will give rise to insufficient sensitivity. On the other hand, a very short constant approaching the pulse duration will give rise to insufficient accuracy.

The *dc/pulse power comparison* technique is based on the concept of first revealing the peak envelope power through a diode sensor, and then comparing the peak to a known dc source with a dual trace scope. A block diagram of an instrument based on this concept is illustrated in Figure 3.40. The peak of the measurand pulse signal terminated on a suitable load is sensed by a peak detector by obtaining a proportional signal V_p . This signal is input to a channel of a dual trace oscilloscope and the envelope peak is displayed. An adjustable dc source is input to the other channel of the scope to obtain a signal V_{dc} to be compared to the envelope peak signal. When the two signals are made equal, a dc voltage meter directly calibrated in peak power measures the output power.

References

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Power Factor Measurement

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4.1	Reasons for Interest in Power Factor	4-2
4.2	Ac Electric Loads	4-3
	Linear Loads • Nonlinear Loads	
4.3	Ac Power Relationships	4-3
	Sinusoidal Voltage and Current • Single-Phase Circuits • Polyphase Circuits • Nonsinusoidal Voltage and Current • Power Factor	
4.4	Power Factor "Measurement"	4-13
	Metering for Linear Loads • Metering for Nonlinear Loads • Metering Applications	
4.5	Instrumentation	4-16

Electricity can be thought of as a means of delivering power from one place to another to do work. The laws and relationships for delivering power were originally developed for direct current. Power delivered, expressed in *watts*, was calculated by multiplying the voltage and current as shown in Equation 4.1.

$$P_{dc} = EI \quad (4.1)$$

The situation becomes more complex when alternating current is used to deliver power. Figure 4.1 shows a sine wave representing either ac current or voltage. Since the instantaneous value of the wave is continually changing, a numerical quantity is defined to represent an average property of this wave. This quantity, the root-mean-square, or **rms** value, calculated by squaring the instantaneous value, integrating it during one cycle, dividing by the period, and taking the square root of the result, is equal to the peak value of the ac wave divided by the square root of 2, or, for ac current, $I_{rms} = i_{peak}/\sqrt{2}$. In the physical world, a sine wave ac current having an rms value of 1 A (A = ampere), passed through a resistive load, produces the same heating effect as 1 A of dc current. Thus, one might expect delivered ac power to be easily calculated in watts using Equation 4.1 and inserting rms values for current and voltage. While this simple relationship holds true for the instantaneous voltage and current values as shown in Equation 4.1a in general, it is not true for the rms quantities except for the special case when the ac current and voltage are restricted to perfect sine waves *and* the load is a pure resistance.

$$p_{inst.} = ei \quad (4.1a)$$

In real-world situations where current and/or voltage waveforms are not perfectly sinusoidal and/or the loads are other than resistive, the relationships are no longer simple and the power delivered, or *active power*, is usually less than the product of rms voltage and current, as shown in Equation 4.2.

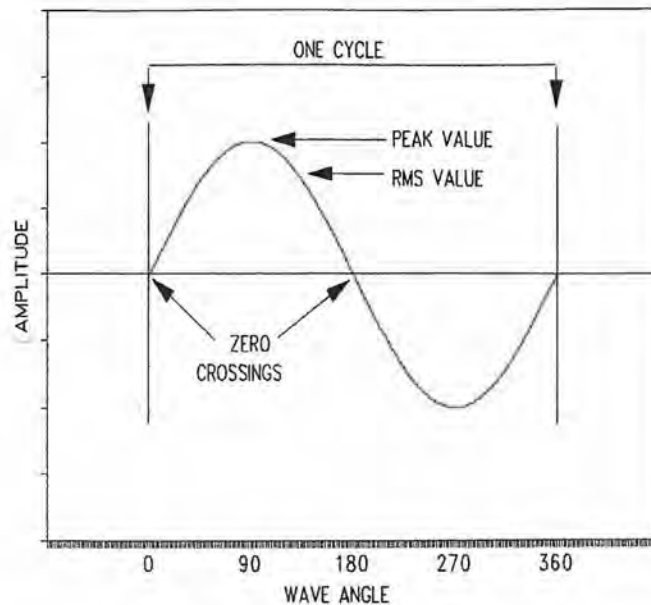


FIGURE 4.1 Sine wave characteristics. One cycle of a continuous wave is shown. The wave begins at a zero-crossing, reaches a positive peak, continues through zero to a negative peak, and back to zero. The wave repeats every 360° . The wave angle can also be expressed as a function of frequency f and time t . For a given frequency, the wave angle is related to the expression, $2\pi ft$.

$$P_{ac} \leq E_{rms} I_{rms} \quad (4.2)$$

The product of rms voltage and rms current does, however, define a quantity termed *apparent power*, U , as shown in Equation 4.3.

$$U = E_{rms} I_{rms} \quad (4.3)$$

A derived term, the *power factor*, F_p , used to express the relationship between delivered or active power, P , and apparent power, U , is defined by Equation 4.4.

$$F_p = \frac{P}{U} \quad (4.4)$$

From Equations 4.2, 4.3, and 4.4, it is clear that the value of F_p must lie in a range between zero and one.

This chapter focuses on: (1) ac power relationships and the calculation of power factor; (2) the physical meaning of these relationships; and (3) measurement techniques and instrumentation for determining these relationships and calculating power factor.

4.1 Reasons for Interest in Power Factor

Power factor is a single number that relates the active power, P , to the apparent power, U . Electric components of a utility distribution system are designed on a kVA basis; i.e., they are designed to operate at a given voltage and carry a rated current without undue temperature rise. Transformer and conductor sizes are chosen on this basis. While active power does useful work, reactive and harmonic powers do no useful work, absorb system capacity, and increase system losses; but reactive and harmonic powers are needed to provide magnetic fields or nonlinear currents. *The capacity of electric systems is limited*

by apparent power, not active power. Power factor expresses, with a single value, the extent to which an electrical distribution system is efficiently and effectively utilized. A low value for the power factor means that much of the system capacity is not available for useful work. From a utility viewpoint, this means reduced ability to deliver revenue-producing active power; from a user viewpoint, a low power factor reduces the available active power or requires increased system size to deliver needed power.

4.2 Ac Electric Loads

Linear Loads

Electric loads in ac power systems with sinusoidal voltages are categorized by the way they draw current from the system. Loads that draw sinusoidal currents, i.e., the current waveshape is the same as the voltage waveshape, are defined as *linear loads*. Historically, a high percentage of electric loads have been linear. Linear loads include: (1) induction motors; (2) incandescent lighting; and (3) heaters and heating elements. Linear loads use ac electric power directly to accomplish their functions.

Nonlinear Loads

Electric loads that draw nonsinusoidal currents, i.e., the current waveshape differs from the voltage waveshape, are defined as *nonlinear loads*. As energy savings and efficient use of electricity are emphasized, an increased percentage of nonlinear electric devices, both new and replacement, are being installed. Nonlinear loads include: (1) adjustable-speed motor drives; (2) fluorescent and arc-discharge lighting; (3) computers and computerized controls; and (4) temperature-controlled furnaces and heating elements. Nonlinear loads, rather than using ac electric power directly, often convert ac power into direct current before it is used to accomplish their functions. A common element in nonlinear loads is some kind of rectifier to accomplish this ac to dc conversion. Rectifiers do not draw sinusoidal currents.

4.3 Ac Power Relationships

Sinusoidal Voltage and Current

Power calculations for sinusoidal ac electric systems require knowledge of the rms voltage, the rms current, and the phase relationships between the two. Figure 4.2 illustrates possible phase relationships between voltage and current. If the positive-going zero-crossing of the voltage is considered the reference point, then the nearest positive-going zero-crossing of the current can occur at a wave angle either less than or greater than this reference. If the current zero-crossing occurs before the reference, the current is said to *lead* the voltage. If the current zero-crossing occurs after the reference, the current is *lagging*. If the zero-crossing for the current coincides with the reference, the two waves are said to be *in phase*. The wave angle, θ , by which the current leads or lags the voltage is called the *phase angle*, in this case, 30° .

Single-Phase Circuits

Power Calculations

The power delivered to do work is easily calculated [1]. Given a sinusoidal voltage of rms magnitude E and sinusoidal current of rms magnitude I , displaced by angle θ , at time t ,

$$\text{Instantaneous voltage} = e = \sqrt{2}E \sin(2\pi ft)$$

$$\text{Instantaneous current} = i = \sqrt{2}I \sin(2\pi ft - \theta) \quad (\text{note that } \theta \text{ can have a positive or negative value}) \quad (4.5)$$

$$\text{Instantaneous power} = p = ei$$

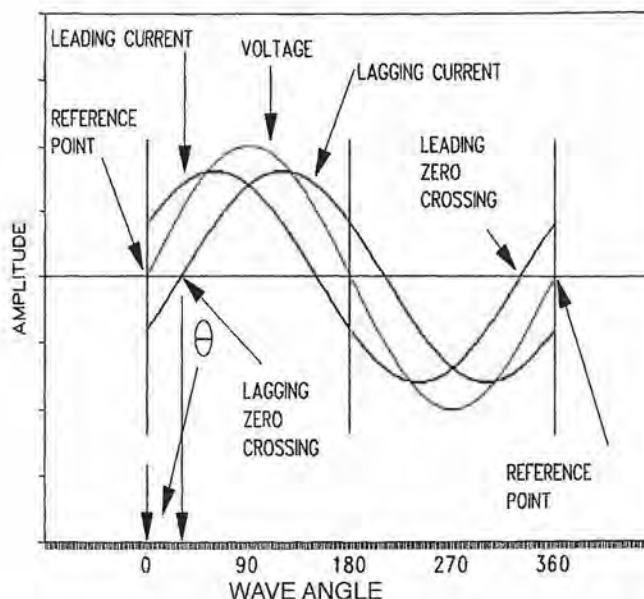


FIGURE 4.2 Sine wave phase angle. Two waves with the same zero-crossing are *in phase*. A sine wave that crosses zero before the reference wave is *leading*, and one that crosses zero after the reference wave is *lagging*. The phase angle θ illustrated is 30° lagging.

$$p = 2EI \sin(2\pi ft) \sin(2\pi ft - \theta)$$

$$p = EI \cos(\theta) - EI \cos(4\pi ft - \theta)$$

$$\text{Average power over an integral number of cycles } P = EI \cos(\theta) \quad (4.5)$$

$$\text{Power factor } F_p = \frac{P}{U} = \frac{EI \cos(\theta)}{EI} = \cos(\theta) \quad (4.5)$$

Equation 4.5 is the fundamental equation that defines power for systems in which the current and voltage are sinusoidal. The application of this equation is illustrated for three cases: (1) the current and voltage are in phase; (2) the current and voltage are out of phase by an angle less than 90° ; and (3) the current and voltage are out of phase by exactly 90° .

Ac Power Examples

Figure 4.3 shows voltage current and power when the voltage and current are in phase and the current displacement angle is zero (0°). (An example would be a resistance space heater.) The power curve is obtained by multiplying together the instantaneous values of voltage and current as the wave angle is varied from 0° to 360° . Instantaneous power, the product of two sine waves, is also a sine wave. There are two zero-crossings per cycle, dividing the cycle into two regions. In region (1) both the voltage and current are positive and the resultant product, the power, is positive. In region (2) both the voltage and current are negative and the power is again positive. The average power in watts, given by Equation 4.5, $EI \cos(0^\circ) = EI$, is the maximum power that can be delivered to do work. When sinusoidal voltage and current are in phase, the delivered power in watts is the same as for dc and is the maximum that can be delivered. The power factor, $F_p = \cos(0^\circ) = 1$, or unity.

Figure 4.4 shows voltage, current, and power when the current lags the voltage by 60° . (An example might be a lightly loaded induction motor.) The power sine wave again is obtained by multiplying together

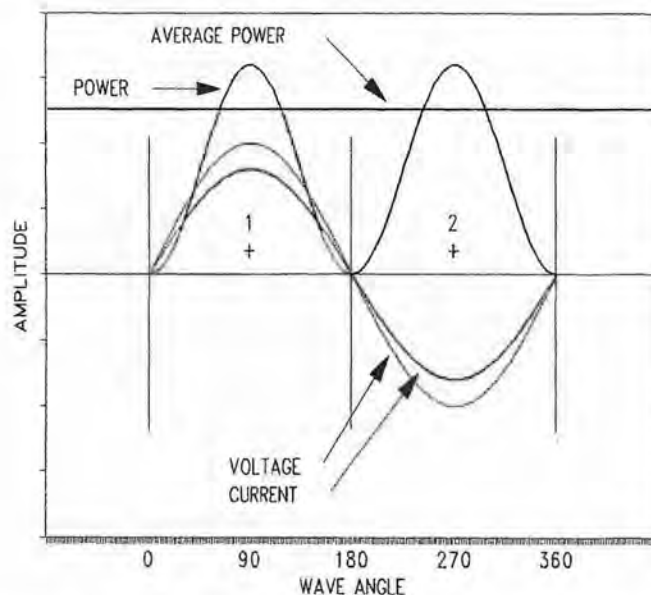


FIGURE 4.3 Voltage, current, and power for sine waves in phase. The vertical scales for voltage and current are equal. The scale for power is relative and is selected to permit the display of all three waves on a single graph. The voltage and current are in phase and both are sinusoidal. The power is everywhere positive, and average power delivered to do work is the maximum power.

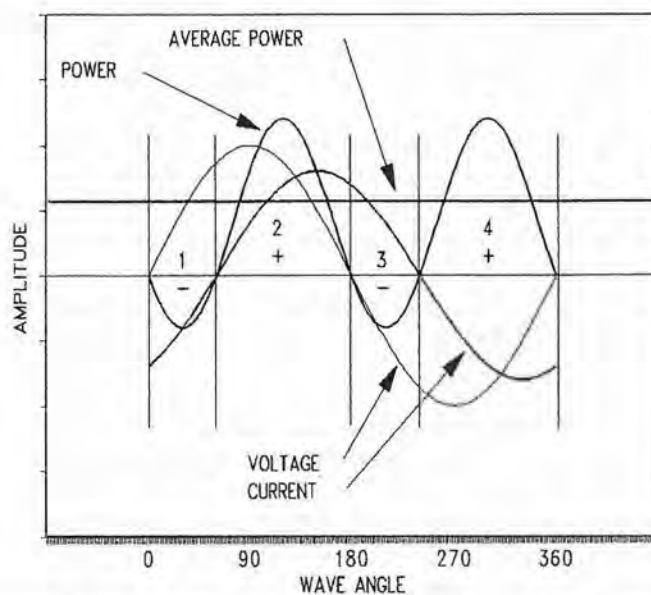


FIGURE 4.4 Voltage, current, and power for sine waves 60° out of phase. The vertical scales for voltage and current amplitudes are the same as those for Figure 4.3. Current lags voltage by 60° and both are sinusoidal. The power is positive in regions (2) and (4), and negative in regions (1) and (3). Average power delivered to do work is less than the maximum power.

the instantaneous values of voltage and current. There are now four zero-crossings per cycle, dividing the cycle into four regions. In regions (2) and (4), voltage and current have the same sign and the power is positive. In regions (1) and (3), voltage and current have opposite signs, resulting in a negative value

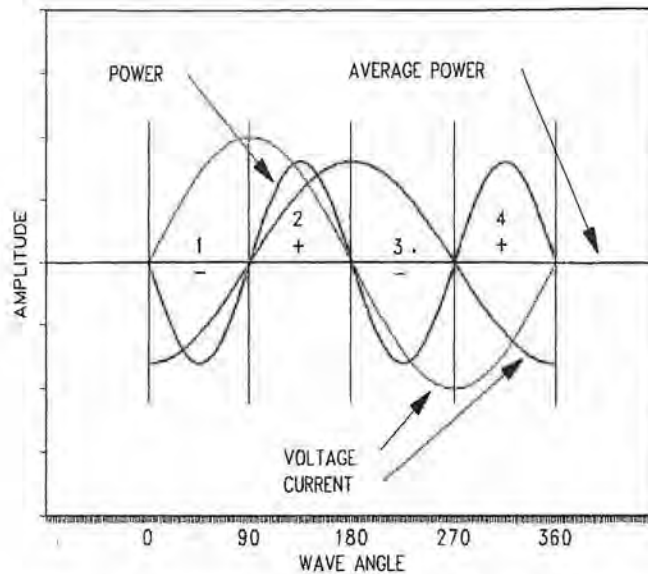


FIGURE 4.5 Voltage, current, and power for sine waves 90° out of phase. The vertical scales for voltage and current amplitudes are the same as those for Figure 4.3. Current lags voltage by 90° and both are sinusoidal. The power is positive in regions (2) and (4), negative in regions (1) and (3), and is of equal absolute magnitude in all four regions. Average power delivered to do work is zero.

for the power. The average power in watts, given by Equation 4.5, $EI\cos(60^\circ) = EI(0.5)$, is less than the maximum that could be delivered for the particular values of voltage and current. *When voltage and current are out of phase, the delivered power in watts is always less than the maximum.* In this example, $F_p = \cos(60^\circ) = 0.5$.

Figure 4.5 shows voltage, current, and power when the current lags the voltage by 90° . (This situation is not attainable in the real world.) The power sine wave again is obtained by multiplying together the instantaneous values of voltage and current. Again, four zero-crossings divide the cycle into four regions. In regions (2) and (4), the power is positive, while in regions (1) and (3), the power is negative. The average power in watts is given by Equation 4.5, $EI\cos(90^\circ) = 0$. No matter what the values of voltage and current, *when voltage and current are exactly 90° out of phase, the delivered power in watts is always zero.* The power factor, $F_p = \cos(90^\circ) = \text{zero}$.

Power Factor

Resolving the current into orthogonal components on a phasor diagram illustrates how delivered power can vary from a maximum to zero, depending on the phase angle between the voltage and the current sine waves. Figure 4.6 shows the voltage vector along with the current resolved into orthogonal components [1]. The current I at an angle θ relative to the voltage can be resolved into two vectors: $I\cos(\theta)$ and $I\sin(\theta)$. The in-phase component $I\cos(\theta)$ multiplied by the voltage gives average power in watts. The current component that is 90° out of phase with the voltage, $I\sin(\theta)$, is not associated with delivered power and does not contribute to work. For want of a better name, this was often termed the *wattless component* of the current. Since this wattless current could be associated with magnetic fields, it was sometimes termed *magnetizing current* because, while doing no work, this current interacts through the inductive reactance of an ac motor winding to provide the magnetic field required for such a motor to operate.

Three types of power have been defined for systems in which both the voltage and current are sinusoidal. Throughout the years, a number of different names have been given to the three power types. The names in present usage will be emphasized.

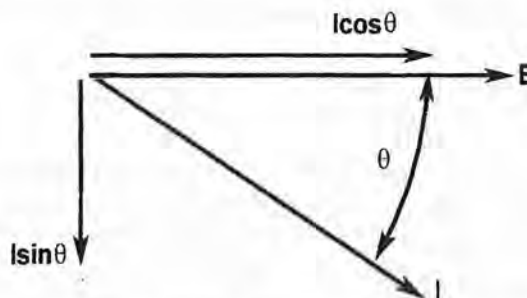


FIGURE 4-6 Phasor diagram for current and voltage. Voltage is the reference phasor. The current I has been resolved into orthogonal components $I \cos \theta$ and $I \sin \theta$. (From Reference [1].)

Active power is given the symbol P and is defined by Equation 4.5:

$$P = EI \cos(\theta) \quad (4.5)$$

Other names for active power include: (1) real power and (2) delivered power. Active power is the power that does work. Note that while all power quantities are volt-ampere products, only *active power* is expressed in watts.

Reactive power is given the symbol Q and is defined by the equation:

$$Q = EI \sin(\theta) \quad (4.6)$$

Other names for reactive power include: (1) imaginary power; (2) wattless power; (3) and magnetizing power. Reactive power is expressed in *voltamperes_{reactive}* or *vars*. If the load is predominantly inductive, current lags the voltage and the reactive power is given a positive sign. If the load is predominantly capacitive, current leads the voltage and the reactive power is given a negative sign.

Phasor power is given the symbol S and is defined by the equation:

$$S = \sqrt{P^2 + Q^2} \quad (4.7)$$

Phasor power was called apparent power for many years, and it will be seen in a later section that phasor power S , for sinusoidal voltages and currents, is identical to what is now called apparent power U . Phasor power is expressed in *voltamperes* or *VA*.

Figure 4.7 is a phasor diagram, often called a *power triangle*, which illustrates the relationships among the three types of power defined above. Reactive power is orthogonal to active power, and is shown as positive for lagging current. It is clear that the definition of phasor power, Equation 4.7, is geometrically derived from active and reactive power.

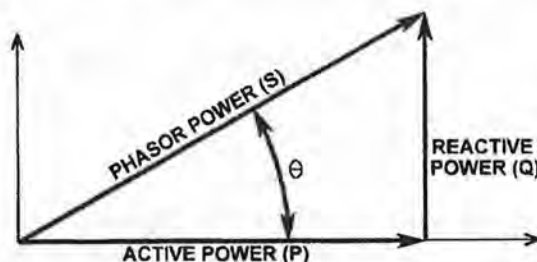


FIGURE 4.7 Power triangle showing the geometric relationships between active, reactive, and phasor power. Power factor is defined geometrically as $\cos \theta$.

Power factor is given the symbol F_p and for sinusoidal quantities is defined by the equation:

$$F_p = \frac{\text{ACTIVE POWER}}{\text{PHASOR POWER}} = \frac{P}{S} = \frac{\text{WATTS}}{\text{VOLTAMPS}} = \cos\theta \quad (4.8)$$

Since the power factor can be expressed in reference to the displacement angle between voltage and current, power factor so defined should be termed *displacement power factor*, and the symbol is often written $F_{p \text{ displacement}}$. Values for displacement power factor range from one (unity) to zero as the current displacement angle varies from 0° (current and voltage in phase) to 90° . Since the cosine function is positive in both the first and fourth quadrants, the power factor is positive for both leading and lagging currents. To completely specify the voltage-current phase relationship, the words *leading* or *lagging* must be used in conjunction with power factor. Power factor can be expressed as a decimal fraction or as percent. For example, the power factor of the case shown in Figure 4.4 is expressed either as 0.5 lagging or 50% lagging.

Polyphase Circuits

Power Calculations

The power concepts developed for single-phase circuits with sinusoidal voltages and currents can be extended to polyphase circuits. Such circuits can be considered to be divided into a group of two-wire sets, with the neutral conductor (or a resistively derived neutral for the case of a delta-connected, three-wire circuit) paired with each other conductor. Equations 4.3 to 4.5 can be rewritten to define power terms equivalent to the single-phase terms. In these equations, k represents a phase number, m is the total number of phases, and α and β are, respectively, the voltage and current phase angles with respect to a common reference frame.

$$P = \sum_{k=1}^m E_k I_k \cos(\alpha - \beta) \quad (4.9)$$

$$Q = \sum_{k=1}^m E_k I_k \sin(\alpha - \beta) \quad (4.10)$$

and, restating Equation 4.7:

$$S = \sqrt{P^2 + Q^2}$$

For example, a three-phase sinusoidal power distribution service, with phases a, b, and c:

$$P = E_a I_a \cos(\alpha_a - \beta_a) + E_b I_b \cos(\alpha_b - \beta_b) + E_c I_c \cos(\alpha_c - \beta_c)$$

$$Q = E_a I_a \sin(\alpha_a - \beta_a) + E_b I_b \sin(\alpha_b - \beta_b) + E_c I_c \sin(\alpha_c - \beta_c)$$

$$S = \sqrt{P^2 + Q^2}$$

Power Factor

Power factor is defined by Equation 4.11. Note that it is no longer always true to say that power factor is equal to the cosine of the phase angle. In many three-phase balanced systems, the phase angles of all three phases are equal and the cosine relationship holds. In unbalanced systems, such as that represented

35

Light-Emitting Diode Displays

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A light-emitting diode (LED) is a particular solid-state p - n junction diode that gives out light upon the application of a bias voltage. The luminescence process in this case is electroluminescence, which is associated with emission wavelengths in the visible and infrared regions of the spectrum. When a forward bias is applied to the p - n junction diode, carriers are injected into the depletion region in large numbers. Because of their physical proximity, the electron-hole pairs undergo a recombination that is associated with the emission of energy. Depending on the semiconductor band-gap characteristics, this emitted energy can be in the form of heat (as phonons) or light (as photons).

The solution of the Schrödinger equation for a typical crystal reveals the existence of Brillouin zones. A plot between the energy E of an electron in a solid and its wave vector k represents the allowed energy bands. It may be noted that the lattice structure affects the motion of an electron when k is close to $n\pi/l$ (where n is any integer and l is the crystal periodicity) and the effect of this constraint is to introduce an energy band gap between the allowed energy bands. Figure 35.1a shows portions of two E vs. k curves for neighboring energy bands within the regions $k = \pi/l$ and $k = -\pi/l$ (also known as the reduced zone).

While the upper band of Figure 35.1 represents the energy of conduction band electrons, the curvature of the lower band can be associated with electrons having negative effective mass. The concept of negative effective mass can readily be identified with the concept of holes in the valence band. While the majority of the electrons are identified with the minima of the upper E - k curve, the majority of the holes are identified with the maxima of the lower E - k curve. The minimum value of the conduction band and the maximum value of the valence band in Figure 35.1a both have identical k values. A semiconductor having such a characteristic is said to have a direct band gap, and the associated recombination in such a semiconductor is referred to as direct.

The *direct recombination* of an electron-hole pair always results in the emission of a photon. In a direct band-gap semiconductor, the emitted photon is not associated with any change in momentum (given by $\hbar k/2\pi$) since $\Delta k = 0$. However, for some semiconducting materials, the E vs. k curve may be somewhat different, as shown in Figure 35.1b. While the minimum conduction band energy can have a nonzero k , the maximum valence band energy can have $k = 0$. The electron-hole recombination in such a semiconductor is referred to as indirect.

An *indirect recombination* process involves a momentum adjustment. Most of the emission energy is thus expended in the form of heat (as phonons). Very little energy is left for the purpose of photon emission, which in most cases is a very slow process. Furthermore, since both photons and phonons are involved in this energy exchange, such transitions are less likely to occur. The interband recombination rate is basically given by

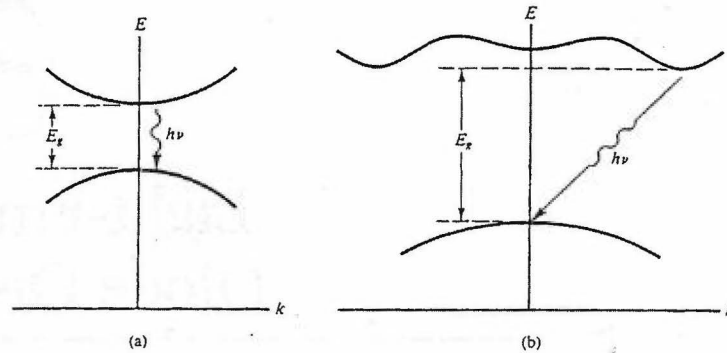


FIGURE 35.1 E vs. k for semiconductors having (a) a direct band gap and (b) an indirect band gap.

$$dn/dt = B_r np \quad (35.1)$$

where B_r is a recombination-dependent constant which for a direct band-gap semiconductor is $\sim 10^6$ times larger than that for an indirect band-gap semiconductor. For direct recombination, B_r value ranges from 0.46×10^{-10} to 7.2×10^{-10} cm³/s.

All semiconductor crystal lattices are alike, being dissimilar only in terms of their band characteristics. Si and Ge both have indirect band transitions, whereas GaAs, for example, is a semiconductor that has a direct band transition. Thus, while Si and Ge are preferred for fabrication of transistors and integrated circuits, GaAs is preferred for the fabrication of LEDs.

The direct recombination (when $k = \text{constant}$) results in a photon emission whose wavelength (in micrometers) is given by

$$\lambda = hc/E_g = 1.24/E_g (\text{eV}) \quad (35.2)$$

where E_g is the band-gap energy. The LEDs under proper forward-biased conditions can operate in the ultraviolet, visible, and infrared regions. For the visible region, however, the spectral luminous efficiency curves of Figure 35.2, which account for the fact that the visual response to any emission is a function of wavelength, should be of concern. It is unfortunate that there is not a single-element semiconductor suitable for fabrication of LEDs, but there are many binary and ternary compounds that can be used for fabrication of LEDs. Table 35.1 lists some of these binary semiconductor materials. The ternary semiconductors include GaAlAs, CdGeP₂, and ZnGeP₂ for infrared region operation, CuGaS₂ and AgInS₂ for visible region operation, and CuAlS₂ for ultraviolet region operation. Ternary semiconductors are used because their energy gaps can be tuned to a desired emission wavelength by picking appropriate composition.

Of the ternary compounds, gallium arsenide-phosphide (written as GaAs_{1-x}P_x) is an example that is basically a combination of two binary semiconductors, namely, GaAs and GaP. The corresponding bandgap energy of the semiconductor can be varied by changing the value of x . For example, when $x = 0$, $E_g = 1.43$ eV. E_g increases with increasing x until $x = 0.44$ and $E_g = 1.977$ eV, as shown in Figure 35.3. However for $x \geq 0.45$, the band gap is indirect. The most common composition of GaAs_{1-x}P_x used in LEDs has $x = 0.4$ and $E_g = 1.3$ eV. This band-gap energy corresponds to an emission of red light. Calculators and watches often use this particular composition of GaAs_{1-x}P_x.

Interestingly, the indirect band gap of GaAs_{1-x}P_x (with $1 \geq x \geq 0.45$) can be used to output light ranging from yellow through green provided the semiconductor is doped with impurities such as nitrogen. The dopants introduced in the semiconductor replace phosphorus atoms which, in turn, introduce electron trap levels very near the conduction band. For example, if $x = 0.5$, the doping of nitrogen increases the LED efficiency from 0.01 to 1%, as shown in Figure 35.4. It must be noted, however, that nitrogen doping

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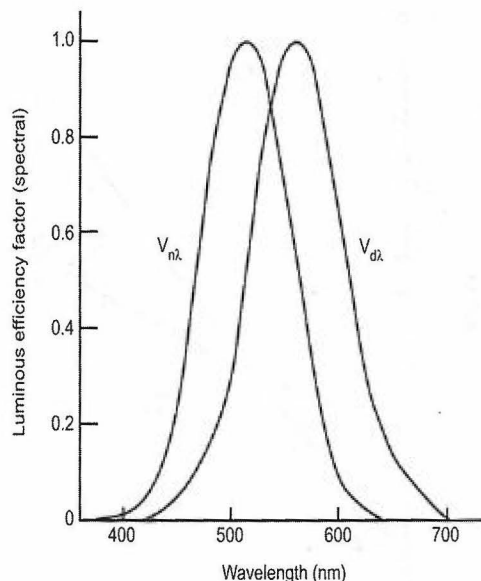


FIGURE 35.2 Spectral luminous efficiency curves. The photopic curve $V_{d\lambda}$ corresponds to the daylight-adapted case while the scotopic curve $V_{n\lambda}$ corresponds to the night-adapted case. (From Boyd, R.W., *Radiometry and the Detection of Optical Radiation*, John Wiley & Sons, New York, 1983. With permission.)

TABLE 35.1 Binary Semiconductors Suitable for LED Fabrication

	Material	E_g (eV)	Emission Type
III-V	GaN	3.5	UV
II-VI	ZnS	3.8	UV
II-VI	SnO ₂	3.5	UV
II-VI	ZnO	3.2	UV
III-VII	CuCl	3.1	UV
II-VI	BeTe	2.8	UV
III-VII	CuBr	2.9	UV — visible
II-VI	ZnSe	2.7	Visible
III-VI	In ₂ O ₃	2.7	Visible
II-VI	CdS	2.52	Visible
II-VI	ZnTe	2.3	Visible
III-V	GaAs	1.45	IR
II-VI	CdSe	1.75	IR — Visible
II-VI	CdTe	1.5	IR
III-VI	GaSe	2.1	Visible

shifts the peak emission wavelength toward the red. The shift is comparatively larger at and around $x = 0.05$ than $x = 1.0$. The energy emission in nitrogen-doped GaAs_{1-x}P_x devices is a function of both x and the nitrogen concentration.

Nitrogen is a different type of impurity from those commonly encountered in extrinsic semiconductors. Nitrogen, like arsenic and phosphorus, has five valence electrons, but it introduces no net charge carriers in the lattice. It provides active radiative recombination centers in the indirect band-gap materials. For an electron, a recombination center is an empty state in the band gap into which an electron falls and, thereafter, falls into the valence band by recombining with a hole. For example, while a GaP LED emits green light (2.23 eV), a nitrogen-doped GaP LED emits yellowish green light (2.19 eV), and a heavily nitrogen-doped GaP LED emits yellow light (2.1 eV).

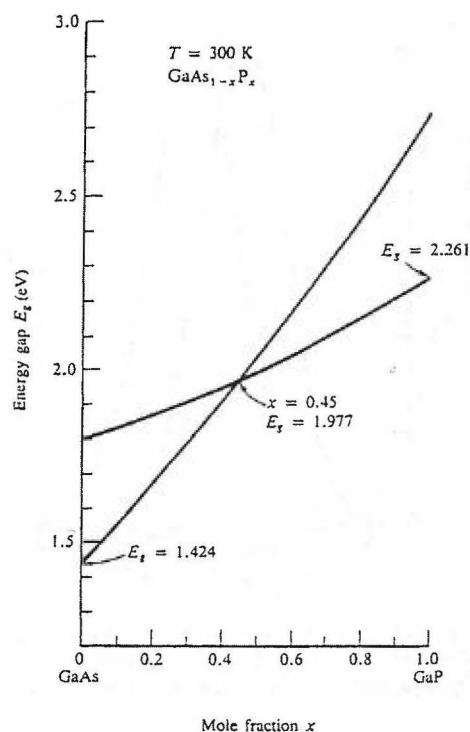


FIGURE 35.3 Band-gap energy vs. x in $\text{GaAs}_{1-x}\text{P}_x$. (From Casey, H.J., Jr. and Parish, M.B., Eds., *Heterostructure Lasers*, Academic Press, New York, 1978. With permission.)

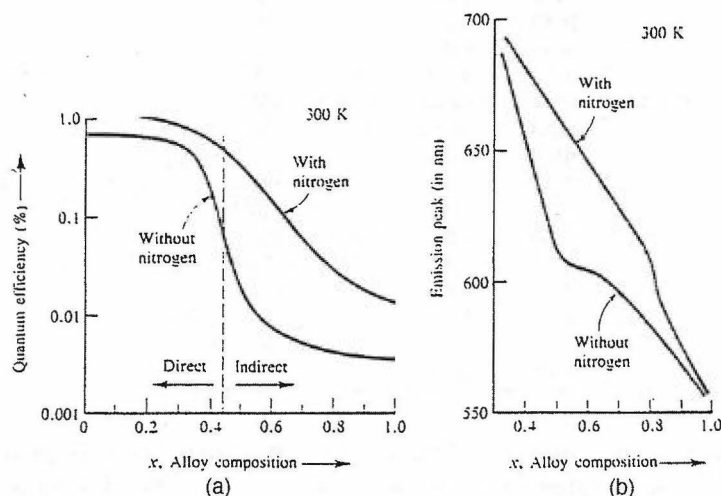


FIGURE 35.4 The effects of nitrogen doping in $\text{GaAs}_{1-x}\text{P}_x$: (a) quantum efficiency vs. x and (b) peak emission wavelength vs. x .

The injected excess carriers in a semiconductor may recombine either radiatively or nonradiatively. Whereas nonradiative recombination generates phonons, radiative recombination produces photons. Consequently, the internal quantum efficiency η , defined as the ratio of the radiative recombination rate R_r to the total recombination rate, is given by

$$\eta = R_r / (R_r + R_{nr}) \quad (35.3)$$

where R_{nr} is the nonradiative recombination rate. However, the injected excess carrier densities return to their value exponentially as

$$\Delta p = \Delta n = \Delta n_0 e^{-t/\tau} \quad (35.4)$$

where τ is the carrier lifetime and Δn_0 is the excess electron density at equilibrium. Since $\Delta n/R_r$ and $\Delta n/R_{nr}$ are, respectively, equivalent to the radiative recombination lifetime τ_r and the nonradiative recombination lifetime τ_{nr} , we can obtain the effective minority carrier bulk recombination time τ as

$$(1/\tau) = (1/\tau_r) + (1/\tau_{nr}) \quad (35.5)$$

such that $\eta = \tau/\tau_r$. The reason that a fast recombination time is crucial is that the longer the carrier remains in an excited state, the larger the probability that it will give out energy nonradiatively. In order for the internal quantum efficiency to be high, the radiative lifetime τ_r needs to be small. For indirect band-gap semiconductors, $\tau_r \gg \tau_{nr}$ so that very little light is generated, and for direct band-gap semiconductors, τ_r increases with temperature so that the internal quantum efficiency deteriorates with the temperature.

As long as the LEDs are used as display devices, it is not too important to have fast response characteristics. However, LEDs are also used for the purpose of optical communications, and for those applications it is appropriate to study their time response characteristics. For example, an LED can be used in conjunction with a photodetector for transmitting optical information between two points. The LED light output can be modulated to convey optical information by varying the diode current. Most often, the transmission of optical signals is facilitated by introducing an optical fiber between the LED and the photodetector.

There can be two different types of capacitances in diodes that can influence the behavior of the minority carriers. One of these is the *junction capacitance*, which is caused by the variation of majority charge in the depletion layer. While it is inversely proportional to the square root of bias voltage in the case of an abrupt junction, it is inversely proportional to the cube root of bias voltage in the case of a linearly graded junction. The second type of capacitance, known as the *diffusion capacitance*, is caused by the minority carriers.

Consider an LED that is forward biased with a dc voltage. Consider further that the bias is perturbed by a small sinusoidal signal. When the bias is withdrawn or reduced, charge begins to diffuse from the junction as a result of recombination until an equilibrium condition is achieved. Consequently, as a response to the signal voltage, the minority carrier distribution contributes to a signal current.

Consider a one-dimensional *p*-type semiconducting material of cross-sectional area A whose excess minority carrier density is given by

$$\delta \Delta n_p / \delta t = D_n \delta^2 \Delta n_p / \delta x^2 - \Delta n_p / \tau \quad (35.6)$$

As a direct consequence of the applied sinusoidal signal, the excess electron distribution fluctuates about its dc value. In fact, we may assume excess minority carrier density to have a time-varying component as described by

$$\Delta n_p(x, t) = \langle \Delta n_p(x) \rangle + n'_p(x) e^{j\omega t} \quad (35.7)$$

where $\langle \Delta n_p(x) \rangle$ is a time-invariant quantity. By introducing Equation 35.7 into Equation 35.6, we get two separate differential equations:

$$\delta^2 / \delta x^2 \langle \Delta n_p(x) \rangle = \langle \Delta n_p(x) \rangle / (L_n)^2 \quad (35.8a)$$

and

$$\delta^2/\delta x^2 [\Delta n'_p(x)] = \Delta n'_p(x) / [L_n^*]^2 \quad (35.8b)$$

where

$$L_n^* = L_n / (1 + j\omega\tau)^{1/2} \quad (35.9a)$$

and

$$L_n = (D_n \tau)^{1/2} \quad (35.9b)$$

The dc solution of Equation 35.8a is well known. Again, the form of Equation 35.8b is similar to that of Equation 35.8a and, therefore, its solution is given by

$$\Delta n'_p(x) = \Delta n'_p(0) e^{-x/L} \quad (35.10)$$

Since the frequency-dependent current $I(\omega)$ is simply a product of eAD_n and the concentration gradient, we find that

$$\begin{aligned} I(\omega) &= |eAD_n \Delta n'_p(x) / dx|_{x=0} \\ &= I(0) / (1 + \omega^2 \tau^2)^{1/2} \end{aligned} \quad (35.11)$$

where $I(0)$ is the intensity emitted at zero modulation frequency. We can determine the admittance next by dividing the current by the perturbing voltage. The real part of the admittance, in this case, will be equivalent to the diode conductance, whereas its imaginary part will correspond to the diffusion capacitive susceptance.

The modulation response as given by Equation 35.11 is, however, limited by the carrier recombination time. Often an LED is characterized by its modulation bandwidth, which is defined as the frequency band over which signal power (proportional to $I^2(\omega)$) is half of that at $\omega = 0$. Using Equation 35.11, the 3-dB modulation bandwidth is given by

$$\Delta\omega \approx 1/\tau_r \quad (35.12)$$

where the bulk lifetime has been approximated by the radiative lifetime. Sometimes the 3-dB bandwidth of the LED is given by $I(\omega) = \frac{1}{2}I(0)$, but this simplification contributes to an erroneous increase in the bandwidth by a factor of 1.732.

Under conditions of thermal equilibrium, the recombination rate is proportional to the product of initial carrier concentrations, n_0 and p_0 . Then, under nonequilibrium conditions, additional carriers $\Delta n = \Delta p$ are injected into the material. Consequently, the recombination rate of injected excess carrier densities is given by initial carrier concentrations and injected carrier densities as

$$\begin{aligned} R_{\Delta r} &= [B_r(n_0 + \Delta n)(p_0 + \Delta p) - B_r n_0 p_0] \\ &= B_r(n_0 + p_0 + \Delta n)\Delta n \end{aligned} \quad (35.13)$$

where B_r is the same constant introduced in Equation 35.1. For p -type GaAs, for example, $B_r = 1.7 \times 10^{-10}$ cm³/s when $p_0 = 2.4 \times 10^{18}$ holes/cm³. Equation 35.13 is used to define the radiative carrier recombination lifetime by

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$$\tau_r = \Delta n / R_{\Delta r} = [B_r(n_o + p_o + \Delta n)]^{-1} \quad (35.14)$$

In the steady-state condition, the excess carrier density can be calculated in terms of the active region width d by

$$\Delta n = J\tau_r / ed \quad (35.15)$$

where J is the injection current density.

The radiative recombination lifetime is found by solving Equation 35.14 after having eliminated Δn from it using Equation 35.15:

$$\tau_r = \left[\left\{ (n_o + p_o)^2 + (4J/B_r ed) \right\}^{1/2} - (n_o + p_o) \right] / (2J/ed) \quad (35.16)$$

Thus, while for the low carrier injection (i.e., $n_o + p_o \gg \Delta n$), Equation 35.16 reduces to

$$\tau_r \approx [B_r(n_o + p_o)]^{1/2} \quad (35.17a)$$

for the high carrier injection (i.e., $n_o + p_o \ll \Delta n$), it reduces to

$$\tau_r \approx (ed/JB_r)^{1/2} \quad (35.17b)$$

Equation 35.17a indicates that in highly doped semiconductors, τ_r is small. But the doping process has its own problem, since in many of the binary LED compounds higher doping may introduce nonradiative traps just below the conduction band, thus nullifying Equation 35.12. In comparison to Equation 35.17a, Equation 35.17b provides a better alternative whereby τ_r can be reduced by decreasing the active region width or by increasing the current density. For the case of p -type GaAs, the radiative lifetimes vary between 2.6 and 0.35 ns, respectively, when p_o varies between 1.0×10^{18} holes/cm³ and 1.5×10^{19} holes/cm³.

Usually, LEDs are operated at low current (≥ 10 mA) and low voltages (≥ 1.5 V), and they can be switched on and off in the order of 10 ns. In addition, because of their small sizes, they can be reasonably treated as point sources. It is, therefore, not surprising that they are highly preferred over other light sources for applications in fiber-optic data links.

Two particular LED designs are popular: *surface emitters* and *edge emitters*. They are shown in Figure 35.5. In the former, the direction of major emission is normal to the plane of the active region, whereas in the latter the direction of major emission is in the plane of the active region. The emission pattern of the surface emitters is very much isotropic, whereas that of the edge emitters is highly directional.

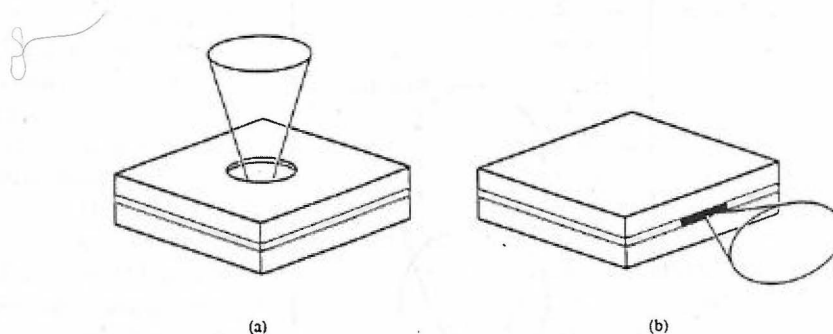


FIGURE 35.5 LED type: (a) surface emitter and (b) edge emitter.

As the LED light originating from a medium of refractive index n_1 goes to another medium of refractive index n_2 ($n_2 < n_1$), only a portion of incident light is transmitted. In particular, the portion of the emitted light corresponds to only that which originates from within a cone of semiapex angle θ_c , such that

$$\theta_c = \sin^{-1}(n_2/n_1) \quad (35.18)$$

In the case of an LED, n_1 corresponds to the refractive index of the LED medium and n_2 corresponds to that of air (or vacuum). Light originating from *beyond* angle θ_c undergoes a total internal reflection. However, the light directed from *within* the cone of the semiapex angle θ_c will be subjected to Fresnel's loss. Thus, the overall transmittance T is given by

$$T = 1 - \left\{ (n_1 - n_2) / (n_1 + n_2) \right\}^2 \quad (35.19)$$

Accordingly, the total electrical-to-optical conversion efficiency in LEDs is given by

$$\begin{aligned} \eta_{LED} &= T \left[(\text{solid angle within the cone}) / (4\pi) \right] \\ &= (T/2)(1 - \cos \theta_c) \\ &= (T/4) \sin^2 \theta_c \\ &= (1/4)(n_2/n_1)^2 \left[1 - \left\{ (n_1 - n_2) / (n_1 + n_2) \right\}^2 \right] \end{aligned} \quad (35.20)$$

Only two schemes increase the electrical-to-optical conversion efficiency in an LED. The first technique involves guaranteeing that most of the incident rays strike the glass-to-air interface at angles less than θ_c . It is accomplished by making the semiconductor-air interface hemispherical. The second method involves schemes whereby the LED is encapsulated in an almost transparent medium of high refractive index. The latter means is comparatively less expensive. If a glass of refractive index 1.5 is used for encapsulation, the LED efficiency can be increased by a factor of 3. Two of the possible encapsulation arrangements and the corresponding radiation patterns are illustrated in Figure 35.6.

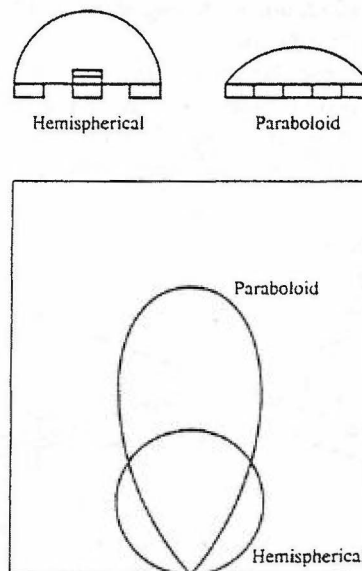


FIGURE 35.6 LED encapsulation geometries and their radiation patterns.

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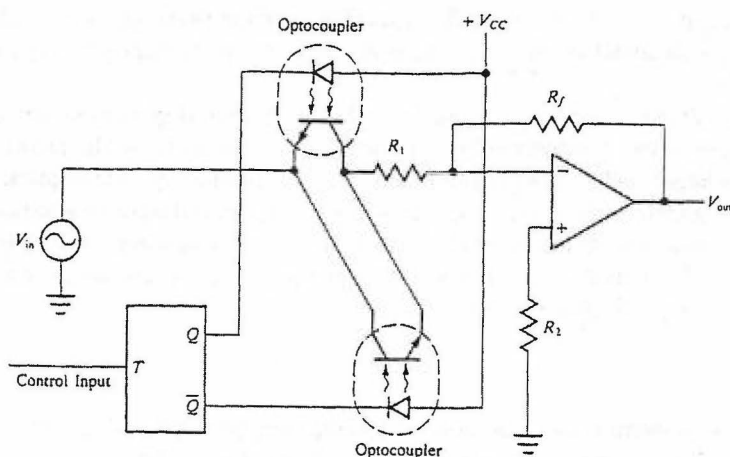
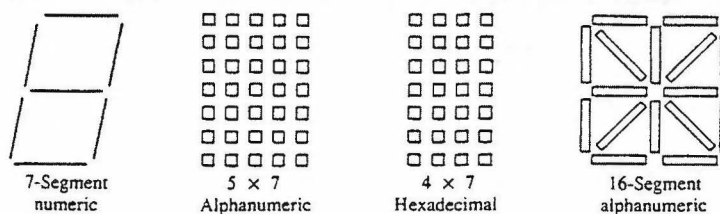


FIGURE 35.7 A chopping circuit with an amplifier.



(a)



(b)

FIGURE 35.8 (a) LED display formats; and (b) displayed alphanumeric characters using 16-segment displays.

LEDs are often used in conjunction with a phototransistor to function as an optocoupler. The optocouplers are used in circumstances when it is desirable to have a transmission of signals between electrically isolated circuits. They are used to achieve noise separation by eliminating the necessity of having a common ground between the two systems. Depending on the type of coupling material, these miniature devices can provide both noise isolation as well as high voltage isolation. Figure 35.7 shows a typical case where two optocouplers are used to attain a chopper circuit. The two optocouplers chop either the positive or the negative portion of the input signals with a frequency of one half that of the control signal that is introduced at the T flip-flop. The operational amplifier provides an amplified version of the chopped output waveform. In comparison, a chopper circuit that uses simple bipolar transistors produces noise spikes in the output because of its inherent capacitive coupling.

The visible LEDs are best known for their uses in displays and indicator lamps. In applications where more than a single source of light is required, an LED array can be utilized. An LED array is a device consisting of a row of discrete LEDs connected together within or without a common reflector cavity. Figure 35.8a shows different LED arrangements for displaying hexadecimal numeric and alphanumeric characters, whereas Figure 35.8b shows, for example, the possible alphanumeric characters using 16-segment

displays. In digital systems, the binary codes equivalent to these characters are usually decoded and, consequently, a specific combination of LED segments are turned on to display the desired alphanumeric character.

The dot matrix display provides the most desirable display font. It gives more flexibility in shaping characters and has a lower probability of being misinterpreted in case of a display failure. However, these displays involve a large number of wires and increased circuit complexity. LED displays, in general, have an excellent viewing angle, high resonance speed (≥ 10 ns), long life, and superior interface capability with electronics with almost no duty cycle limitation. LEDs with blue emission are not available commercially. When compared with passive displays, LED displays consume more power and involve complicated wiring with at least one wire per display element.

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