12. Pressure, Flow and Level measurement

bellows
diaphragm
LTD (linear differential transformer)

Pressure
- vortex shedding
- turbine
- magnetic
- differential

Flow
- bellows
- floats
- differential pressure

13. Encoding

Linear and rotary

resistive potentiometers
- linear and rotary

resolvers
optical
- quadrature + zero
- multiple track ie. number of digital bits
- Gray Scale
- Gray to Binary code converter

tachometers
- DC
- AC
- optical

Sydney Institute of Technology

School of Electrotechnology

Industrial Sensors

Theory Notes

Area ______________________

Topic ______________________

Session No.

For further information on this module, or this subject contact Jim Hafford, (02) 217 3620, Bld. K. Ultimo, S.I.T.

SUMMARY

- Electrical input transducers and sensors are classified according to the physical variables to which they are sensitive. The four main categories are thermal, optical, magnetic, and electromechanical. The following tables give a brief summary of the operating characteristics of the transducers and sensors discussed in this chapter.

Thermal Transducers

	Operation	*Advantages*	*Disadvantages*
Thermistor	Device resistance changes inversely with variations in temperature.	Large nominal R Large R variation Inexpensive Fast	Nonlinear Narrow temperature range Requires power Self-heats
RTD	Device resistance changes directly with variations in temperature.	Linear R variation Most stable Most accurate	Expensive Small R variation Requires power
Thermocouple	Device produces voltage or current proportional to temperature.	Nearly linear output Wide temperature range Used at high temperatures Self-powered	Low output Least sensitive Requires reference

Temperature Sensors

	Operation	*Advantages*	*Disadvantages*
IC Temperature Sensor	Device produces voltage or current numerically equivalent to absolute temperature.	Linear output Small packages Inexpensive Fast	Low temperature range Low output
Temperature Transmitter	Device accepts thermocouple or RTD input. Output current is proportional to temperature.	Linear output Facilitates remote sensing	Accepts only thermocouples or RTDs as inputs
Optical Pyrometer	Focuses infrared energy on IC temperature sensor in probe. Output current is proportional to temperature.	Linear output Used at high temperatures Noncontact	Expensive

Optical Transducers

	Operation	Advantages	Disadvantages
Photoconductive Cell	Device resistance changes inversely with variations in EMR intensity.	Large *R* variation	Slow response Requires power Temperature sensitive
Photovoltaic (Solar) Cell	Converts EMR into electrical current. Current is proportional to EMR intensity.	Self-powered Linear output current	Large surface area Inefficient Slow response Nonlinear output voltage
Photodiode	Converts EMR into electrical current. Current is proportional to EMR intensity.	Small Fast	Low output Requires external power supply
Phototransistor	Converts EMR into current, which is injected into the base. Device then responds as typical transistor.	Sensitive High output Fast	Requires external power supply

Optical Sensing Methods

	Operation	Advantages	Disadvantages
Opposed Sensing	EMR beam is directed from transmitter to receiver. Object is detected when beam is interrupted.	Long sensing range	Transmitter and receiver must be precisely aligned. Separate receiver and transmitter required
Retroreflective Sensing	EMR beam is directed from transmitter to a reflector and back to receiver. Object is detected when beam is interrupted.	Transmitter and receiver may be housed in the same unit.	Sensing range limited to about 30 ft Requires remote reflector
Diffuse Sensing	EMR beam is directed at sensed object, which acts as reflector. Object is sensed when beam is reflected back to receiver.	Transmitter and receiver may be housed in the same unit. No reflector required	Sensing range limited to less than 10 ft Sensed object must be reflective.

Electromechanical Transducers

	Operation	Advantages	Disadvantages
Limit Switch	Electrical contact is maintained or broken according to position of actuator.	Usually requires no external power supply Available in a variety of styles Durable	Must contact the sensed object
Potentiometer	Resistance (or voltage) between wiper and either terminal changes as shaft is turned.		Requires external power supply Requires some sort of gearing or linkage
Wire Strain Gauge	Device resistance changes as gauge is compressed or elongated.	Durable	Requires external power supply Low gauge factor
Semiconductor Strain Gauge	Device resistance changes as gauge is compressed or elongated.	High gauge factor	Requires external power supply

Magnetic Transducers

	Operation	Advantages	Disadvantages
Inductive	Relative motion of magnetic field and coil induces voltage in coil.	Requires no power supply	Detects motion only
Reluctive	Proximity of a low-reluctance object alters magnetic field surrounding device and induces voltage in coil.	Requires no power supply	Detects motion only
Hall-Effect	Presence of a magnetic field produces voltage proportional to strength of field.	Senses stationary fields Inexpensive	Requires external power supply

Flow Sensors (*continued*)

	Operation	Advantages	Disadvantages
Ultrasonic	Acoustic energy is transmitted upstream into fluid, at a given frequency. Frequency of reflected energy is proportional to fluid velocity.	No moving parts Flow may be sensed through pipe walls. Does not impede fluid flow	Most expensive
Magnetic	Voltage is induced in fluid as it passes through magnetic field. Electrodes sense voltage, which is proportional to flow.	No moving parts Relatively inexpensive Does not impede fluid flow	Fluid must be electrically conductive.

Level Sensors

	Operation	Advantages	Disadvantages
Float	When rising liquid level reaches float, float rises and actuates either a magnetic or a mechanical switch.	Inexpensive	Senses liquids only
Conductive Probe	When rising liquid level contacts the probes, current flows and is available at output.	No moving parts Inexpensive	Senses liquids only Liquid must be electrically conductive.
Ultrasonic Contact	Acoustic energy is directed across a gap toward receiver. Liquid within gap completes transmission path.	No moving parts Senses any liquid	Expensive
Ultrasonic Noncontact	Acoustic energy is transmitted toward surface to be sensed. Reflection time determines surface distance.	No moving parts Surface contact not required Measures absolute height	Most expensive

Level Sensors (*continued*)

	Operation	Advantages	Disadvantages
Proximity	Frequency change in oscillator indicates object proximity. Output is activated when oscillator changes frequency.	No moving parts	Sensing range limited to about 10 cm

Flow Sensors

	Operation	Advantages	Disadvantages
Turbine	Fluid flowing past turbine causes blades to spin. Rotation is detected by magnetic sensor.	Relatively inexpensive	Impedes fluid flow
Vortex	As fluid flows past a vortex shedder, vortices are produced downstream. Resultant vibrations are sensed piezoelectrically.	Minimal moving parts Accurate	Expensive Impedes fluid flow
Thermal	Probe is heated to constant temperature. Thermal transducer measures rate of cooling, which is proportional to flow rate.	No moving parts Relatively inexpensive	Slow response Impedes fluid flow Fluid must be thermally conductive.

8.7 Glossary of Terms

Alignment
Transmitter and receiver of thru-beam units should be properly aligned along a common axis, otherwise sensitivity is reduced.
Maximum intensity of the radiated energy is obtained when the axis of the radiated light hits the view field of the receiver.

Amplifier
A device or circuit which amplifies signals. The control unit which supplies DC power and solid-state output or relay-drive output.

Amplifier Built-in Sensor
Amplifier is incorporated in the sensor unit, which brings the unit to supply relay-drive or solid-state output.

Amplifier-separated Sensor
A sensor without an amplifer in it but only a light emitting and receiving element. For sensing operation, the sensor has to be combined to either amplifer or control unit.

Contact Output
Output is obtained in the form of relay contact output.

Cross Beam
The optical lens are aligned in such a manner that the emitted light beam axis and the reflected light axis cross each other. This enables a sensing area to be limited in some field at a certain distance from the tip of a sensor. The advantage of this is that the sensor cannot be influenced by the reflection from the background of an object.

Dark Mode
The output of the sensor is energised when no light falls on the receiver.

Effective Beam Diameter
The effective beam diameter is the 'core' of the beam in which the receiver, the reflector or the target may be placed to operate.

External Synchronisation
A sensing signal is controlled in timing with other signals coming from outside.

Extraneous Light
Ambient light which affects the operation of a beam sensor.

Hysterisis
Distance between turn-on and turn-off points of a sensor when an object is detected.

Infa-red (I.R.)
An invisible light beyond the range of the visible spectrum in the red region.

Input or Supply Voltage
Operating voltage required to sufficiently maintain power to a device.

Mark Sensor
Sensor designed to detect marks and/or colour tones on the surface of objects.

Modulated Light
The infa-red LED is pulsed rapidly at a specific frequency so that the photodetector turned to the same pulse frequency, detects only the desired light source, reducing interference from ambient light.

Non-contact Voltage Output
Output voltage generated in either NPN or PNP solid-state output circuit.

Operating Mode / Switch Mode
State of sensor which energises the output. Light 'On' or Dark-'On' is selected with an operation mode selector.

Operating Range
Maximum distance in which the target can be detected from the photoelectric sensor. (Example: Maximum distance between emitter-receiver or sensor and target, or sensor and reflector.)

Optic-Fibre
Glass or plastic fibre enables to transmit and lead a light beam to form Thru-beam or Proximity scanning.

Output
A switching device which opens or closes a circuit enabling or disabling current flow.

Output Mode
Indicates in what forms an output is obtained: relay output or solid-state output.

Output Rating
Maximum contact capacity of a relay or limit value of voltage/ current of a solid-state output circuit.

Repeatability / Repeat Accuracy
The accuracy of a sensor to detect an object repeatedly.

Response Time
The time required between the presense of the target (presence or absense of received beam) and the corresponding output change.

Self-Contained Beam Sensor
A beam sensor which has power supply, signal conditioning and output in one enclosure. The opposite of this type of beam sensor is component systems which have sources and detectors remote from power.

Slot Sensor
A U-frame scanner with opposing emitter and receiver in the same unit. May be used for small object detection revolution counting, notch/space detection on toothed wheels, thread break detection and edge guide control.

Timing
A method of enhancing or varying a signal. May be used to delay an output from energising, or to keep an output energised longer.

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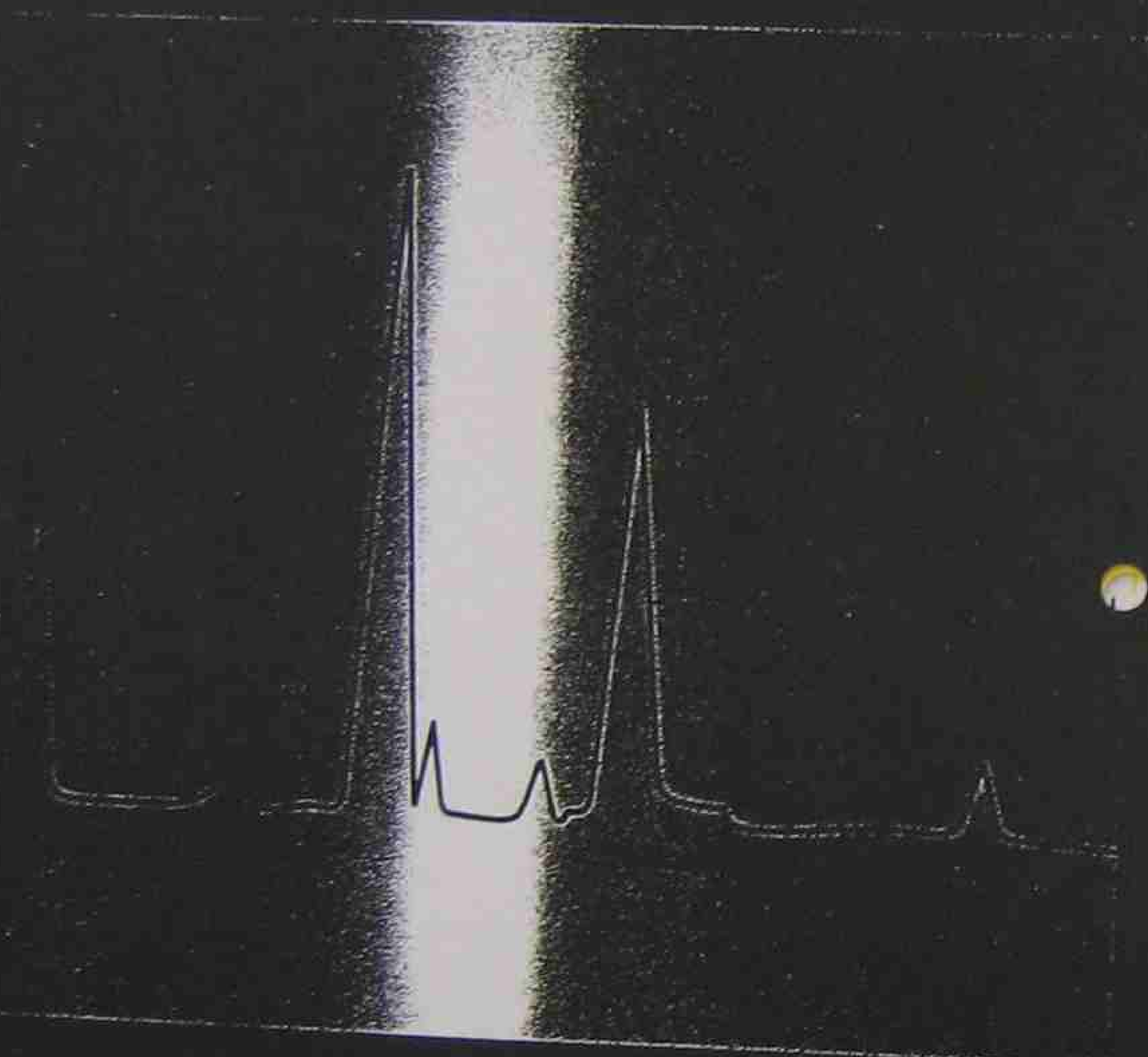
LIGHT



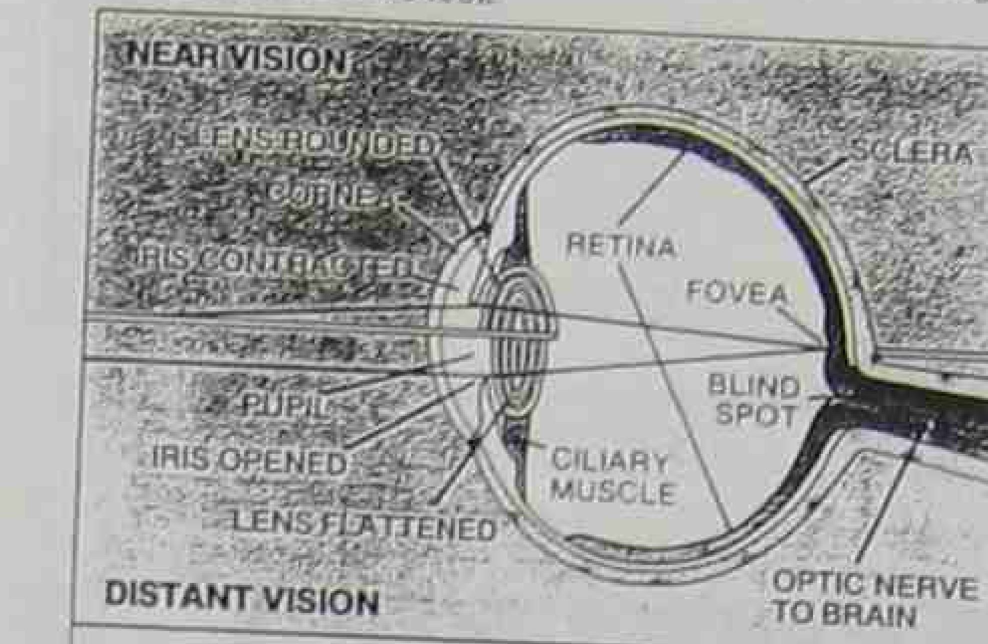
Sight • Science • Sources

PHILIPS

The Eye and how we see

Without the eye, light would have no meaning. Sight gives us the most detailed information about our surroundings. We depend on the eye forming images on nerve cells which are stimulated into sending signals along the optic nerve to the visual centres of the brain.

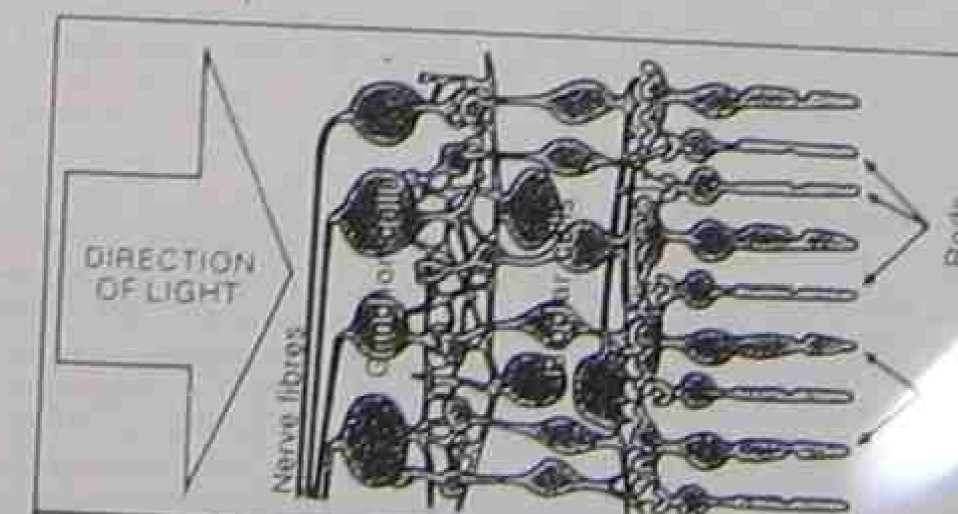
Our eyes are roughly spherical and 25 mm in diameter. They can swivel in any direction and six positioning muscles ensure that the visual axes of both eyes converge no matter where we look.



Sectional diagram of human eye

The tough outer layer of the eye, the sclera or "white", has a transparent area, called the cornea, through which light enters. The amount of light entering is controlled by the iris, a coloured diaphragm (the "colour" of your eyes). The hole in the diaphragm (called the pupil) can vary in size from 2 to 8 mm in diameter. As the brightness of the field of view increases, the pupil size decreases. When we are in the dark, our pupils enlarge to their maximum size.

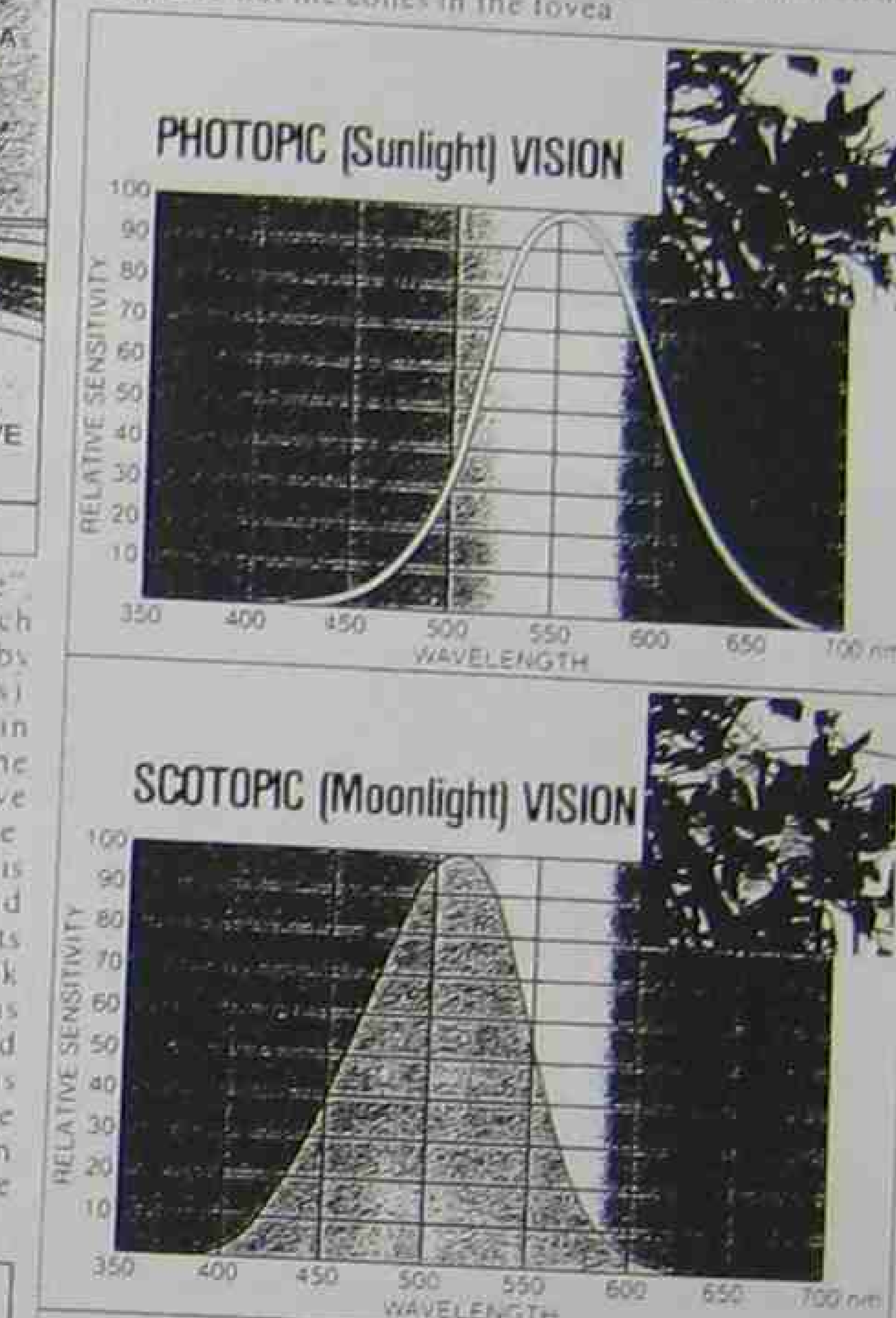
The lens, which focuses images onto the retina, is behind the iris. It is made like an onion in many layers, and is capable of changing its shape. Radial ligaments normally keep the lens fairly flat. However when we look at close objects, the ligaments contract to make the lens more rounded. This process of changing focus is called accommodation. The retina is the start of the nervous system leading to the brain and covers almost ⅘ of the inner surface of the eye. Each retina has about 100 million light sensitive nerve endings called rods and cones because of their shape.



Cross section through retina

There are nearly ten times as many rods as cones. Rods are spread fairly evenly over the retina; cones are concentrated on the visual axis in an area called the fovea. Rods are very sensitive to light but cannot distinguish colour. Cones are less [illegible] but can distinguish colours. Because they [illegible] closely packed and have individual nerve connec[illegible] cones enable us to see very fine detail.

We therefore have two kinds of vision – one depending on cones, the other on rods. As the light level changes so our eyes are constantly adapting to suit. The full range of sensitivity of the eye is greater than ten billion to one. In bright light, the cones provide photopic vision with high definition and colour discrimination using the area of the retina in the fovea. In moonlight however, the rods give us scotopic vision which is in shades of grey with less resolution of detail. So if you want to see very weak lights, don't look straight at them. The image will then fall on the rods and not the cones in the fovea.



Relative spectral sensitivity of human eye

The change from scotopic to photopic vision is gradual and takes place between one hundredth and ten candela/m^2. Vision in this region relies on both rods and cones and is called mesopic vision.

Light is turned into nerve signals by a pigment called rhodopsin found in the rods and three similar pigments in the cones. It is believed that there are three separate sets of cones responding to the primary colours – red, green and blue. Light causes the pigment molecules to split, triggering a nerve impulse which passes along the optic nerve and out of the eye through the blind spot and on to the brain. The chemical change is reversed when the molecules are not stimulated.

The actual mechanism of vision is made more complicated because our eyes are never still. Even when we stare at a fixed object there is an imperceptible movement so that the images on the retina are never static. These movements lend weight to the argument that we see more by *changes* in the illumination of the rods and cones than by the illumination itself.

Light - The Basic Principles

Scientists were puzzled by the true nature of light and it wasn't until 1864 that the Scottish mathematician James Clerk Maxwell put forward the electro-magnetic theory of light.

Light is one form of electro-magnetic radiation within the electro-magnetic spectrum, and occupies only a small part of this spectrum. In fact our eyes have evolved simply to detect the strongest parts of the sun's radiation. (See illustrations on pages 4–5.)

The whole electro-magnetic spectrum ranges from very short wavelengths, such as X-rays, through to very long wavelength radio waves. The visible part of the spectrum covers wavelengths from 380 to 760 nm (1nm, or nanometre = 10^{-9}m) and our eyes discriminate the various wavelengths in terms of colours. The blue end of our spectrum has short wavelengths while the red has long. On either side of the visible spectrum are the ultra-violet and infra-red radiations which we cannot see, but which we detect by other effects. UV for example is responsible for sunburn, and IR is detected as heat.

All electro-magnetic radiation travels in straight lines at a speed of 300,000 km/s in vacuum. Any wave has a velocity which is a product of its wavelength and frequency ($v = f\lambda$). When passing through materials other than vacuum the velocity is reduced by a factor known as refractive index. At the boundary of two different media the incident lightwaves are either reflected, transmitted or absorbed. If the light strikes the surface at right angles, the transmitted light continues in a straight line. However, if it strikes the surface at an angle it will be bent, or refracted. (See Fig. 1.)

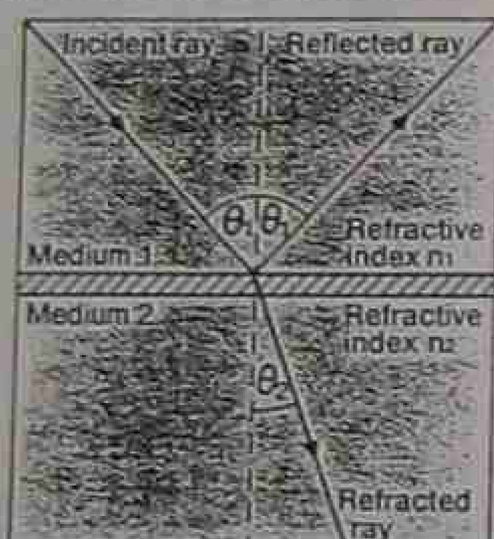


Fig 1 — Refraction/Reflection

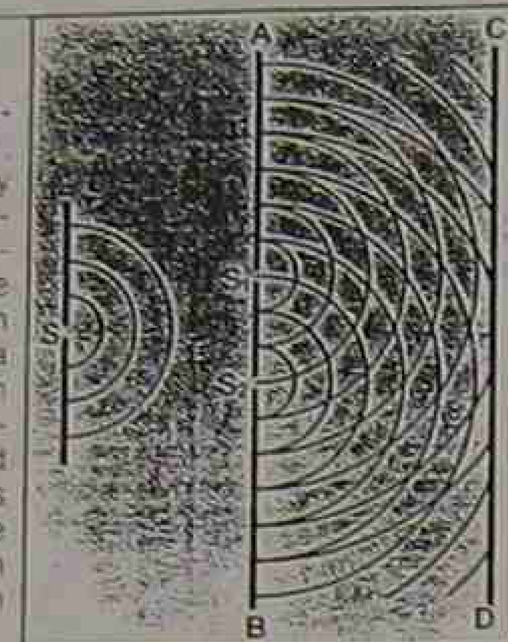
For rays passing from a high to a low refractive index medium, e.g. glass to air, a refracted ray only exists if the angle of incidence is below a critical value otherwise a condition known as total internal reflection takes place.

Fibre optics take advantage of this phenomenon enabling light to be channelled along glass rods and fibres. The technique enables surgeons to examine the insides of our stomachs and lungs and is also being developed as a way of sending many telephone messages on light beams along fibres instead of using wires and cables.

The wave nature of light gives rise to an effect called interference. If two sources having the same wavelength are combined, an interference pattern of light and dark is formed. The pattern is explained by the two light waves being in step in the bright areas and out of step (or opposite) in the dark areas. Interference is used in dichroic filters which only transmit parts of the spectrum. Interference is caused by building up many coated layers whose thickness is a function of the wavelength to be filtered. The rest of the spectrum passes with very little loss. The dichroic filter is, therefore, much more efficient than conventional pigmented filters.

Another phenomenon resulting from the wave nature of light, is known as diffraction — the bending of light rays around the edges of obstacles. A combination of diffraction and interference is used in the diffraction grating commonly used for examining the spectra of light sources. (See Fig. 2.)

Fig. 2 — **Young's Experiment.** In 1801, Young gave a very convincing demonstration of the diffraction of light when he allowed sunlight from a pinhole, S to fall on a screen, AB, in which there were two adjacent pinholes, S and S. The light waves spread out from the latter to produce an interference pattern on the screen CD.



Later experimenters replaced the three pinholes by narrow and parallel slits and used monochromatic light instead of sunlight.

LIGHT POLARISATION

When light waves vibrate in one plane only, such light is said to be polarised. (See Fig. 3.) The vibrations which cause the wave motion in a ray of light are at right angles to the direction the light is travelling, and in a beam of ordinary light these vibrations occur in all possible directions in that plane. By passing light through a material with a crystalline structure, such that it transmits waves only vibrating in a certain direction, it is possible to produce polarised light, all of whose vibrations are parallel. High quality sunglasses use this principle to reduce glare.

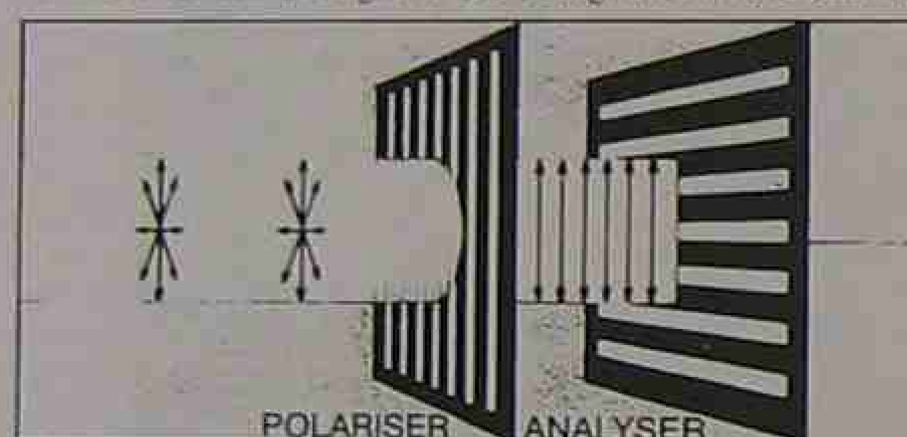


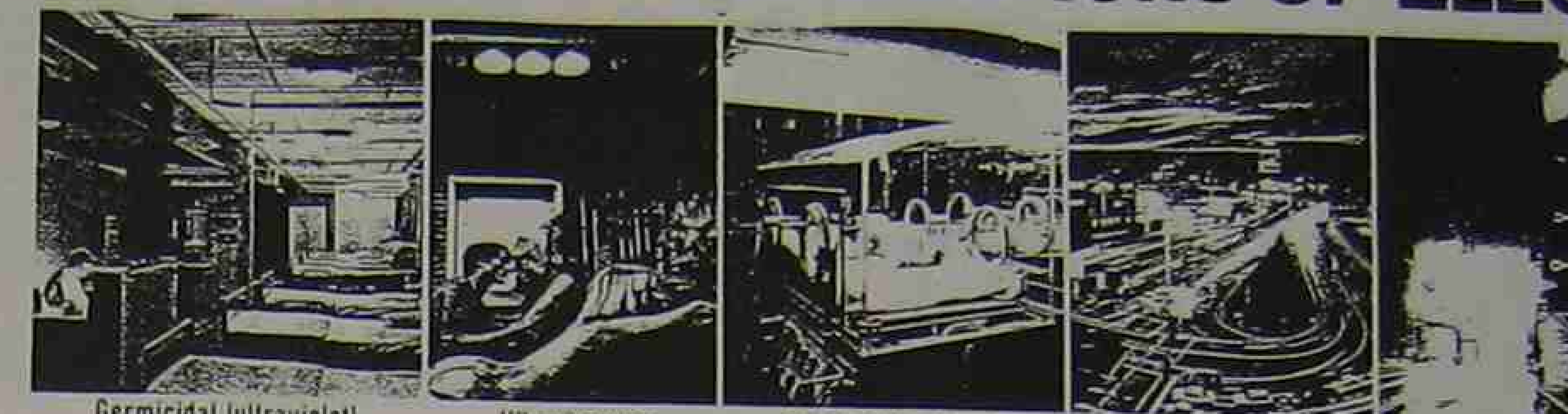
Fig 3 — Light polarisation

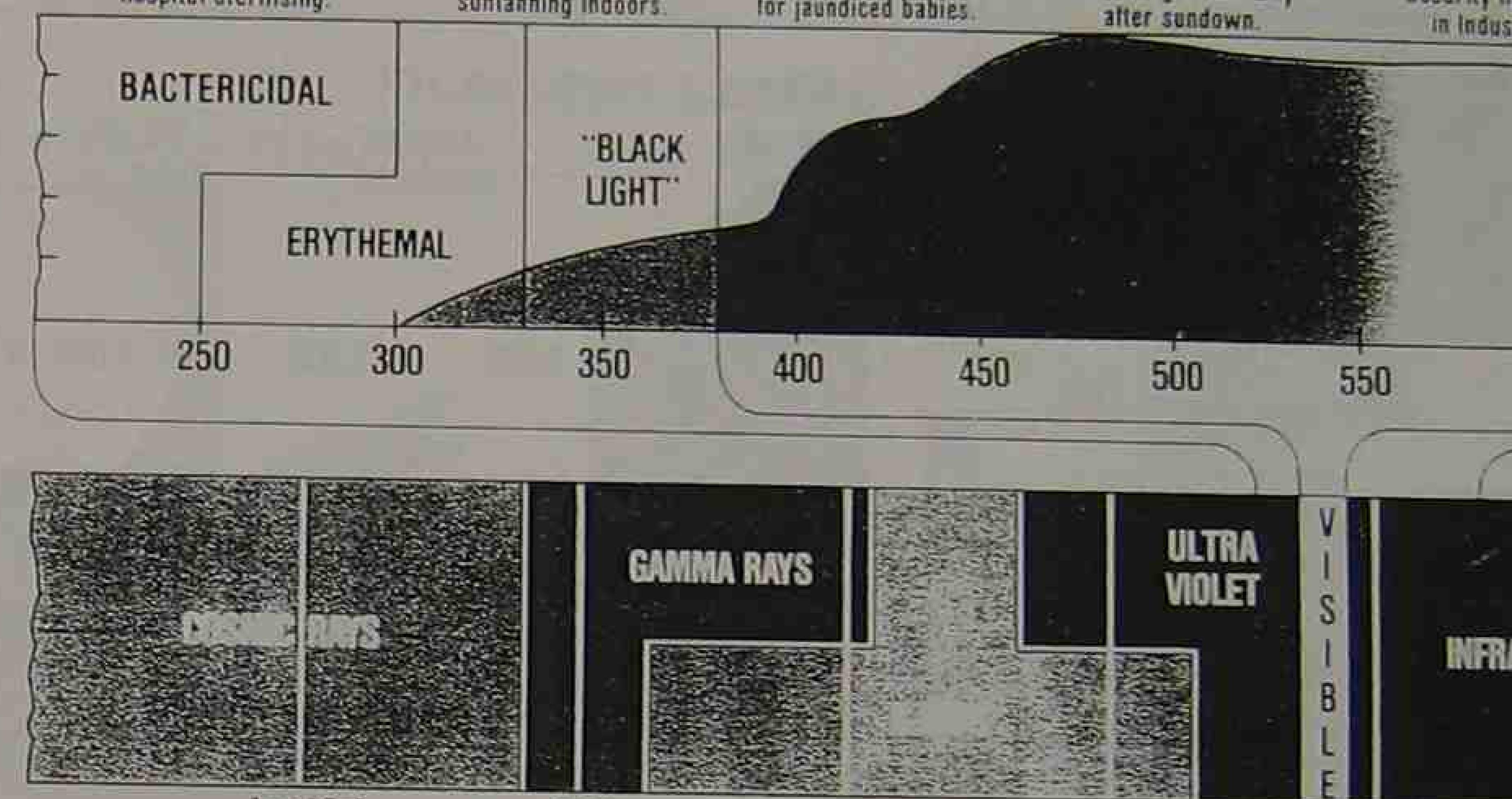
THE NATURE OF COLOUR

In his famous experiments around 1700, Newton demonstrated how a beam of white light could be split up into a spectrum by means of a prism. He identified seven distinct hues but observed that they merged into one another "so that there appeared as many degrees of colours as there were sorts of rays differing in refrangibility". Newton also observed that the colours could be recombined to give white light. He also showed that colours produced by mixing separate wavelengths could give the same visual effect as an intermediate wavelength. His experiments laid the foundation for the modern science of colorimetry.

A fundamental of colorimetry is that any colour of light can be exactly imitated by a combination of not more than three pure spectral wavelengths of light. Thus the primaries — red, blue and green — could be mixed in different proportions to give a particular colour $c_1C_1 = r_1R + b_1B + g_1G$ where C_1 is the colour to be matched.

APPLICATIONS OF ELEC


Germicidal (ultraviolet) hospital sterilising.
Ultraviolet for suntanning indoors.
Blue light treatment for jaundiced babies.
Switching on the City after sundown.
Security ligh[t] in Industry



ELECTROMAGI

R,B,G are the primaries and r_1, b_1, g_1 and c_1 are the amounts of the respective colours. Another mix could give a colour $c_2C_2 = r_2R + b_2B + g_2G$.

If we now mix c_1C_1 and c_2C_2 to give $c_3C_3 = r_3 + b_3B + g_3G$ then another principle of additive colour mixing says that $r_3 = r_2 + r_1$ and $b_3 = b_2 + b_1$ and $g_3 = g_2 + g_1$.

The equation $cC = rR + bB + gG$ involves both quality and quantity. For a match, the quantities of light must be equal so $c = r + b + g$. Dividing both sides by c give $C = RR + BB + GG$

where $R = \dfrac{r}{r + b + g}$ etc.

This in turn leads to a further equation $R + G + B = 1$ and R, G and B are called the chromaticity co-ordinates of C and can be plotted on a two dimensional chart known as the chromaticity diagram or colour triangle. (See Fig. 4.)

COLOUR TEMPERATURE

When a black body is heated it radiates energy or radiation in a characteristic manner which is dependent upon and related to its temperature. Certain materials for example such as those used as lamp filaments behave in a similar way and it has been found useful to classify the colour of certain "white" light sources by specification of their "colour temperature".

At sufficiently high temperatures they emit red light and as the temperature increases the emission becomes whiter. The colour coordinates of a black body at various temperatures can be plotted in the chromaticity diagram and the resulting line is called the "black body locus". (Fig 4)

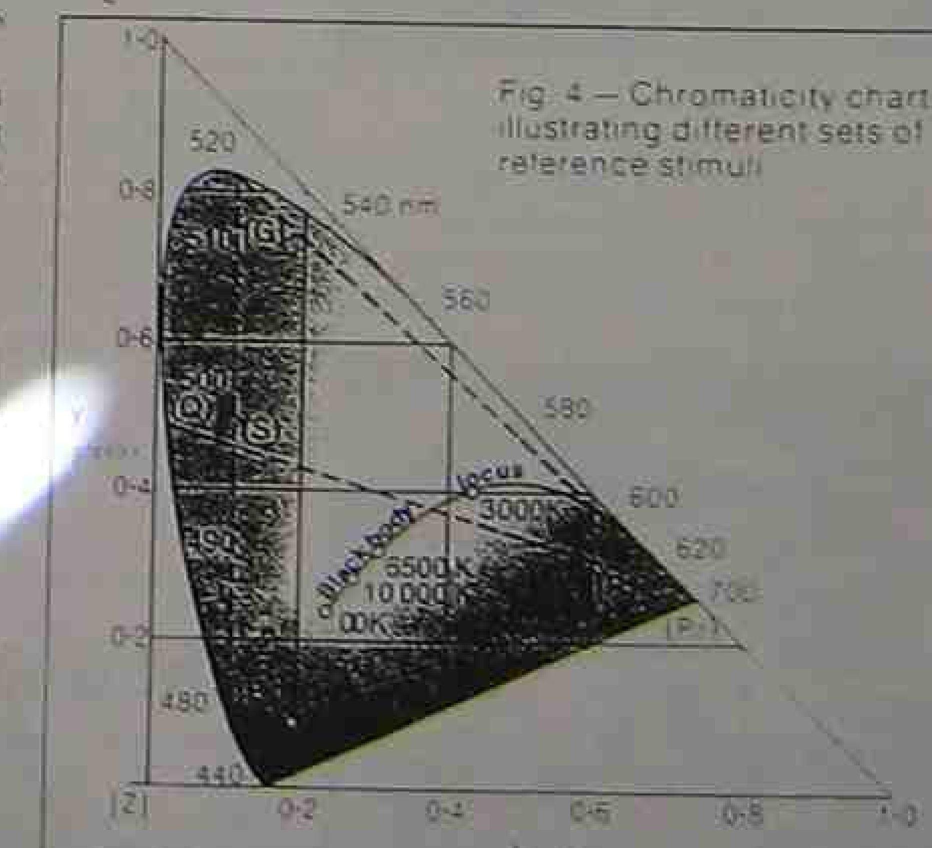
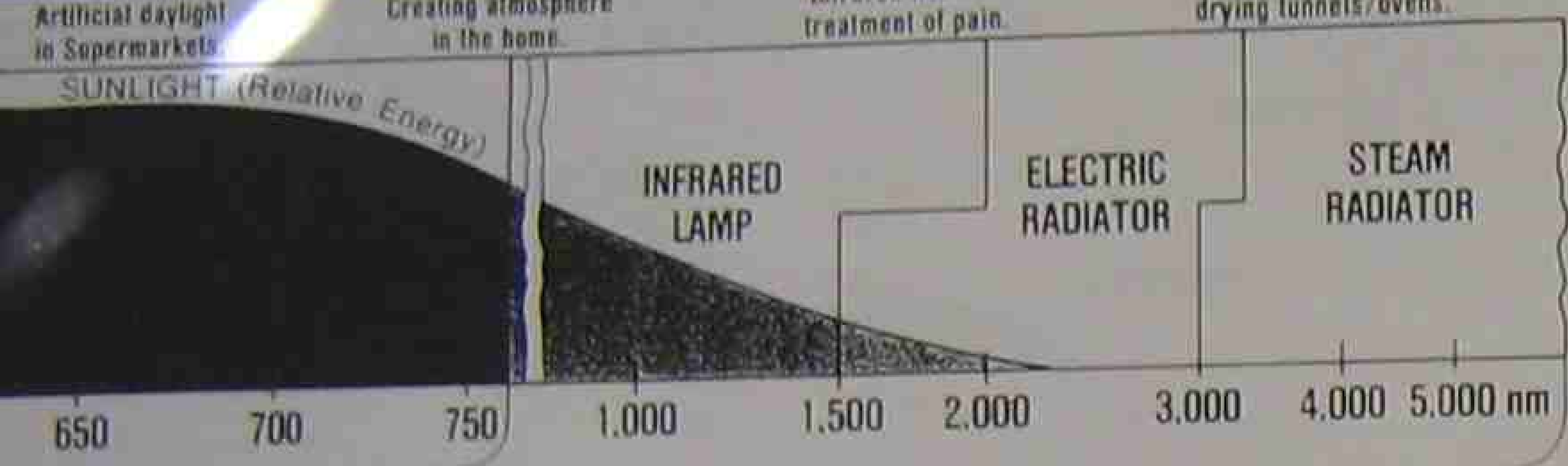
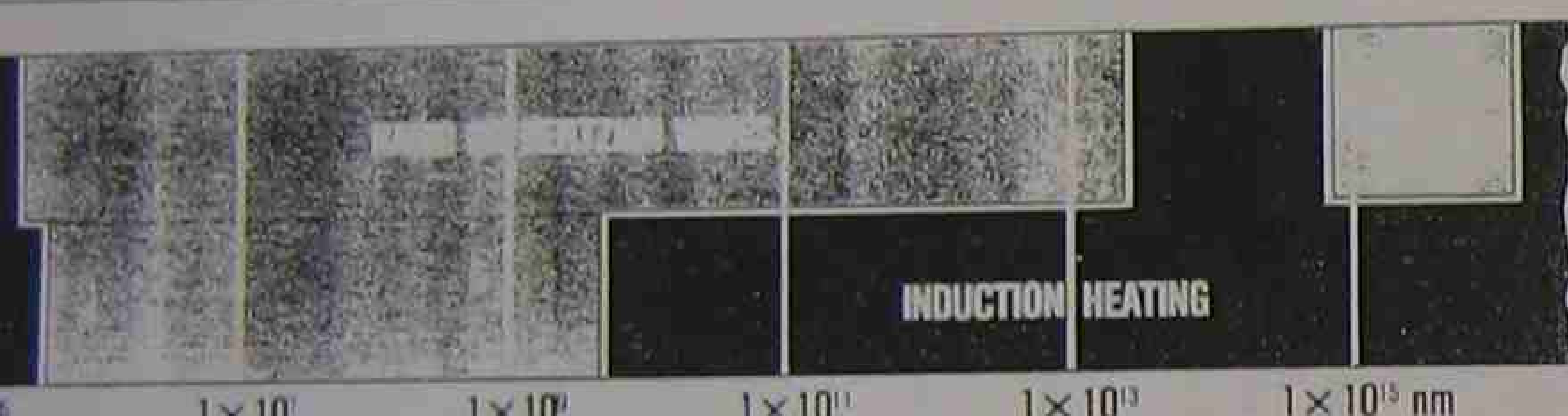


Fig. 4 — Chromaticity chart illustrating different sets of reference stimuli

So far we have considered light sources, but we must also consider the colour of light reflected from surfaces like this paper for example. The colour of any object is

OMAGNETIC RADIATION





IC SPECTRUM

commonly the hue of the most strongly reflected wavelengths. We call this colour **RED** because the ink absorbs the other wavelengths when we view it in white light. However it is important to consider the spectrum of the source illuminating the page. If you were standing under a sodium street light this page would look mainly brown and black – why?

Surface colours therefore depend on subtractive processes. If we mix pigments we cannot apply the principles of colour addition. Thus if we mix blue and yellow pigments we produce green because that is the only part of the spectrum reflected by both pigments (Fig. 5). If we did the same thing with two light sources, what colour would we get?

The most useful subtractive primaries are the complementaries of the additive primaries. They are cyan, magenta and yellow which if mixed in the right amounts produce black and were used in printing this sheet. If you examine the illustrations with a powerful magnifying glass, you will see how the different colours are obtained by varying the size of dots of cyan, magenta and yellow. Black is also used to improve the result and overcome the limitations of the printing process.

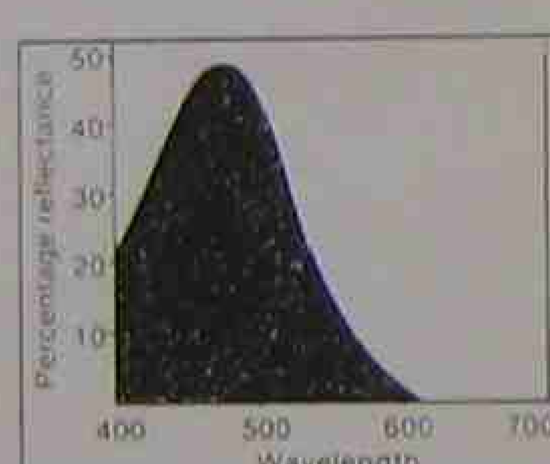


Fig. 5 – Spectral reflectance curves for blue and yellow coloured pigments and their subtractive mixture

Basic light units & terminology

Luminous Intensity (I). Early attempts to measure light were by comparison with the commonest source of artificial light – the candle. The luminous intensity of a candle perpendicular to the vertical flame was called one candlepower. It is a directional concept equivalent to the amount of light contained in a small solid angle in the direction under consideration (i.e. lumens/steradian). The present unit, the candela (cd), is approximately equal to 1 candlepower. The standard of light is a black body radiator at the temperature of freezing platinum which has a luminance of 6×10^5 cd/m².

Luminous Flux (F) describes the quantity of light emitted by a source or received by a surface. The unit – the lumen – is the luminous flux emitted within a unit of solid angle (steradian) by a point source having a uniform intensity of one candela in all directions.

Luminance (L) is a measure of the physical brightness of a surface. It is related to the amount of light emitted by a source or reflected by an object. The unit is the candela per square metre (cd/m²) and again is a directional concept as it may vary with the direction being considered.

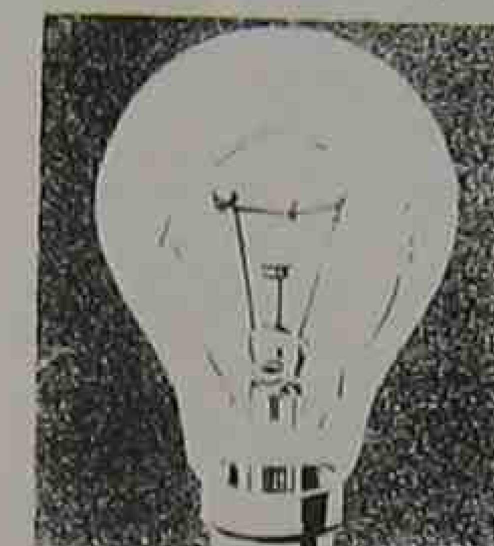
Illuminance (E) is a measure of the light falling on a surface expressed in lumens/m² or lux. The old "foot candle" is still used in some parts of the world and is approx. 0.1 lux.

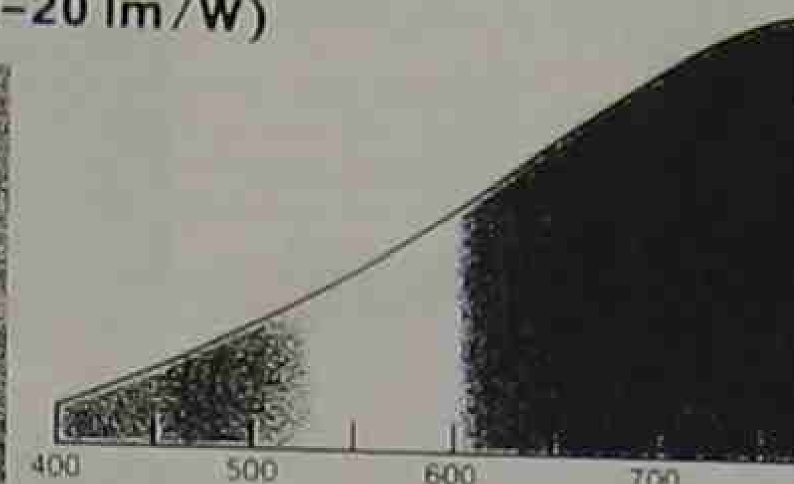
Using Light Today

Although our domestic filament lamp looks superficially like Edison's lamp (see back page), there have been many improvements, notably the replacement of the delicate carbonised thread filament with a strong one made of tungsten. Lighting engineers are constantly striving to increase the efficiency of lamps and the quality of light they produce. The efficient conversion of electrical power to light grows in importance as world energy sources are limited. In many applications we can compromise between the quantity and quality of light – street lighting is a good example. Many different types of lamp are made. They fall into two main categories – filament and discharge. Some typical Philips lamps are shown here. You can compare their efficiencies by the number of lumens per watt they produce, and the quality of light from their spectra.

FILAMENT LAMPS

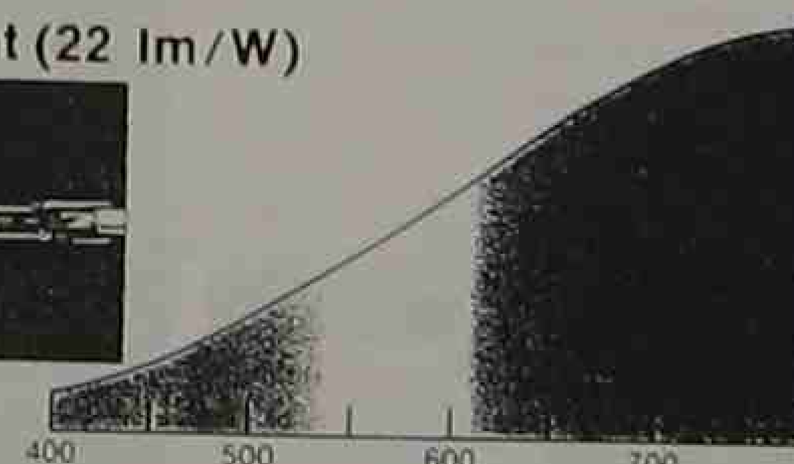
Incandescent (10–20 lm/W)





Incandescent bulbs have low efficacy, but good colour properties. Mass-production means low cost, and therefore extensive use.

Halogen Floodlight (22 lm/W)

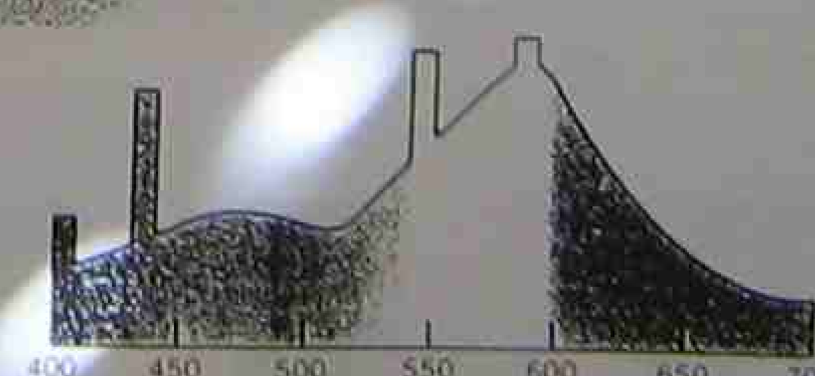


The halogen lamp is a tungsten filament lamp which uses a small amount of a halogen element. Evaporated tungsten from the filament and halogen atoms combine, and when the compound returns to the proximity of the filament the tungsten is redeposited and the halogen released.

DISCHARGE LAMPS

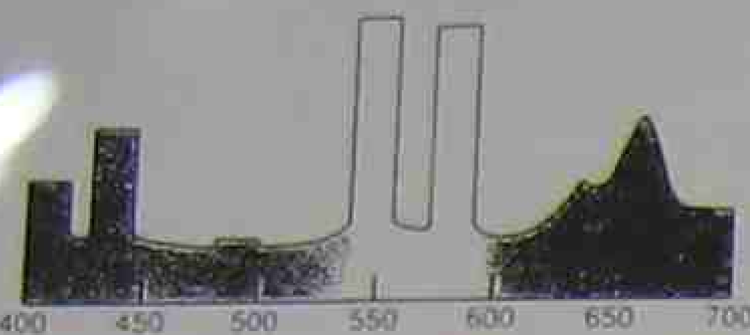
When an electric current is passed through a gas or vapour an electrical discharge or arc takes place. Electro-magnetic radiation, including light, is produced in an arc, and the wavelength depends on the gas or vapour and their pressures. If the radiation produced is ultraviolet radiation a fluorescent coating on the lamp envelope may be used to convert it to visible radiation, or light. Different phosphor blends are available to produce different coloured white light, ranging from "warm" to "cool".

Tubular Fluorescent (50–90 lm/W)



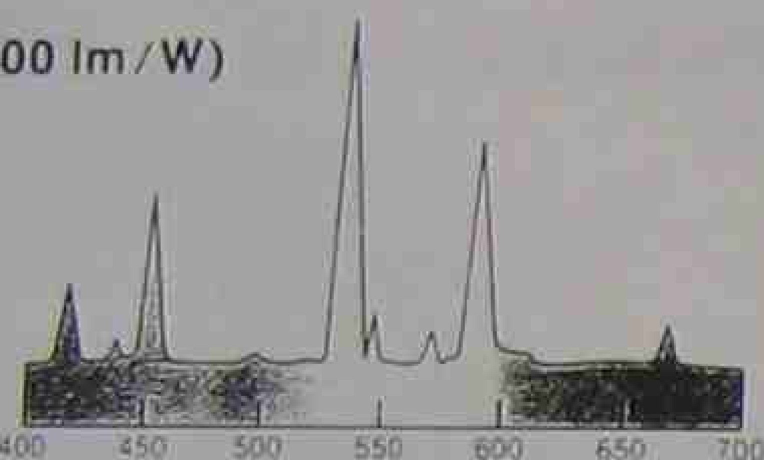
The fluorescent lamp uses an electric discharge in mercury vapour at very low pressure to produce ultraviolet radiation. This radiation is converted to light by a fluorescent powder coating on the lamp wall.

High Pressure Mercury Fluorescent (50–60 lm/W)



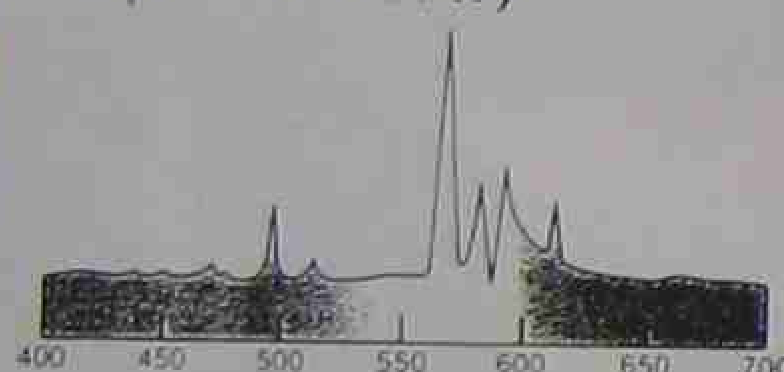
This lamp is a high pressure mercury vapour source which produces light both directly and by using a phosphor coating. It has good efficiency and reasonable colour.

Metal Halide (75–100 lm/W)



The addition of metal halides (such as sodium, thallium or indium iodides) has almost doubled the light output of mercury discharge lamps as well as improving colour quality.

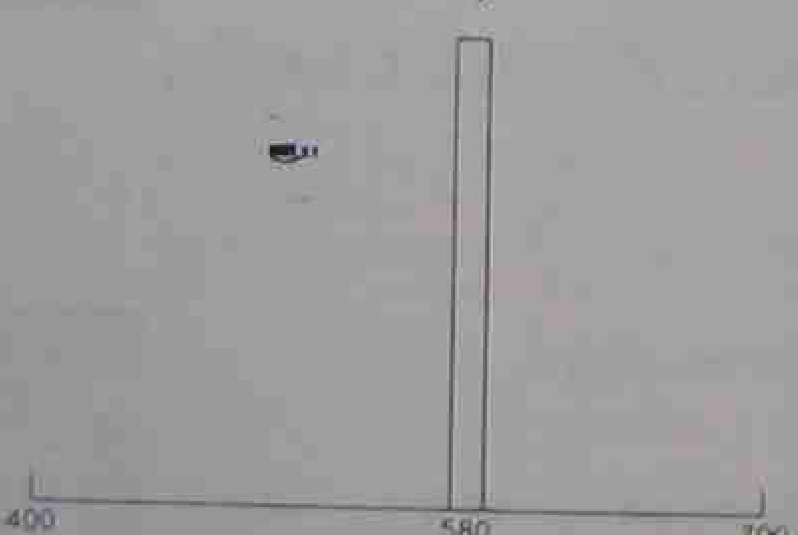
High Pressure Sodium (100–150 lm/W)

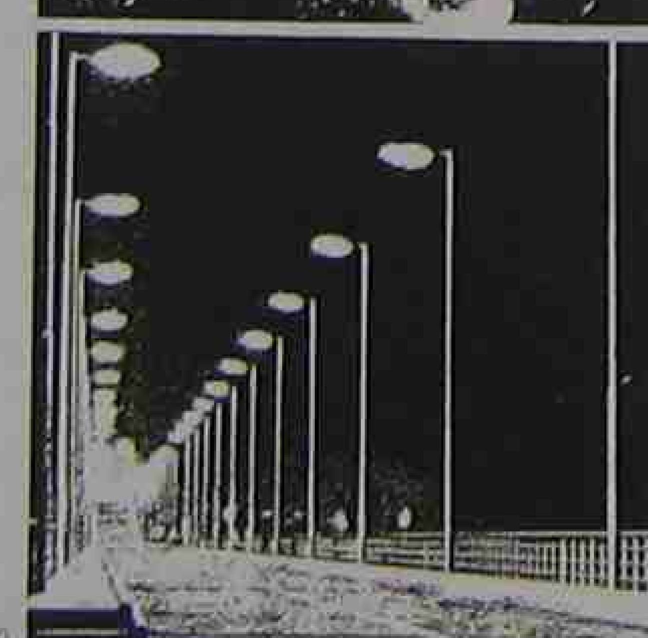


Sodium can be used as a light producing vapour. Under high pressure conditions, the colour rendering is reasonable but efficiency is much higher than for mercury lamps.

Low Pressure Sodium (150–200 lm/W)

The output is at a single wavelength (i.e. monochromatic radiation) to which the eye is extremely sensitive. This source is ideal where colour discrimination is not as important as efficiency.





LIGHT

The History of Lighting

Man has battled with the dark since time began, yet it is only in the past 100 years that we have been able to produce artificial light without relying on a flame. Having produced fire, prehistoric man struggled to keep the flame burning brightly for as long as he could. It didn't take him long to discover that animal fat burned brightly and the first lamps, consisting of hollowed stones with vegetable fibre wicks, appeared over 80,000 years ago.

Until some 200 years ago, progress was almost non-existent. The candle is mentioned in the Old Testament of the Bible. The Romans and Greeks probably knew as much about lighting as was known in the early 19th century.

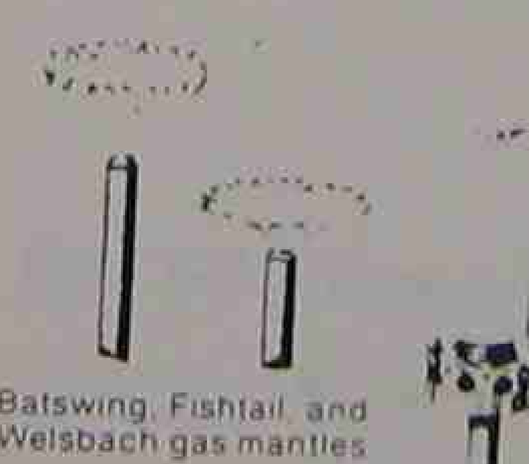
The next major step was to gas lighting in the late 1700s. In 1765, Lord Lonsdale in England piped coal gas from his mine to his office for lighting. Lack of purification and the meagre light delayed the popularity of gas lighting until Welsbach developed the incandescent mantle in the late 1880s. Here the lighting effect depends on the mantle being heated to incandescence rather than the flame itself.

Early experiments with electric lighting were hindered by the lack of large-scale permanent sources of electricity. The carbon arc was demonstrated by Sir Humphrey Davy in 1810, but it wasn't until Faraday discovered the principle of electro-magnetism that the electrical industry was born. The first permanent installation of electric arc lighting was in a British lighthouse in 1862.

Many scientists attempted to produce light from the heating effect of an electric current, but were hampered by the rapid oxidation of the filaments they used which were mainly of platinum.

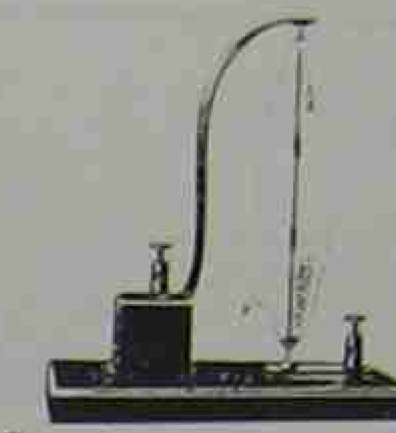
The first patent was granted to an Englishman, Frederick De Moleyns, in 1840 for an ingenious lamp relying on charcoal bridging a small gap between two platinum wires. However, the lamp proved impractical due to rapid blackening from the incandescent charcoal.

The Russian, Lodyguine, produced an incandescent lamp using graphite operating in a nitrogen-filled globe. Two hundred of his globes were installed in the St. Petersburg Dockyard in 1872, but the lamps only had a life of twelve hours.



Batswing, Fishtail, and Welsbach gas mantles

The first practical incandescent lamps were produced by two scientists working independently, Edison in America, and Swan in England. Both were prolific inventors. Edison is probably best known for his invention of the phonogram or record player, while Swan was responsible for the bromide process used in photography.



Swan Carbon filament lamp.

Edison's lamp, 1879.

It is probably true to say that while Swan made the first practical lamp, it was Edison who had the vision to see how electricity could be distributed so that maximum use could be made of the electric lamp. The lamp itself was only part of Edison's concept.

Edison reasoned that from a distribution point of view, a high resistance lamp was needed. Other attempts had followed the arc concept relying on low voltages and high currents. Edison worked on the opposite idea and his first lamp burned steadily for 45 hours on October 21st 1879, leading to a patent on January 27th 1880. The principles he outlined then are practically the same as those used today with improvements only in the construction of the filament.

Philips first lamp factory, Eindhoven, 1891

It was eleven years later that Gerard Philips bought a buck-skin factory in Eindhoven in the Netherlands, and equipped it to make electric lamps. Manufacture started in 1892, and by 1895 production had reached 109,000 lamps. By the turn of the century, Philips had become the third largest manufacturer of lamps in Europe and in 1916 was producing 80,000 lamps per day. The company established its own research department in 1914, and after the first world war started to expand its range of products into the making of X-ray tubes and radio valves. Today Philips is possibly best known for its involvement in the broadcasting and television industry, but lighting still forms a major part of its activities.

Topic - LIGHT SOURCES - ASSORTED

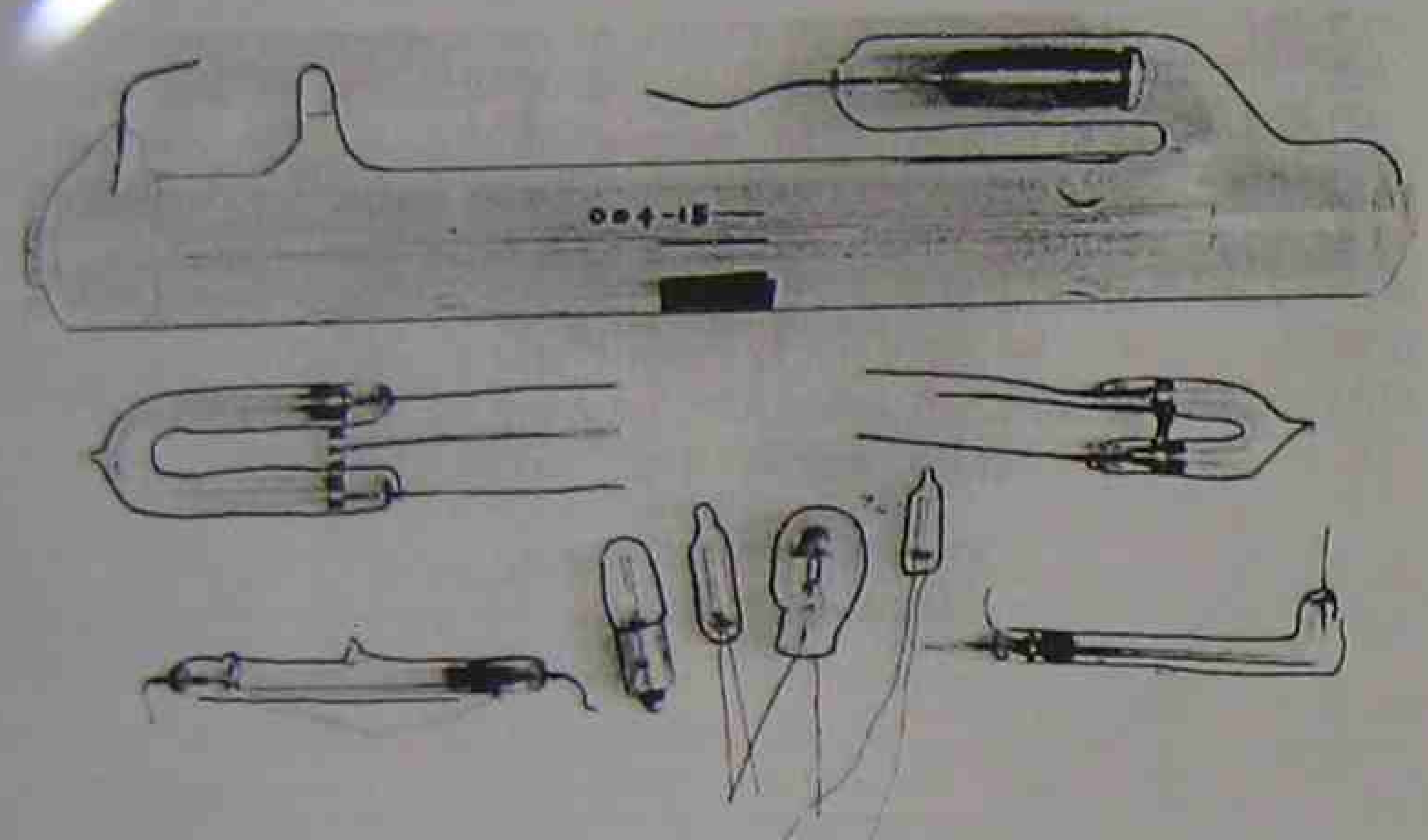
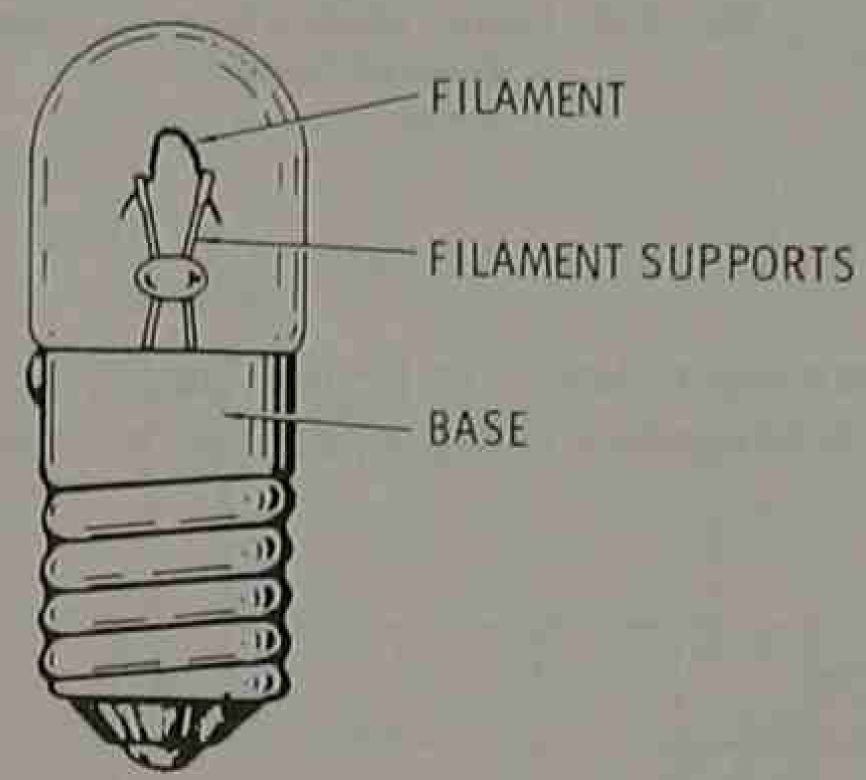


Fig. 3-18. Helium-neon laser tube (top) and assortment of neon, argon, and xenon lamps.

Topic - INCANDESCENT LAMP



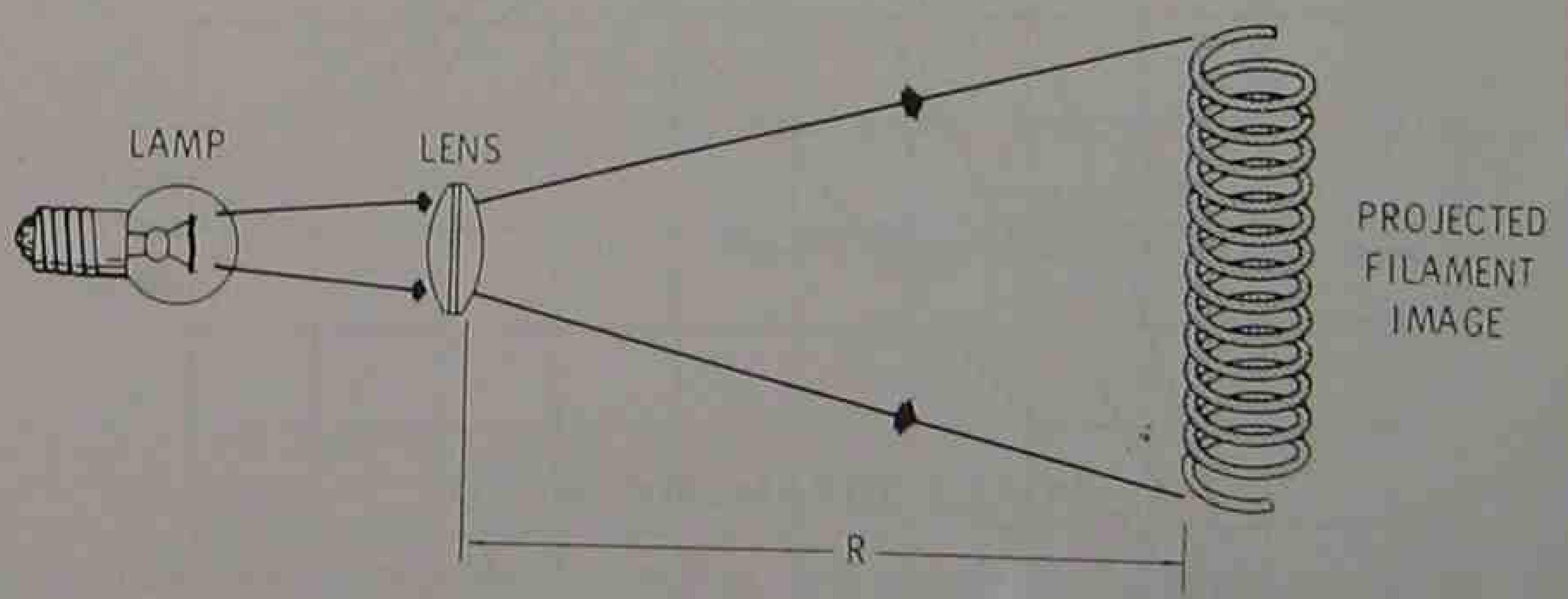


Fig. 3-3. Filament image projected by focused incandescent source.

Industrial Sensors.

Topic - INCANDESCENT LAMP MODULATION

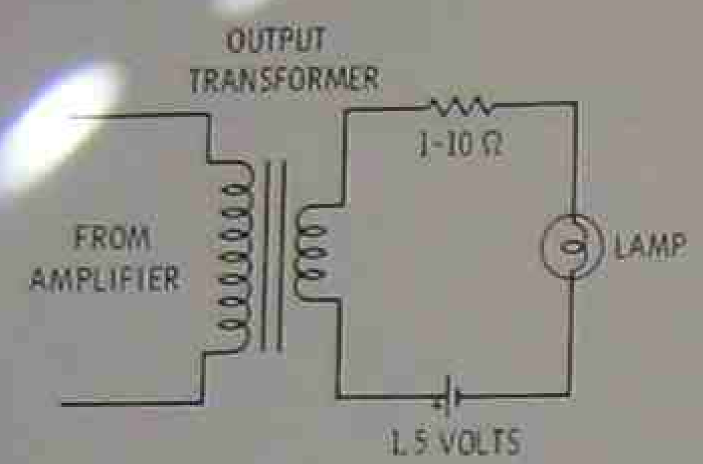


Fig. 3-4. Direct modulation of incandescent lamp.

The spectral output of a tungsten lamp is broad band, and Fig. 3-5 shows the output of a tungsten lamp operated at several color temperatures.

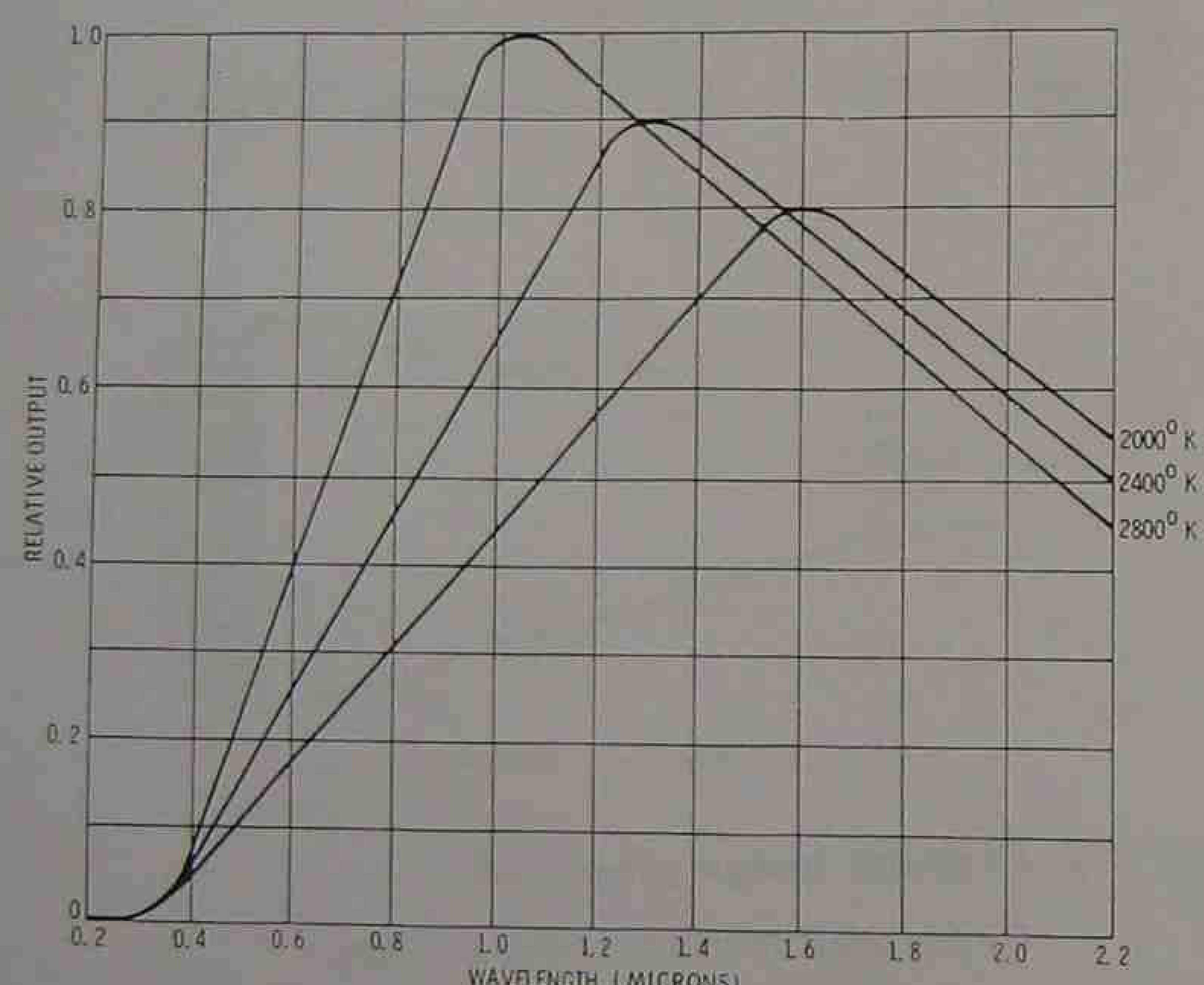


Fig. 3-5. Tungsten lamp spectral output versus filament color temperature.

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Topic - GLOW DISCHARGE LAMPS

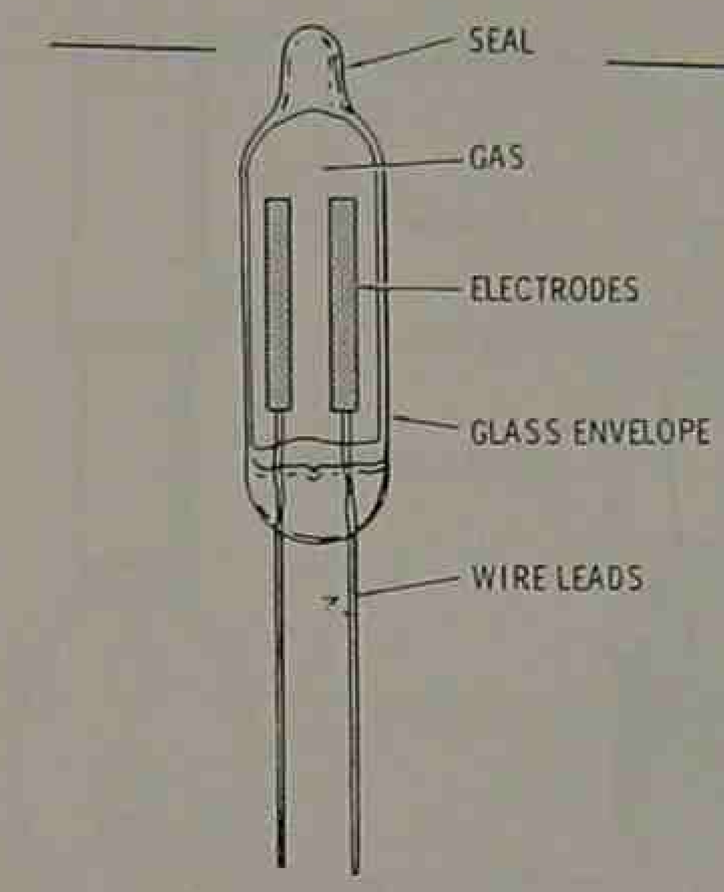

Fig. 3-6. Glow-lamp construction.

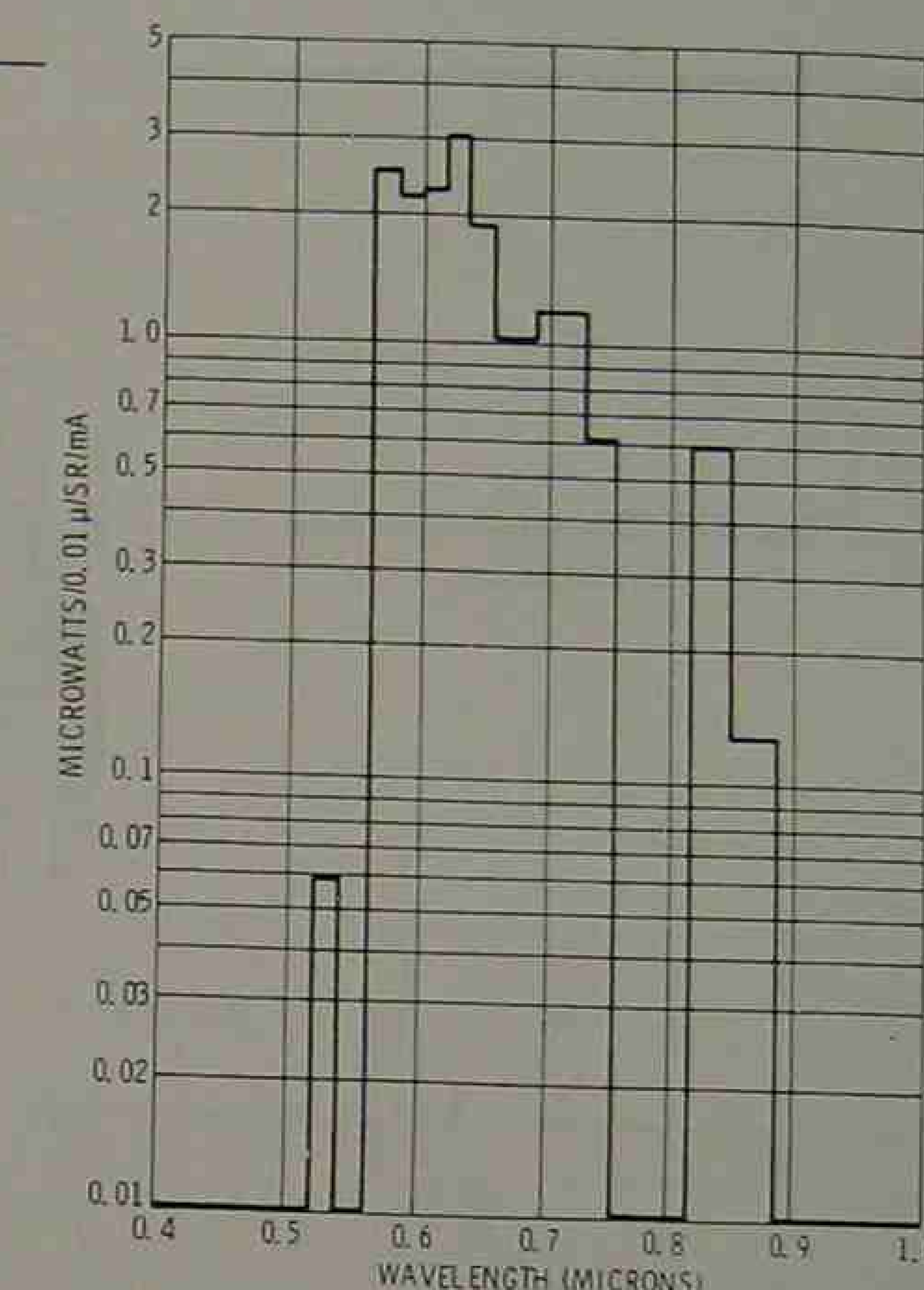

Fig. 3-7. Spectral output of a neon lamp.

GLOW-DISCHARGE LAMPS

The glow lamp shown in Fig. 3-6 consists of a glass envelope containing two or more electrodes and filled with a gas capable of being ionized. The most common glow lamps contain neon or argon. Neon produces a yellow-red range of wavelengths

Industrial Sensors.

Topic - ARC DISCHARGE LAMP

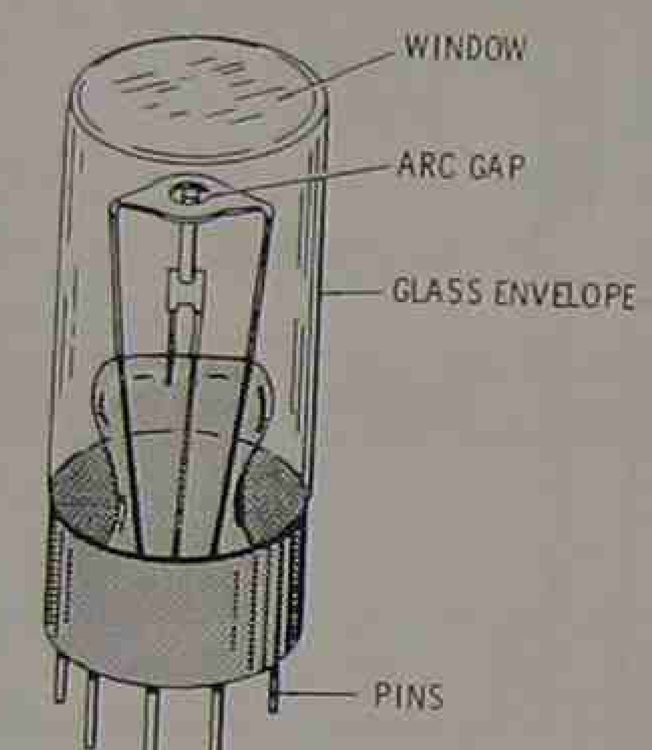

Fig. 3-9. Zirconium arc lamp construction.

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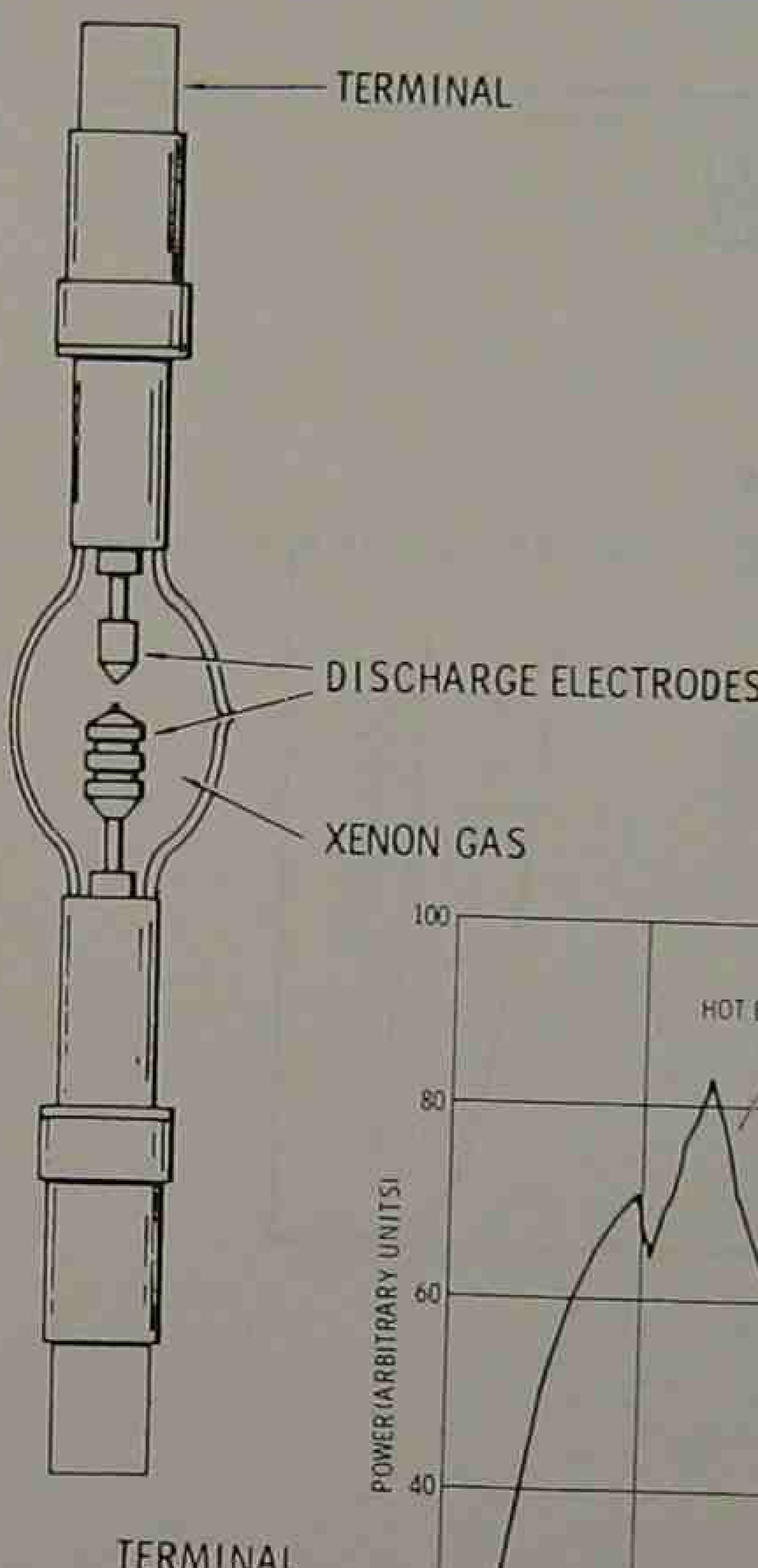
Topic - XENON DISCHARGE LAMP



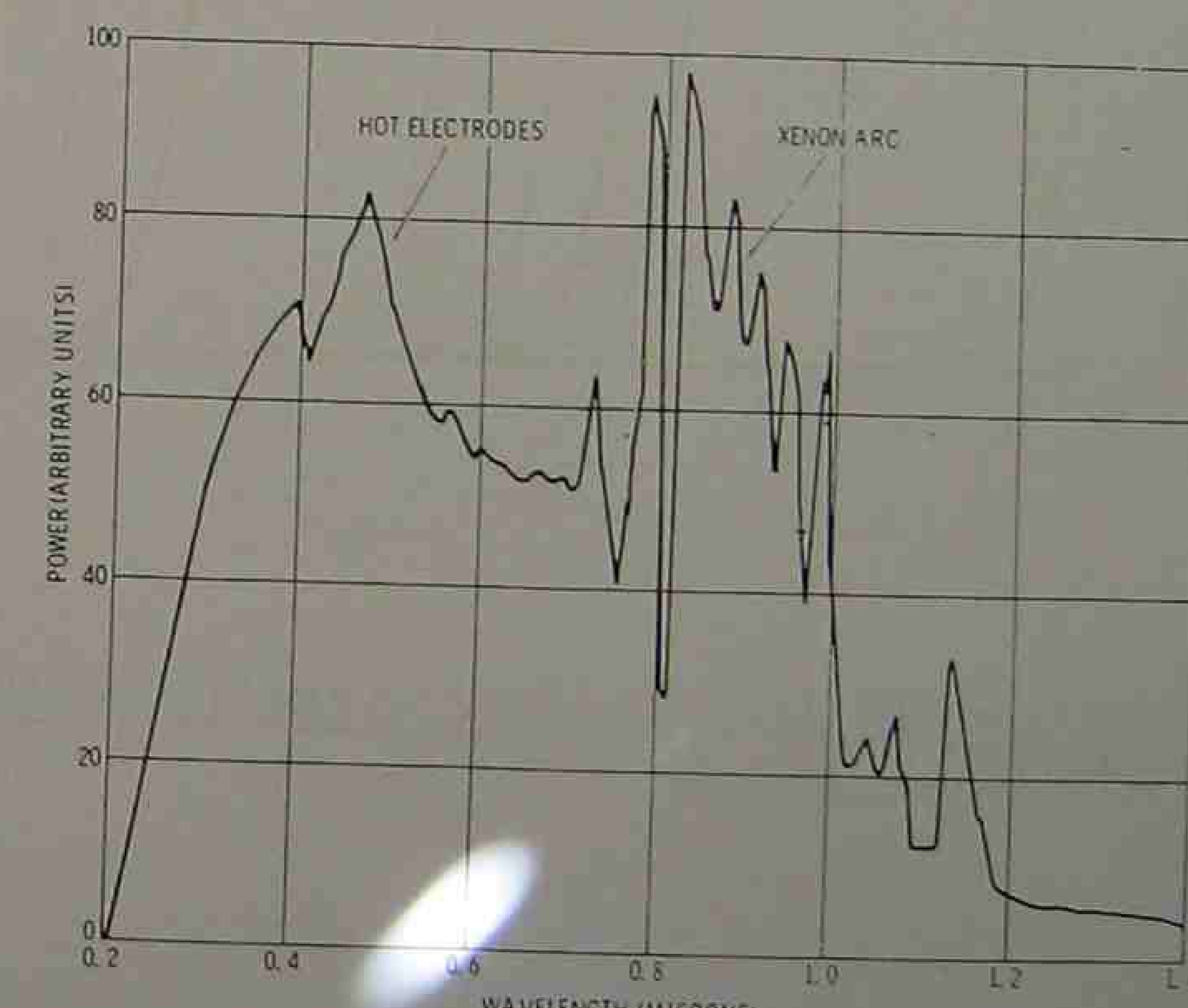
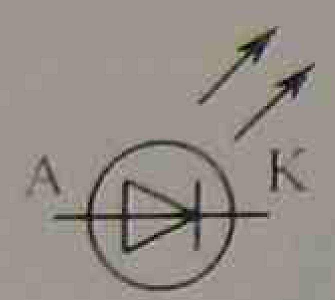


Fig. 3-11. Typical xenon arc lamp spectrum.

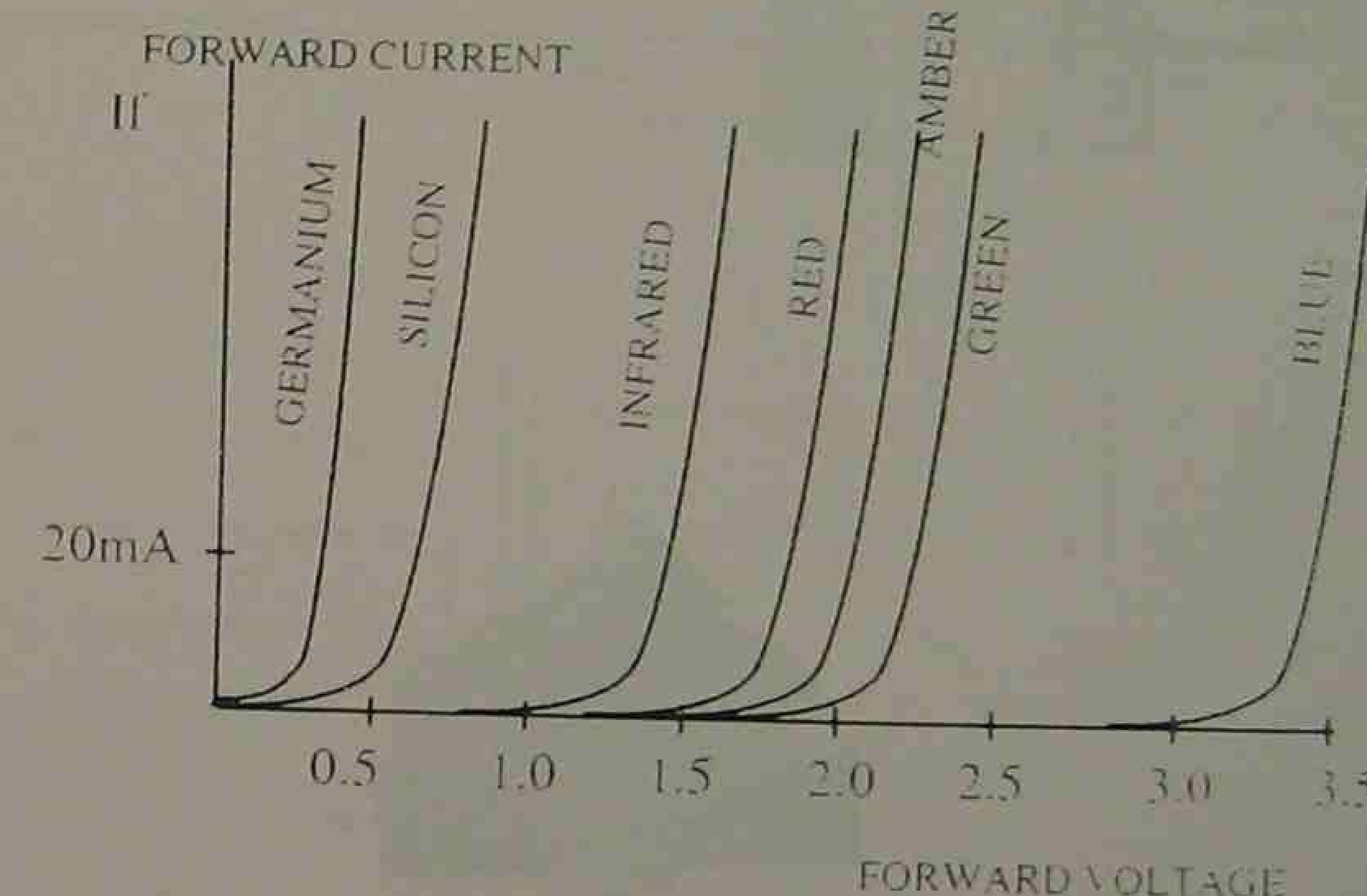
LED

LIGHT EMITTING DIODE

LED SYMBOL



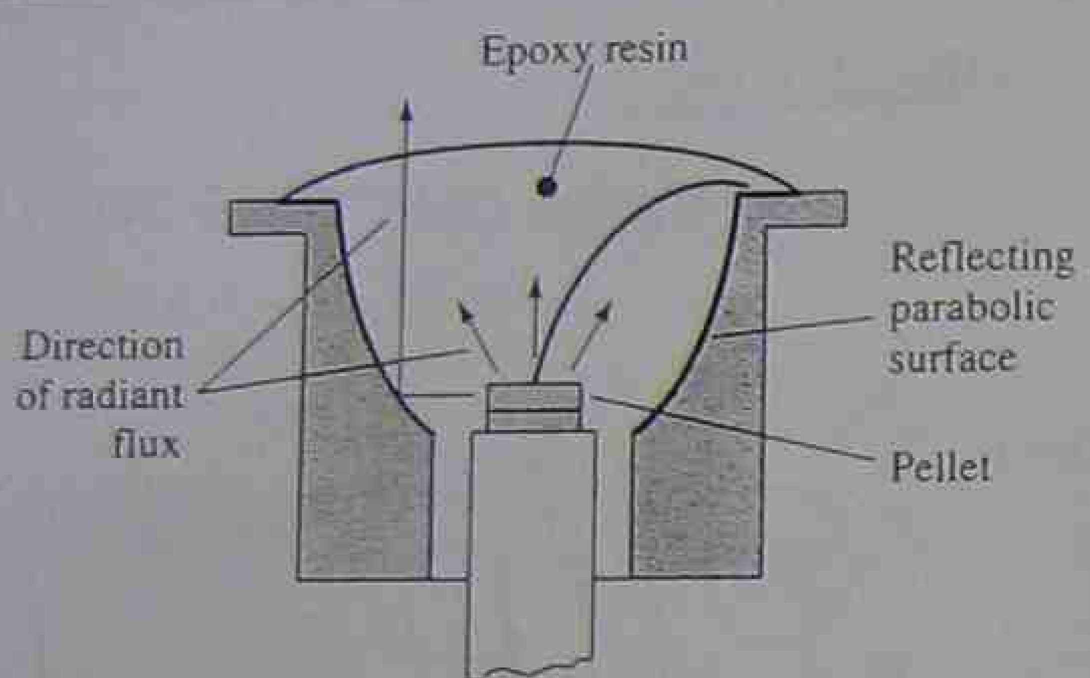
LEDs (cont.)



FORWARD CHARACTERISTICS OF LEDs AND STANDARD Si AND Ge DIODES

Industrial Sensors.

Topic - LED CONSTRUCTION



(a)

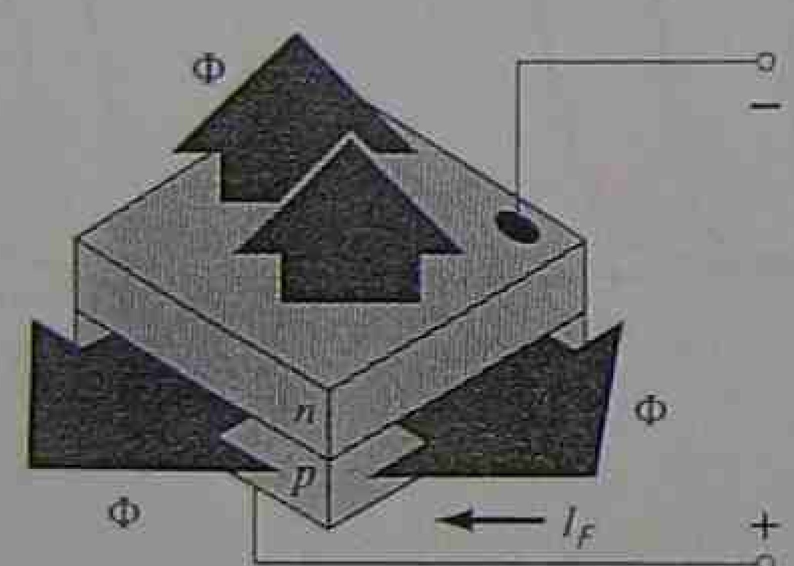
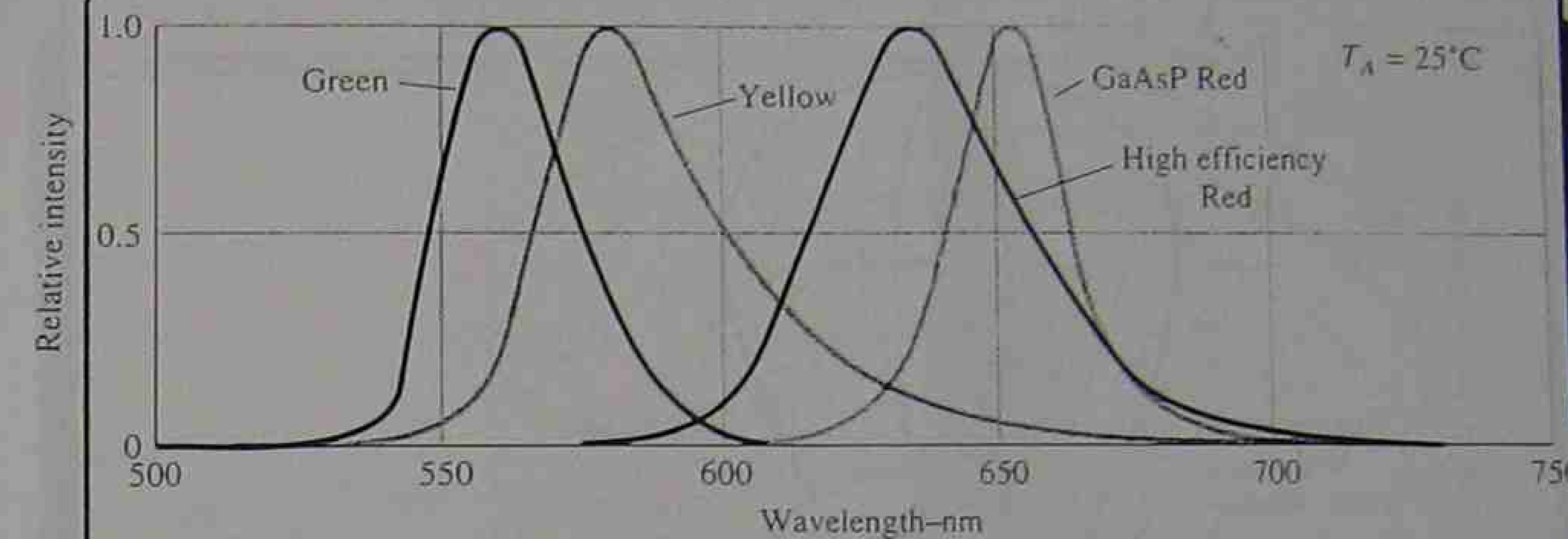
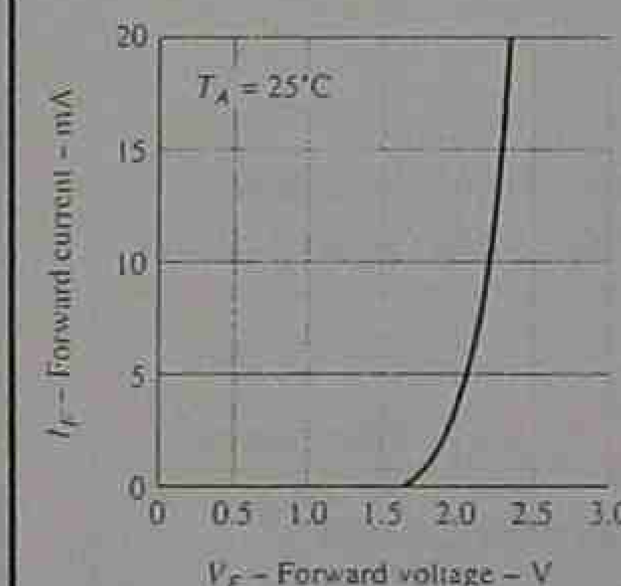


Figure 20.31 General structure of a semiconductor IR-emitting diode. (Courtesy RCA Solid State Division.)

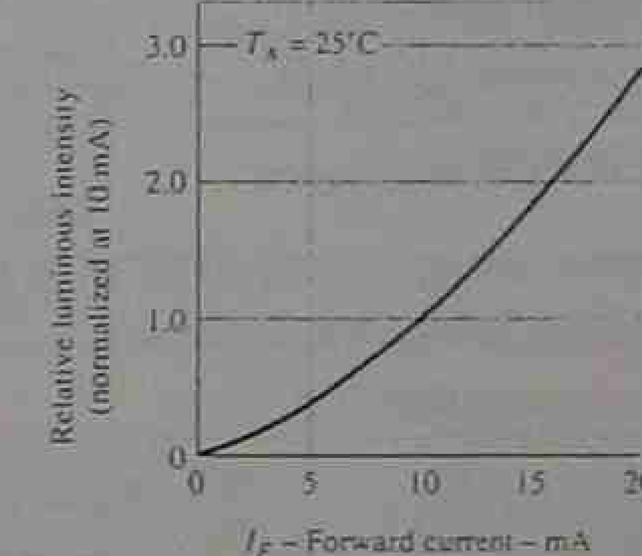
Industrial Sensors.

Topic - LED CHARACTERISTICS

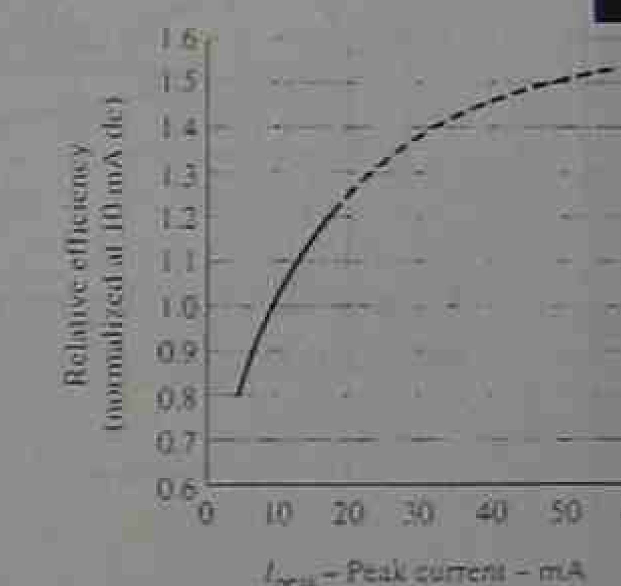




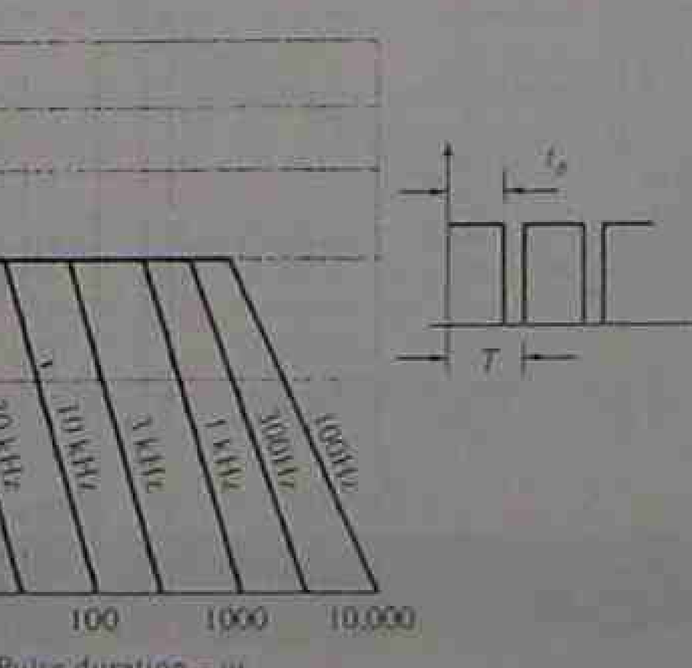
(e)



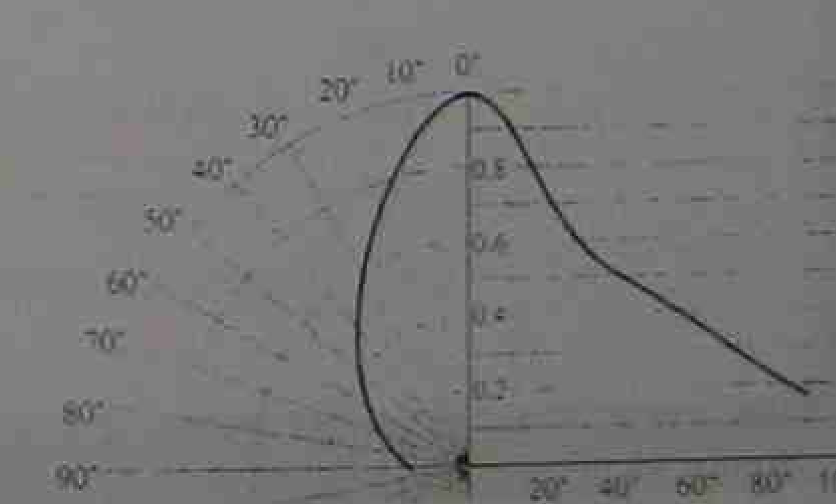
(f)



(g)



(h)



(i)

Figure 1.55 Continued

Industrial Sensors.

Topic - LED - I.R. EMITTER

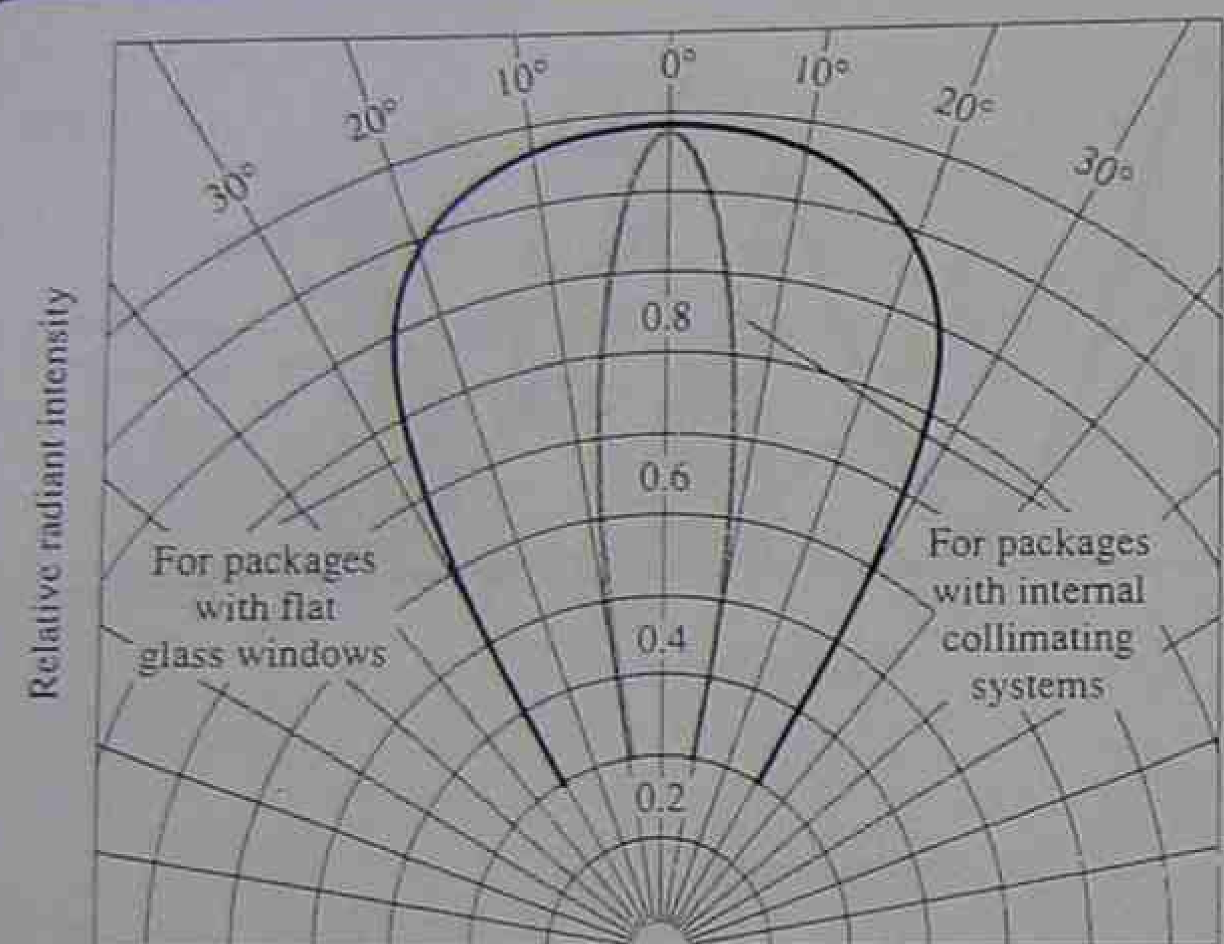
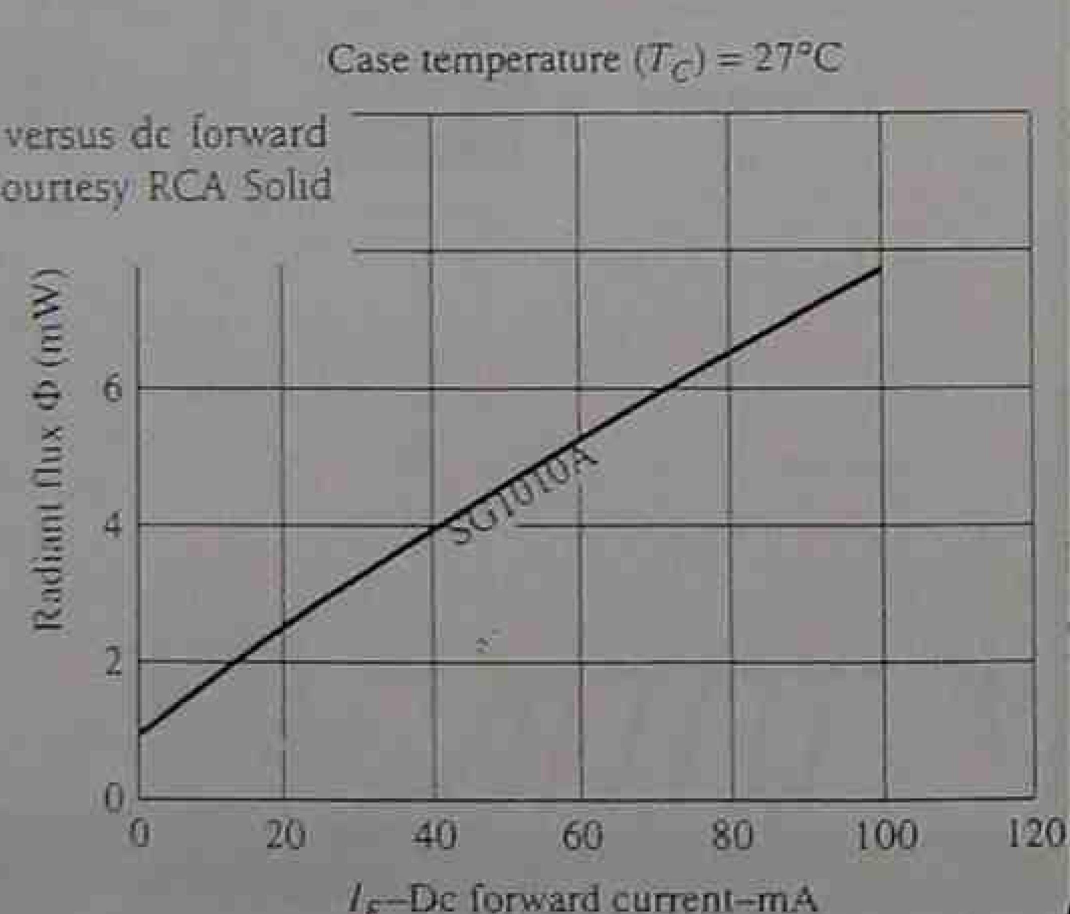


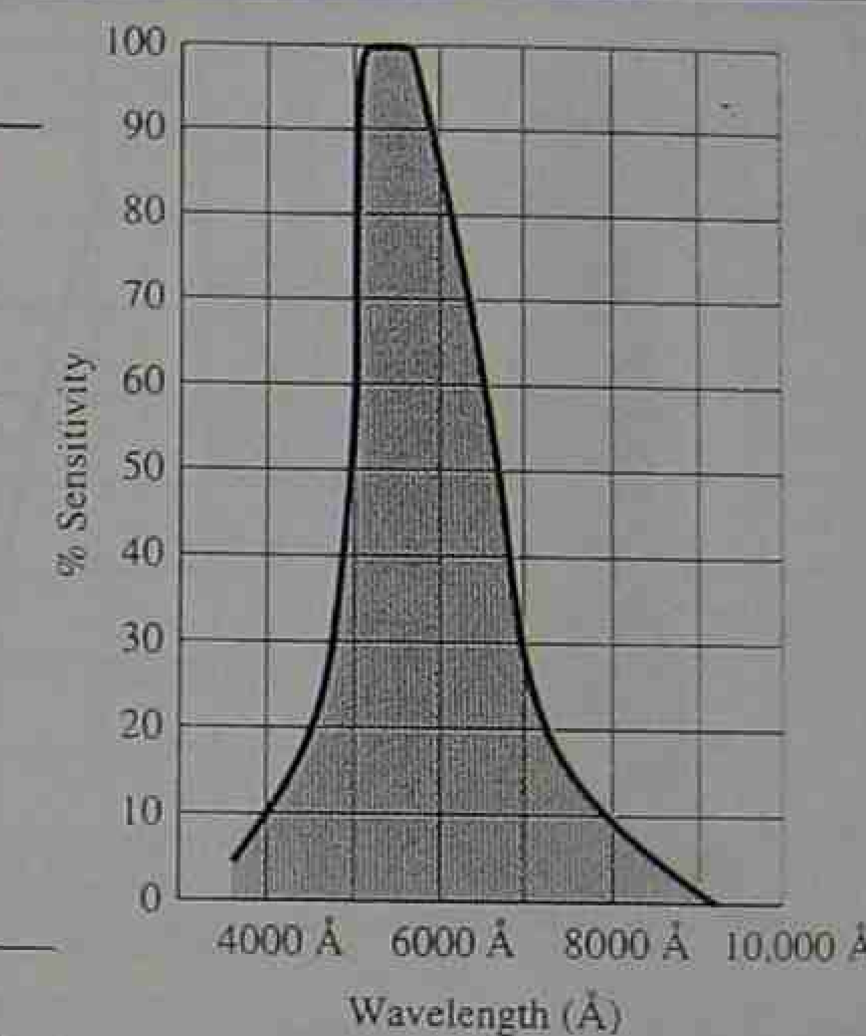
Figure 20.33 Typical radiant intensity patterns of RCA IR-emitting diodes. (Courtesy RCA Solid State Division.)

Figure 20.32 Typical radiant flux versus dc forward current for an IR-emitting diode. (Courtesy RCA Solid State Division.)



Industrial Sensors.

Topic - LED IR EMITTER



Variation of Conductance With Temperature and Light

Footcandles	*.01*	*0.1*	*1.0*	*10*	*100*
Temperature			*% Conductance*		
−25°C	103	104	104	102	106
0	98	102	102	100	103
25°C	100	100	100	100	100
50°C	98	102	103	104	99
75°C	90	106	108	109	104

Response Time Versus Light

Footcandles	*0.01*	*0.1*	*1.0*	*10*	*100*
Rise (seconds)	0.5	0.095	0.022	0.005	0.002
Decay (seconds)	0.125	0.021	0.005	0.002	0.001

Figure 20.30 Characteristics of a Clairex CdS photoconductive cell. (Courtesy Clairex Electronics.)

Industrial Sensors.

Topic - PHOTODIODES

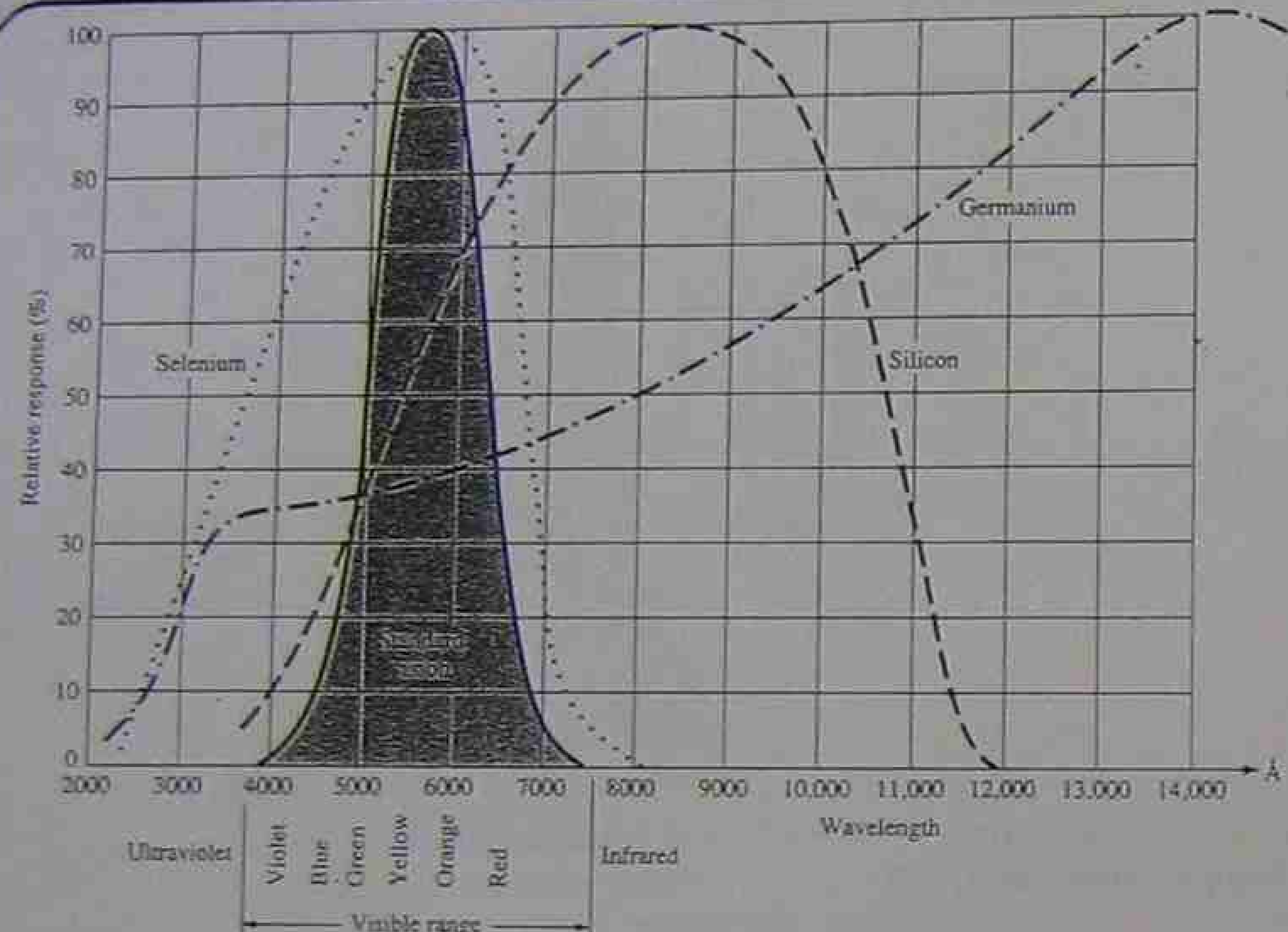
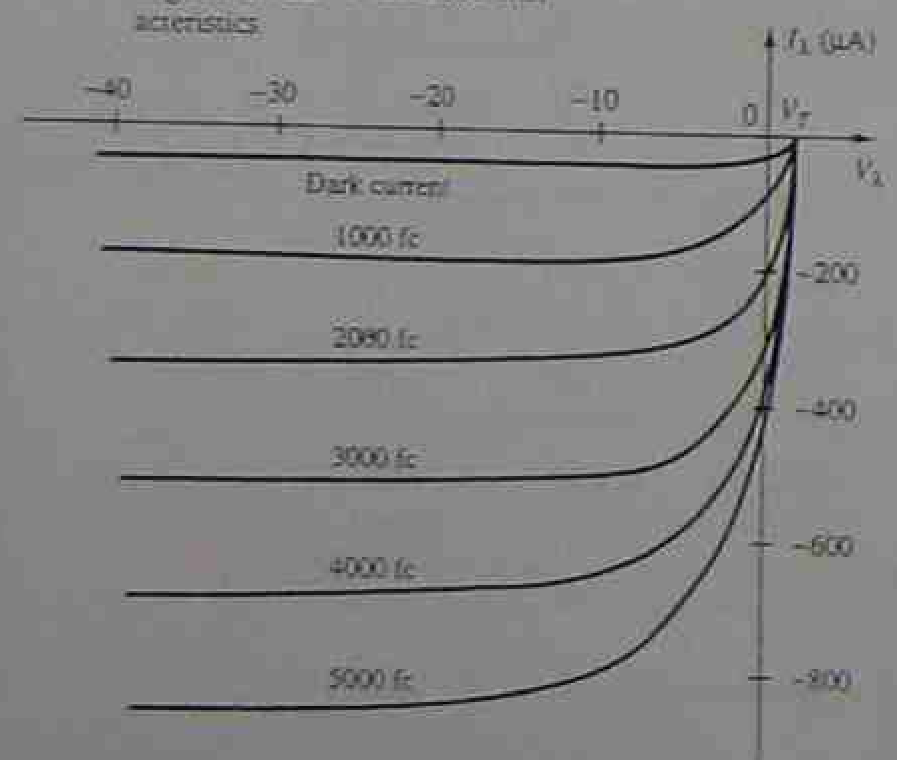


Figure 20.20 Relative spectral response for Si, Ge, and selenium as compared to the human eye.

Figure 20.22 Photodiode characteristics



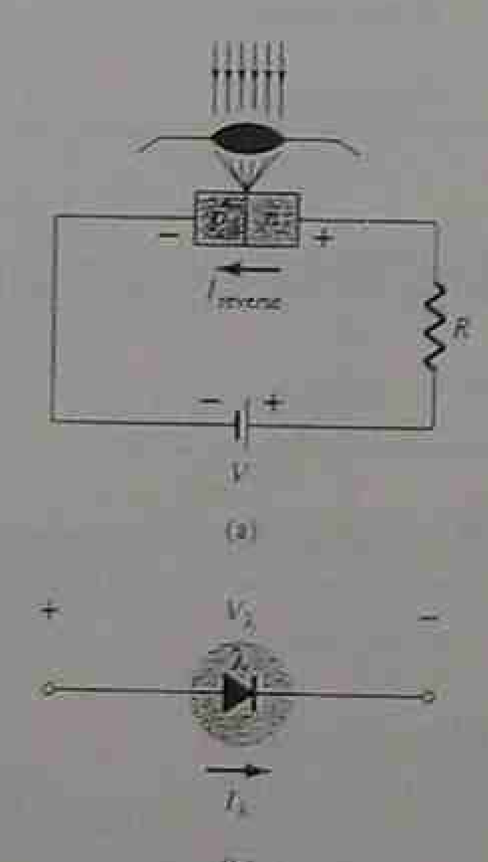
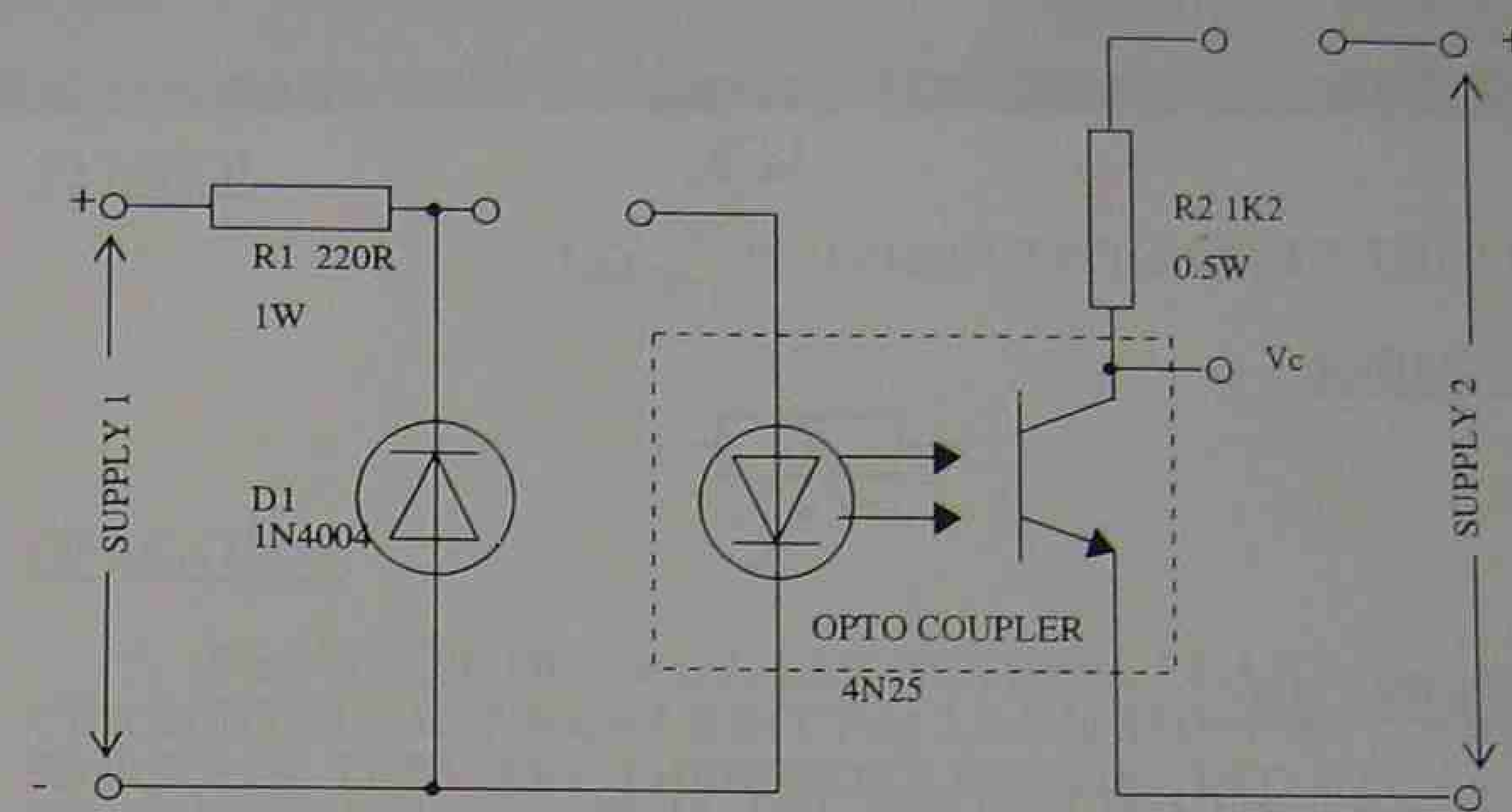


Figure 20.21 Photodiode. (a) basic-biasing arrangement and construction. (b) symbol



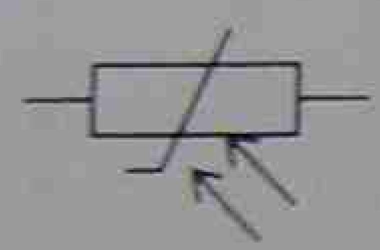
OPTO COUPLER PRACTICAL

Topic - Photoresistor.

LDR

LIGHT DEPENDENT RESISTOR

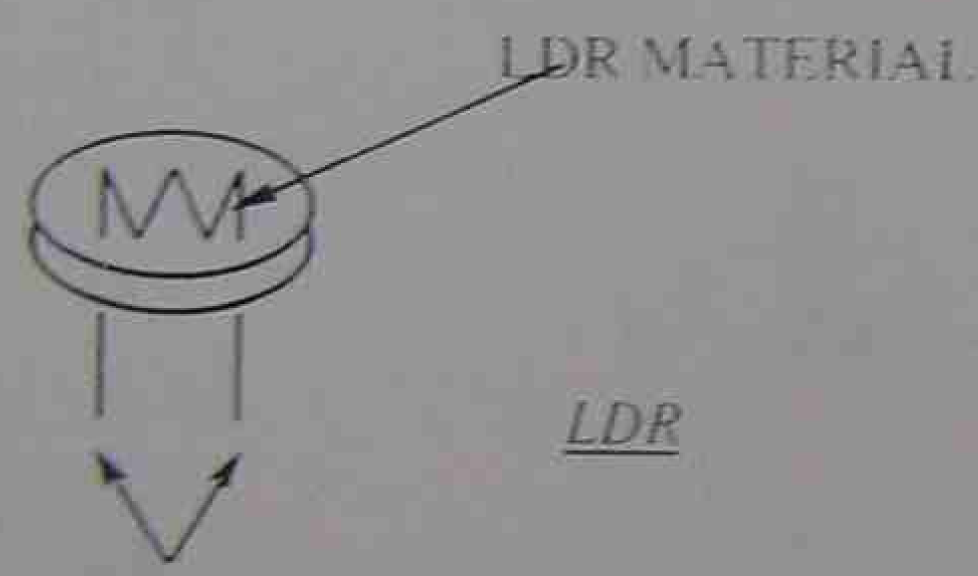
SYMBOL



OPERATION

LIGHT FALLING ON THE SURFACE OF THE LDR CAUSES A DECREASE IN RESISTANCE .

DARK RESISTANCE IS ***VERY HIGH*** AROUND 1M TO 10M.



PEAK SPECTRAL RESPONSE

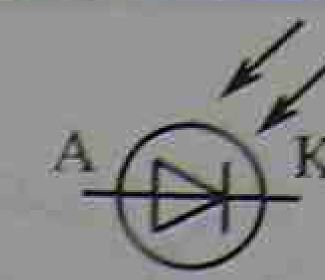
CADMIUM SELENIDE CdS = 735nm(INFRARED)
CADMIUM SULPHIDE CdSe = 530nm to 600nm (EYE)

RESPONSE TIME

VERY SLOW IN SECONDS FROM DARK TO LIGHT

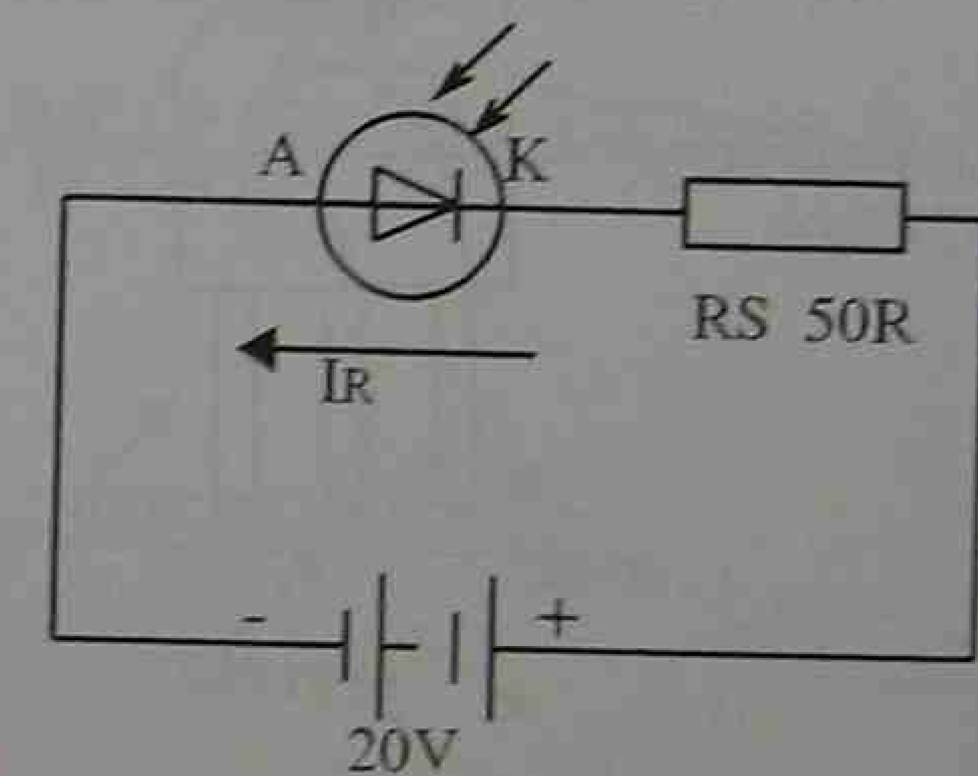
Topic - Photodiode.

SYMBOL



OPERATION

PHOTO DIODES WILL LET REVERSE LEAKAGE CURRENT FLOW WHEN LIGHT FALLS ON THE P.N. JUNCTION. THEY ARE THEREFORE CONNECTED IN REVERSE BIAS.



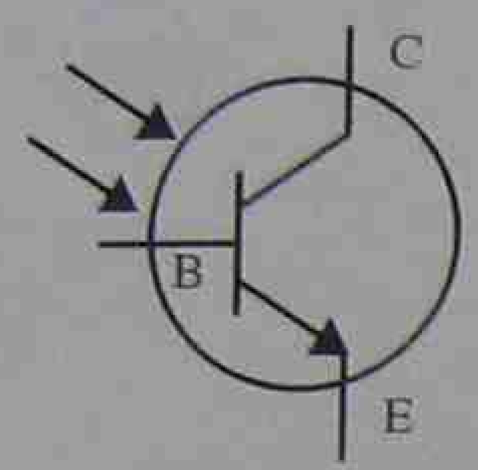
TYPICAL CONNECTION OF PHOTO DIODE

SPECTRAL RESPONSE IS 400nm to 1100nm BUT PEAKING IN INFRARED

RESPONSE TIME IS EXTREMELY FAST = 1nS

Topic - Phototransistor.

SYMBOL



CONSTRUCTION

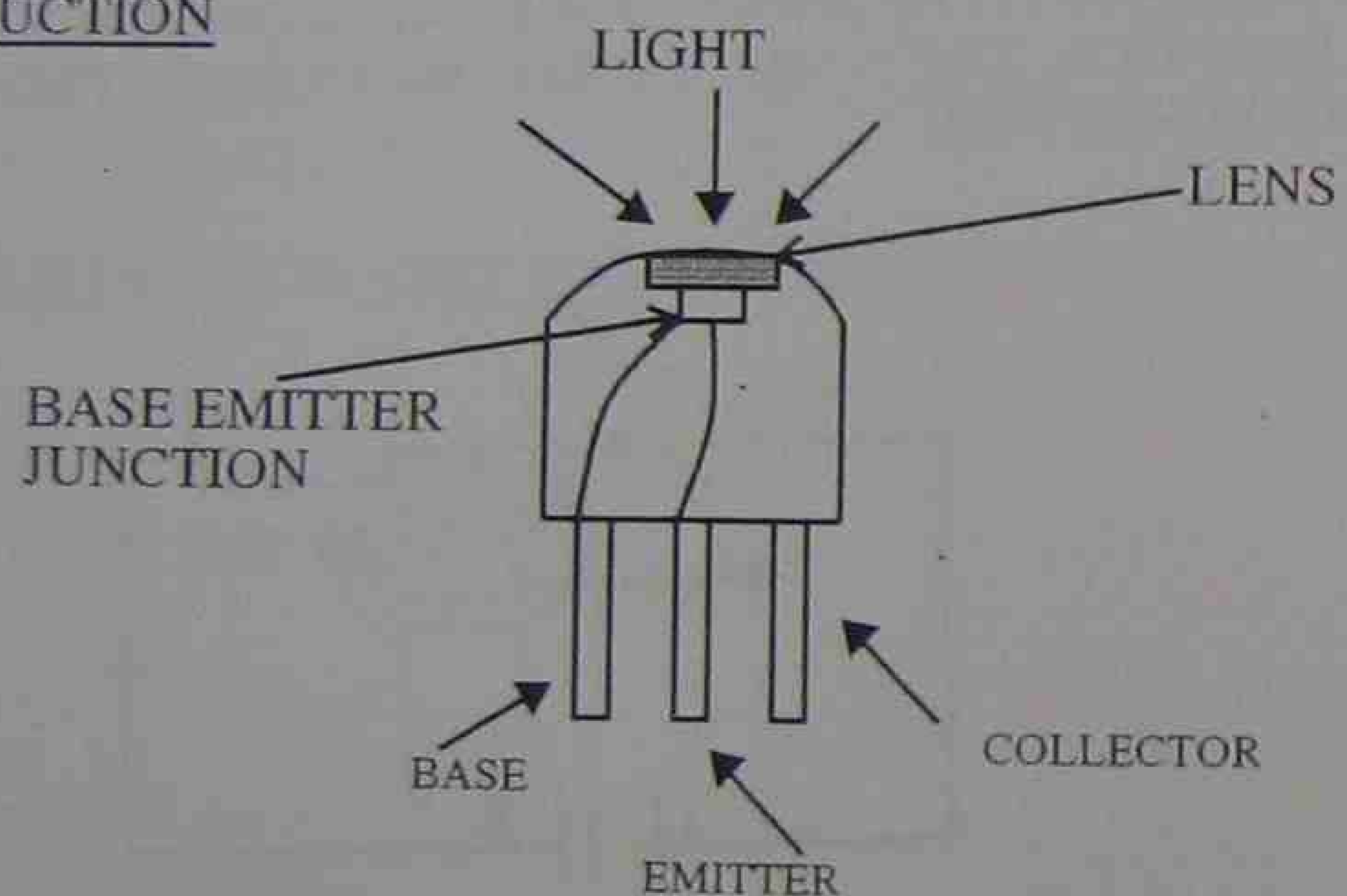


PHOTO TRANSISTOR

OPERATION

LIGHT ENERGY FALLING ON THE BASE / EMITTER JUNCTION CAUSES CURRENT TO FLOW FROM BASE TO EMITTER SWITCHING ON THE TRANSISTOR.

Topic - Phototransistor. (CONT.)

RESPONSE TIME
IS ABOUT 1μS

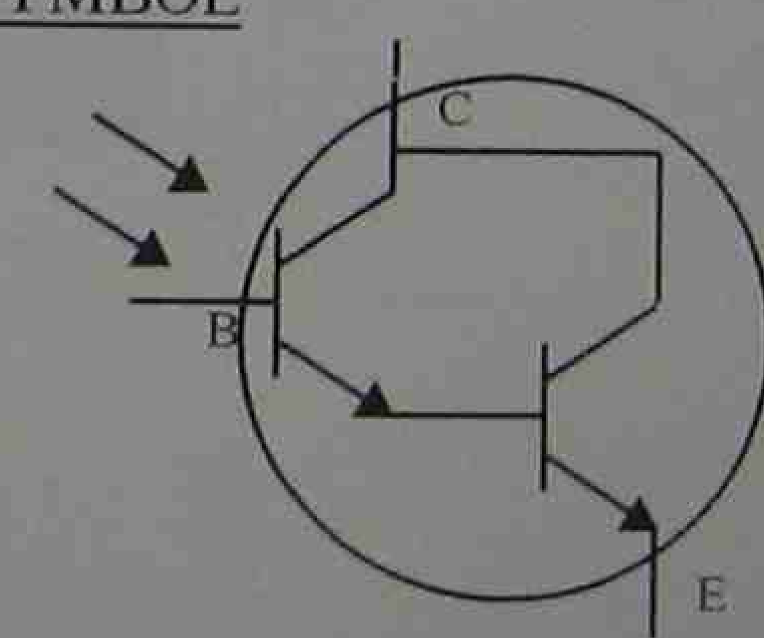
LENSES ARE USED TO FOCUS AND THEREFORE INCREASE THE LIGHT FALLING ON THE BASE / EMITTER JUNCTION.

PEAK SPECTRAL RESPONSE IS TO INFRARED ≈900nm

IF THE LIGHT LEVELS TO BE DETECTED ARE VERY LOW A PHOTO DARLINGTON MAY BE USED.
HOWEVER RESPONSE TIME IS REDUCED.

RESPONSE TIME OF PHOTO DARLINGTON
IS ABOUT 75μS

PHOTO DARLINGTON SYMBOL



Topic - Photo SCR -LASCR-.

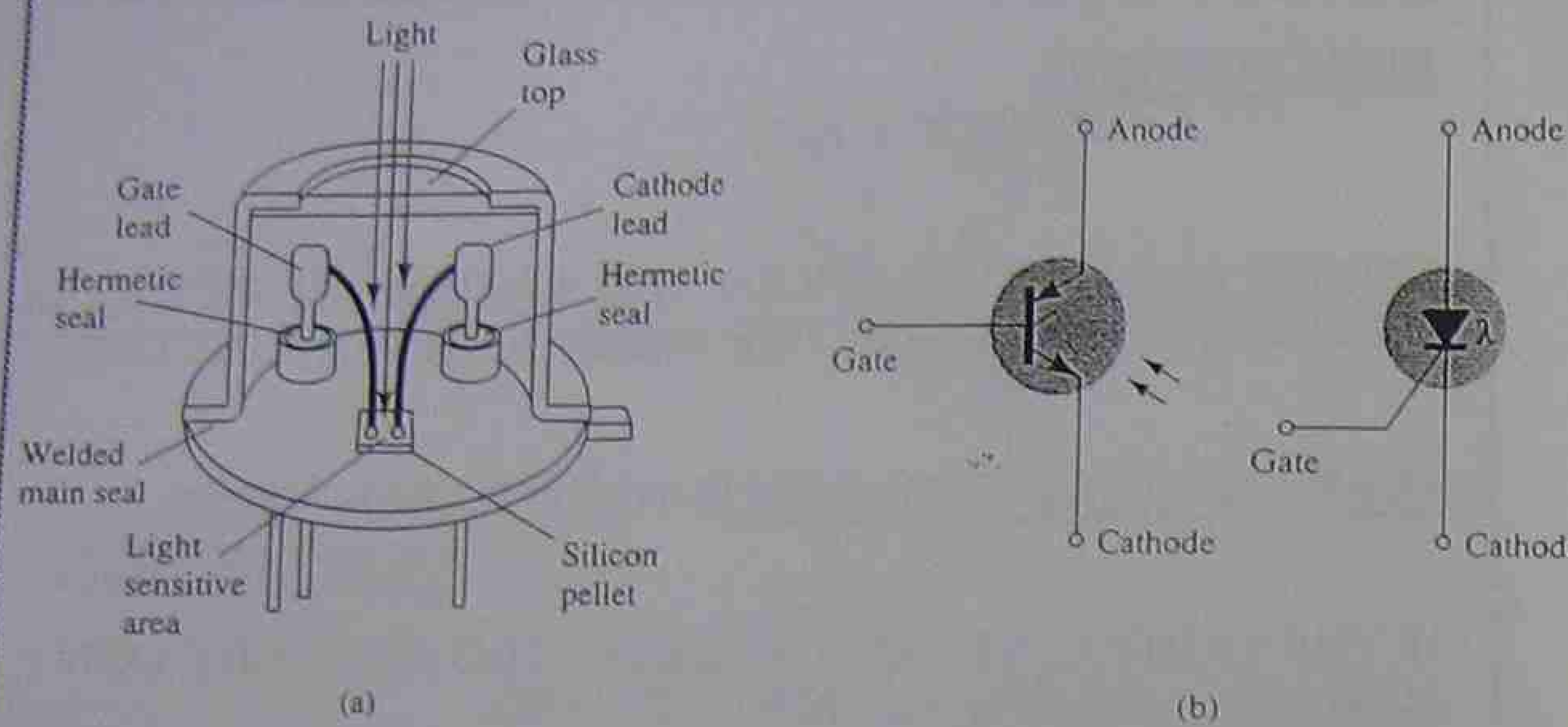


Figure 21.24 Light-activated SCR (LASCR): (a) basic construction; (b) symbols.

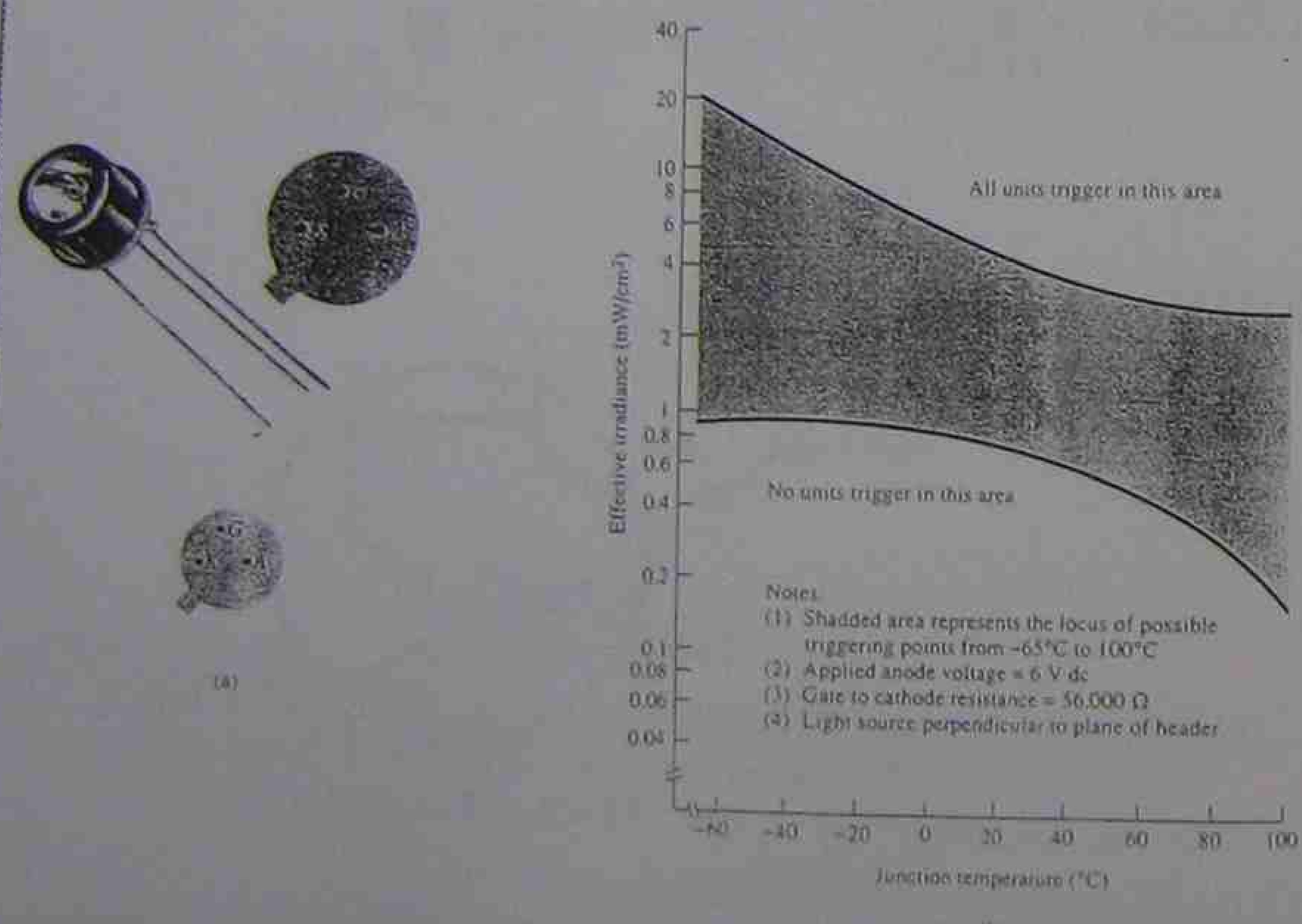


Figure 21.25 LASCR: (a) appearance and terminal identification; (b) light-triggering characteristics. (Courtesy General Electric Company.)

Topic - Photo SCR -LASCR-. (CONT.)

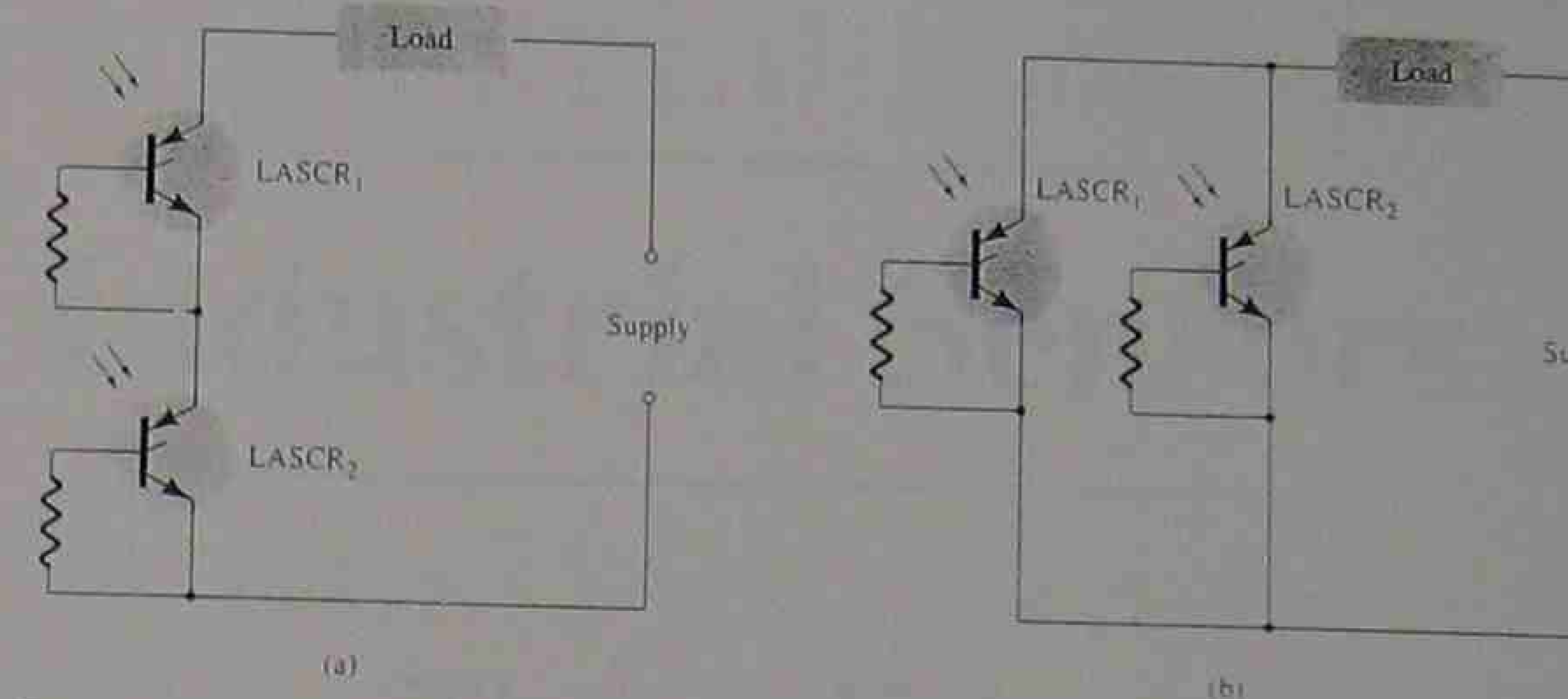


Figure 21.26 LASCR optoelectronic logic circuitry: (a) AND gate: input to $LASCR_1$ and $LASCR_2$ required for energization of the load; (b) OR gate: input to either $LASCR_1$ or $LASCR_2$ will energize the load.

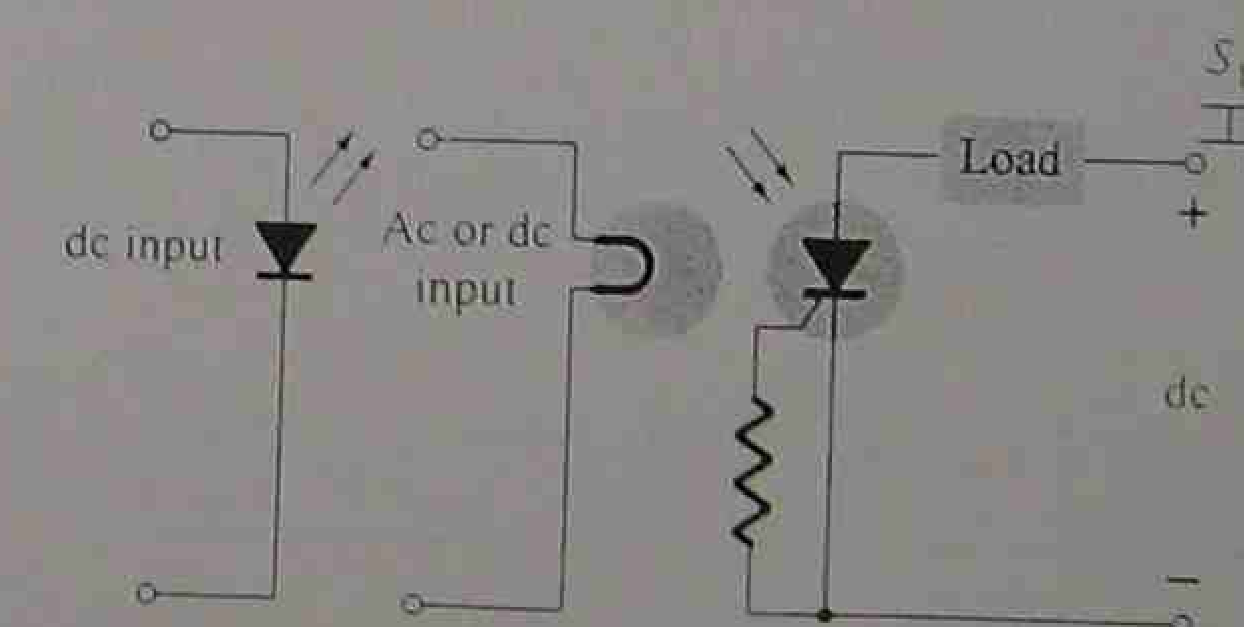


Figure 21.27 Latching relay. (Courtesy Powerex Inc.)

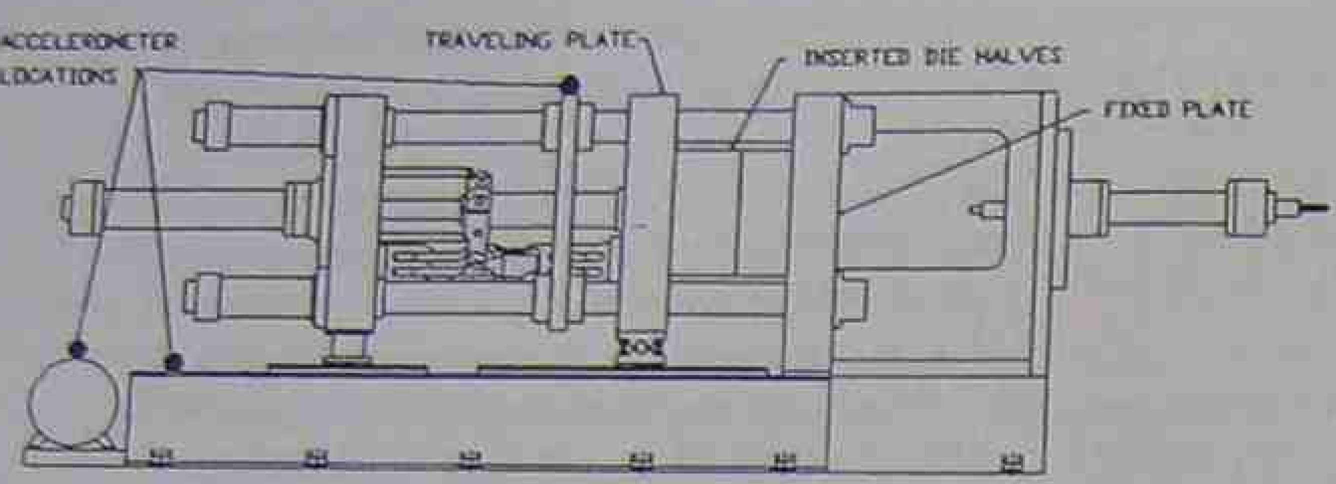


Figure 5-22 Production machine outfitted with an accelerometer for troubleshooting. (Adapted from Wayne Alofs and James R. Carstens, *Mechanical Maintenance and Evaluation of Die Casting Machines*, copyright 1987. Reprinted by permission of the North American Die Casting Association, River Grove, IL.)

Figure 5-23 Using accelerometer probe to detect vibrations in an electric motor.

Figure 5-24 Crash testing a vehicle equipped with accelerometers. (Courtesy of Chrysler Corp., Detroit, MI.)

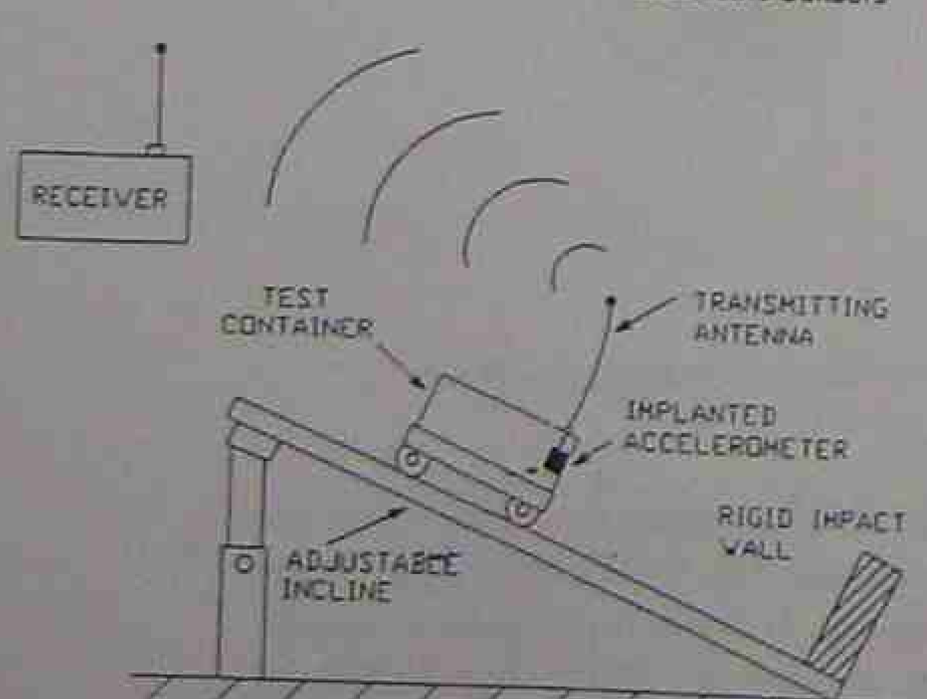


Figure 5-25 Shipping simulator used for the testing of product shipping methods.

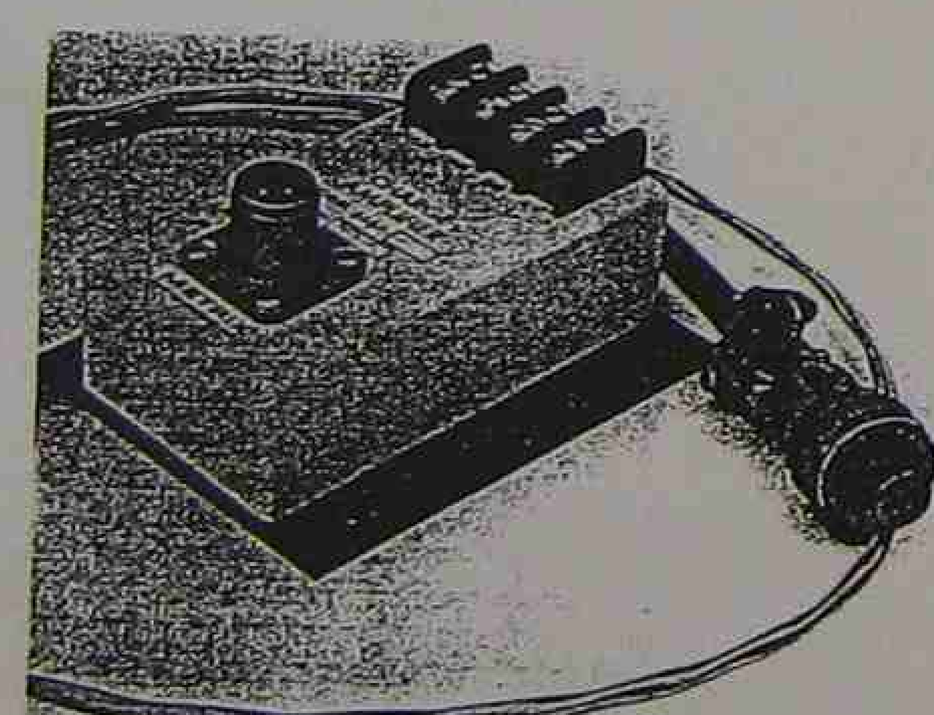


Figure 5-26 Charge amplifier used in conjunction with piezoelectric accelerometers. (Courtesy of Metrix Instrument Co., Houston, TX.)

Sydney Institute of Technology

School of Electrotechnology

Industrial Sensors

Theory Notes

Area ____________________

Topic ____________________

Session No.

For further information on this module, or this subject contact Jim Hafford, (02) 217 3620, Bld. K. Ultimo, S.I.T.

ANSI T/C	Symbol Single	Generic and Trade Names	Color Coding: Single	Color Coding: Overall T/C Wire	Color Coding: Overall Extension Grade Wire	Magnetic Yes	Magnetic No	Maximum Usful Temp. Range	EMF (MV) Over Useful Temp. Range	Average sensitivity $\mu V/°C$	Environment (Bare Wire)
T	TP TN	Copper Constantan, Cupron, Advance	Blue Red	Brown	Blue		X X	°F −328 – 662 °C −200 – 350	−5.602 – 17.816	40.5	Mild oxidizing, reducing. Vacuum or inert. Good where moisture is present.
J	JP JN	Iron Constantan, Cupron, Advance	White Red	Brown	Black	X	 X	°F 32 – 1382 °C 0 – 750	0 – 42.283	52.6	Reducing. Vacuum, inert. Limited use in oxidizing at high temperatures. Not recommended for low temps.
E	EP EN	Chromel, Tophel, T[1] Thermokanthal KP Constantan, Cupron, Advance	Purple Red	Brown	Purple		X X	°F −328 – 1652 °C −200 – 900	−8.824 – 68.783	67.9	Oxidizing or inert. Limited uses in vacuum or reducing.
K	KP KN	Chromel, Tophel, T[1] Thermokanthal KP Alumel, Nial T[2] Thermokanthal KN	Yellow Red	Brown	Yellow	 X	X	°F −328 – 2282 °C −200 – 1250	−5.973 – 50.633	38.8	Clean oxidizing and inert. Limited in Vacuum or reducing.
S	SP SN	Platinum 10% rhodium Pure platinum	Black Red		Green		X X	°F 32 – 2642 °C 0 – 1450	0 – 14.973	10.6	Oxidizing or inert. Atmos. Do not insert in metal tubes. Beware of contaminations.
R	RP RN	Platinum 13% rhodium Pure platinum	Black Red		Green		X X	°F 32 – 2642 °C 0 – 1450	0 – 16.741	12.0	
B	BP BN	Platinum 30% rhodium Platinum 6% rhodium	Gray Red		Gray		X X	°F 32 – 3092 °C 0 – 1700	0 – 12.426	7.6	
C*	CP* CN*	Tungsten 5% rhenium Tungsten 26% rhenium	White/red trace Red		White/red trace		X X	°F 32 – 4208 °C 0 – 2320	0 – 37.066	16.6	Vacuum, inert, hydrogen atmospheres. Beware of embrittlement.
G*	GP* GN*	Tungsten Tungsten 26% rhenium	White/blue trace Red		White/blue trace		X X	°F 32 – 4208 °C 0 – 2320	0 – 38.564	16.0	
D*	DP* DN*	Tungsten 3% rhenium Tungsten 25% rhenium	White/yellow trace Red		White/yellow trace		X X	°F 32 – 4208 °C 0 – 2320	0 – 39.506	17.0	

* Not ANSI symbol

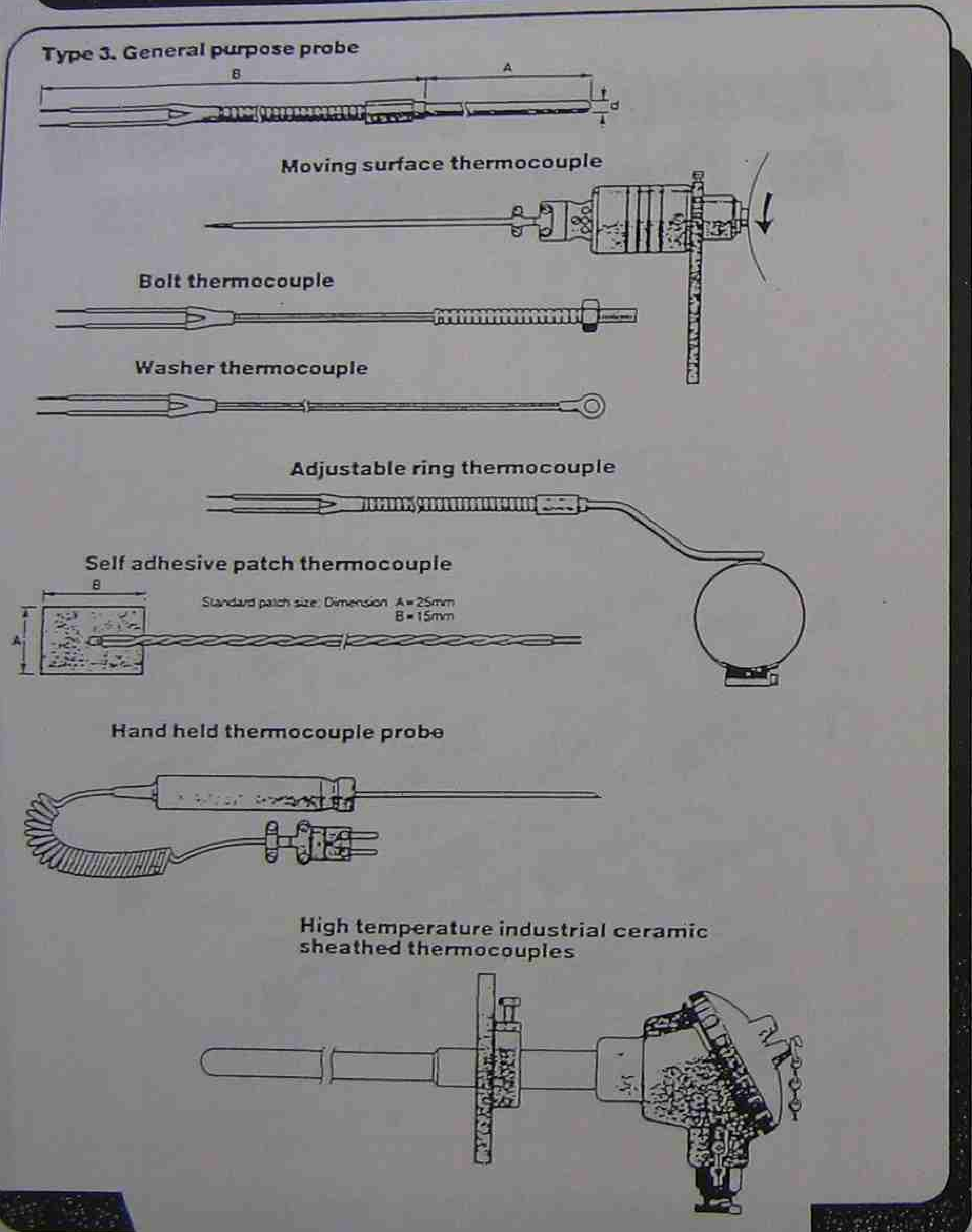
FIGURE 7–10
Characteristics of Standard Thermocouple Types

International Colour Coding for Thermocouple Wires

CODE	AMERICAN to ANSI/MC96.1	BRITISH to bs 1843	GERMAN to DIN 43714	JAPANESE to JIS C 1610-1981	FRENCH to NF C 42-323
K	+YELLOW −RED	+BROWN −BLUE	+RED −GREEN	+RED −WHITE	+YELLOW −PERPLE
T	+BLUE −RED	+WHIT −BLUE	+RED −BROWN	+RED −WHITE	+YELLOW −BLUE
J	+WHITE −RED	+YELLOW −BLUE	+RED −BLUE	+RED −WHITE	+YELLOW −BLACK
E	+PERPLE −RED	+BROWN −BLUE	+RED −BLACK	+RED −WHITE	
N	+BROWN −RED	+BROWN −BLUE			

Industrial Sensors.

Topic -



A selection of thermocouple types (courtesy TC Ltd).

THERMOCOUPLE REVERSAL AND DOUBLE REVERSAL.

This is an installation problem where incorrect thermocouple and lead identification results in polarity reversal of connections - at the thermocouple head or/and at the lead connections to the back of the instrument.

(a) for single reversal
at the thermocouple head OR lead connections
to the back of the instrument

(b) for double reversal
at the thermocouple head AND lead
connections to the back of the instrument

The double reversal will cause the "installed" instrument to indicate high or low after being correctly calibrated.

A single reversal will cause the pointer to drive downscale with increasing temperature.

To check for reversal at the instrument disconnect the leads at the thermocouplehead. Join/hold/twist the lead ends together to form a junction. Application of heat at this junction should drive the instrument pointer upscale if lead connections are correct.

ASSIGNMENT.

Thermocouple tables

The thermocouple tables simply give the voltage that results for a particular type of thermocouple when the reference junctions are at a particular reference temperature, and the measurement junction is at a temperature of interest. Referring to the tables, for example, we see that for a type J thermocouple at 210°C with a 0°C reference, the voltage is

$$V(210°C) = 11.34 \text{ mV} \qquad \text{(type J, 0°C ref.)}$$

Conversely, if we measured a voltage of 4.768 mV with a type S and a 0°C reference, we find from the table

$$T(4.768 \text{ mV}) = 555°C \qquad \text{(type S, 0°C ref.)}$$

In most cases, the measured voltage does not fall exactly on a table value as in this case. When this happens, it is necessary to *interpolate* between table values that bracket the desired value. In general, the value of temperature can be found using the following interpolation equation:

$$T_M = T_L + \left[\frac{T_H - T_L}{V_H - V_L}\right](V_M - V_L) \tag{4.14}$$

The measured voltage V_M lies between a higher voltage V_H and a lower voltage V_L, which are in the tables. The temperatures corresponding to these voltages are T_H and T_L, respectively, as shown in Example 4.10.

Example 4.10

A voltage of 23.72 mV is measured with a type K thermocouple (TC) at a 0°C reference. Find the temperature of the measurement junction.

Solution From the table we find that $V_M = 23.72$ lies between $V_L = 23.63$ mV and $V_H = 23.84$ mV with corresponding temperatures of $T_L = 570°C$ and $T_H = 575°C$, respectively. The junction temperature is found from Equation (4.14).

The reverse situation, although not as common in practice, may occur when the voltage for a particular temperature T_M, which is not in the table, is desired. Again, an interpolation equation can be used, such as

$$V_M = V_L + \left[\frac{V_H - V_L}{T_H - T_L}\right](T_M - T_L) \tag{4.15}$$

where all terms are as defined for Equation (4.14).

Example 4.11

Find the voltage of a type J TC with a 0°C reference if the junction temperature is −172°C.

Solution We do not let the signs bother us but merely apply the interpolation relation directly. From the tables, we see that the junction temperature lies between a high (algebraically) $T_H = -170°C$ and a low $T_L = -175°C$. The corresponding voltages are $V_H = -7.12$ mV, $V_L = -7.27$ mV. The TC voltage will be

(4.15)

Change of table reference

It has already been pointed out that thermocouple tables are prepared for a particular junction temperature. It is possible to use these tables with a thermocouple (TC) that has a different reference temperature by an appropriate shift in the table scale. The key point to remember is that the voltage is proportional to the difference between the reference and measurement junction temperature. Thus, if a new reference is greater than the table reference, all voltages of the table will be less for this TC. The amount less will be just the voltage of the new reference as found on the table. Perhaps a few examples are in order here. Suppose we have a type J TC with a 30°C reference. On the 0°C reference table, a type J at 30°C will produce 1.54 mV. This means that any temperature with this TC will generate a voltage 1.54 mV less than those in the table. Thus, referring to the table

$$400°C \text{ results in } V = 21.85 - 1.54 = 20.31 \text{ mV (type J, 30°C)}$$

150°C results in $V = 8.00 - 1.54 = 6.46$ mV (type J, 30°C)

-90°C results in $V = -4.21 - 1.54 = -5.75$ mV (type J, 30°C)

In a similar fashion, if the new reference is lower than the reference, all of the table voltages will be larger. For example, consider a type K thermocouple with a reference at -26°C. First, by interpolation, we find the voltage that this corresponds to on the 0°C reference tables

$$V(-26°\mathrm{C}) = -1.14 + \frac{-0.95 + 1.14}{-25 + 30}(-26 + 30)$$

$$V(-26°\mathrm{C}) = -0.98 \text{ mV (type K, 0°C ref)}$$

Thus, every voltage on the table must be increased by 0.98 mV, so

400°C results in $V =$

150°C results in $V =$

-90°C results in $V =$

In effect, we are sliding the curves of TC voltage versus temperature along the temperature axis to give a zero voltage at the reference being used. This is shown in Figure 4.10. The shifts made, as in the previous examples, are not exact because

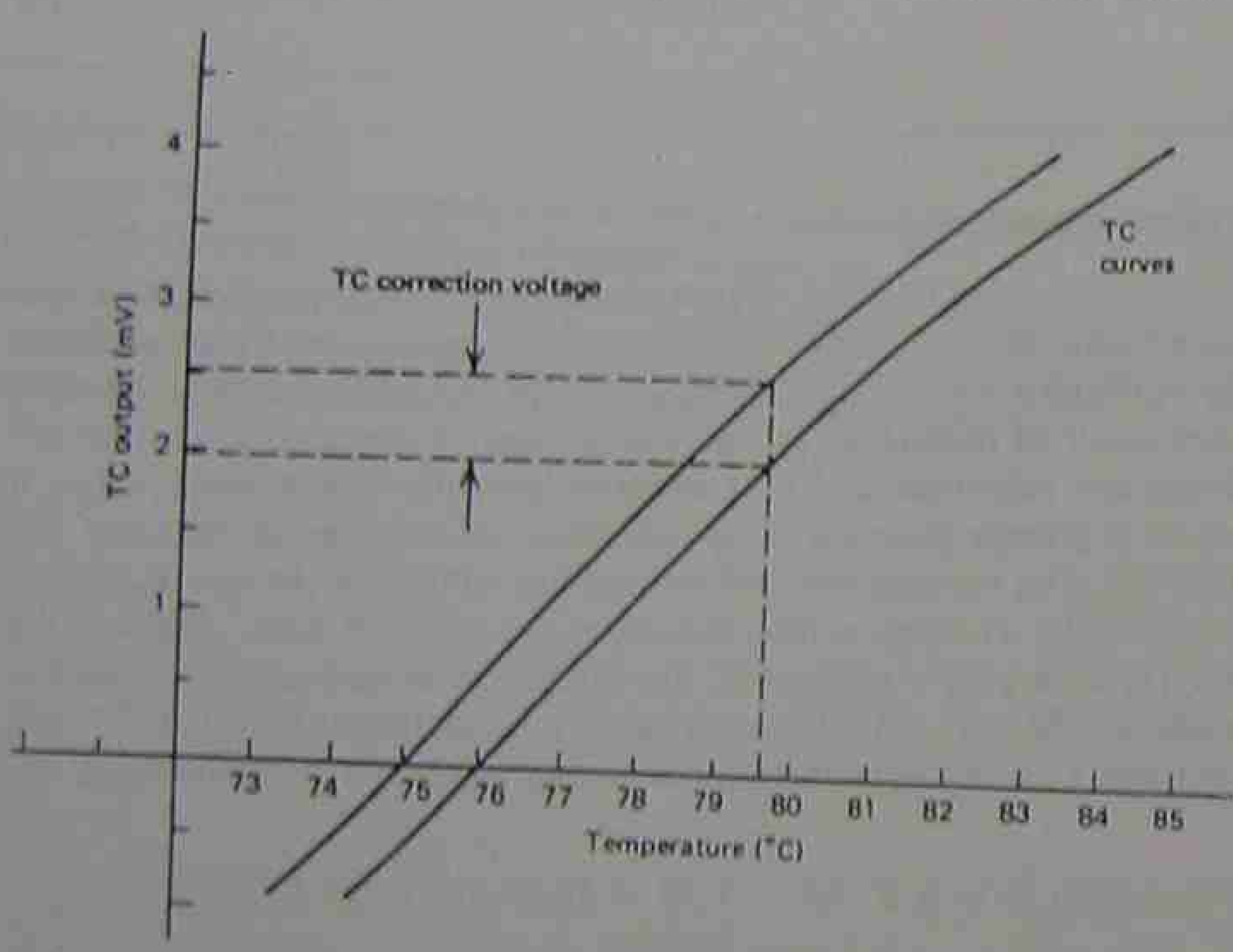


Figure 4.10 When a TC reference is changed, a correction voltage must be applied. If the tables are for 75°C and the actual reference is 76°C, a correction of 0.5 mV is needed.

of the dependence on temperature of the metallic thermoelectric constants. If a very large difference in temperature exists between the table reference temperature and the reference being used, inaccuracies will probably exist.

TYPE K THERMOCOUPLES

TEMPERATURES IN DEGREES C (IPTS 1968). REFERENCE JUNCTION AT 0 C

THERMOELECTRIC VOLTAGE IN ABSOLUTE MILLIVOLTS

DEG C	0	1	2	3	4	5	6	7	8	9	10
[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]

TYPE K THERMOCOUPLES

TEMPERATURES IN DEGREES C (IPTS 1968). REFERENCE JUNCTION AT 0 C

THERMOELECTRIC VOLTAGE IN ABSOLUTE MILLIVOLTS

DEG C	0	1	2	3	4	5	6	7	8	9	10
[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]

THERMOCOUPLE TEMPERATURE CONTROL CIRCUIT

from Maloney "Devices & Systems"
p464-p4655

Brief circuit explanation

Figure 12.4a is the set point , temperature detection and error signal pre amplifier

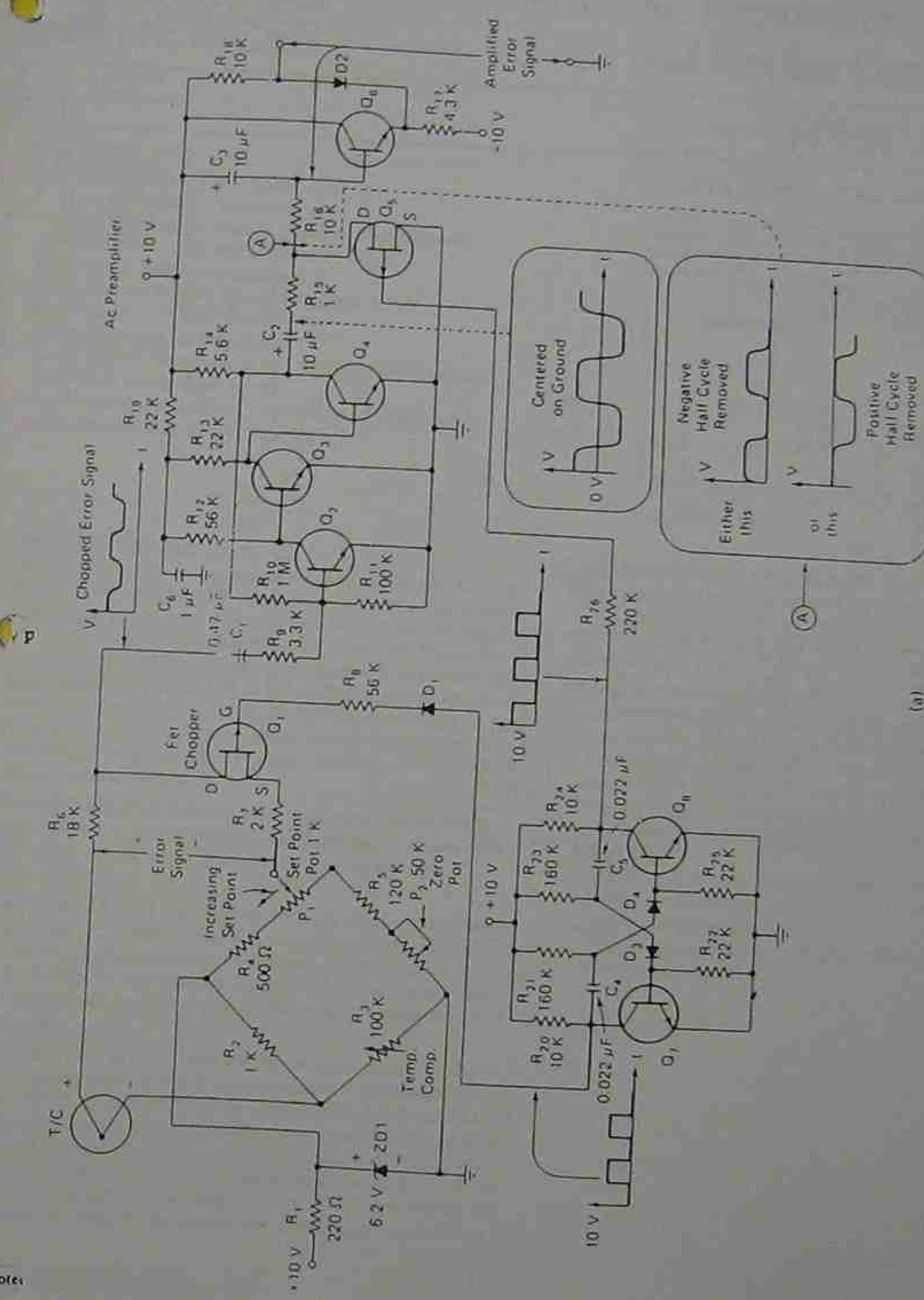
Figure 12.4b is the valve position control circuitry which includes the proportional and integral control of the process.

Figure 12.4a

- The temperature detection is via the thermocouple T/C.
- The set point of the control is determined by the position of the pot P1. P1 is the dial controlled by the operator.
- The wheatstone bridge is initially balanced when the voltages at the top of R5 & R3 are equal. The error value can be provided by the variation in the set point and the output of the T/C. Once the set point is established the thermocouple is the automatic input control error.
- Transistors Q7 & Q8 for part of an astable oscillator. The output from Q7 is used to chop the error signal into a square wave via the FET Q1.
- The transistor Q2 Q3 & Q4 amplify the error signal.
- The FET Q5 is used to remove the positive or negative depending on the error signal polarity.
- The amplified error signal is fed into the valve position control. Figure 12.4b

Figure 12.4b

- Op-Amp 1 is a summing amplifier. It sums the proportional control (feedback from the valve position indicator pot P6) , Integral control (reset control which returns the process completely to the set value) and the amplified error signal from the thermocouple control circuit.
- OP-Amp 2 is a non inverting amplifier to the output stage.
- Op-Amp3 is an integrator (it converts a square wave into a ramp)
  Op-Amp4 is an inverter.
- Transistors Q9 & Q10 are the output driver to the valve motor relay R1
  Transistors Q11 & Q12 are the output drivers to the valve motor relay R2
- The relays R1 & R2 are selected dependant on the polarity of the output voltage Vout2
- The motor is a capacitor start and run split single phase motor. The relays R1 & R2 connect the capacitor C5 in series with the appropriate winding to cause the motor to run in the desired direction.



(a)

Figure 12-4. A thermocouple temperature control circuit. (a) The thermocouple bridge input, chopper, preamplifier, and demodulator. (b) The proportional plus integral control circuit which positions the valve.

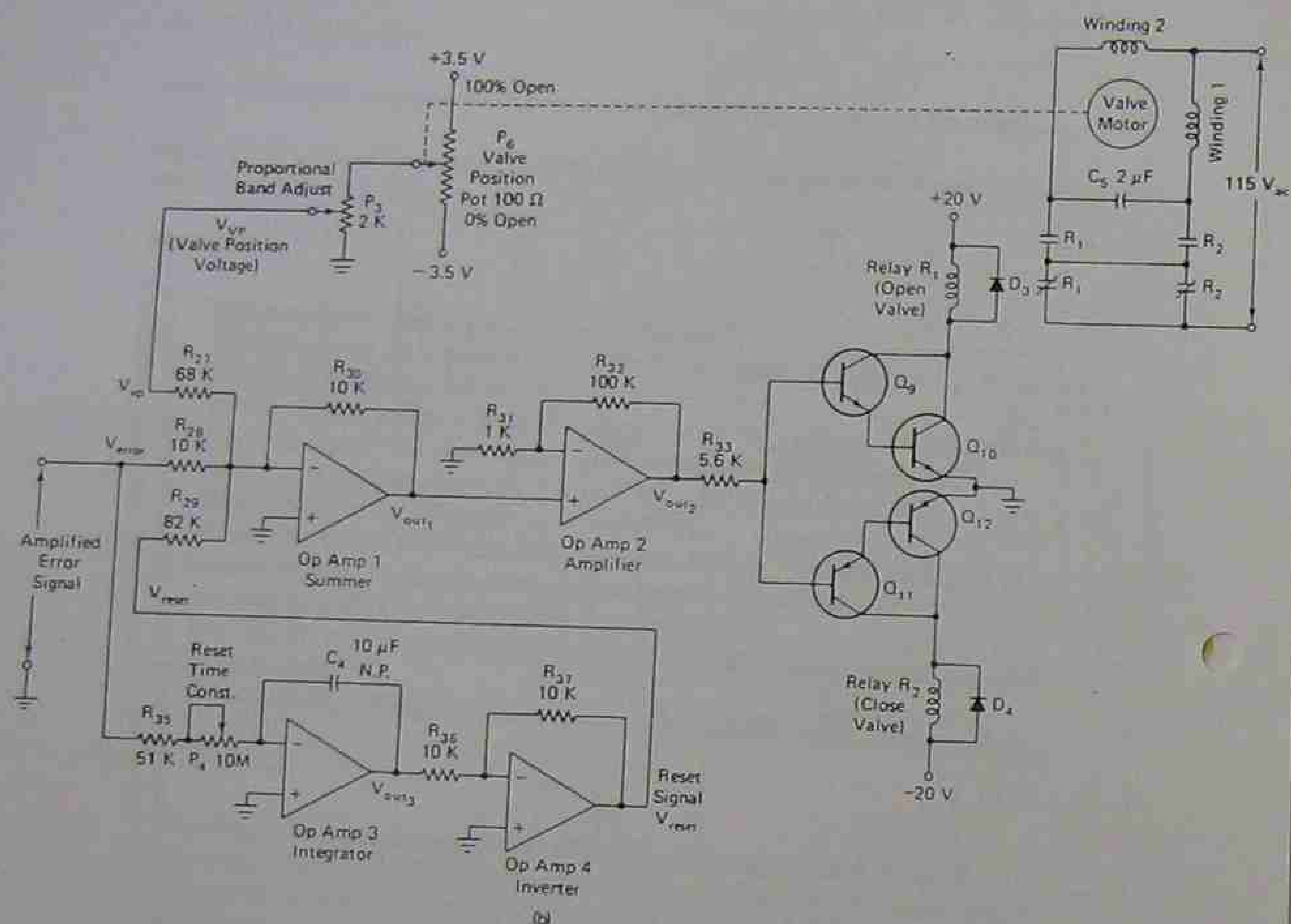


Figure 12-4. (Cont.)

ANALOG DEVICES

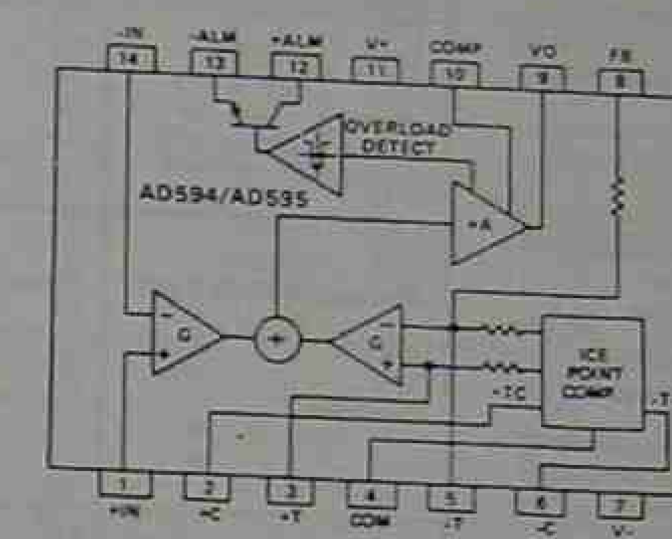
Monolithic Thermocouple Amplifier with Cold Junction Compensation

AD594*/AD595*

FEATURES

- Pretrimmed for Type J (AD594) or Type K (AD595) Thermocouples
- Can Be Used with Type T Thermocouple Inputs
- Low Impedance Voltage Output: 10mV/°C
- Built-In Ice Point Compensation
- Wide Power Supply Range: +5V to ±15V
- Low Power: <1mW typical
- Thermocouple Failure Alarm
- Laser Wafer Trimmed to 1°C Calibration Accuracy
- Set-Point Mode Operation
- Self-Contained Celsius Thermometer Operation
- High Impedance Differential Input

AD594/AD595 BLOCK DIAGRAM



PRODUCT DESCRIPTION

The AD594/AD595 is a complete instrumentation amplifier and thermocouple cold junction compensator on a monolithic chip. It combines an ice point reference with a precalibrated amplifier to produce a high level (10mV/°C) output directly from a thermocouple signal. Pin-strapping options allow it to be used as a linear amplifier-compensator or as a switched output set-point controller using either fixed or remote set-point control. It can be used to amplify its compensation voltage directly, thereby converting it to a stand-alone Celsius transducer with a low-impedance voltage output.

The AD594/AD595 includes a Thermocouple Failure Alarm that indicates if one or both thermocouple leads become open. The alarm output has a flexible format which includes TTL drive capability.

The AD594/AD595 can be powered from a single ended supply (including +5V) and by including a negative supply, temperatures below 0°C can be measured. To minimize self-heating, an unloaded AD594/AD595 will typically operate with a total supply current of 160μA, but is also capable of delivering in excess of ±5mA to a load.

The AD594 is precalibrated by laser wafer trimming to match the characteristic of type J (iron-constantan) thermocouples and the AD595 is laser trimmed for type K (chromel-alumel) inputs. The temperature transducer voltages and gain control resistors are available at the package pins so that the circuit can be recalibrated for other thermocouple types by the addition of two or three resistors. These terminals also allow more precise calibration for both thermocouple and thermometer applications.

*Protected by U.S. Patent No. 4,029,974.

The AD594/AD595 is available in two performance grades. The C and the A versions have calibration accuracies of ±1°C and ±3°C, respectively. Both are designed to be used from 0 to +50°C, and are available in a 14-pin, hermetically sealed, side-brazed ceramic DIP.

PRODUCT HIGHLIGHTS

1. The AD594/AD595 provides cold junction compensation, amplification, and an output buffer in a single IC package.
2. Compensation, zero, and scale factor are all precalibrated by laser wafer trimming (LWT) of each IC chip.
3. Flexible pin-out provides for operation as a set-point controller or a stand-alone temperature transducer calibrated in degrees Celsius.
4. Operation at remote application sites is facilitated by low quiescent current and a wide supply voltage range of +5V to dual supplies spanning 30V.
5. Differential input rejects common-mode noise voltage on the thermocouple leads.

SPECIFICATIONS (@ +25°C and V_S = 5V, Type J (AD594), Type K (AD595) Thermocouple, unless otherwise noted)

Model	AD594A Min	AD594A Typ	AD594A Max	AD594C Min	AD594C Typ	AD594C Max	AD595A Min	AD595A Typ	AD595A Max	AD595C Min	AD595C Typ	AD595C Max	Units
ABSOLUTE MAXIMUM RATINGS													
$+V_S$ to $-V_S$			36			36			36			36	Volts
Common-Mode Input Voltage	$-V_S-0.15$		$+V_S$	$-V_S-0.15$		$+V_S$	$-V_S-0.15$		$+V_S$	$-V_S-0.15$		$+V_S$	Volts
Differential Input Voltage	$-V_S$		$+V_S$	$-V_S$		$+V_S$	$-V_S$		$+V_S$	$-V_S$		$+V_S$	Volts
Alarm Voltages													
+ALM	$-V_S$		$-V_S+36$	$-V_S$		$-V_S+36$	$-V_S$		$-V_S+36$	$-V_S$		$-V_S+36$	Volts
−ALM	$-V_S$		$+V_S$	$-V_S$		$+V_S$	$-V_S$		$+V_S$	$-V_S$		$+V_S$	Volts
Operating Temperature Range	−55		+125	−55		+125	−55		+125	−55		+125	°C
Output Short Circuit to Common	Indefinite			Indefinite			Indefinite			Indefinite			
TEMPERATURE MEASUREMENT (Specified Temperature Range 0 to +50°C)													
Calibration Error at +25°C[1]			**±3**			**±1**			**±3**			**±1**	°C
Stability vs. Temperature[2]			**±0.05**			**±0.025**			**±0.05**			**±0.025**	°C/°C
Gain Error			**±1.5**			**±0.75**			**±1.5**			**±0.75**	%
Nominal Transfer Function		10			10			10			10		mV/°C
AMPLIFIER CHARACTERISTICS													
Closed Loop Gain[3]		193.4			193.4			247.3			247.3		
Input Offset Voltage		(Temperature in °C) × 51.70μV/°C			(Temperature in °C) × 51.70μV/°C			(Temperature in °C) × 44.44μV/°C			(Temperature in °C) × 44.44μV/°C		μV
Input Bias Current		0.1			0.1			0.1			0.1		μA
Differential Input Range	−10		+50				−10		+50	−10		+50	mV
Common Mode Range	−4 to $(-V_S-0.15)$		$+V_S$	−4 to $(-V_S-0.15)$		$+V_S$	−4 to $(-V_S-0.15)$		$+V_S$	−4 to $(-V_S-0.15)$		$+V_S$	Volts
Common Mode Sensitivity–RTO			**10**			**10**			**10**			**10**	mV/V
Power Supply Sensitivity–RTO			**10**			**10**			**10**			**10**	mV/V
Output Voltage Range													
Dual Supplies	$-V_S+2.5$		$+V_S-2$	$-V_S+2.5$		$+V_S-2$	$-V_S+2.5$		$+V_S-2$	$-V_S+2.5$		$+V_S-2$	Volts
Single Supply	0		$+V_S-2$	0		$+V_S-2$	0		$+V_S+2$	0		$+V_S-2$	Volts
Usable Output Current[4]		±5			±5			±5			±5		mA
3dB Bandwidth		15			15			15			15		kHz
ALARM CHARACTERISTICS													
$V_{CE(SAT)}$ at 2mA		0.3			0.3			0.3			0.3		Volts
Leakage Current			±1			±1			±1			±1	μA max
Operating Voltage at −ALM			$+V_S-4$			$+V_S-4$			$+V_S-4$			$+V_S-4$	Volts
Short Circuit Current		20			20			20			20		mA
POWER REQUIREMENTS													
Specified Performance		$+V_S=5, -V_S=0$			$+V_S=5, -V_S=0$			$+V_S=5, -V_S=0$			$+V_S=5, -V_S=0$		Volts
Operating		$+V_S$ to $-V_S \le 30$			$+V_S$ to $-V_S \le 30$			$+V_S$ to $-V_S \le 30$			$+V_S$ to $-V_S \le 30$		Volts
Quiescent Current (No Load)													
$+V_S$		160	300		160	300		160	300		160	300	μA
$-V_S$		100			100			100			100		μA
PACKAGE OPTION[5] (D14A)		AD594AD			AD594CD			AD595AD			AD595CD		

NOTES

[1] Calibrated for minimum error at +25°C using a thermocouple sensitivity of 51.7μV/°C. Since a J type thermocouple deviates from this straight line approximation, the AD594 will normally read 3.1mV when the measuring junction is at 0°C. The AD595 will similarly read 2.7mV at 0°C.

[2] Defined as the slope of the line connecting the AD594/AD595 errors measured at 0°C and 50°C ambient temperature.

[3] Pin 8 shorted to pin 9.

[4] Current Sink Capability in single supply configuration is limited to current drawn to ground through a 50kΩ resistor at output voltages below 2.5V.

[5] See Section 19 for package outline information.

Specifications subject to change without notice.

Specifications shown in boldface are tested on all production units at final electrical test. Results from those tests are used to calculate outgoing quality levels. All min and max specifications are guaranteed, although only those shown in boldface are tested on all production units.

ORDERING GUIDE

Model	Maximum Cal. Error
AD594AD	±3°C
AD594CD	±1°C
AD595AD	±3°C
AD595CD	±1°C

Thermocouple Temperature °C	Type J Voltage mV	AD594 Output mV	Type K Voltage mV	AD595 Output mV	Thermocouple Temperature °C	Type J Voltage mV	AD594 Output mV	Type K Voltage mV	AD595 Output mV
−200	−7.890	−1523	−5.891	−1454	500	27.388	5300	20.640	5107
−180	−7.402	−1428	−5.550	−1370	520	28.511	5517	21.493	5318
−160	−6.821	−1316	−5.141	−1269	540	29.642	5736	22.346	5529
−140	−6.159	−1188	−4.669	−1152	560	30.782	5956	23.198	5740
−120	−5.426	−1046	−4.138	−1021	580	31.933	6179	24.050	5950
−100	−4.632	−893	−3.553	−876	600	33.096	6404	24.902	6161
−80	−3.785	−729	−2.920	−719	620	34.273	6632	25.751	6371
−60	−2.892	−556	−2.243	−552	640	35.464	6862	26.599	6581
−40	−1.960	−376	−1.527	−375	660	36.671	7095	27.445	6790
−20	−.995	−189	−.777	−189	680	37.893	7332	28.288	6998
−10	−.501	−94	−.392	−94	700	39.130	7571	28.128	7206
0	0	3.1	0	2.7	720	40.382	7813	29.965	7413
10	.507	101	.397	101	740	41.647	8058	30.799	7619
20	1.019	200	.798	200	750	42.283	8181	31.214	7722
25	1.277	250	1.000	250	760	–	–	31.629	7825
30	1.536	300	1.203	300	780	–	–	32.455	8029
40	2.058	401	1.611	401	800	–	–	33.277	8232
50	2.585	503	2.022	503	820	–	–	34.095	8434
60	3.115	606	2.436	605	840	–	–	34.909	8636
80	4.186	813	3.266	810	860	–	–	35.718	8836
100	5.268	1022	4.095	1015	880	–	–	36.524	9035
120	6.359	1233	4.919	1219	900	–	–	37.325	9233
140	7.457	1445	5.733	1420	920	–	–	38.122	9430
160	8.560	1659	6.539	1620	940	–	–	38.915	9626
180	9.667	1873	7.338	1817	960	–	–	39.703	9821
200	10.777	2087	8.137	2015	980	–	–	40.488	10015
220	11.887	2302	8.938	2213	1000	–	–	41.269	10209
240	12.998	2517	9.745	2413	1020	–	–	42.045	10400
260	14.108	2732	10.560	2614	1040	–	–	42.817	10591
280	15.217	2946	11.381	2817	1060	–	–	43.585	10781
300	16.325	3160	12.207	3022	1080	–	–	44.349	10970
320	17.432	3374	13.039	3327	1100	–	–	45.108	11158
340	18.537	3588	13.874	3434	1120	–	–	45.863	11345
360	19.640	3801	14.712	3641	1140	–	–	46.612	11530
380	20.743	4015	15.552	3849	1160	–	–	47.356	11714
400	21.846	4228	16.395	4057	1180	–	–	48.095	11897
420	22.949	4441	17.241	4266	1200	–	–	48.828	12078
440	24.054	4655	18.088	4476	1220	–	–	49.555	12258
460	25.161	4869	18.938	4686	1240	–	–	50.276	12436
480	26.272	5084	19.788	4896	1250	–	–	50.635	12524

Table I. Output Voltage vs. Thermocouple Temperature (Ambient +25°C, V_S = −5V, +15V)

INTERPRETING AD594/AD595 OUTPUT VOLTAGES

To achieve a temperature proportional output of 10mV/°C and accurately compensate for the reference junction over the rated operating range of the circuit, the AD594/AD595 is gain trimmed to match the transfer characteristic of J and K type thermocouples at 25°C. For a type J output in this temperature range the TC is 51.70μV/°C, while for a type K it is 40.44μV/°C. The resulting gain for the AD594 is 193.4 (10mV/°C divided by 51.7μV/°C) and for the AD595 is 247.3 (10mV/°C divided by 40.44μV/°C). In addition, an absolute accuracy trim induces an input offset to the output amplifier characteristic of 16μV for the AD594 and 11μV for the AD595. This offset arises because the AD594/AD595 is trimmed for a 250mV output while applying a 25°C thermocouple input.

Because a thermocouple output voltage is nonlinear with respect to temperature, and the AD594/AD595 linearly amplifies the compensated signal, the following transfer functions should be used to determine the actual output voltages:

AD594 output = (Type J Voltage + 16μV) × 193.4

AD595 output = (Type K Voltage + 11μV) × 247.3

or conversely:

Type J voltage = (AD594 output / 193.4) − 16μV

Type K voltage = (AD595 output / 247.3) − 11μV

Table I above lists the ideal AD594/AD595 output voltages as a function of Celsius temperature for type J and K ANSI standard thermocouples, with the package and reference junction at 25°C. As is normally the case, these outputs are subject to calibration, gain and temperature sensitivity errors. Output values for intermediate temperatures can be interpolated, or calculated using the output equations and ANSI thermocouple voltage tables referred to zero degrees Celsius. Due to a slight variation in alloy content between ANSI type J and DIN Fe–CuNi thermocouples Table I should not be used in conjunction with European standard thermocouples. Instead the transfer function given previously and a DIN thermocouple table should be used. ANSI type K and DIN NiCr–Ni thermocouples are composed of identical alloys and exhibit similar behavior. The upper temperature limits in Table I are those recommended for type J and type K thermocouples by the majority of vendors.

SINGLE AND DUAL SUPPLY CONNECTIONS

The AD594/AD595 is a completely self-contained thermocouple conditioner. Using a single +5V supply the interconnections shown in Figure 1 will provide a direct output from a type J thermocouple (AD594) or type K thermocouple (AD595) measuring from 0 to +300°C.

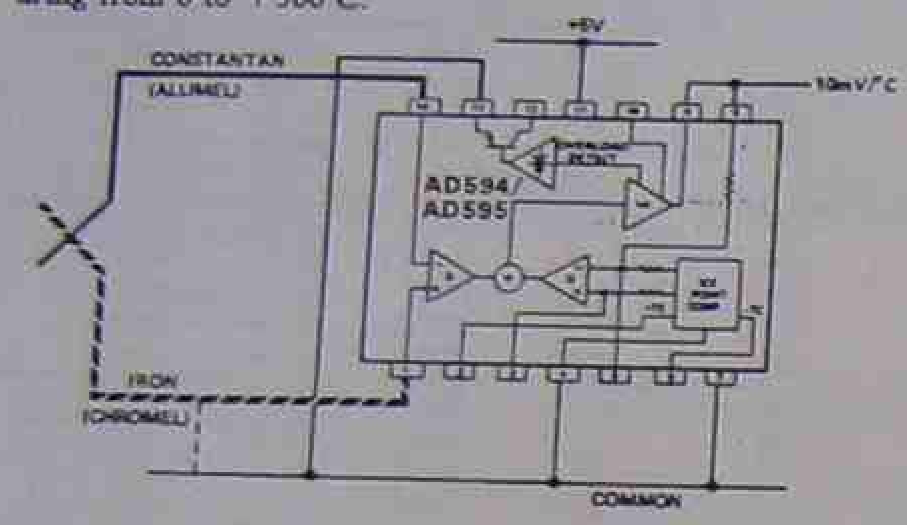


Figure 1. Basic Connection, Single Supply Operation

Any convenient supply voltage from +5V to +30V may be used, with self-heating errors being minimized at lower supply levels. In the single supply configuration the +5V supply connects to pin 11 with the V− connection at pin 7 strapped to power and signal common at pin 4. The thermocouple wire inputs connect to pins 1 and 14 either directly from the measuring point or through intervening connections of similar thermocouple wire type. When the alarm output at pin 13 is not used it should be connected to common or −V. The precalibrated feedback network at pin 8 is tied to the output at pin 9 to provide a 10mV/°C nominal temperature transfer characteristic.

By using a wider ranging dual supply, as shown in Figure 2, the AD594/AD595 can be interfaced to thermocouples measuring both negative and extended positive temperatures.

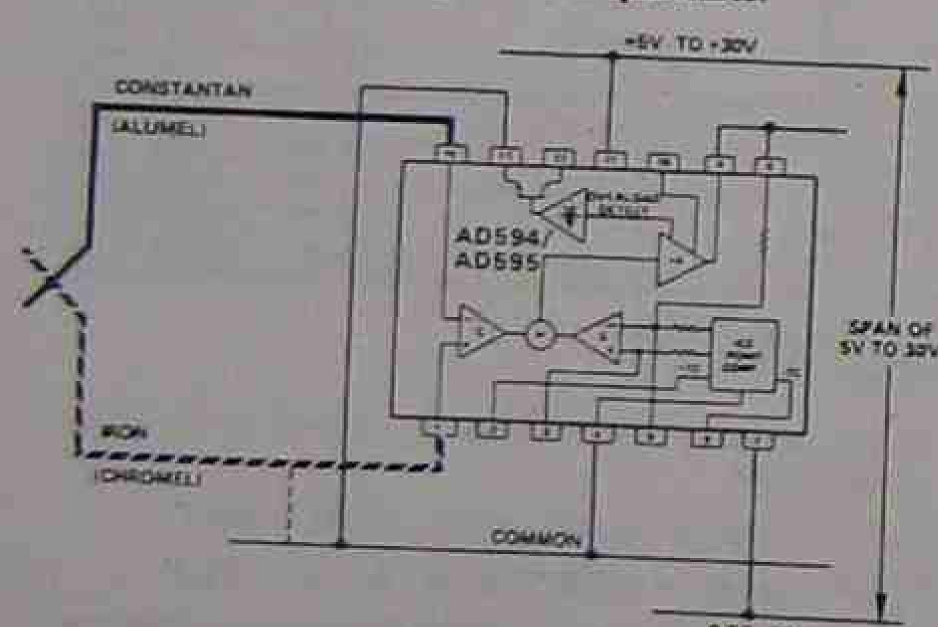


Figure 2. Dual Supply Operation

With a negative supply the output can indicate negative temperatures and drive grounded loads or loads returned to positive voltages. Increasing the positive supply from 5V to 15V extends the output voltage range well beyond the 750°C temperature limit recommended for type J thermocouples (AD594) and the 1250°C for type K thermocouples (AD595).

Common-mode voltages on the thermocouple inputs must remain within the common-mode range of the AD594/AD595, with a return path provided for the bias currents. If the thermocouple is not remotely grounded, then the dotted line connections in Figures 1 and 2 are recommended. A resistor may be needed in this connection to assure that common mode voltages induced in the thermocouple loop are not converted to normal mode.

THERMOCOUPLE CONNECTIONS

The isothermal terminating connections of a pair of thermocouple wires forms an effective reference junction. This junction must be kept at the same temperature as the AD594/AD595 for the internal cold junction compensation to be effective.

A method that provides for thermal equilibrium is the printed circuit board connection layout illustrated in Figure 3.

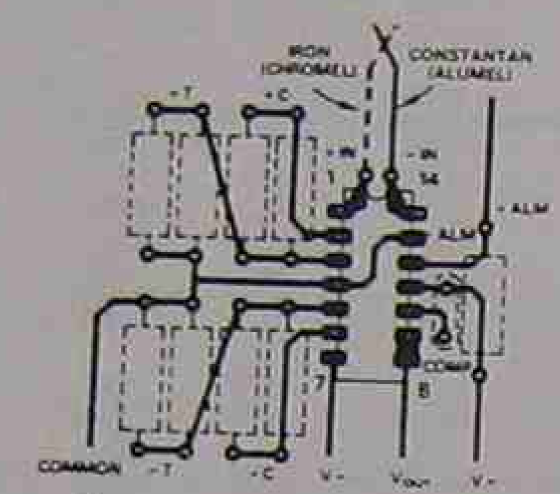


Figure 3. PCB Connections

Here the AD594/AD595 package temperature and circuit board are thermally contacted in the copper printed circuit board tracks under pins 1 and 14. The reference junction is now composed of a copper-constantan (or copper-alumel) connection and copper-iron (or copper-chromel) connection, both of which are at the same temperature as the AD594/AD595.

The printed circuit board layout shown also provides for placement of optional alarm load resistors, recalibration resistors and a compensation capacitor to limit bandwidth.

To ensure secure bonding the thermocouple wire should be cleaned to remove oxidation prior to soldering. Noncorrosive rosin flux is effective with iron, constantan, chromel and alumel and the following solders: 95% tin-5% antimony, 95% tin-5% silver or 90% tin-10% lead.

FUNCTIONAL DESCRIPTION

The AD594 behaves like two differential amplifiers. The outputs are summed and used to control a high-gain amplifier, as shown in Figure 4.

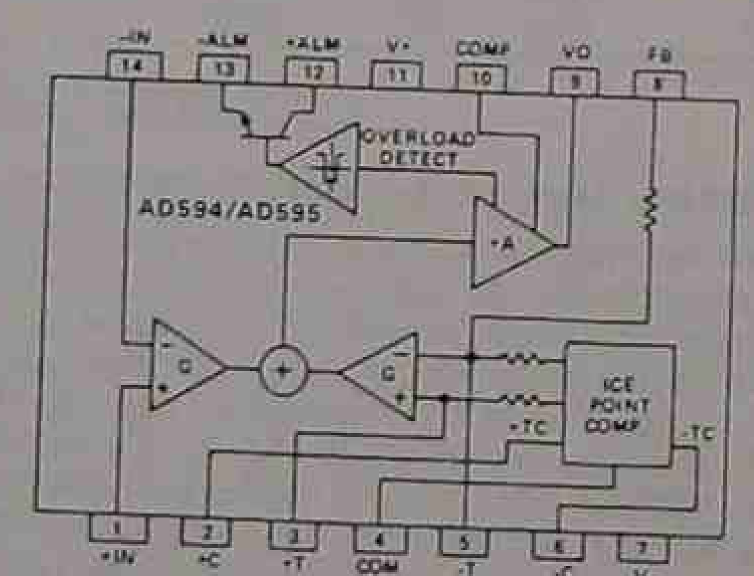


Figure 4. AD594/AD595 Block Diagram

In normal operation the main amplifier output, at pin 9, is connected to the feedback network, at pin 8. Thermocouple signals applied to the floating input stage, at pins 1 and 14, are amplified by gain G of the differential amplifier and are then further amplified by gain A in the main amplifier. The output of the main amplifier is fed back to a second differential stage in an inverting connection. The feedback signal is amplified by this stage and is also applied to the main amplifier input through a summing circuit. Because of the inversion, the amplifier causes

the feedback to be driven to reduce this difference signal to a small value. The two differential amplifiers are made to match and have identical gains, G. As a result, the feedback signal that must be applied to the right-hand differential amplifier will precisely match the thermocouple input signal when the difference signal has been reduced to zero. The feedback network is trimmed so that the effective gain to the output, at pins 8 and 9, results in a voltage of 10mV/°C of thermocouple excitation.

In addition to the feedback signal, a cold junction compensation voltage is applied to the right-hand differential amplifier. The compensation is a differential voltage proportional to the Celsius temperature of the AD594/AD595. This signal disturbs the differential input so that the amplifier output must adjust to restore the input to equal the applied thermocouple voltage.

The compensation is applied through the gain scaling resistors so that its effect on the main output is also 10mV/°C. As a result, the compensation voltage adds to the effect of the thermocouple voltage a signal directly proportional to the difference between 0°C and the AD594/AD595 temperature. If the thermocouple reference junction is maintained at the AD594/AD595 temperature, the output of the AD594/AD595 will correspond to the reading that would have been obtained from amplification of a signal from a thermocouple referenced to an ice bath.

The AD594/AD595 also includes an input open circuit detector that switches on an alarm transistor. This transistor is actually a current-limited output buffer, but can be used up to the limit as a switch transistor for either pull-up or pull-down operation of external alarms.

The ice point compensation network has voltages available with positive and negative temperature coefficients. These voltages may be used with external resistors to modify the ice point compensation and recalibrate the AD594/AD595 as described in the next column.

The feedback resistor is separately pinned out so that its value can be padded with a series resistor, or replaced with an external resistor between pins 5 and 9. External availability of the feedback resistor allows gain to be adjusted, and also permits the AD594/AD595 to operate in a switching mode for set-point operation.

CAUTIONS:

The temperature compensation terminals (+C and −C) at pins 2 and 6 are provided to supply small calibration currents only. The AD594/AD595 may be permanently damaged if they are grounded or connected to a low impedance.

The AD594/AD595 is internally frequency compensated for feedback ratios (corresponding to normal signal gain) of 75 or more. If a lower gain is desired, additional frequency compensation should be added in the form of a 300pF capacitor from pin 10 to the output at pin 9. As shown in Figure 5 an additional 0.01μF capacitor between pins 10 and 11 is recommended.

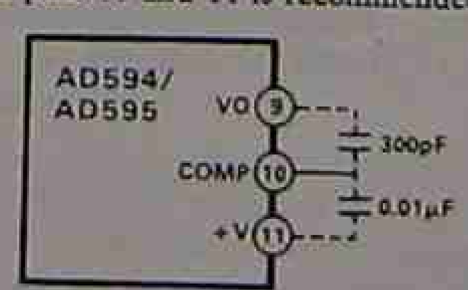


Figure 5. Low Gain Frequency Compensation

RECALIBRATION PRINCIPLES AND LIMITATIONS

The ice point compensation network of the AD594/AD595 produces a differential signal which is zero at 0°C and corresponds to the output of an ice referenced thermocouple at the temperature of the chip. The positive TC output of the circuit is proportional to Kelvin temperature and appears as a voltage at +T. It is possible to decrease this signal by loading it with a resistor from +T to COM, or increase it with a pull-up resistor from +T to the larger positive TC voltage at +C. Note that adjustments to +T should be made by measuring the voltage which tracks it at −T. To avoid destabilizing the feedback amplifier the measuring instrument should be isolated by a few thousand ohms in series with the lead connected to −T.

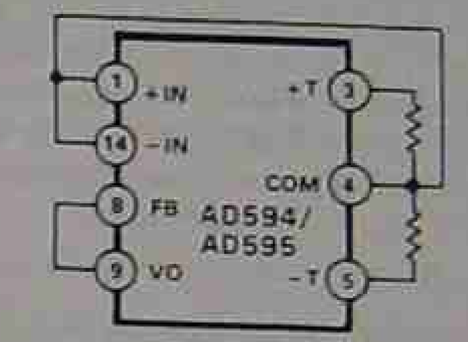


Figure 6. Decreased Sensitivity Adjustment

Changing the positive TC half of the differential output of the compensation scheme shifts the zero point away from 0°C. The zero can be restored by adjusting the current flow into the negative input of the feedback amplifier, the −T pin. A current into this terminal can be produced with a resistor between −C and −T to balance an increase in +T, or a resistor from −T to COM to offset a decrease in +T.

If the compensation is adjusted substantially to accommodate a different thermocouple type, its effect on the final output voltage will increase or decrease in proportion. To restore the nominal output to 10mV/°C the gain may be adjusted to match the new compensation and thermocouple input characteristics. When reducing the compensation the resistance between −T and COM automatically increases the gain to within 0.5% of the correct value. If a smaller gain is required, however, the nominal 47kΩ internal feedback resistor can be paralleled or replaced with an external resistor.

Fine calibration adjustments will require temperature response measurements of individual devices to assure accuracy. Major reconfigurations for other thermocouple types can be achieved without seriously compromising initial calibration accuracy, so long as the procedure is done at a fixed temperature using the factory calibration as a reference. It should be noted that intermediate recalibration conditions may require the use of a negative supply. An example using a type E thermocouple and an AD594 is given on the next page.

EXAMPLE: TYPE E RECALIBRATION – AD594/AD595

Both the AD594 and AD595 can be configured to condition the output of a type E (chromel-constantan) thermocouple. Temperature characteristics of type E thermocouples differ less from type J, than from type K, therefore the AD594 is preferred for recalibration.

While maintaining the device at a constant temperature follow the recalibration steps given here. First, measure the device temperature by tying both inputs to common (or a selected common mode potential) and connecting FB to V_O. The AD594 is now in the stand alone Celsius thermometer mode. For this example assume the ambient is 24°C and the initial output V_O is 240mV. Check the output at V_O to verify that it corresponds to the temperature of the device.

Next, measure the voltage –T at pin 5 with a high impedance DVM (capacitance should be isolated by a few thousand ohms of resistance at the measured terminals). At 24°C the –T voltage will be about 8.3mV. To adjust the compensation of an AD594 to a type E thermocouple a resistor, R1, should be connected between +T and +C, pins 2 and 3, to raise the voltage at –T by the ratio of thermocouple sensitivities. The ratio for converting a type J device to a type E characteristic is:

$$r\ (AD594) = (60.9\mu V/^\circ C)/(51.7\mu V/^\circ C) = 1.18$$

Thus, multiply the initial voltage measured at –T by r and experimentally determine the R1 value required to raise –T to that level. For the example the new –T voltage should be about 9.8mV. The resistance value should be approximately 1.8kΩ.

The zero differential point must now be shifted back to 0°C. This is accomplished by multiplying the original output voltage V_O by r and adjusting the measured output voltage to this value by experimentally adding a resistor, R2, between –C and –T, pins 5 and 6. The target output value in this case should be about 283mV. The resistance value of R2 should be approximately 240kΩ.

Finally, the gain must be recalibrated such that the output V_O indicates the device's temperature once again. Do this by adding a third resistor, R3, between FB and –T, pins 8 and 5. V_O should now be back to the initial 240mV reading. The resistance value of R3 should be approximately 280kΩ. The final connection diagram is shown in Figure 7. An approximate verification of the effectiveness of recalibration is to measure the differential gain to the output. For type E it should be 164.2.

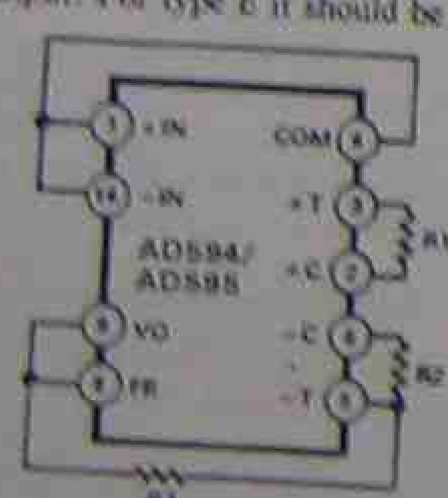


Figure 7. Type E Recalibration

When implementing a similar recalibration procedure for the AD595 the values for R1, R2, R3 and r will be approximately 650Ω, 84kΩ, 93kΩ and 1.51, respectively. Power consumption will increase by about 50% when using the AD595 with type E inputs.

Note that during this procedure it is crucial to maintain the AD594/AD595 at a stable temperature because it is used as the temperature reference. Contact with fingers or any tools not at ambient temperature will quickly produce errors. Radiational heating from a change in lighting or approach of a soldering iron must also be guarded against.

USING TYPE T THERMOCOUPLES WITH THE AD595

Because of the similarity of thermal EMFs in the 0 to 50°C range between type K and type T thermocouples, the AD595 can be directly used with both types of inputs. Within this ambient temperature range the AD595 should exhibit no more than an additional 0.2°C output calibration error when used with type T inputs. The error arises because the ice point compensator is trimmed to type K characteristics at 25°C. To calculate the AD595 output values over the recommended –200 to 350°C range for type T thermocouples, simply use the ANSI thermocouple voltages referred to 0°C and the output equation given on page 3 for the AD595. Because of the relatively large nonlinearities associated with type T thermocouples the output will deviate widely from the nominal 10mV/°C. However, cold junction compensation over the rated 0 to 50°C ambient will remain accurate.

STABILITY OVER TEMPERATURE

Each AD594/AD595 is tested for error over temperature with the measuring thermocouple at 0°C. The combined effects of cold junction compensation error, amplifier offset drift and gain error determine the stability of the AD594/AD595 output over the rated ambient temperature range. Figure 8 shows an AD594/AD595 drift error envelope. The slope of this figure has units of °C/°C.

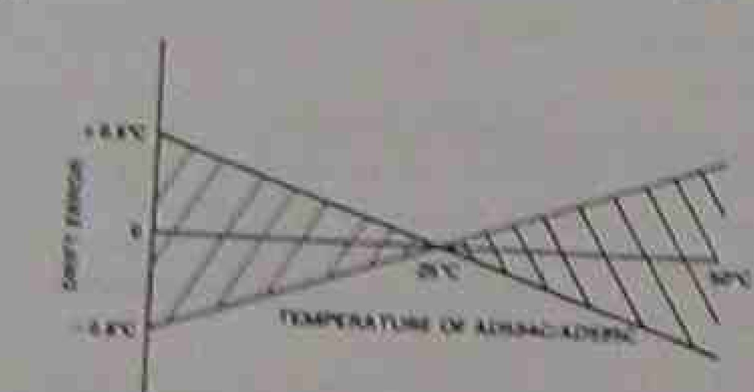


Figure 8. Drift Error vs. Temperature

THERMAL ENVIRONMENT EFFECTS

The inherent low power dissipation of the AD594/AD595 and the low thermal resistance of the package make self-heating errors almost negligible. For example, in still air the chip to ambient thermal resistance is about 80°C/watt. At the nominal dissipation of 800μW the self-heating in free air is less than 0.065°C. Submerged in fluorinert liquid (unstirred) the thermal resistance is about 40°C/watt, resulting in a self-heating error of about 0.032°C.

SET-POINT CONTROLLER

The AD594/AD595 can readily be connected as a set-point controller as shown in Figure 9.

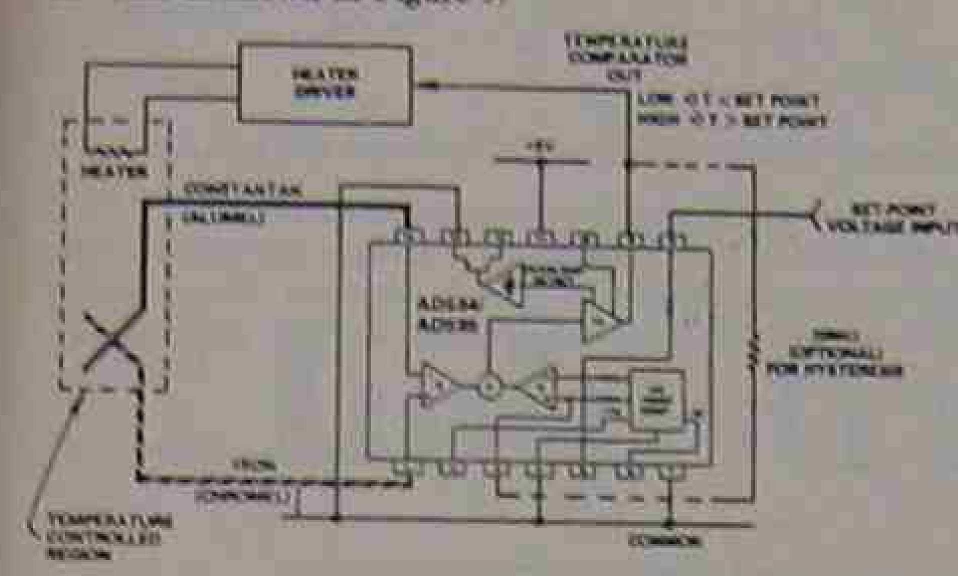


Figure 9. Set-Point Controller

The thermocouple is used to sense the unknown temperature and provide a thermal EMF to the input of the AD594/AD595. The signal is cold junction compensated, amplified to 10mV/°C and compared to an external set-point voltage applied by the user to the feedback at pin 8. Table I lists the correspondence between set-point voltage and temperature, accounting for the nonlinearity of the measurement thermocouple. If the set-point temperature range is within the operating range (–55°C to 125°C) of the AD594/AD595, the chip can be used as the transducer for the circuit by shorting the inputs together and utilizing the nominal calibration of 10mV/°C. This is the centigrade thermometer configuration as shown in Figure 13.

In operation if the set-point voltage is above the voltage corresponding to the temperature being measured the output swings low to approximately zero volts. Conversely, when the temperature rises above the set-point voltage the output switches to the positive limit of about 4 volts with a +5V supply. Figure 9 shows the set-point comparator configuration complete with a heater element driver circuit being controlled by the AD594/AD595 toggled output. Hysteresis can be introduced by injecting a current into the positive input of the feedback amplifier when the output is toggled high. With an AD594 about 200nA into the +T terminal provides 1°C of hysteresis. When using a single 5V supply with an AD594, a 20MΩ resistor from V_O to +T will supply the 200nA of current when the output is forced high (about 4V). To widen the hysteresis band decrease the resistance connected from V_O to +T.

ALARM CIRCUIT

In all applications of the AD594/AD595 the –ALM connection, pin 13, should be constrained so that it is not more positive than (V+) – 4V. This can be most easily achieved by connecting pin 13 to either common at pin 4 or V– at pin 7. For most applications that use the alarm signal, pin 13 will be grounded and the signal will be taken from +ALM on pin 12. A typical application is shown in Figure 10.

In this configuration the alarm transistor will be off in normal operation and the 20k pull up will cause the +ALM output on pin 12 to go high. If one or both of the thermocouple leads are interrupted, the +ALM pin will be driven low. As shown in Figure 10 this signal is compatible with the input of a TTL gate which can be used as a buffer and/or inverter.

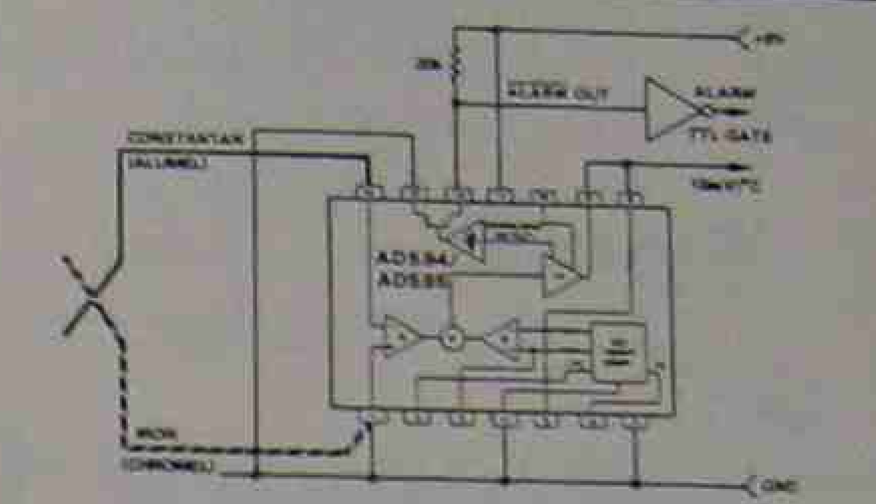


Figure 10. Using the Alarm to Drive a TTL Gate ("Grounded" Emitter Configuration)

Since the alarm is a high level output it may be used to directly drive an LED or other indicator as shown in Figure 11.

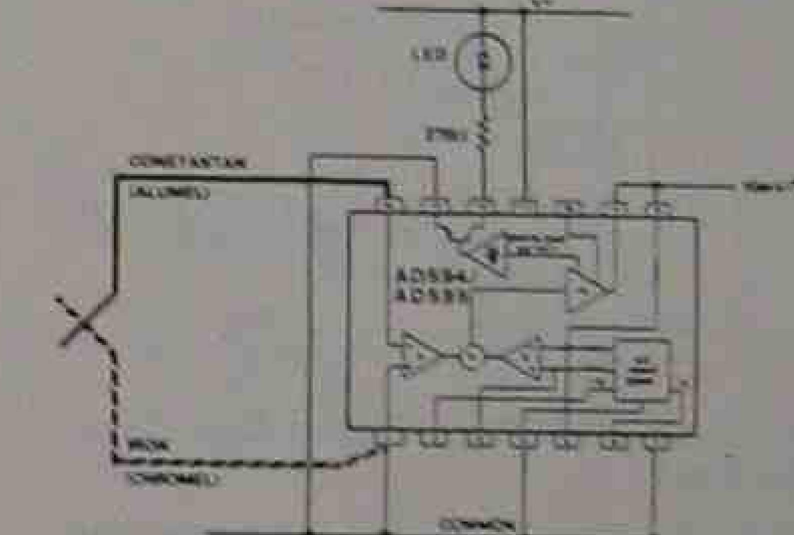


Figure 11. Alarm Directly Drives LED

A 270Ω series resistor will limit current in the LED to 10mA, but may be omitted since the alarm output transistor is current limited at about 20mA. The transistor, however, will operate in a high dissipation mode and the temperature of the circuit will rise well above ambient. Note that the cold junction compensation will be affected whenever the alarm circuit is activated. The time required for the chip to return to ambient temperature will depend on the power dissipation of the alarm circuit, the nature of the thermal path to the environment and the alarm duration.

The alarm can be used with both single and dual supplies. It can be operated above or below ground. The collector and emitter of the output transistor can be used in any normal switch configuration. As an example a negative referenced load can be driven from –ALM as shown in Figure 12.

The collector (+ALM) should not be allowed to become more positive than (V–) +36V, however, it may be permitted to be more positive than V+. The emitter voltage (–ALM) should be constrained so that it does not become more positive than 4 volts below the V+ applied to the circuit.

Additionally, the AD594/AD595 can be configured to produce an extreme upscale or downscale output in applications where an extra signal line for an alarm is inappropriate. By tying either of the thermocouple inputs to common most runaway control conditions can be automatically avoided. A +IN to common connection creates a downscale output if the thermocouple opens, while connecting –IN to common provides an upscale output.

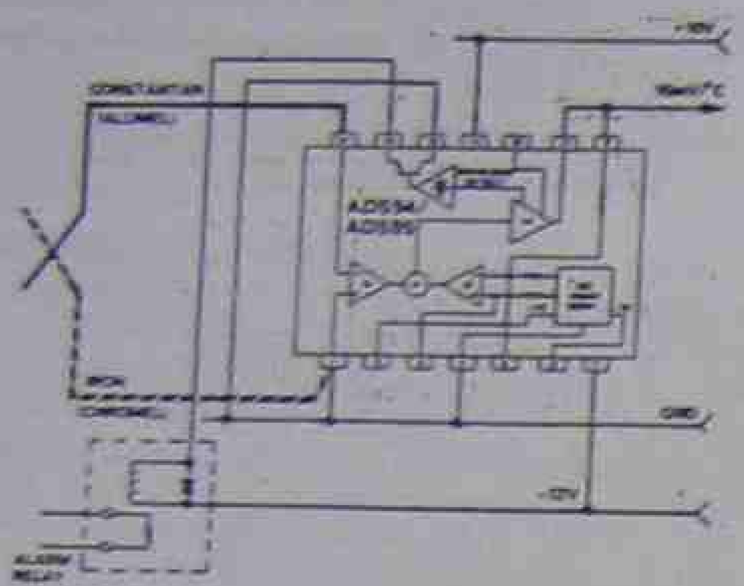


Figure 12. —ALM Driving A Negative Referenced Load

CELSIUS THERMOMETER

The AD594/AD595 may be configured as a stand-alone celsius thermometer as shown in Figure 13.

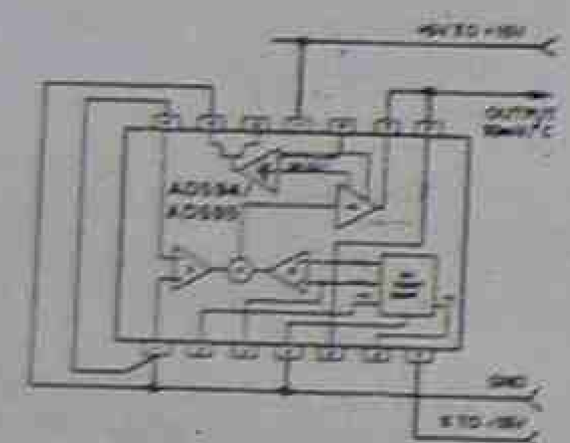


Figure 13. AD594/AD595 as a Stand-Alone Celsius Thermometer

Simply omit the thermocouple and connect the inputs (pins 1 and 14) to common. The output now will reflect the compensation voltage and hence will indicate the AD594/AD595 temperature with a scale factor of 10mV/°C. In this three terminal, voltage output, temperature sensing mode, the AD594/AD595 will operate over the full military −55°C to +125°C temperature range.

THERMOCOUPLE BASICS

Thermocouples are economical and rugged; they have reasonably good long-term stability. Because of their small size, they respond quickly and are good choices where fast response is important. They function over temperature ranges from cryogenics to jet-engine exhaust and have reasonable linearity and accuracy.

Because the number of free electrons in a piece of metal depends on both temperature and composition of the metal, two pieces of dissimilar metal in isothermal contact will exhibit a potential difference that is a repeatable function of temperature, as shown in Figure 14. The resulting voltage depends on the temperatures, T1 and T2, in a repeatable way.

Since the thermocouple is basically a differential rather than absolute measuring device, a known reference temperature is required for one of the junctions if the temperature of the other is to be inferred from the output voltage. Thermocouples made of specially selected materials have been exhaustively characterized in terms of voltage versus temperature compared to primary temperature standards. Most notably the water-ice point of 0°C is used for tables of standard thermocouple performance.

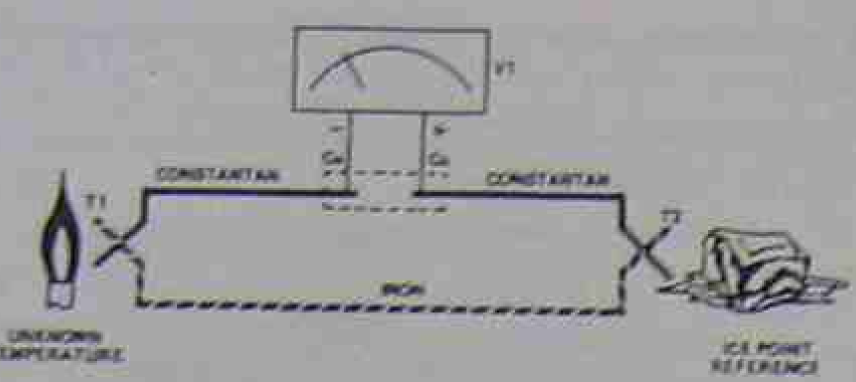


Figure 14. Thermocouple Voltage with 0°C Reference

An alternative measurement technique, illustrated in Figure 15, is used in most practical applications where accuracy requirements do not warrant maintenance of primary standards. The reference junction temperature is allowed to change with the environment

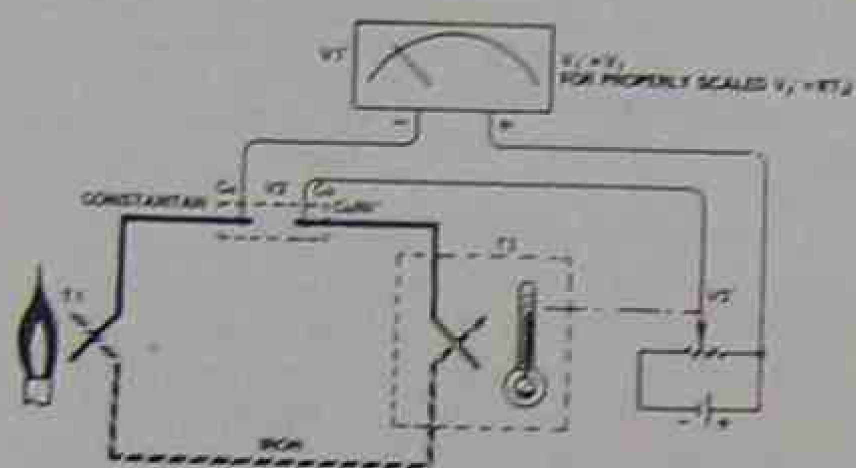


Figure 15. Substitution of Measured Reference Temperature for Ice Point Reference

of the measurement system, but it is carefully measured by some type of absolute thermometer. A measurement of the thermocouple voltage combined with a knowledge of the reference temperature can be used to calculate the measurement junction temperature. Usual practice, however, is to use a convenient thermoelectric method to measure the reference temperature and to arrange its output voltage so that it corresponds to a thermocouple referred to 0°C. This voltage is simply added to the thermocouple voltage and the sum then corresponds to the standard voltage tabulated for an ice-point referenced thermocouple.

The temperature sensitivity of silicon integrated circuit transistors is quite predictable and repeatable. This sensitivity is exploited in the AD594/AD595 to produce a temperature related voltage to compensate the reference or "cold" junction of a thermocouple as shown in Figure 16.

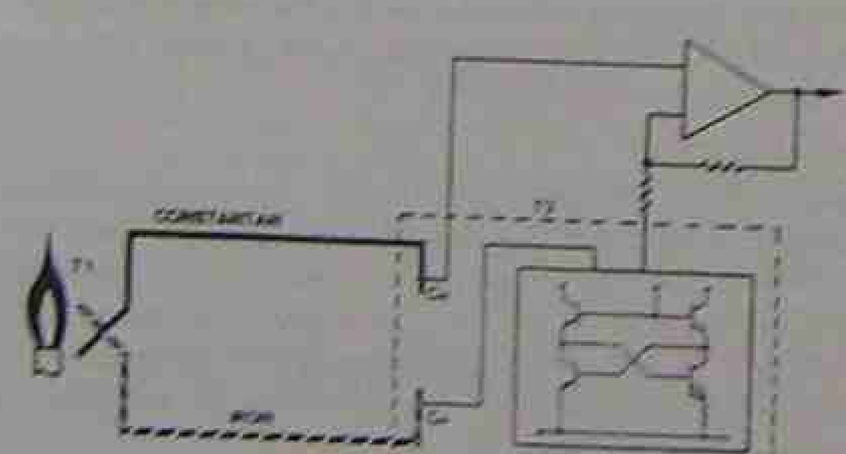


Figure 16. Connecting Isothermal Junctions

Since the compensation is at the reference junction temperature, it is often convenient to form the reference "junction" by connecting directly to the circuit wiring. So long as these connections and the compensation are at the same temperature no error will result.

Sydney Institute of Technology

School of Electrotechnology

Industrial Sensors

Theory Notes

Area ______________________

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Session No.

For further information on this module, or this subject contact Jim Hafford, (02) 217 3620, Bld. K. Ultimo, S.I.T.

13-1 INTRODUCTION

A *thermoresistive sensor* is a device whose internal electrical dc resistance is a function of temperature. There are a number of different types of such devices and we attempt to discuss them all in this chapter. Unlike the thermocouple, however, the thermoresistive sensor does not generate its own voltage source; an external voltage source is necessary to "bring out" this change in resistance within the sensor. We begin our discussion, as usual, describing the measurands that are generally detected. We then describe the various resistivity detectors used today beginning with the earliest known thermoresistive device, the resistance temperature detector.

13-2 COMMONLY SENSED MEASURANDS

The primary measurand sensed by thermoresistive sensors is temperature. Only one secondary measurand has been detected with any practicality—fluid velocity.

13-3 RESISTANCE TEMPERATURE DETECTOR

The *resistance temperature detector* (RTD) operates on a very simple principle but requires the most sophisticated instrumentation for its proper operation. Although the term *resistance temperature detector* can actually refer to a relatively broad area of sensors whose dc resistances change with temperature, the term has recently become more closely associated with only a portion of all the resistive sensors. This portion comprises those sensors made from a pure wire-wound metal having a positive temperature coefficient. To understand what is meant by this term, however, we must next study a frequently observed relationship that exists between temperature and the electrical resistance for certain metals.

13-3.1 Temperature Coefficients of Metals

Virtually all the "earth-mined" metals are conductors of electricity, some being better than others. To state this in another way, virtually all metals display a certain amount of dc resistance to current flow. Some metals have more or less resistance than other metals, depending on their molecular structure. This resistance varies with the metal's temperature, and in almost all cases where we are dealing with metals in their most pure elementary form, this resistance varies directly with temperature. That is, as the temperature increases, the resistance also increases at a fairly linear, proportional rate, and vice versa. Hence the term *positive coefficient* has been affixed to these metals to describe their resistance behavior.

To understand why a metal's resistance increases with an increase in its temperature, we must first understand the nature of the metal's molecular structure and how its behavior relates to temperature. Temperature and energy are interrelated quantities. The thermal energy of a molecule, E_{th}, is related to its absolute temperature T through the equation

$$E_{th} = (1.5)(k)(T) \qquad (13\text{-}1)$$

where E = thermal energy (J)

k = Boltzmann's constant, 1.38×10^{-23} J/K

T = absolute temperature (K)

At absolute zero the molecules of a metal sample have no kinetic energy. The molecules are "stationary" in that they are not vibrating from outside heat energy sources contrary to those metals whose temperatures are above absolute zero and whose molecules do vibrate as a result. Because of this lack of vibrating electrons at the absolute zero condition, a current flow of electrons can pass relatively easily between these molecules without encountering any appreciable resistance (friction). However, with the addition of heat (energy) the metal's molecules vibrate about their locations. This, then, increases the chances of electron collisions from a current flow trying to move between the metallic molecules. Any collisions will produce friction, that is, resistance. The amount of resistance will vary proportionally with the amount of energy (temperature) input.

With some metals the linear rate of resistance change to temperature change is quite predictable. These few metals are generally very stable, chemically. That is, they react very little, if any, with other elements over time. Chemical activity is a problem with using metals for the purpose of temperature sensing. This activity causes the metals' dc resistance to change over a period of time for a given temperature. Figure 13-1 shows the linearity of various metals for a wide temperature range versus their dc resistance. The data acquired for producing this curve were obtained from metal samples having the same length and cross-sectional areas.

As stated earlier, for any given metal, its dc resistance varies directly with, and is proportional to, temperature. However, let's first investigate the relationship between resistance and a metal's cross-sectional area and length. We assume that the temperature is held at a constant room-temperature value. That relationship is

$$R = \frac{\rho l}{A_{CM}} \qquad (13\text{-}2)$$

where R = resistance of wire (Ω)

ρ = resistivity constant (Ω-mil/ft)

l = length (ft)

A_{CM} = cross-sectional area of wire (mil)

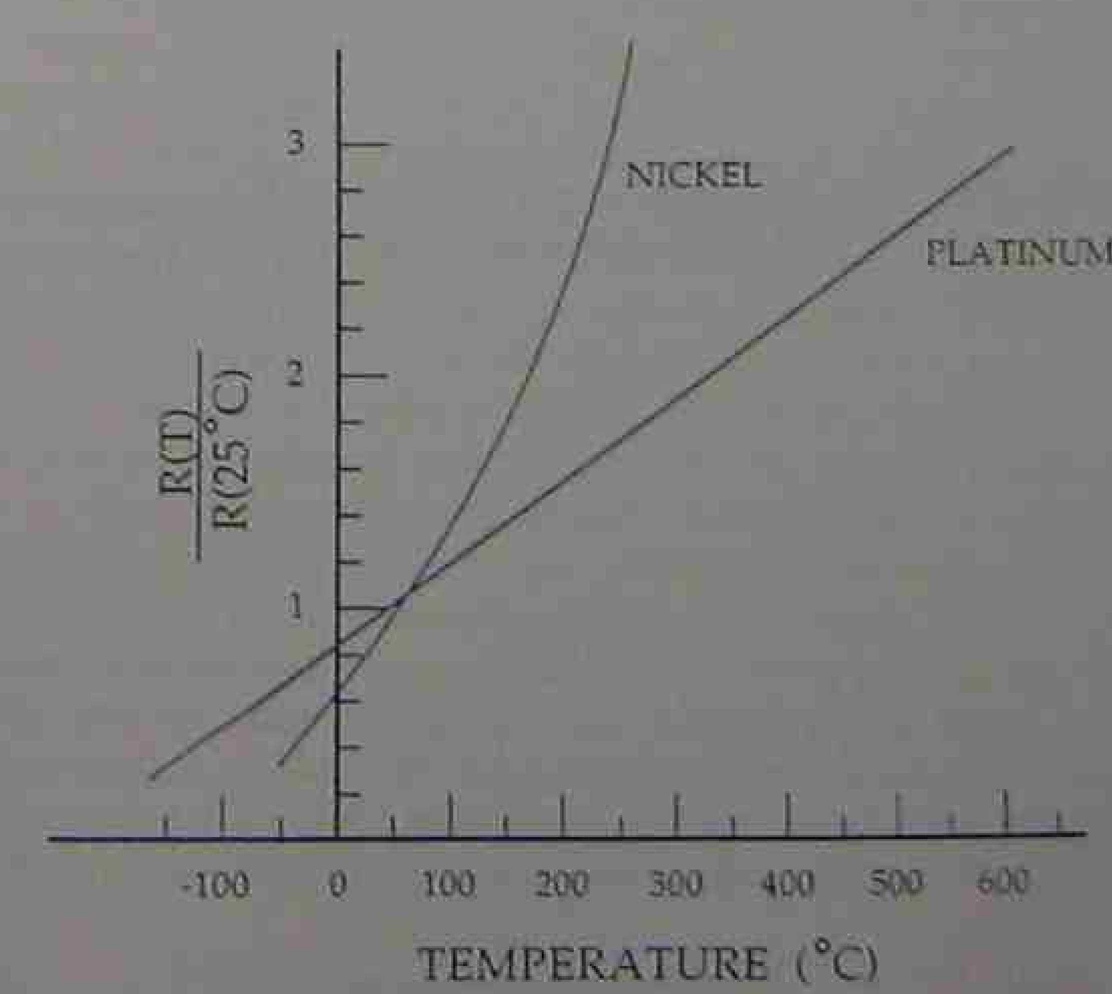


Figure 13-1 Graph of dc resistance for various metals versus temperature

The value for ρ is good only for a specific temperature. The following two examples demonstrate this.

Example 13-1

Find the resistance of a coil of copper wire that is 750 ft in length and having a diameter of 25 mils. Assume a temperature of 20°C and a ρ value of 10.37 Ω-CM/ft for copper.

Solution: Using eq. (13-2) with the values given, we get

$$R =$$

$$=$$

Keep in mind that this value is correct only for a 20°C temperature, however. To find the resistance of the coil for any other temperature, you must use the equation

$$R = \frac{\rho l(1 + \alpha T)}{A_{CM}} \qquad (13\text{-}3)$$

where α = temperature coefficient between the operating temperature and 20°C (1/°C)

T = difference between the operating temperature and 20°C

Example 13-2

Continuing with Example 13-1, calculate the resistance of the same copper wire coil, but do this now for a temperature of 35°C. Assume an α value of 0.00392/°C.

Solution: Using eq. (13-3) with the values given above, we now get

$$R =$$

$$=$$

$$=$$

As can be seen from the two calculations above, the difference between the two calculated resistances is 0.73 Ω. This difference is the result of the 15° increase in temperature in Example 13-2.

13-3.2 Metals Used for Temperature Sensing

The metals most frequently used for temperature sensing, that is, metals that are used for their relatively linear resistance/temperature response characteristics, are nickel, platinum, and certain alloyed forms of copper. Of these three metals,

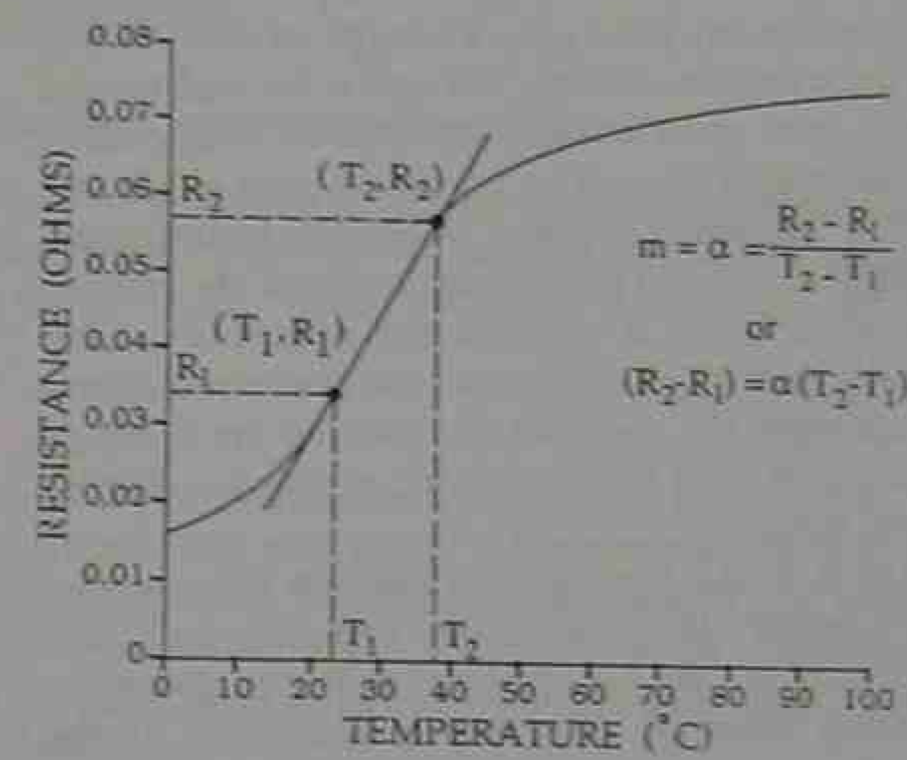


Figure 13-2 Temperature versus dc resistance for platinum.

platinum is probably used most frequently. Figure 13-2 shows platinum to have a very linear temperature versus resistance response over a narrow range of temperatures. Also, platinum is extremely stable in that it reacts with very few substances. This makes it a very reliable and predictable material to work with as a temperature sensor. However, it should be mentioned here that different platinum sensors may exhibit different resistance values for the same temperature. This is due to differing manufacturing and processing techniques used, and it also has to do with strain within the metal itself. Because strain also varies the resistance of metals, it is very important to be aware of this fact during the manufacturing and installation of the platinum sensor. This is true for any metal that is used for temperature detecting.

Because of the possible variations of resistance versus temperature for the same materials, attempts have been made to standardize the resistance/temperature ratios. Unfortunately, neither U.S. or foreign manufacturers have been able to agree on a ratio standard that satisfies everyone. For instance, one American standardizing organization stipulates that platinum sensors must have a resistance ratio, R_{100}/R_0, of 1.3924, where R_{100} is the resistance at 100°C and R_0 is the resistance at 0°C. On the other hand, a British standardizing organization requires a value for R_{100}/R_0 of 1.3850. These ratio figures are nothing more than an indirect means of referring to the slope of the sensor's resistance–temperature curve over a temperature range. While differing resistance ratios do not necessarily affect the performance of this kind of sensor, it does create somewhat of a problem. If you wanted to make a direct replacement, or if you wanted to swap a sensor between instruments, you must make certain that you are using sensors having the same R_{100}/R_0 ratios.

To summarize, to keep the temperature-indicating errors to a minimum, it is necessary to pick a sensor for a specific temperature range. Doing this will maximize the utilization of the sensor's resistance versus temperature linear response characteristics.

13-3.3 Typical Construction

RTDs come in two configurations. One is a wire-wound construction. The coiled construction allows for greater resistance variations for a given temperature change compared to single-strand detectors used many years ago. This type of construction increases the sensing device's sensitivity and resolution. The coil is formed around a nonthermally conductive material such as a ceramic to reduce temperature response time. A sheath is used to surround the coil to protect it from abuse and environmental reactions. The sheath is constructed from a highly conductive material so as not to create a thermal barrier between the sensing coil and the outside temperature. Usually, a stainless steel covering is used for this application. This sheath is hermetically sealed to prevent outside moisture and other elements from affecting the wire sensor inside. In addition, the sheath is filled with a dry thermal-conducting gas to increase thermal contact with the sheath; all of this is a further effort to reduce the sensor's response time.

The RTD's second configuration is in the form of a woven thin-film metallic layer that has been deposited on a nonthermally conductive substrate, such as a ceramic. This process allows for considerable miniaturization and also allows for doing away with an external sheathing material; the ceramic itself acts as the protective sheath. Figure 13-3 shows typical package arrangements used for both construction types.

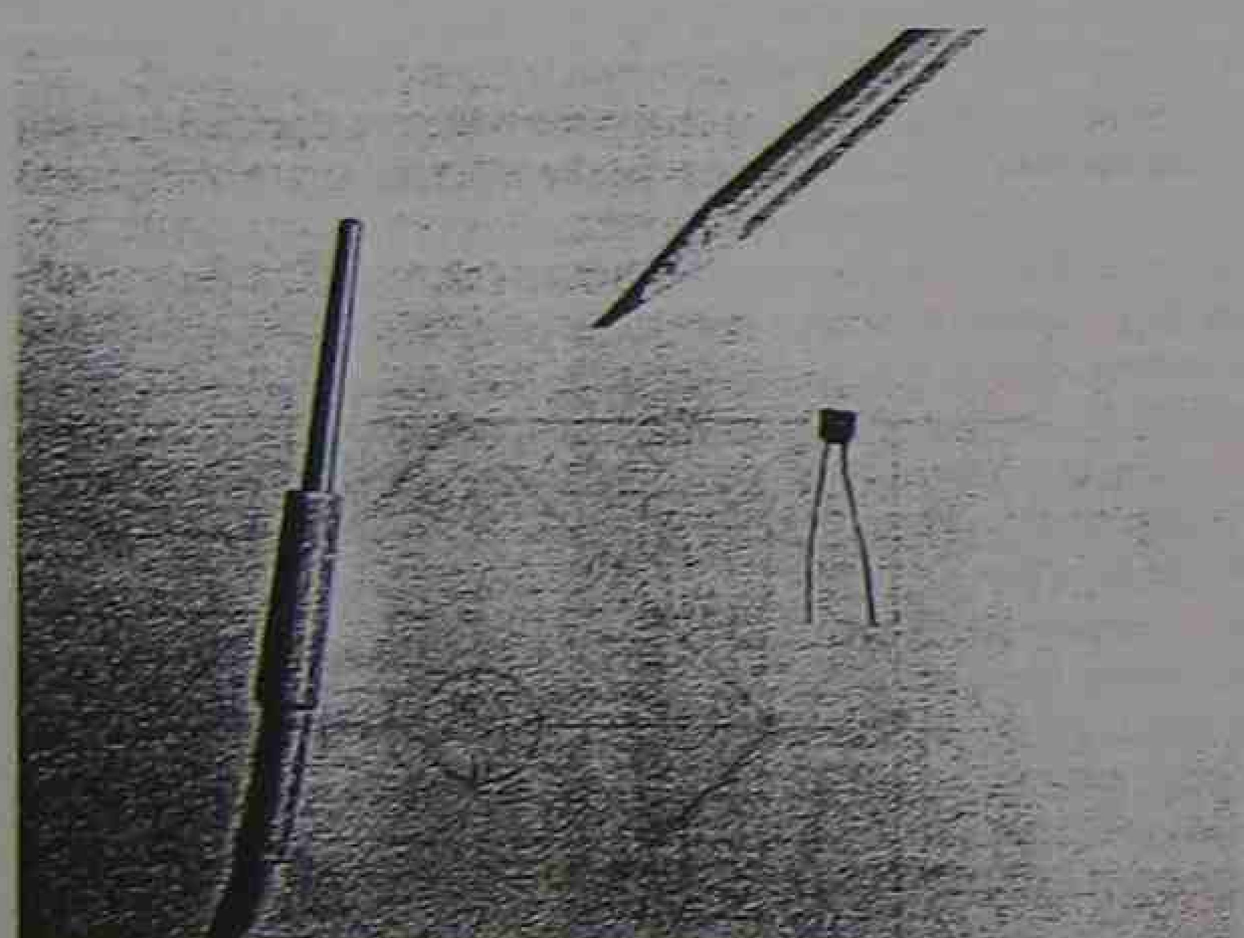


Figure 13-3 Shown at left, wire-wound platinum RTD. At right, deposited-film RTD.

13-3.4 RTD Signal-Conditioning Circuits

Because of the very small change in resistance for a given change in temperature, the Wheatstone bridge, or some variation of this circuit, is used for making these measurements.

Two-wire uncompensated RTD circuit

Figure 13-4 shows a basic two-wire RTD system for temperature measurement. The resistors, R_{L1} and R_{L2}, are the equivalent lead resistances for the two wire leads going to the RTD, R_3. When the bridge is balanced, the following condition is achieved:

$$\frac{R_1}{R_2} = \frac{R_{L1} + R_{L2} + R_3}{R_4} \tag{13-4}$$

where the R notations are as given in Figure 13-5. The circuit also possesses maximum sensitivity to a change in temperature when $R_1 = R_2 = (R_{L1} + R_{L2} + R_3) = R_4$. Notice, however, that the lead resistances play a significant role in determining the overall resistance of the RTD for a given temperature. This is true for lead lengths in excess of several feet or so, and it is certainly true for lengths of several hundreds of feet.

Two-wire compensated RTD circuit

One way to partially compensate for the additional lead resistances added by the wire leads going to the sensor is to add a similar resistance, R_C, to the leg containing R_4 of the Wheatstone bridge, as shown in Figure 13-5(a). Further temperature compensating can be obtained by placing this added compensating resistor with its own set of leads near the sensing leads of the RTD, as shown in Figure 13-5(b). In this way, any change in the RTD lead resistance due to temperature is partially nullified by a similar change in resistance in R_C. This type of circuit is called a *two-wire lead circuit*.

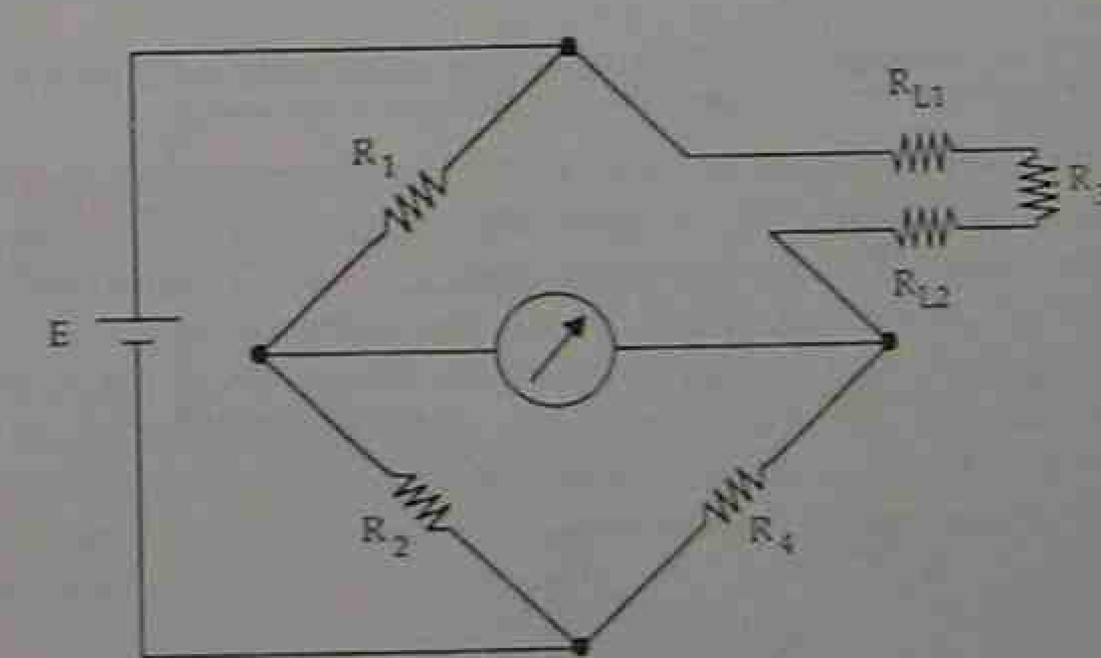


Figure 13-4 Two-wire uncompensated RTD circuit.

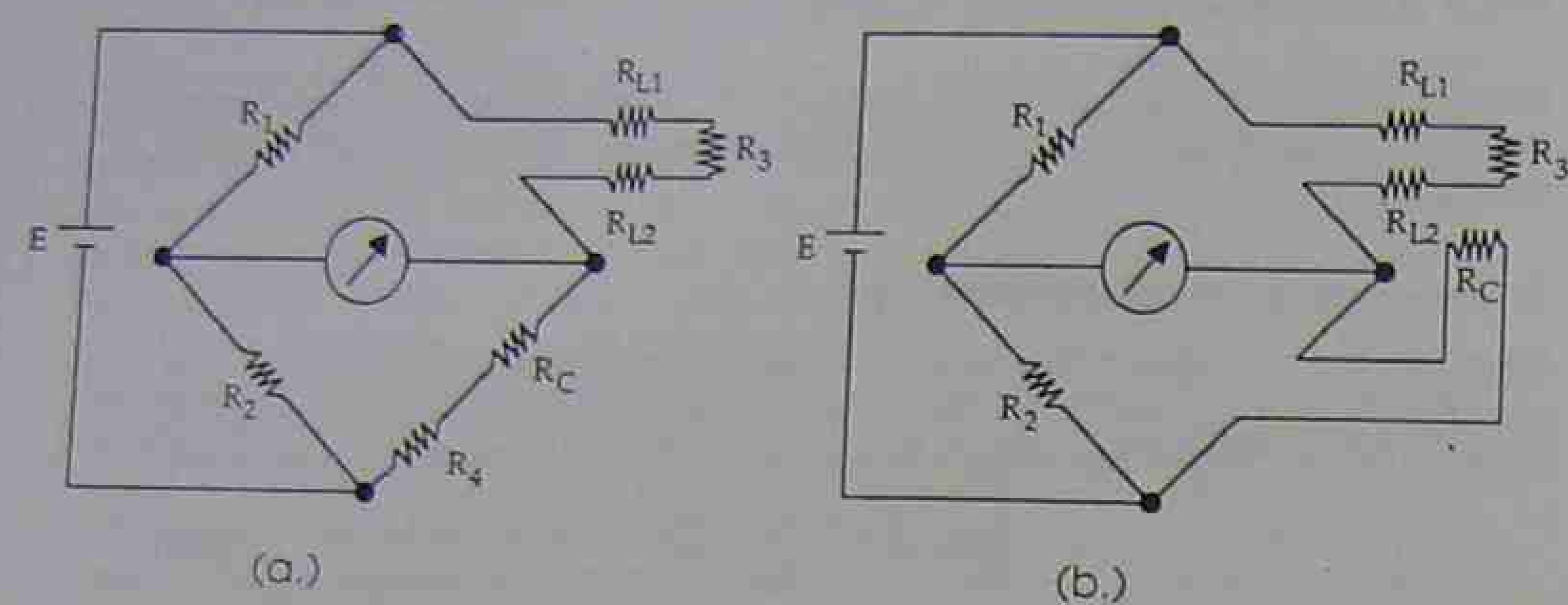


Figure 13-5 Two-wire compensated RTD circuits.

Three-wire RTD circuit

The *three-wire lead circuit* is one that should be used in those instances where the lead resistance is significant compared to the measured resistance of the RTD. Figure 13-6 demonstrates this particular hookup. Notice that lead resistance R_{L1} is in one leg of the bridge, while R_{L2} is in the other leg beneath the first. As a result of this configuration, the two resistances cancel each other out. Also, during the balanced condition of the bridge, there is no current flowing in R_{L3} since it is located in the center leg.

Four-wire RTD circuit

The *four-wire lead* RTD circuit is used for extremely sensitive temperature detection where precision laboratory-type measurements are necessary. The circuitry for this particular configuration is quite complicated and is not discussed further here only because of its extremely limited application.

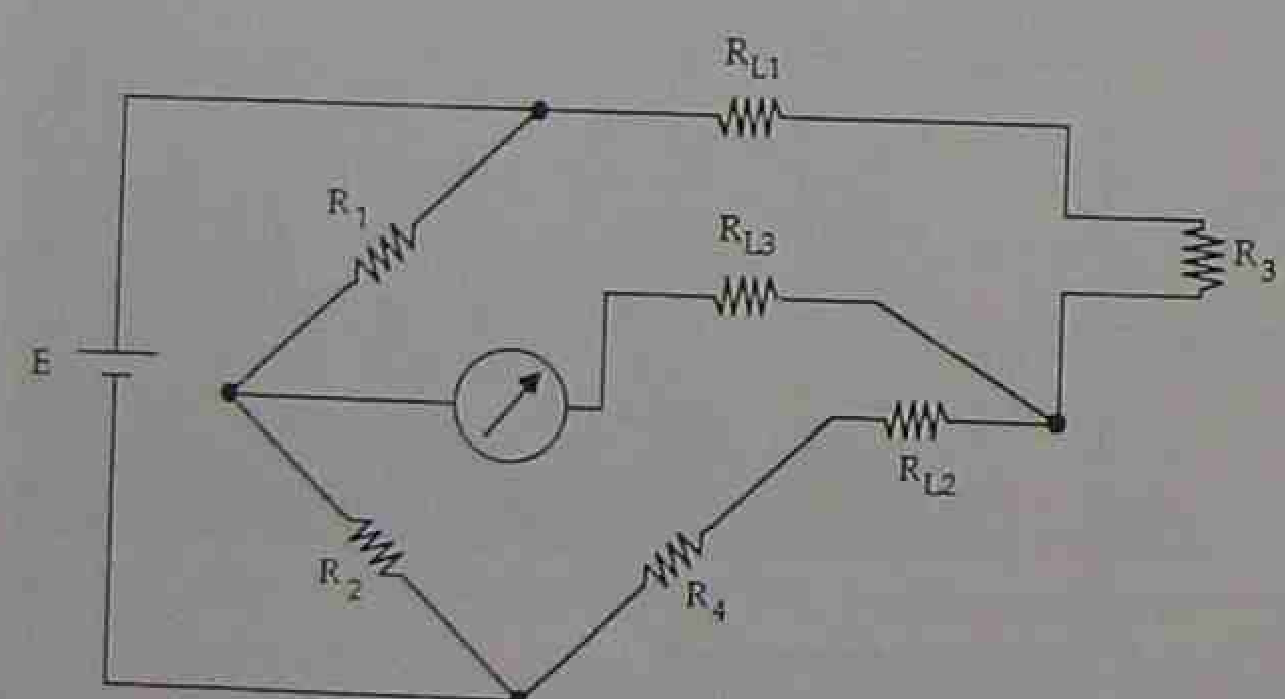


Figure 13-6 Three-wire lead circuit.

13-3.5 Self-Heat

One of the difficulties in using an RTD is to ascertain just how much current should be allowed to travel through it. After all, what is really being monitored when using an RTD to measure temperature is the voltage drop created by a current traveling through the RTD's internal resistance. The amount of current of course is a function of the voltage supply amount applied across the leads of the RTD. Obviously, there is a practical limit to the amount of voltage that can be supplied without damaging the RTD itself. Too-high voltage amounts would cause the RTD to generate excessive or possible damaging heat due to the power being dissipated by its internal resistance ($P = I^2R$). As a matter of fact, for any voltage amount the resulting current flow will cause heat to be dissipated by the RTD. The RTD will not be able to differentiate between this *self-heat* and the heat energy to which it is supposed to respond. The question then becomes: What sort of temperature rise results from this self-heat characteristic?

Fortunately, manufacturers have anticipated this problem by publishing a self-heating error value in their catalog data literature for each RTD that they manufacture. A typical self-heat error figure may be 0.18°C/mW in still air, or 0.07°C/mW in moving air. In other words, for this particular RTD measuring a temperature of a heat source in still air, for every milliwatt of power dissipated by the RTD's internal resistance, there will be a 0.18°C increase in its temperature. This amount would have to be subtracted for each milliwatt of power consumed from the RTD's overall indicated temperature. Therefore,

$$T_{SH} = P_{SHE} \times P_{RTD} \tag{13-5}$$

where T_{SH} = temperature rise due to self-heat (°C)

P_{SHE} = self-heat error (°C/mW)

P_{RTD} = power consumed by RTD (mW)

Let's investigate an example of self-heat with an RTD to become more familiar with how this quantity is handled.

SEE EXAMPLE 13-3

13-3.6 Typical Characteristics and Design Specifications

Typical temperature ranges for the RTD are the following. For copper RTDs, the range is −200 to 260°C; for nickel, the range is −80 to 300°C, and for platinum, the range is −260 to 630°C. Most RTDs require some sort of protective shield to protect the sensor wire from corrosion and general abuse. As a result, the RTD tends to be somewhat fragile in its overall construction, more fragile than the thermocouple. Because of the required sheathing (see Figure 13-3) the thermal response time is significantly increased. A typical response curve for a platinum wire RTD is illustrated in Figure 13-7. Compared to other temperature sensors, this response is somewhat slow.

Example 13-3

A platinum temperature sensor has a published self-heat error of 0.12°C/mW. The sensor's measured resistance for a particular temperature is found to be 129.78 Ω. The voltage supplied to the sensor is 1.88 V dc. According to a temperature versus resistance chart accompanying the sensor, the temperature for 206.66 Ω is supposed to be 280.0°C. What is the *actual* temperature existing at the 129.78 Ω condition after taking self-heating into account?

Solution: We must first calculate the amount of power dissipation in the sensor. This can be done by using the equation $P = E^2/R$. Therefore,

$$\begin{aligned} P &= \frac{1.88^2}{129.78} \\ &= 0.027 \text{ W} \\ &= 27 \text{ mW} \end{aligned}$$

Now, using eq. (13-5),

$$\begin{aligned} T_{SH} &= (0.12)(27) \\ &= 3.24\text{°C rise due to self-heat} \end{aligned}$$

We must next determine what the temperature *appears* to be based on the 129.78 Ω measurement. Because of the linear characteristics of an RTD, we can make a simple proportional ratio based on the supplied manufacturer's data to determine this apparent temperature. That is,

$$\frac{206.66\ \Omega}{280.0\text{°C}} = \frac{129.78\ \Omega}{x}$$

$$\begin{aligned} x &= \frac{(129.78\ \Omega)(280.0\text{°C})}{206.66\ \Omega} \\ &= 175.84\text{°C} \end{aligned}$$

Because this is a temperature that includes the self-heat of the RTD; in other words, this is a temperature obtained with data (the RTD's measured resistance) influenced by self-heat. We can now compensate for this effect by performing the following step. We subtract the temperature amount calculated with eq. (13-5) from our apparent temperature. That is,

$$\text{actual temperature} = \text{apparent temperature} - T_{SH} \qquad (13\text{-}6)$$

Therefore,

$$\begin{aligned} \text{actual temperature} &= 175.84\text{°C} - 3.24\text{°C} \\ &= 172.60\text{°C} \end{aligned}$$

13-3.7 Practical Applications

Because of the compactness furnished by modern solid-state circuitry, it has recently been possible to construct hand-held RTD transducers complete with signal conditioning and digital temperature readout displays. These devices employ three- and four-wire balanced bridge circuitry for excellent temperature reading accuracy. Figure 13-8 shows a complete hand-held RTD thermometer using a platinum sensor and having a temperature range from −220 to 630°C. This is a remarkable feat when monitoring and controlling temperature in a hydraulic press used for molding plastics. Certain plastics require a particular curing temperature and curing rate. By implanting RTDs into the press' platens (Figure 13-9) it becomes possible to monitor and control these temperatures. Each RTD is located in a separately controlled heating zone so that each zone is controlled by that particular RTD.

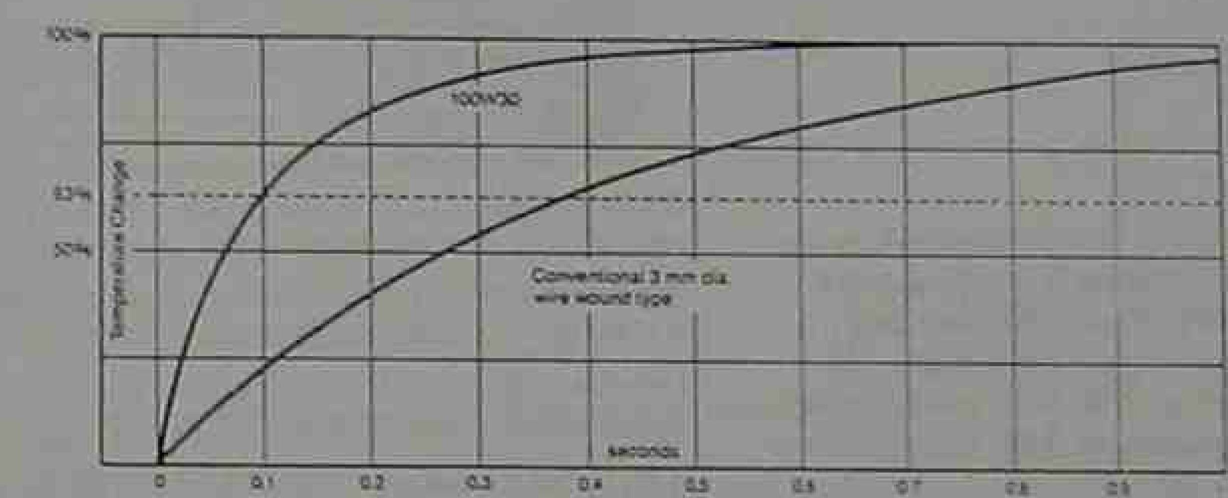


Figure 13-7 Time-response curve for a platinum wire RTD. (REPRODUCED WITH THE PERMISSION OF OMEGA ENGINEERING, INC.)

you compare this device to a typical RTD setup of, say, 15 years ago. At that time the instrumentation needed would have covered the area of roughly 4 ft² and its bulk would have weighed in excess of 30 lb or so.

RTDs have found their way into many laboratory applications where extreme temperature accuracy is needed along with their very linear, durable, and high-temperature-range characteristics. In industry a typical application would be in

Figure 13-8 Hand-held RTD thermometer. (REPRODUCED WITH THE PERMISSION OF OMEGA ENGINEERING, INC.)

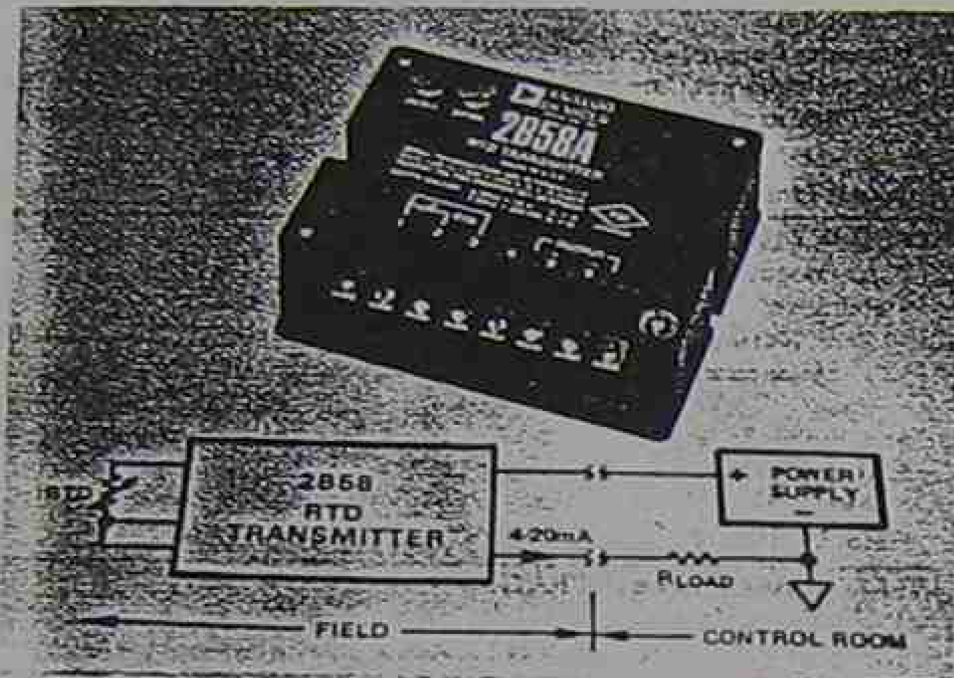
ANALOG DEVICES

Two-Wire, Linearized RTD Temperature Transmitter

MODEL 2B58

FEATURES
Platinum RTD Input
Linearized 4-20mA Output
High Accuracy: ±0.1%
Low Drift: ±0.01°C/°C Max
RFI Immunity
Low Cost
FM Approved

APPLICATIONS
RTD Temperature Transmission in:
- Process Control
- Factory Automation
- Energy Management



9

GENERAL DESCRIPTION
The model 2B58 is a high accuracy, two-wire temperature transmitter designed to accept a platinum RTD (Resistance Temperature Detector) input and produce a proportional standard 4-20mA output. The RTD signal is internally linearized to provide an output which is linear with temperature. Four precalibrated ranges are available for RTD measurements from −100°C to +400°C.

The 2B58 features high accuracy of 0.1%, low drift of ±0.01°C/°C, high noise rejection and RFI immunity. Both two-wire and three-wire 100Ω sensors may be used. Lead wire compensation is provided for three-wire RTDs. The 2B58 is approved by Factory Mutual for intrinsically safe use in hazardous locations.

A rugged metal enclosure, suitable for field mounting, offers environmental protection and screw terminal input and output connections. This enclosure may be either surface or standard relay track mounted.

APPLICATIONS
The 2B58 has been specifically designed to provide high-performance two-wire transmission of measured temperatures using RTD sensors.

Two-wire current transmission permits remote mounting of the transmitter near the sensor to minimize the effects of noise and signal degradation to which low level sensor outputs are susceptible. Transmission of the proportional current output may be accomplished by means of inexpensive copper wires. These factors make the 2B58 ideally suited for applications where accuracy, stability, and low cost installation are desired.

DESIGN FEATURES AND USER BENEFITS
High Accuracy: The 2B58 offers high calibration accuracy, linearized output and conformity with the standard DIN 43760 (α = 0.00385) RTD sensors.

Low Cost: The 2B58 combines low price with a two-wire transmission, lowering total installation cost.

High Noise Rejection: The transmitter features internal filtering circuitry to assure protection against RFI and line frequency pickup.

Standard Loop Compatibility: The two-wire output structure conforms to the ISA Standard S50.1 "Compatibility of Analog Signals for Electronic Industrial Process Instruments".

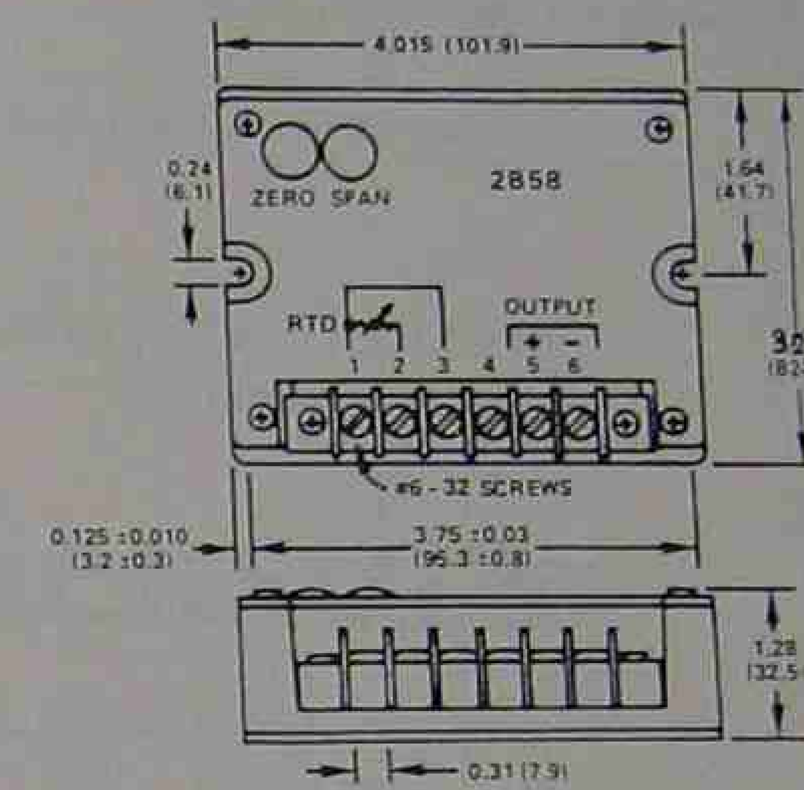
Wide Operating Temperature Range: The 2B58 has been designed to operate over −30°C to +85°C ambient temperature range.

SPECIFICATIONS (typical @ +25°C and V_S = +24V dc unless otherwise noted)

Model	2B58A
INPUT SPECIFICATIONS	
Sensor Type	Platinum, 100Ω @ 0°C, α = 0.00385 2 or 3 Wire
Normal Mode Rejection	56dB @ 60Hz
Sensor Excitation Current	0.5mA
Zero and Span Adjustment Range	±5% of Span
OUTPUT SPECIFICATIONS	
Output Span	4-20mA
Minimum Output Current	3.5mA, typ
Maximum Output Current	40mA, typ
Load Resistance Range Equation	R_L max = (+V_S −16V)/20mA
@ +24V Supply	0 to 400Ω
Output Protection[1]	±60V
ACCURACY	
Total Output Error[2]	±0.1%
Stability vs. Ambient Temperature	
Zero, Measurement Range 01 through 03[3]	±0.01°C/°C max (±0.005°C/°C typ)
Measurement Range 04[3]	±0.01°C/°C
Span	±0.005%/°C
Stability vs. Time[2]	±50ppm/Month
Lead Resistance Effect, to 40Ω per Lead	
Span Error	±0.5%
Warm-Up Time to Rated Performance	3 Minutes
INTRINSICALLY SAFE OPERATION	
Use in Class I, Division 1, Groups A,B,C, and D Hazardous Locations	FM Approved
RESPONSE TIME	
To 90% of Span	0.4 sec
POWER SUPPLY	
Voltage, Operating Range	+16V to +60V dc
Supply Change Effect, % of Span	
on Zero	±0.005%/V
on Span	±0.01%/V
ENVIRONMENTAL	
Temperature Range, Rated Performance	−30°C to +85°C
Storage Temperature Range	−55°C to +125°C
Humidity, Effect[4]	
Error	±0.6% of Span
RFI Effect (5W @ 470MHz @ 3 ft.)	
Error	±0.5% of Span
PHYSICAL	
Case Size	4" X 3.25" X 1.25"
Weight	8 oz (227 g)

NOTES
[1] Protected for reverse polarity and for any input/output connection combination.
[2] Accuracy is specified as a percent of output span (16mA). Accuracy spec includes combined effects of transmitter repeatability, hysteresis and sensor linearization conformity. Does not include sensor error.
[3] See ordering information for measurement temperature ranges 01 through 04.
[4] Per MIL-STD-202E Method 103B.
Specifications subject to change without notice.

OUTLINE DIMENSIONS (MAX)
Dimensions shown in inches and (mm).



ORDERING INFORMATION
Example: Model 2B58A - 1 - 1 - 01

Enter Model
2B58A

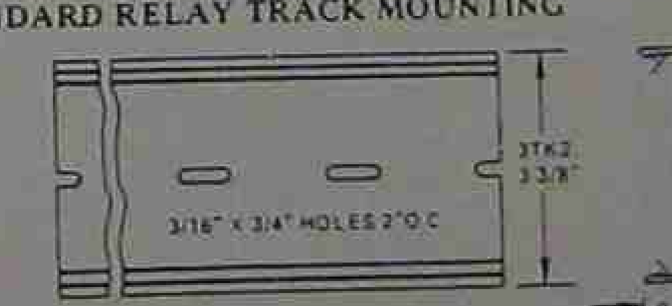
Enter Housing
1-Standard Enclosure

Enter Sensor Type
1-100Ω Platinum, α = 0.00385

Enter Temperature Range
01 Through 04

Range in °C (°F)	No.
−100 to +100 (−148 to +212)	01
0 to +100 (+32 to +212)	02
0 to +200 (+32 to +392)	03
0 to +400 (+32 to +752)	04

STANDARD RELAY TRACK MOUNTING



Model 2B58 may be conveniently mounted in a standard relay mounting channel (3.25" wide) such as Reed Devices Inc. (RDI) model 3TK2-6 or equivalent.

TEMPERATURE CONTROLS PTY LTD

A.C.N. 003 512 294

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THERMOCOUPLES • RTD SENSORS • THERMOWELLS • EXTENSION CABLES

Reference table for platinum resistance elements

R.T.D.

Resistance values in Ohm from 0°C to +400°C

°C	0	1	2	3	4	5	6	7	8	9
0	100.00	100.39	100.78	101.17	101.56	101.95	102.34	102.73	103.12	103.51
10	103.90	104.29	104.68	105.07	105.46	105.85	106.24	106.63	107.02	107.40
20	107.79	108.18	108.57	108.96	109.35	109.73	110.12	110.51	110.90	111.28
30	111.67	112.06	112.45	112.83	113.22	113.61	113.99	114.38	114.77	115.15
40	115.54	115.93	116.31	116.70	117.08	117.47	117.85	118.24	118.62	119.01
50	119.40	119.78	120.16	120.55	120.93	121.32	121.70	122.09	122.47	122.86
60	123.24	123.62	124.01	124.39	124.77	125.16	125.54	125.92	126.31	126.69
70	127.07	127.45	127.84	128.22	128.60	128.98	129.37	129.75	130.13	130.51
80	130.89	131.27	131.66	132.04	132.42	132.80	133.18	133.56	133.94	134.32
90	134.70	135.08	135.46	135.84	136.22	136.60	136.98	137.36	137.74	138.12
100	138.50	138.88	139.26	139.64	140.02	140.39	140.77	141.15	141.53	141.91
110	142.29	142.66	143.04	143.42	143.80	144.17	144.55	144.93	145.31	145.68
120	146.06	146.44	146.81	147.19	147.57	147.94	148.32	148.70	149.07	149.45
130	149.82	150.20	150.57	150.95	151.33	151.70	152.08	152.45	152.83	153.20
140	153.58	153.95	154.32	154.70	155.07	155.45	155.82	156.19	156.57	156.94
150	157.31	157.69	158.06	158.43	158.81	159.18	159.55	159.93	160.30	160.67
160	161.04	161.42	161.79	162.16	162.53	162.90	163.27	163.65	164.02	164.39
170	164.76	165.13	165.50	165.87	166.24	166.61	166.98	167.35	167.72	168.09
180	168.46	168.83	169.20	169.57	169.94	170.31	170.68	171.05	171.42	171.79
190	172.16	172.53	172.90	173.26	173.63	174.00	174.37	174.74	175.10	175.47
200	175.84	176.21	176.57	176.94	177.31	177.68	178.04	178.41	178.78	179.14
210	179.51	179.88	180.24	180.61	180.97	181.34	181.71	182.07	182.44	182.80
220	183.17	183.53	183.90	184.26	184.63	184.99	185.36	185.72	186.09	186.45
230	186.82	187.18	187.54	187.91	188.27	188.63	189.00	189.36	189.72	190.09
240	190.45	190.81	191.18	191.54	191.90	192.26	192.63	192.99	193.35	193.71
250	194.07	194.44	194.80	195.16	195.52	195.88	196.24	196.60	196.96	197.33
260	197.69	198.05	198.41	198.77	199.13	199.49	199.85	200.21	200.57	200.93
270	201.29	201.65	202.01	202.36	202.72	203.08	203.44	203.80	204.16	204.52
280	204.88	205.23	205.59	205.95	206.31	206.67	207.02	207.38	207.74	208.10
290	208.45	208.81	209.17	209.52	209.88	210.24	210.59	210.95	211.31	211.66
300	212.02	212.37	212.73	213.09	213.44	213.80	214.15	214.51	214.86	215.22
310	215.57	215.93	216.28	216.64	216.99	217.35	217.70	218.05	218.41	218.76
320	219.12	219.47	219.82	220.18	220.53	220.88	221.24	221.59	221.94	222.29
330	222.65	223.00	223.35	223.70	224.06	224.41	224.76	225.11	225.46	225.81
340	226.17	226.52	226.87	227.22	227.57	227.92	228.27	228.62	228.97	229.32
350	229.67	230.02	230.37	230.72	231.07	231.42	231.77	232.12	232.47	232.82
360	233.17	233.52	233.87	234.22	234.56	234.91	235.26	235.61	235.96	236.31
370	236.65	237.00	237.35	237.70	238.04	238.39	238.74	239.09	239.43	239.78
380	240.13	240.47	240.82	241.17	241.51	241.86	242.20	242.55	242.90	243.24
390	243.59	243.93	244.28	244.62	244.97	245.31	245.66	246.00	246.35	246.69
400	247.04									

PROUDLY MANUFACTURING IN AUSTRALIA

Sydney Institute of Technology

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Industrial Sensors

Theory Notes

Area Industrial Sensors

Topic Industrial Sensors

Session No.

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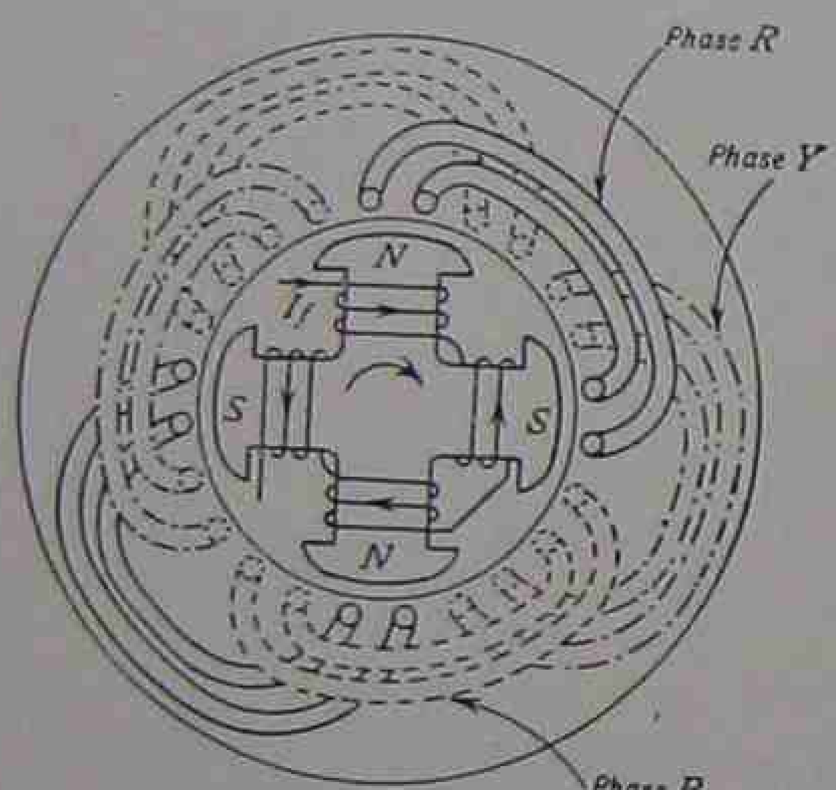
13-4 THERMISTOR

We now look at another thermoresistive sensor, the *thermistor*. This device has entirely different characteristics compared to the RTD. For one thing, the thermistor is made from a human-made substance rather than being produced from naturally occurring materials, as in the case of the RTD. This human-made substance is called a semiconductor. Second, the thermistor's behavior with respect to temperature is entirely different from that of the RTD. These and other characteristics are discussed extensively in the sections that follow.

13-4.1 Theory of Operation

To understand how thermistors operate, we must first understand what a semiconductor is. Recalling our discussion concerning why metals conduct better at lower temperatures rather than at higher temperatures, there are certain types of materials that conduct very little, if any at all, at extremely low temperatures. The reason for this is because of the degree of bonding of the outer electrons in the shells of these materials. In metals these electrons experience very little attraction from the nucleus and, consequently, have little difficulty in being dislodged from the parent atom.



AN IMPORTANT USE OF THERMISTORS IS IMPLANTATION INTO SCIM WINDINGS FOR USE IN OVER TEMPERATURE PREVENTION.

FIG. 188.—End connections of a three-phase single-layer winding.

However, in certain nonmetallic compounds, the bonding is much greater, and unless the outer electrons are agitated by heat, the bonding is so high that no free electrons are free to roam at the coldest temperatures. Only through the addition of heat can these electrons become agitated enough because of their vibrating that they can break their bonds and travel on to similarly ionized atoms. Compounds that show this type of behavior are called semiconductors.

13-4.2 Characteristics of the Thermistor

Because a semiconductor's resistance decreases with an increase in temperature, which is a behavior pattern opposite to that of metals, semiconductors are said to possess a *negative temperature coefficient*. Figure 13-10 depicts a typical resistance versus temperature curve for a thermistor semiconductor. You can conclude from observing this curve that a thermistor is not considered a linear device, especially when compared to the RTD curve also shown in Figure 13-10. Both curves shown are only typical curves. Thermistors can be manufactured so that different resistance values can be formulated to occur at a particular temperature, say at 0 or 25°C. These values can range from 100 Ω to as high as 1 MΩ for these temperatures. If the resistance versus temperature curves of these other thermistors were plotted alongside the one in Figure 13-10, you would observe a family of curves paralleling the one shown in Figure 13-10.

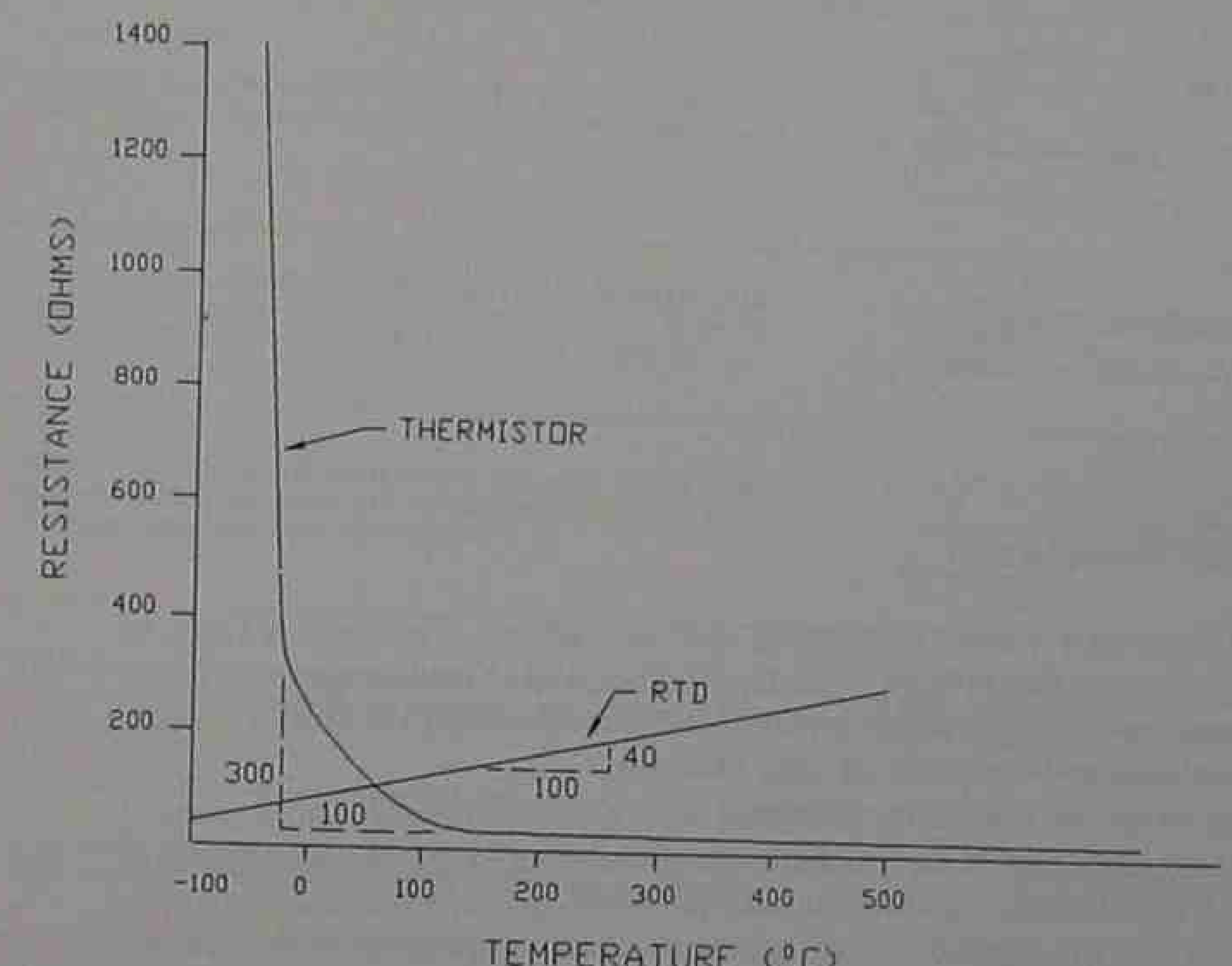


Figure 13-10 Temperature versus resistance curve for a thermistor.

Because the thermistor is a semiconducting material, the temperature range of this type of sensor is somewhat limited. Semiconductors can rarely withstand temperatures above about 300°C before melting. However, at this temperature the thermistor conducts current very well, approaching the conduction qualities of a poor conductor. In addition, the slope of the response curve approaches zero, meaning that the thermistor's sensitivity becomes almost nonexistent. The thermistor loses its effectiveness as a sensing device at this point.

Below −50°C the thermistor's resistance approaches that of a poor insulator. Resistances of several megohms are likely to be found. At −100°C the practical sensing limit of a thermistor is probably reached. However, again noting the shape of a thermistor's curve, you can see that the thermistor is most linear in the cold region of operation, say below 0°C. Also, its sensitivity is highest in this region.

The thermistor is generally a more sensitive sensor than the RTD. Since the sensitivity of any sensor is found by *dividing its output signal by its input signal*, this quantity is reflected in the amount of the slope in its plotted response curve. In other words, for the thermistor, we can observe the slope of its temperature versus resistance response curve (refer again to Figure 13-10), compare it to the slope of the RTD curve, and see immediately that the thermistor is far more sensitive than the RTD to a given change in temperature (assuming that we neglect the very high temperature end of the thermistor's response curve).

Example 13-4

Using the slope information shown in Figure 13-10, compare the sensitivities of the thermistor and RTD.

Solution: The sensitivity amount for the thermistor obtained from the slope information in Figure 13-10 is the following.

13-4.3 Typical Construction

Thermistors come in a variety of physical sizes and shapes. The various types of shapes or styles are the following: (1) beads, (2) disks, and (3) rods or probes. There are other shapes and styles, but the ones listed above are probably the most common. Figure 13-11 shows an illustration of each example.

In some cases the thermistor is coated with a protective substance, usually glass, to prevent oxidation of the solid-state material used in the thermistor's construction. The glass furnishes a hermetic seal that wards off oxidation and corrosion that may occur due the thermistor's environment. Another purpose of the coating is to bond the leads of the thermistor more firmly to the thermistor's body, thus acting as a strain relief for the leads.

The glass-coated beads on thermistors range in size from approximately 0.1 to 2 mm in diameter. On the other hand, the disk-type thermistors typically have diameters ranging from 1 to 3 mm and are often not coated. These are often cheaper in cost. During manufacturing, the thermistors are easily adjusted to the desired resistance by grinding away a portion of the thermistor's disk. Thermistor rods are typically manufactured in diameters of about 0.5 mm to as much as 50 mm.

The most common application for thermistors is for thermometry (Figure 13-12). Because of the wide variety of shapes available, it is possible to find a thermometer probe that can fit into the tightest of temperature-measuring physical restrictions. This is discussed further in the next section concerning applications.

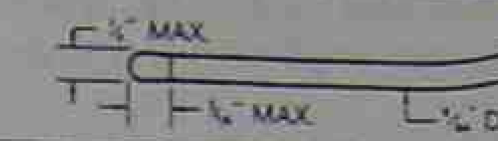
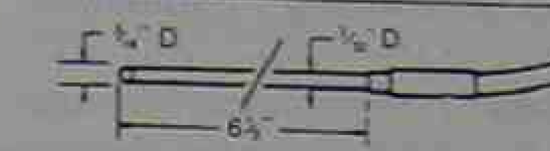
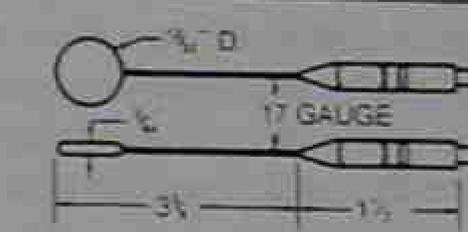
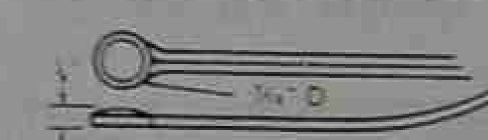
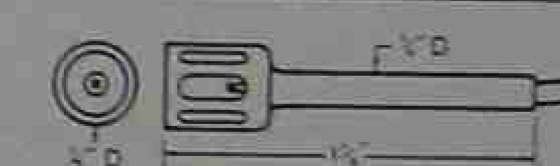
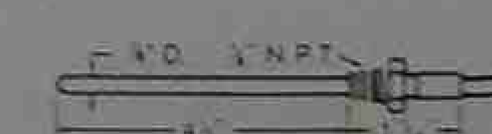
Description	Configuration
GENERAL PURPOSE. Vinyl tipped, most rugged probe. Used for short-term water and sub-soil readings. Waterproof construction now standard.	
SMALL FLEXIBLE. Vinyl sheath and tip. Cuvette temperatures. General purpose measurement.	
GENERAL PURPOSE. Non-immersible, epoxy tipped probe. Can be potted in place. Probe is suitable for temperature measurements on surfaces.	
"BANJO" SURFACE TEMPERATURE. Skin, oral, axillary, water bath, air, surface temperatures. Stainless steel.	
ATTACHABLE SURFACE TEMPERATURE. Stainless steel cup, epoxy backed. Easy to tape on flat surfaces. Good for heat loss or compression efficiency study of piping.	
SMALL SURFACE TEMPERATURE. Cuvette, water bath, leaf and other surfaces. 24" Teflon-covered flexible wire. Stainless steel disc with epoxy back.	
AIR TEMPERATURE. Stainless steel probe suitable for test rooms, incubators, remote air readings, monitoring of gas streams, etc.	
TUBULAR. Stainless steel probe for rugged duty. Often used for liquid immersion. Probe is immersible only to cap.	
TUBULAR-GLASS. Chemically inert for liquid immersion use. Thermometric titration. Freezing point determination. Pyrex. 5" long.	
TUBULAR WITH FITTING. Rugged, stainless steel probe with pipe fitting. Suitable for taking readings in pipes or inside closed vessels.	

Figure 13-11 Various forms of thermistors. (REPRODUCED WITH THE PERMISSION OF OMEGA ENGINEERING, INC.)

13-4.4 Thermistor Self-Heat

Like the RTD the thermistor is susceptible to the problem of self-heat. Unlike the RTD, however, the thermistor is placed into a potential self-destructive mode when self-heat becomes excessive. That is, as the thermistor becomes warmer, its internal resistance becomes lowered, allowing an increase in current flow which creates additional heat, and so on. It is especially critical to limit the current flow through a thermistor to prevent this situation from happening.

13-4.5 Linearizing Thermistors

Because of the nonlinear characteristic of thermistors, various circuit schemes have been devised to make the thermistor behave in a linear fashion. One such scheme is shown in Figure 13-13. Two thermistors of identical response are so wired that one is parallel to another having a resistor in series with it. Through considerable experimentation it is found that this particular circuit achieves at least a 10-fold decrease in the amount of temperature deviating from linearity. For instance, for a temperature range of, say, 0 to 100°C, a maximum deviation of only 0.15°C from linearity is achieved, whereas normally, a single thermistor system yields an amount of 1.5°C. By adding the additional matching parallel thermistor with its series resistor, a dramatic linearization for the entire system can be obtained.

13-4.6 Practical Applications

Thermistors are usually associated with thermometry-type instrumentation, that is, instrumentation used solely for the reading of temperature. Figure 13-12 in Section 13-4.3 is a good example. Despite their nonlinear characteristics, thermistors still enjoy widespread use. The reason for this is due to the relative ease in linearizing an otherwise nonlinear sensor through the use of lookup tables that have been stored

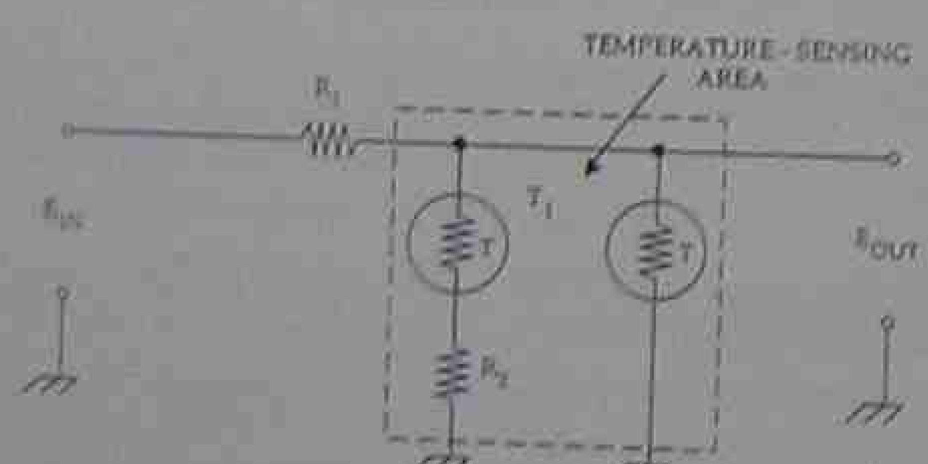


Figure 13-13 Linearizing circuit for a thermistor.

in firmware. Many thermistor-equipped electronic thermometers contain IC chips that can convert a thermistor's nonlinear output into a very linear response, resulting in extremely accurate output data. The temperature probe containing the thermistor is also made so that it is replaceable in case it is damaged.

There are at least two major forms of signal-conditioning circuitry that are used in converting a thermistor's resistance change to a signal that eventually drives a digital or analog readout. One form uses a resistance-measuring circuit such as a Wheatstone bridge to determine the temperature (refer to Figure 13-5). In this circuit R_1 is adjusted for a null according to the center-leg indicator. Instead of calibrating R_1 in ohms to read the resistance of R_3, in this case the thermistor, it is calibrated in units of temperature instead.

A less sophisticated resistance-measuring circuit employs a basic ohmmeter circuit (Figure 13-14). This is where the voltage drop occurring across resistance, R_{IND}, produces the necessary voltage signal for creating the desired temperature reading.

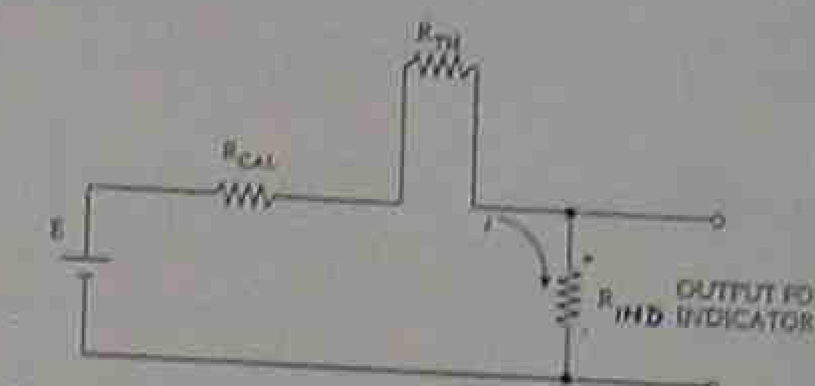


Figure 13-14 Using a simple ohmmeter circuit to read a thermistor's resistance.

The second type of circuit used for thermometry purposes is illustrated in Figure 13-15. In this example we see a thermistor, R_{TH}, whose varying resistance controls the frequency of oscillation of an oscillator (the 555 timer IC, R, and C). The output of this system is in the form of a square wave with varying frequency. The frequency of this wave can then be measured and easily converted to a temperature display. In this circuit the frequency of oscillation varies inversely with the change in resistance of R_{TH}.

Example 13-5

In Figure 13-15 the frequency of oscillation of the circuit shown is determined by

$$f = \frac{0.722}{(R_{TH} + R)C} \qquad (13\text{-}7)$$

Assuming that capacitor C has a value of 0.005 μF and resistor R a value of 1200 Ω, determine the values of f for the following three values of R_{TH}: 750 Ω, 1500 Ω, and 300 Ω.

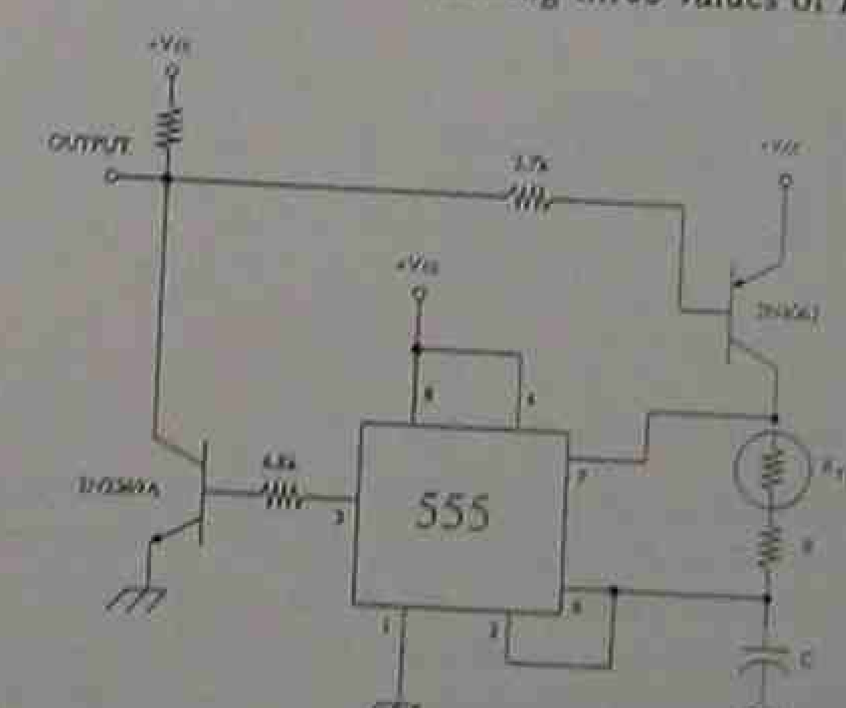


Figure 13-15 Using a thermistor to control the frequency output of an oscillator. (From Howard M. Berlin, *The 555 Timer Applications Sourcebook*, copyright 1976, p. 65. Reprinted by permission of the author.)

Solution: Placing the values of C, R, and the values of R_{TH} into eq. (13-7), we get

f_1

f_2

f_3

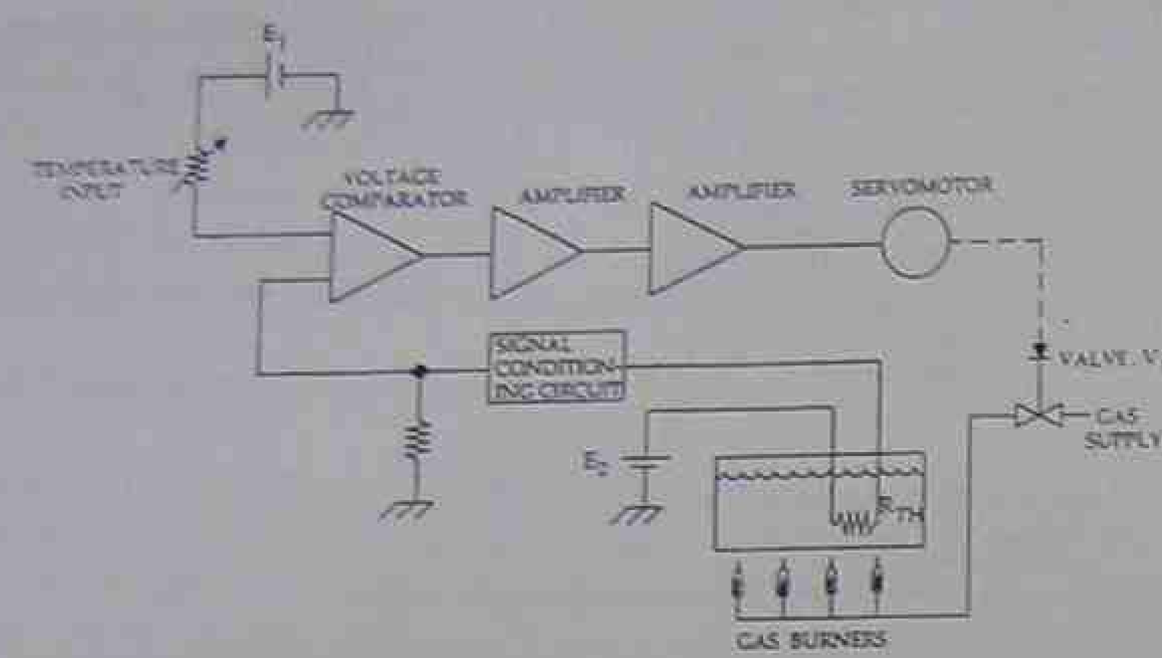


Figure 13-16 Thermistor being used in a process control system for controlling the temperature of a liquid in a heated container.

In addition to being used in thermometry, thermistors are also often used for temperature control applications. A typical application is seen in Figure 13-16. In this circuit we see a controller being used to maintain a desired water temperature in a food-processing system. The operator adjusts the temperature controller to the desired temperature, while the thermistor detector senses the water's temperature. The sensed temperature is then compared to the desired temperature. Any difference in the two values (this difference is called the *error signal*) causes the gas valve, V_1, either to close or to open depending on the arithmetic sign of this difference.

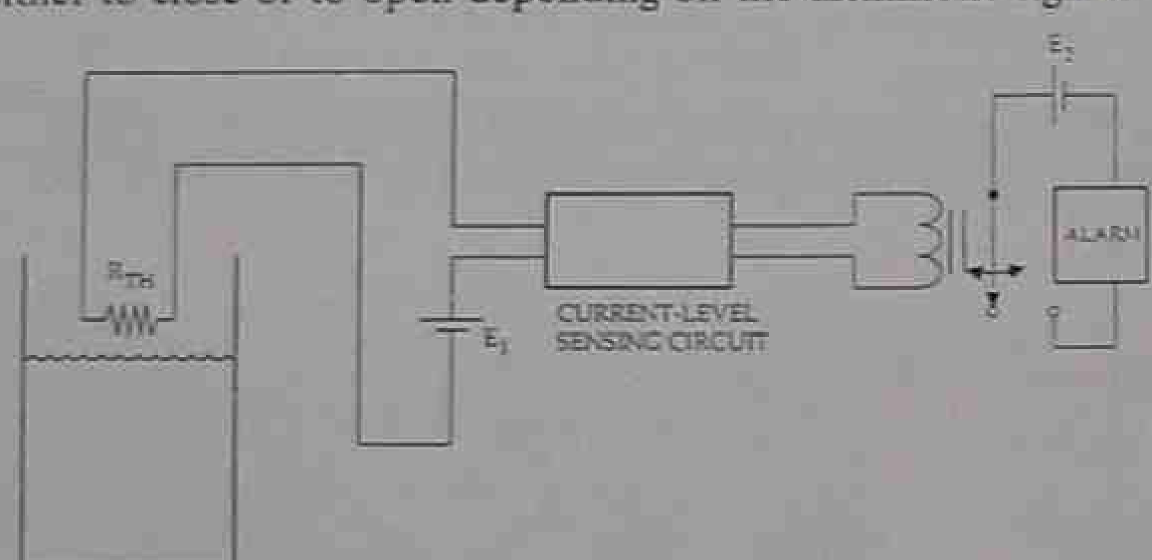


Figure 13-17 Thermistor being used as a level-indicating device.

It is the self-heat characteristic of thermistors that has created this next application. Thermistors are often used as level-indicating alarms, as depicted in Figure 13-17. A thermistor is adjusted for a particular height in a liquid-filled container that is being filled. The thermistor's height represents the maximum liquid height that is desirable. Since the thermistor is suspended in air, out of the liquid, its self-heat is sufficient enough to create a current flow in a current-sensing network to keep a holding coil energized on an alarm system. The alarm is off in this position. Should the liquid cover the thermistor, the self-heat will become dissipated within the liquid, causing the thermistor to cool, thereby increasing its resistance and lowering the current flow through it. This lower current is not enough to keep the holding coil to the alarm system energized; therefore, the relay opens, triggering the alarm circuit. The alarm circuit may also be wired to a control valve that opens a dump valve to the container so that the liquid will partially drain, causing the thermistor once again to become "self-heated," causing the alarm's relay coil to close once again.

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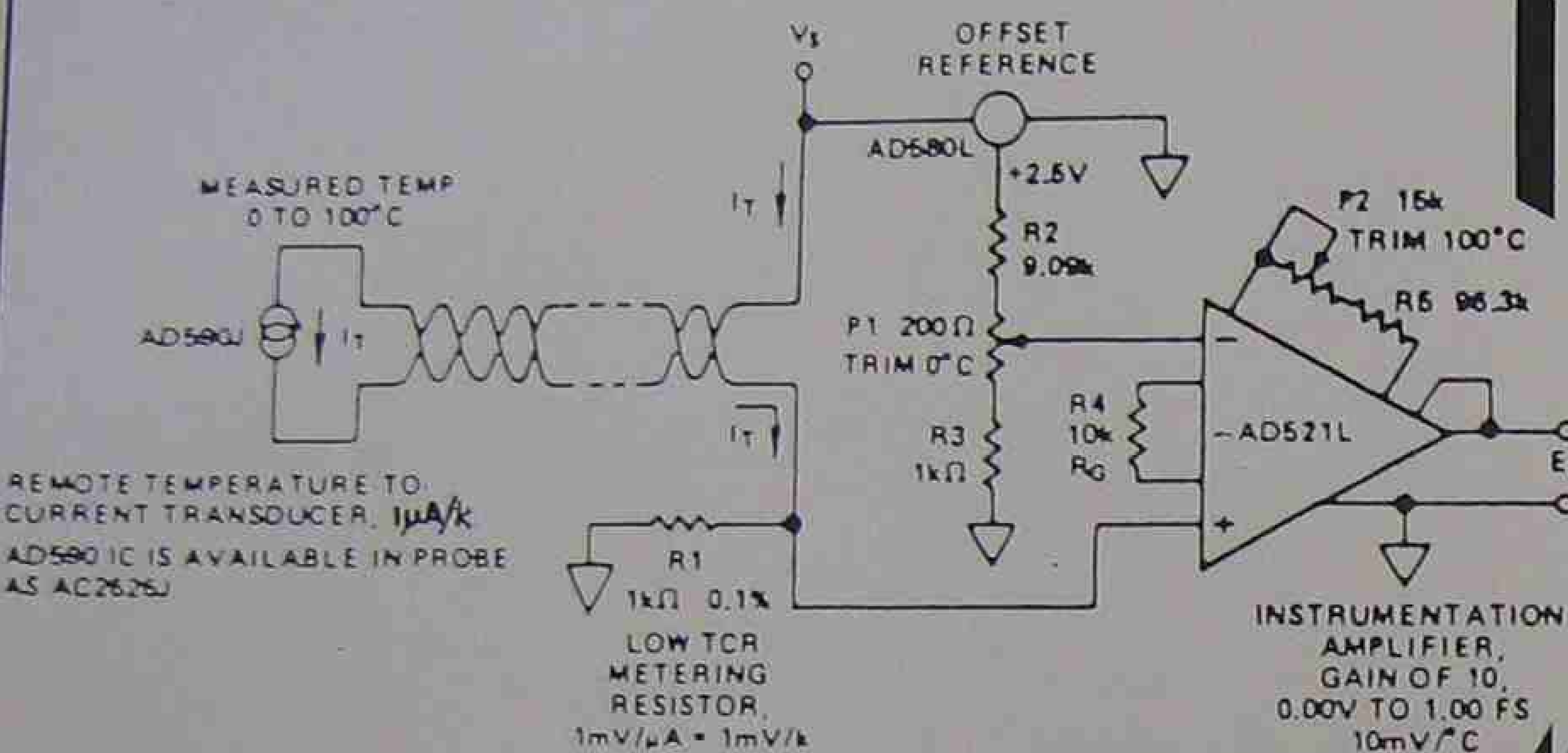


Figure 1. Thermometer Circuit

Orientation
Temperature Measurement Components

The AD590 and AD592 are two-terminal integrated circuit temperature transducers which produce an output current proportional to absolute temperature. The AD590 and AD592 have a standard 1μA/K output current which is inherently linear, therefore, no linearization is required.

Attention to all the detail is the key to success in most interface criteria. The following application is provided to illustrate the problems involved in designing circuits which measure physical pheonomenon.

In this application, there is a need to measure temperatures from 0 to +100°C, to within 1.0°C, at low cost, at a remote location several hundred feet from the instrumentation. The ambient temperature in the vicinity of the instrumentation is expected to be 25°C ±15°. A number of possible transducers will operate over the specified range, but the requirement for a remote measurement suggests the use of the current-output two-wire AD590 or AD592 semiconductor temperature sensor, because the current is unaffected by voltage drops and induced voltages.

Consulting the "Accuracies of the AD590"[1] we find that the AD590J, with two external trims, would be suitable; its maximum error over the 0 to 100°C range is 0.3°. This permits an allowance of 0.7° for all other errors. If a tighter tolerance were required, it would be worthwhile to consider using the AD590M with two trims, for an error *below* 0.05°C.

Since AD590 measures absolute temperature (its nominal output is 1μA/K), the output must be offset by 273.2μA in order to read out in degrees Celsius. The output of the AD590 flows through a 1kΩ resistance, developing a voltage of 1mV/K (Figure 1). The output of an AD580 2.5-volt reference is divided down by resistors to provide a 273.2mV offset, which is subtracted from the voltage across the 1kΩ resistor by an AD521 instrumentation amplifier. The AD521 provides a gain of 10.0, so that the output range, corresponding to 0 to 100°C, is 0 to 1.00V (10mV/°C).

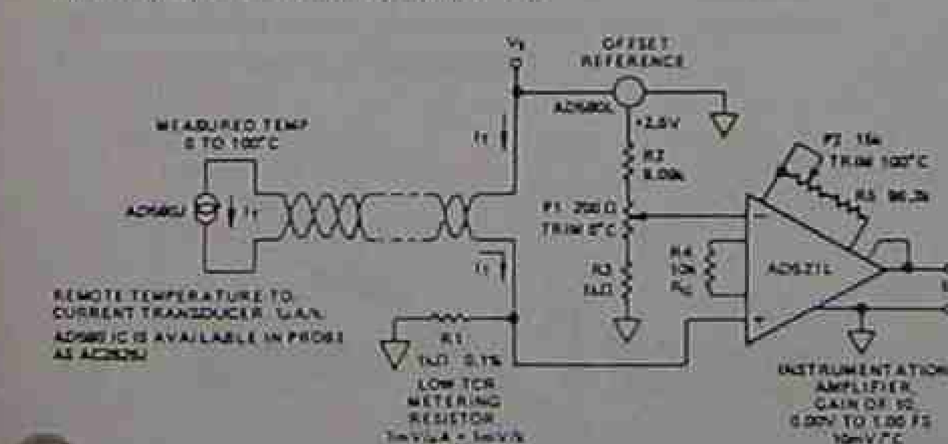


Figure 1. Thermometer Circuit

The desired system accuracy is to within 1.0°C; as noted, all errors other than that of the AD590 must contribute the equivalent of less than 0.7°. It will be helpful to assemble an error budget for the circuit, assessing the contributions of each of the elements (Table I). Errors will be expressed in degrees Celsius.

AD590 regulation. If the AD590 is excited by a voltage source of between 5 and 10V, the typical regulation is 0.2μA/V (0.2°C/V). With 1% source regulation, this contribution will be 0.01°C

AD590 linearity error. Total error for AD590J, over the 0 to 100°C range, with two trims, is 0.3°C. Those trims will be the gain and offset trims for the whole circuit, accounting for resistor and ratio errors, AD521L gain, offset and bias-current errors, AD580L voltage error, and the AD590J's calibration error 0.3°C

R1 temperature coefficient. Since R1 is responsible for the conversion of the AD590's current to voltage, high absolute accuracy is important. Consequently, we would expect to use a device having 10ppm/°C or less in this spot. For ±15°C, the maximum error is $373.2\mu A \times 10^{-5}/°C \times 15° = 0.06\mu A$ 0.06°C

(typical at 25°C and rated supply voltage unless noted otherwise)

Parameter	Condition	Specification
AD580L 2.5V VOLTAGE REFERENCE		
Output voltage	V_S = +15V	2.450V min, 2.550V max
Input voltage, operating		30V max, 7V min
Line regulation	$7V \le V_{IN} \le 30V$	2mV max
Temperature sensitivity	0 to 70°C	4.3mV max, 25ppm/°C typ
Noise	0.1 to 10Hz	60μV, p-p
Stability (drift with time)	long term	250μV (0.01%)
	per month	25μV (10ppm)
AD590J 1μA/K TEMPERATURE TRANSDUCER		
Output current	Nominal at 25°C (298.2k)	298.2μA
Input voltage, operating		30V max, 4V min
Calibration error	25°C, V_S = 5V	±5°C max
Linearity error	Two trims, 0 to 100°C range	0.1°C max
Repeatability	per month	0.1°C max
Long-term drift		0.1°C max
Noise spectral density		40pA/√Hz
Power-supply rejection	$+5V \le V_S \le +15V$	0.2μA/V
Operating range		−55°C to +150°C
AD521L DIFFERENTIAL INSTRUMENTATION AMPLIFIER		
Gain equation (volts/volt)	Nominal	$G = R_S/R_G$
Error from equation	Untrimmed	(±0.25 − 0.004G)%
Nonlinearity	±9V output	0.1% max
Gain tempco	0 to 70°C	±(1±0.05G)ppm/°C
Voltage offset	Input	1.0mV max
	Output	100mV max
Voltage offset tempco	Input, 0 to 70°C	2μV/°C max
	Output, 0 to 70°C	75μV/°C max
Voltage offset vs. supply	Input	3μV/%
	Output, untrimmed*	0.5mV/%
Bias current	25°C	40nA max
Bias current tempco	0 to 70°C	500pA/°C
Input impedance	Common mode	6 × 10^10Ω‖3.0pF
Common-mode rejection	G = 10, dc to 60Hz, 1kΩ source unbalance	94dB min
Voltage noise	G = 10, 0.1Hz to 10Hz, p-p, RTO	325μV

*Can be reduced by trimming the output offset.

Table I. Device Specifications Pertinent to the Analysis in the Text

[1] Accuracies of the AD590 Application Note, Analog Devices.

8

AD580 temperature coefficient. The specified tempco for the AD580L is 25ppm/°C typical (61ppm/°C max over the range 0 to 70°C). Since operation is over a narrow range, the typical value is most useful, unless the AD580 has a critical effect on the overall error. $25 \times 10^{-6}/°C \times 273mV \times 15° = 0.1mV$ — 0.10°C

Resistive divider tempco. The absolute values of R2 and R3 are of considerably less importance than their ability to track. 10ppm/°C is a reasonable value for tracking tempco. $10^{-5}/°C \times 273mV \times 15° = 0.04mV$ — 0.04°C

Common-mode error. At a gain of 10, the minimum common-mode error of the AD521L amplifier is 94dB, one part in 50,000 of the common-mode voltage (273mV), or 5μV (negligible) — 0.0°C

AD521 temperature coefficient. The specified input offset tempco for the AD521L is 2μV/°C max, and the output offset tempco is 75μV/°C max (7.5μV/°C, referred to the input), for a total of 9.5μV/°C R.T.I. $9.5\mu V/°C \times 15° = 143\mu V$ — 0.14°C

AD521 bias-current tempco. The maximum bias-current change is $500pA/°C \times 30°$ (range) = 15nA. The equivalent offset-voltage change is $15nA \times 1k\Omega = 15\mu V$ — 0.02°C

AD521 gain tempco. The circuit will be calibrated for correct output at 100°C by trimming of the gain of the AD521 at a 25°C ambient temperature. Variation of gain will cause output errors. The specified gain tempco at a gain of 10 for the AD521L is 3.5ppm/°C typical. If max is arbitrarily assumed to be ten times worse, and the resistors contribute 15ppm/°C additional, the maximum error will be $50 \times 10^{-6}/°C \times 100° \times 15° = 0.075°$ — 0.075°C

AD521 nonlinearity. The 0.1% nonlinearity specification applies for a ±9V output swing; for a 1V full-scale swing, it may be reasonable to expect a tenfold improvement, or a 1mV linearity error, equivalent to 0.1°C — 0.1°C

Total error (worst case) — 0.84°C

This means that, once the circuit has been calibrated at 0°C and 100°C (25°C ambient), the maximum error at any combination of measured and ambient temperatures can reasonably be expected to be less than 1°C.

If the summation were root-sum-of-squares, instead of worst-case, the error would come to less than 0.4°. This suggests that the design is quite conservative, since the probability of worst-case error is low; also (with some risk), it suggests that if an AD590M were used in the same design, temperature could be measured to within 0.25°C over the range. Naturally, every precaution should be taken to avoid additional errors attributable to either Murphy's or Natural Law. Aside from errors attributable to ambient temperature variations, this simple interface will require some form of protection from extraneous signals. Shielding and grounding should follow the practice suggested earlier in this book. In addition, capacitance across R1 will help reduce the effects of any ac currents induced in the twisted pair. Power supplies must be chosen to minimize error due to sensitivity of any of the elements to power-supply voltage changes, and bypassed to minimize coupling of interference through the power-supply leads.

ADDITIONAL INFORMATION

COLD JUNTION COMPENSATION AMPLIFIER – AD594/AD595

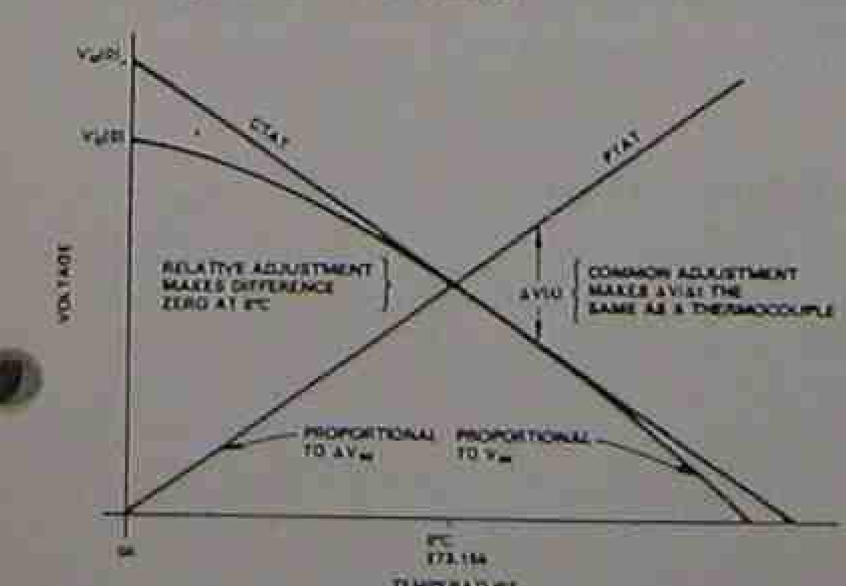
MONOLITHIC THERMOCOUPLE AMPLIFIER WITH COLD JUNCTION COMPENSATION

The AD594/AD595 is a complete instrumentation amplifier and thermocouple cold junction compensator on a monolithic chip. It combines an ice-point reference with a precalibrated amplifier to produce a high level (10mV/°C) output directly from a thermocouple signal. Pin-strapping options allow it to be used as a linear amplifier-compensator or as a switched output set-point controller using either fixed or remote set-point control. It can be used to amplify its compensation voltage directly, thereby converting it to a stand-alone Celsius transducer with a low-impedance voltage output.

The AD596 is a low cost instrumentation amplifier and thermocouple cold junction compensator for set-point control applications. The AD596 is packaged in a rugged hermetic 10 pin TO-100 metal can and its cold junction compensation circuit is trimmed so that it will remain accurate over a wide ambient temperature range, internal architecture.

It is commonly known that the characteristics of bipolar junction transistors are temperature sensitive, and it is a usual object of linear design to suppress this sensitivity. In the case of the AD594/AD595, however, certain well behaved and repeatable temperature dependent parameters are exploited to produce the cold junction compensation voltage. When two transistors are operated at different emitter current densities, the difference in their base-emitter voltages will be *proportional to absolute temperature* or PTAT. The base-emitter voltages of a single transistor falls with rising temperature in a way that can be extrapolated to a known voltage at absolute zero. This voltage *complements* a PTAT voltage with respect to the known bandgap voltage and is referred to as CTAT.

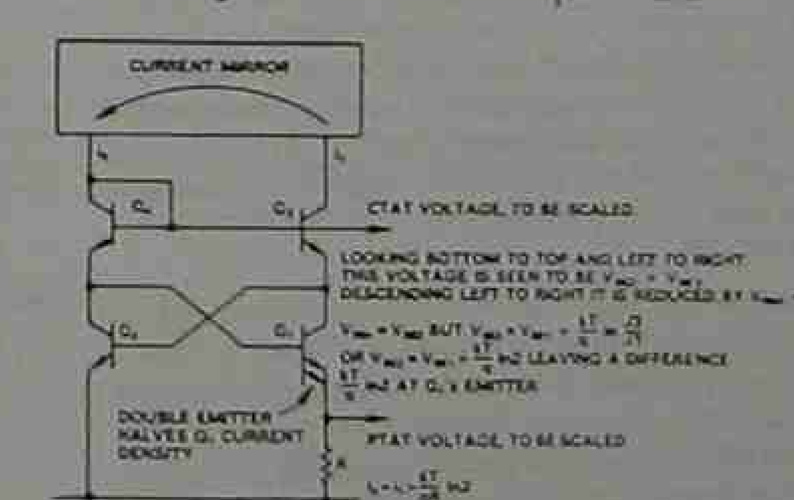
Although these two voltages are predictably related to absolute temperature, their difference can be related to Celsius temperature as shown here.



Ice-Point Compensation from the Difference of a PTAT and a CTAT Voltage

Two temperature sensitive voltages can be derived from the transistor base-emitter characteristics which can be scaled so that their difference approximates the output of an ice referenced thermocouple measuring the IC temperature. This difference is zero at zero Celsius and increases more-or-less linearly with temperature. These voltages are produced by four transistors in the AD594/AD595. A current mirror is used to force a pair of series connected transistors (Q_2, Q_4, in the figure below) to operate at the same current as another series connected pair (Q_1, Q_3). Three of these transistors are the same size and therefore operate at equal current densities. Consequently, they have the same base-emitter voltage. The fourth transistor is larger than the others so that at the same current it operates at lower current density. This implies that it has a lower base-emitter voltage. The base-emitter junctions of the four transistors connect in a loop which is completed by a resistor. Two of the voltages are connected to subtract from the others so that the net voltage across the resistor is just the difference between the base-emitter voltage of the differently sized transistors.

As noted before, this voltage will be PTAT and is scaled to the proper magnitude by a thin film network in the AD594/AD595. It is also possible to extract the sum of the two base-emitter voltages from this loop. This sum is CTAT and when properly scaled makes up the other temperature sensitive voltage for the Ice-Point Compensation.



A Cross-Connected Transistor Quad Provides CTAT Voltage in the Form of $2V_{BE}S$ and PTAT Voltage from the Difference of $V_{BE}S$

8

ANALOG DEVICES

Two-Terminal IC Temperature Transducer

AD590*

FEATURES
- Linear Current Output: 1µA/K
- Wide Range: -55°C to +150°C
- Probe Compatible Ceramic Sensor Package
- Two-Terminal Device: Voltage In/Current Out
- Laser Trimmed to ±0.5°C Calibration Accuracy (AD590M)
- Excellent Linearity: ±0.3°C Over Full Range (AD590M)
- Wide Power Supply Range: +4V to +30V
- Sensor Isolation from Case

AD590 FUNCTIONAL BLOCK DIAGRAM

TO-52
BOTTOM VIEW

8

PRODUCT DESCRIPTION

The AD590 is a two-terminal integrated circuit temperature transducer which produces an output current proportional to absolute temperature. For supply voltages between +4V and +30V the device acts as a high impedance, constant current regulator passing 1µA/K. Laser trimming of the chip's thin film resistors is used to calibrate the device to 298.2µA output at 298.2K (+25°C).

The AD590 should be used in any temperature sensing application below +150°C in which conventional electrical temperature sensors are currently employed. The inherent low cost of a monolithic integrated circuit combined with the elimination of support circuitry makes the AD590 an attractive alternative for many temperature measurement situations. Linearization circuitry, precision voltage amplifiers, resistance measuring circuitry and cold junction compensation are not needed in applying the AD590.

In addition to temperature measurement, applications include temperature compensation or correction of discrete components, biasing proportional to absolute temperature, flow rate measurement, level detection of fluids and anemometry. The AD590 is available in chip form making it suitable for hybrid circuits and fast temperature measurements in protected environments.

The AD590 is particularly useful in remote sensing applications. The device is insensitive to voltage drops over long lines due to its high impedance current output. Any well-insulated twisted pair is sufficient for operation hundreds of feet from the receiving circuitry. The output characteristics also make the AD590 easy to multiplex: the current can be switched by a CMOS multiplexer or the supply voltage can be switched by a logic gate output.

PRODUCT HIGHLIGHTS

1. The AD590 is a calibrated two terminal temperature sensor requiring only a dc voltage supply (+4V to +30V). Costly transmitters, filters, lead wire compensation and linearization circuits are all unnecessary in applying the device.
2. State-of-the-art laser trimming at the wafer level in conjunction with extensive final testing insures that AD590 units are easily interchangeable.
3. Superior interference rejection results from the output being a current rather than a voltage. In addition, power requirements are low (1.5mW's @ 5V @ +25°C). These features make the AD590 easy to apply as a remote sensor.
4. The high output impedance (>10MΩ) provides excellent rejection of supply voltage drift and ripple. For instance, changing the power supply from 5V to 10V results in only a 1µA maximum current change, or 1°C equivalent error.
5. The AD590 is electrically durable: it will withstand a forward voltage up to 44V and a reverse voltage of 20V. Hence, supply irregularities or pin reversal will not damage the device.

*Covered by Patent No. 4,123,698

SPECIFICATIONS (@ +25°C and V_S = 5V unless otherwise noted)

Model	AD590I Min	AD590I Typ	AD590I Max	AD590J Min	AD590J Typ	AD590J Max	AD590K Min	AD590K Typ	AD590K Max	Units
ABSOLUTE MAXIMUM RATINGS										
Forward Voltage (E+ to E−)			+44			+44			+44	Volts
Reverse Voltage (E+ to E−)			−20			−20			−20	Volts
Breakdown Voltage (Case to E+ or E−)			±200			±200			±200	Volts
Rated Performance Temperature Range[1]	−55		+150	−55		+150	−55		+150	°C
Storage Temperature Range[1]	−65		+155	−65		+155	−65		+155	°C
Lead Temperature (Soldering, 10 sec)			+300			+300			+300	°C
POWER SUPPLY										
Operating Voltage Range	+4		+30	+4		+30	+4		+30	Volts
OUTPUT										
Nominal Current Output @ +25°C (298.2K)		298.2			298.2			298.2		µA
Nominal Temperature Coefficient		1			1			1		µA/K
Calibration Error @ +25°C			±10			±5.0		—	±2.5	°C
Absolute Error (over rated performance temperature range)										
Without External Calibration Adjustment			±20			±10			±5.5	°C
With +25°C Calibration Error Set to Zero			±5.8			±3.0			±2.0	°C
Nonlinearity			±3.0			±1.5			±0.8	°C
Repeatability[2]			±0.1			±0.1			±0.1	°C
Long Term Drift[3]			±0.1			±0.1			±0.1	°C
Current Noise		40			40			40		$pA/\sqrt{Hz}$
Power Supply Rejection										
$+4V \le V_S \le +5V$		0.5			0.5			0.5		µA/V
$+5V \le V_S \le +15V$		0.2			0.2			0.2		µA/V
$+15V \le V_S \le +30V$		0.1			0.1			0.1		µA/V
Case Isolation to Either Lead		10^{10}			10^{10}			10^{10}		Ω
Effective Shunt Capacitance		100			100			100		pF
Electrical Turn-On Time		20			20			20		µs
Reverse Bias Leakage Current[4] (Reverse Voltage = 10V)		10			10			10		pA
PACKAGE OPTION[5]										
"H" Package: TO-52		AD590IH			AD590JH			AD590KH		
"F" Package: Flat Pack (F2A)		AD590IF			AD590JF			AD590KF		

NOTES

[1] The AD590 has been used at −100°C and +200°C for short periods of measurement with no physical damage to the device. However, the absolute errors specified apply to only the rated performance temperature range.

[2] Maximum deviation between +25°C readings after temperature cycling between −55°C and +150°C; guaranteed not tested.

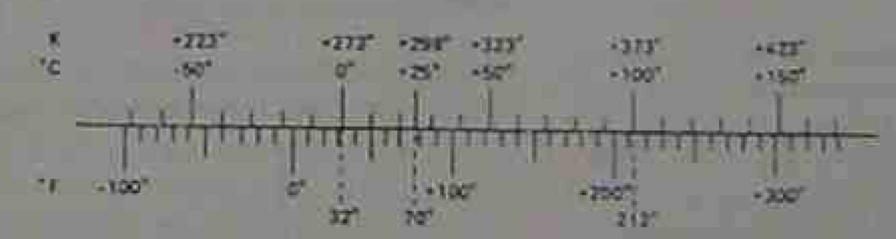
[3] Conditions: constant +5V, constant +125°C; guaranteed, not tested.

[4] Leakage current doubles every 10°C.

[5] See Section 19 for package outline information.

Specifications subject to change without notice.

Specifications shown in boldface are tested on all production units at final electrical test. Results from those tests are used to calculate outgoing quality levels. All min and max specifications are guaranteed, although only those shown in boldface are tested on all production units.



TEMPERATURE SCALE CONVERSION EQUATIONS

$$°C = \frac{5}{9}(°F - 32) \qquad K = °C + 273.15$$

$$°F = \frac{9}{5}°C + 32 \qquad °R = °F + 459.7$$

Model	AD590L Min	AD590L Typ	AD590L Max	AD590M Min	AD590M Typ	AD590M Max	Units
ABSOLUTE MAXIMUM RATINGS							
Forward Voltage (E+ to E−)			+44			+44	Volts
Reverse Voltage (E+ to E−)			−20			−20	Volts
Breakdown Voltage (Case to E+ or E−)			±200			±200	Volts
Rated Performance Temperature Range	−55		+150	−55		+150	°C
Storage Temperature Range	−65		+155	−65		+155	°C
Lead Temperature (Soldering, 10sec)			+300			+300	°C
POWER SUPPLY							
Operating Voltage Range	+4		+30	+4		+30	Volts
OUTPUT							
Nominal Current Output @ +25°C (298.2K)		298.2			298.2		μA
Nominal Temperature Coefficient		1			1		μA/K
Calibration Error @ +25°C			±1.0			±0.5	°C
Absolute Error (over rated performance temperature range)							
Without External Calibration Adjustment			±3.0			±1.7	°C
With +25°C Calibration Error Set to Zero			±1.6			±1.0	°C
Nonlinearity			±0.4			±0.3	°C
Repeatability			±0.1			±0.1	°C
Long Term Drift			±0.1			±0.1	°C
Current Noise		40			40		pA/√Hz
Power Supply Rejection							
+4V ≤ V_S ≤ +5V		0.5			0.5		μA/V
+5V ≤ V_S ≤ +15V		0.2			0.2		μA/V
+15V ≤ V_S ≤ +30V		0.1			0.1		μA/V
Case Isolation to Either Lead		10^{10}			10^{10}		Ω
Effective Shunt Capacitance		100			100		pF
Electrical Turn-On Time		20			20		μs
Reverse Bias Leakage Current (Reverse Voltage = 10V)		10			10		pA
PACKAGE OPTIONS							
"H" Package TO-52		AD590LH			AD590MH		
"F" Package Flat Pack (F2a)		AD590LF			AD590MF		

CIRCUIT DESCRIPTION[3]

The AD590 uses a fundamental property of the silicon transistors from which it is made to realize its temperature proportional characteristic: if two identical transistors are operated at a constant ratio of collector current densities, r, then the difference in their base-emitter voltages will be (kT/q)(ln r). Since both k, Boltzman's constant and q, the charge of an electron, are constant, the resulting voltage is directly proportional to absolute temperature (PTAT).

In the AD590, this PTAT voltage is converted to a PTAT current by low temperature coefficient thin film resistors. The total current of the device is then forced to be a multiple of this PTAT current. Referring to Figure 1, the schematic diagram of the AD590, Q8 and Q11 are the transistors that produce the PTAT voltage. R5 and R6 convert the voltage to current. Q10, whose collector current tracks the collector currents in Q9 and Q11, supplies all the bias and substrate leakage current for the rest of the circuit, forcing the total current to be PTAT. R5 and R6 are laser trimmed on the wafer to calibrate the device at +25°C.

Figure 2 shows the typical V–I characteristic of the circuit at +25°C and the temperature extremes.

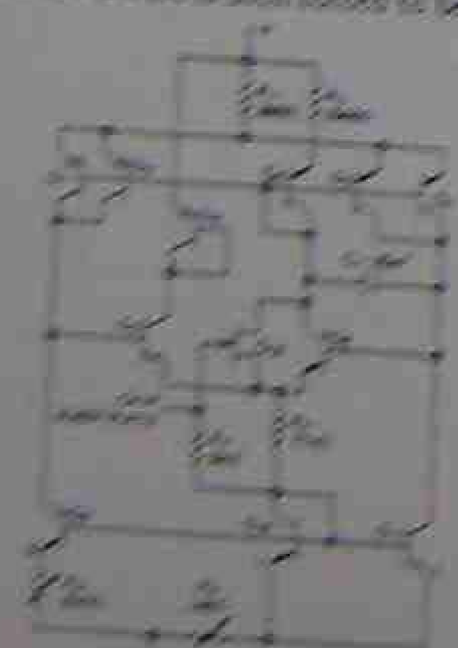


Figure 1. Schematic Diagram

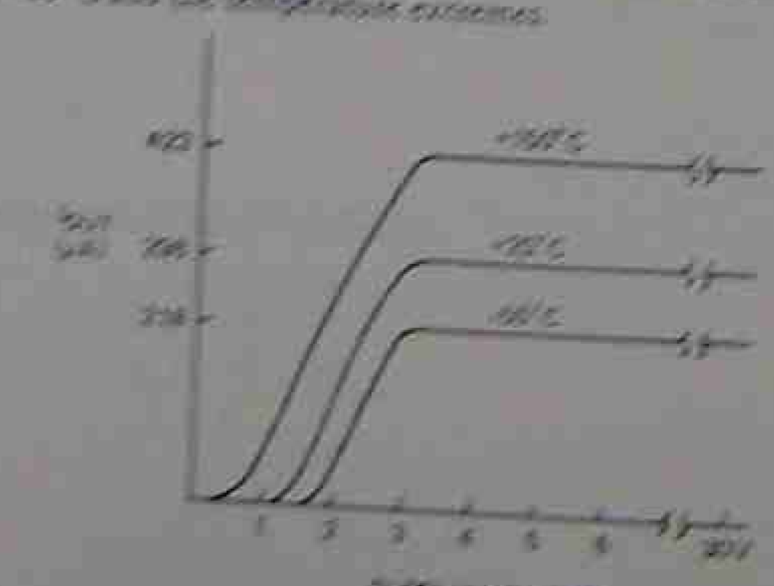


Figure 2. V–I Plot

[3]For a more detailed circuit description see M.P. Timko, "A Two-Terminal IC Temperature Transducer," IEEE J. Solid State Circuits, Vol. SC-11, p. 784-788, Dec. 1976.

8

EXPLANATION OF TEMPERATURE SENSOR SPECIFICATIONS

The way in which the AD590 is specified makes it easy to apply in a wide variety of different applications. It is important to understand the meaning of the various specifications and the effects of supply voltage and thermal environment on accuracy.

The AD590 is basically a PTAT (proportional to absolute temperature)[4] current regulator. That is, the output current is equal to a scale factor times the temperature of the sensor in degrees Kelvin. This scale factor is trimmed to 1μA/K at the factory, by adjusting the indicated temperature (i.e. the output current) to agree with the actual temperature. This is done with 5V across the device at a temperature within a few degrees of 25°C (298.2K). The device is then packaged and tested for accuracy over temperature.

CALIBRATION ERROR

At final factory test the difference between the indicated temperature and the actual temperature is called the calibration error. Since this is a scale factor error, its contribution to the total error of the device is PTAT. For example, the effect of the 1°C specified maximum error of the AD590L varies from 0.73°C at −55°C to 1.42°C at 150°C. Figure 3 shows how an exaggerated calibration error would vary from the ideal over temperature.

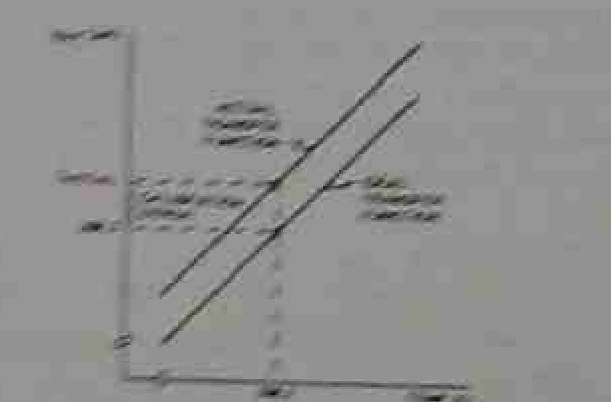


Figure 3. Calibration Error vs. Temperature

The calibration error is a primary contributor to maximum total error in all AD590 grades. However, since it is a scale factor error, it is particularly easy to trim. Figure 4 shows the most elementary way of accomplishing this. To trim this circuit the temperature of the AD590 is measured by a reference temperature sensor and R is trimmed so that V_T = 1mV/K at that temperature. Note that when this error is trimmed out at one temperature, its effect is zero over the entire temperature range. In most applications there is a current to voltage conversion resistor (or, as with a current input ADC, a reference) that can be trimmed for scale factor adjustment.

Figure 4. One Temperature Trim

[4]T(°C) = T(K) − 273.2; Zero on the Kelvin scale is "absolute zero"; there is no lower temperature.

ERROR VERSUS TEMPERATURE: WITH CALIBRATION ERROR TRIMMED OUT

Each AD590 is also tested for error over the temperature range with the calibration error trimmed out. This specification could also be called the "variance from PTAT" since it is the maximum difference between the actual current over temperature and a PTAT multiplication of the actual current at 25°C. This error consists of a slope error and some curvature, mostly at the temperature extremes. Figure 5 shows a typical AD590K temperature curve before and after calibration error trimming.

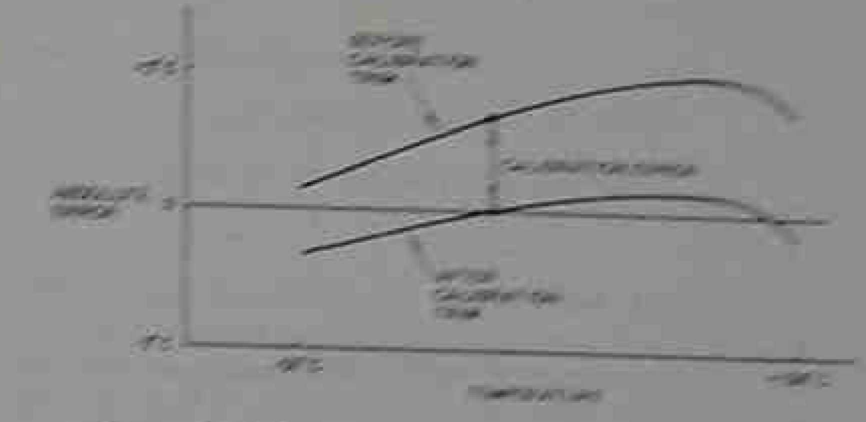


Figure 5. Effect of Scale Factor Trim on Accuracy

ERROR VERSUS TEMPERATURE: NO USER TRIMS

Using the AD590 by simply measuring the current, the total error is the "variance from PTAT" described above plus the effect of the calibration error over temperature. For example the AD590L maximum total error varies from 2.33°C at −55°C to 3.02°C at 150°C. For simplicity, only the larger figure is shown on the specification page.

NONLINEARITY

Nonlinearity as it applies to the AD590 is the maximum deviation of current over temperature from a best-fit straight line. The nonlinearity of the AD590 over the −55°C to +150°C range is superior to all conventional electrical temperature sensors such as thermocouples, RTD's and thermistors. Figure 6 shows the nonlinearity of the typical AD590K from Figure 5.

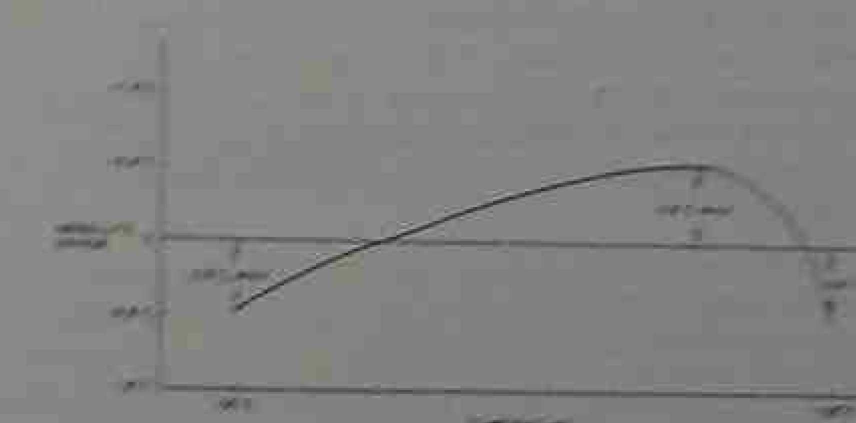


Figure 6. Nonlinearity

Figure 7A shows a circuit in which the nonlinearity is the major contributor to error over temperature. The circuit is trimmed by adjusting R_1 for a 0V output with the AD590 at 0°C. R_2 is then adjusted for 10V out with the sensor at 100°C. Other pairs of temperatures may be used with this procedure as long as they are measured accurately by a reference sensor. Note that for +15V output (150°C) the V+ of the op amp must be greater than 17V. Also note that V− should be at least −4V: if V− is ground there is no voltage applied across the device.

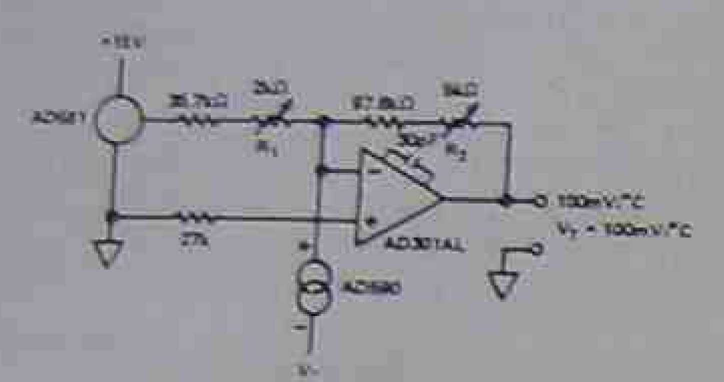


Figure 7A. Two Temperature Trim

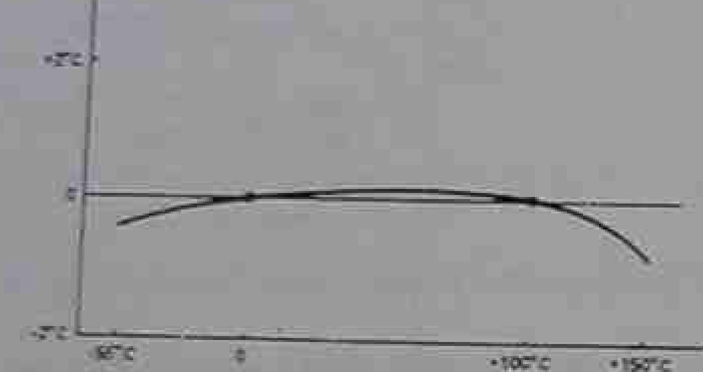


Figure 7B. Typical Two-Trim Accuracy

VOLTAGE AND THERMAL ENVIRONMENT EFFECTS

The power supply rejection specifications show the maximum expected change in output current versus input voltage changes. The insensitivity of the output to input voltage allows the use of unregulated supplies. It also means that hundreds of ohms of resistance (such as a CMOS multiplexer) can be tolerated in series with the device.

It is important to note that using a supply voltage other than 5V does not change the PTAT nature of the AD590. In other words, this change is equivalent to a calibration error and can be removed by the scale factor trim (see previous page).

The AD590 specifications are guaranteed for use in a low thermal resistance environment with 5V across the sensor. Large changes in the thermal resistance of the sensor's environment will change the amount of self-heating and result in changes in the output which are predictable but not necessarily desirable.

The thermal environment in which the AD590 is used determines two important characteristics: the effect of self heating and the response of the sensor with time.

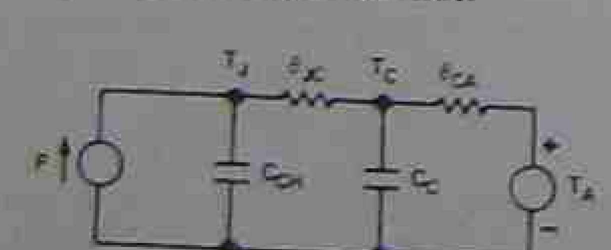


Figure 8. Thermal Circuit Model

Figure 8 is a model of the AD590 which demonstrates these characteristics. As an example, for the TO-52 package, θ_{JC} is the thermal resistance between the chip and the case, about 26°C/watt. θ_{CA} is the thermal resistance between the case and its surroundings and is determined by the characteristics of the thermal connection. Power source P represents the power dissipated on the chip. The rise of the junction temperature, T_J, above the ambient temperature T_A is:

$$T_J - T_A = P(\theta_{JC} + \theta_{CA}). \qquad \text{Eq. 1}$$

Table I gives the sum of θ_{JC} and θ_{CA} for several common thermal media for both the "H" and "F" packages. The heatsink used was a common clip-on. Using Equation 1, the temperature rise of an AD590 "H" package in a stirred bath at +25°C, when driven with a 5V supply, will be 0.06°C. However, for the same conditions in still air the temperature rise is 0.72°C. For a given supply voltage, the temperature rise varies with the current and is PTAT. Therefore, if an application circuit is trimmed with the sensor in the same thermal environment in which it will be used, the scale factor trim compensates for this effect over the entire temperature range.

MEDIUM	$\theta_{JC} + \theta_{CA}$ (°C/watt) H	$\theta_{JC} + \theta_{CA}$ (°C/watt) F	τ (sec) (Note 3) H	τ (sec) (Note 3) F
Aluminum Block	30	10	0.6	0.1
Stirred Oil[1]	42	60	1.4	0.6
Moving Air[2]				
With Heat Sink	45	–	5.0	–
Without Heat Sink	115	190	13.5	10.0
Still Air				
With Heat Sink	191	–	108	–
Without Heat Sink	480	650	60	30

[1] Note: τ is dependent upon velocity of oil; average of several velocities listed above.
[2] Air velocity ≈ 9ft/sec.
[3] The time constant is defined as the time required to reach 63.2% of an instantaneous temperature change.

Table I. Thermal Resistances

The time response of the AD590 to a step change in temperature is determined by the thermal resistances and the thermal capacities of the chip, C_{CH}, and the case, C_C. C_{CH} is about 0.04 watt-sec/°C for the AD590. C_C varies with the measured medium since it includes anything that is in direct thermal contact with the case. In most cases, the single time constant exponential curve of Figure 9 is sufficient to describe the time response, T(t). Table I shows the effective time constant, τ, for several media.

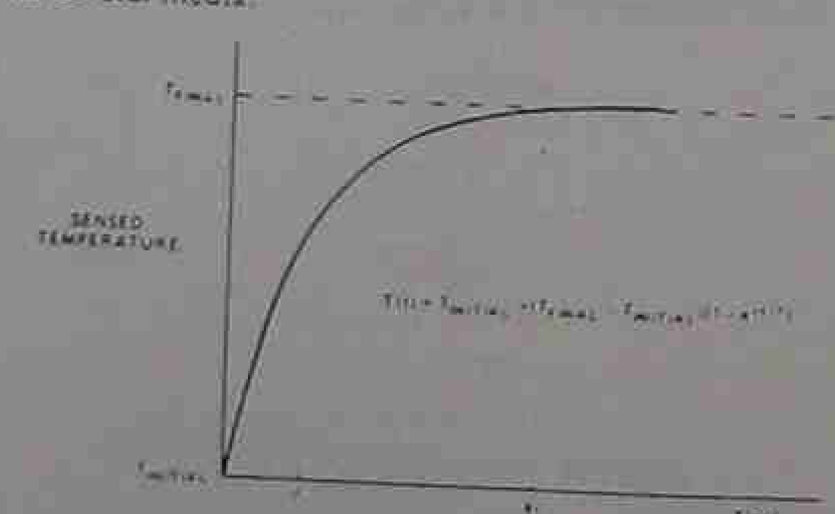


Figure 9. Time Response Curve

GENERAL APPLICATIONS

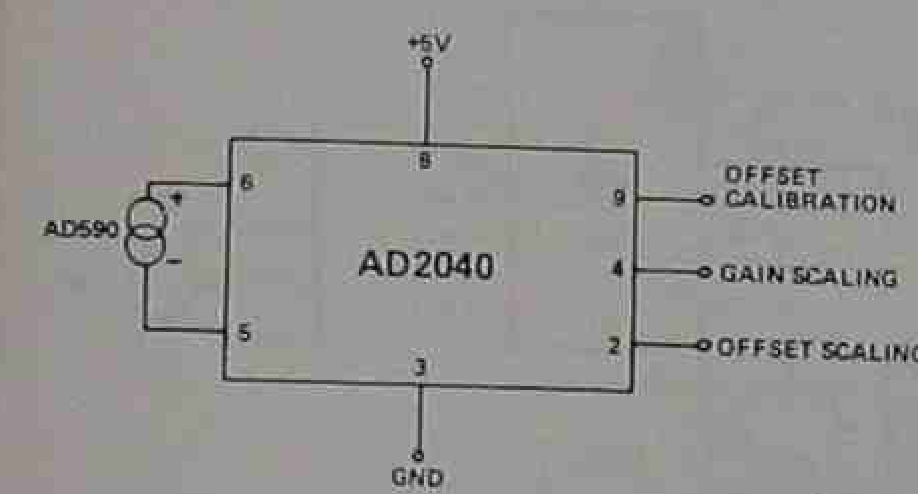


Figure 10. Variable Scale Display

Figure 10 demonstrates the use of a low-cost Digital Panel Meter for the display of temperature on either the Kelvin, Celsius or Fahrenheit scales. For Kelvin temperature Pins 9, 4 and 2 are grounded; and for Fahrenheit temperature Pins 4 and 2 are left open.

The above configuration yields a 3 digit display with 1°C or 1°F resolution, in addition to an absolute accuracy of ±2.0°C over the -55°C to +125°C temperature range if a one-temperature calibration is performed on an AD590K, L, or M.

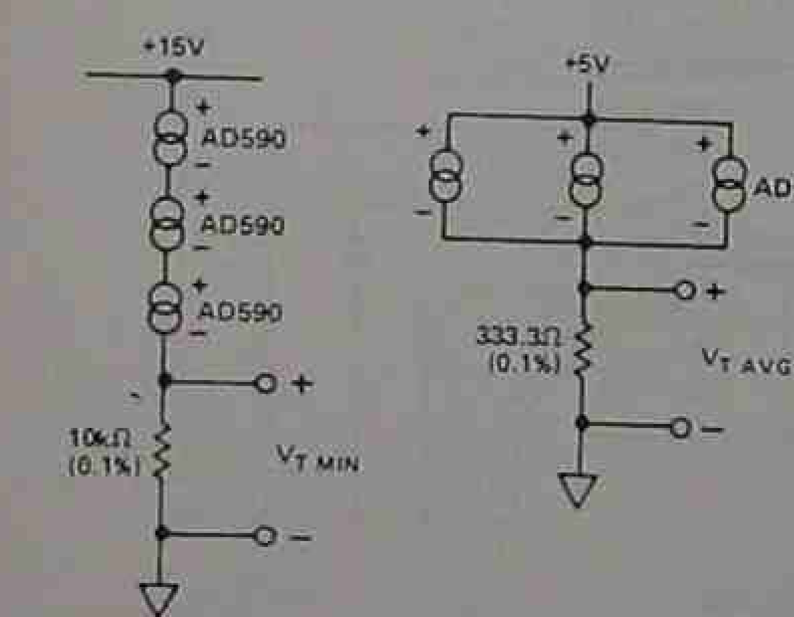


Figure 11. Series & Parallel Connection

Connecting several AD590 units in series as shown in Figure 11 allows the minimum of all the sensed temperatures to be indicated. In contrast, using the sensors in parallel yields the average of the sensed temperatures.

The circuit of Figure 12 demonstrates one method by which differential temperature measurements can be made. R_1 and R_2 can be used to trim the output of the op amp to indicate a desired temperature difference. For example, the inherent offset between the two devices can be trimmed in. If V+ and V– are radically different, then the difference in internal dissipation will cause a differential internal temperature rise. This effect can be used to measure the ambient thermal resistance seen by the sensors in applications such as fluid level detectors or anemometry.

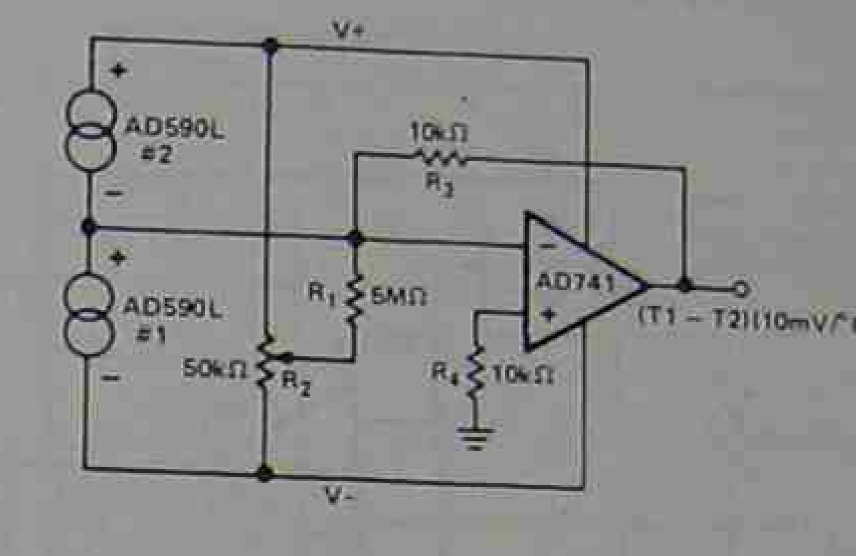


Figure 12. Differential Measurements

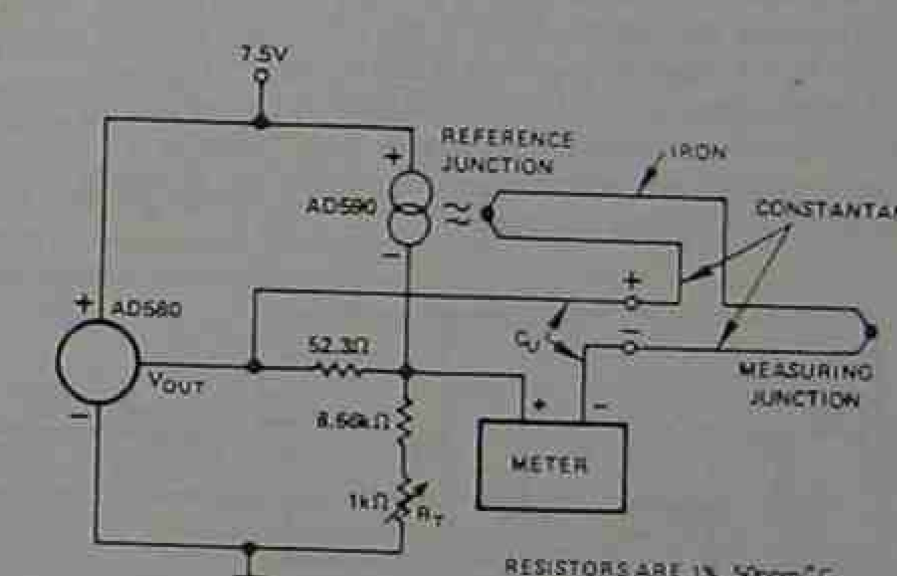


Figure 13. Cold Junction Compensation Circuit for Type J Thermocouple

Figure 13 is an example of a cold junction compensation circuit for a Type J Thermocouple using the AD590 to monitor the reference junction temperature. This circuit replaces an ice-bath as the thermocouple reference for ambient temperatures between +15°C and +35°C. The circuit is calibrated by adjusting R_T for a proper meter reading with the measuring junction at a known reference temperature and the circuit near +25°C. Using components with the T.C.'s as specified in Figure 13, compensation accuracy will be within ±0.5°C for circuit temperatures between +15°C and +35°C. Other thermocouple types can be accommodated with different resistor values. Note that the T.C.'s of the voltage reference and the resistors are the primary contributors to error.

Applying the AD590

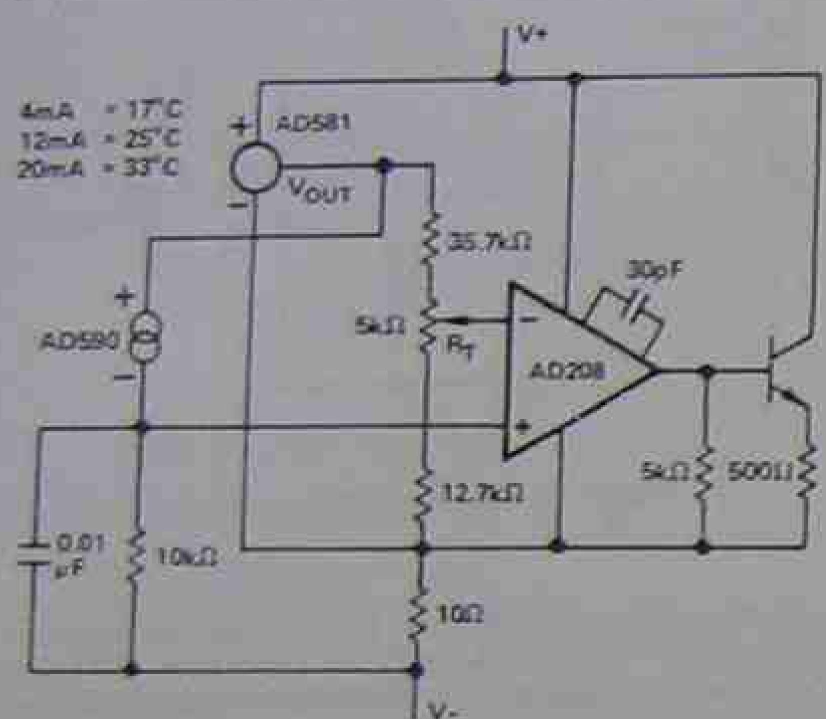


Figure 14. 4 to 20mA Current Transmitter

Figure 14 is an example of a current transmitter designed to be used with 40V, 1kΩ systems; it uses its full current range of 4mA to 20mA for a narrow span of measured temperatures. In this example the 1µA/K output of the AD590 is amplified to 1mA/°C and offset so that 4mA is equivalent to 17°C and 20mA is equivalent to 33°C. R_T is trimmed for proper reading at an intermediate reference temperature. With a suitable choice of resistors, any temperature range within the operating limits of the AD590 may be chosen.

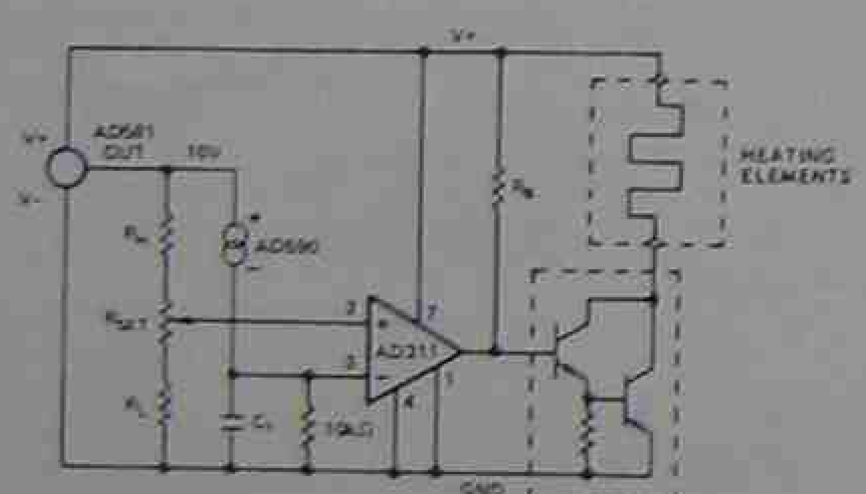


Figure 15. Simple Temperature Control Circuit

Figure 15 is an example of a variable temperature control circuit (thermostat) using the AD590. R_H and R_L are selected to set the high and low limits for R_{SET}. R_{SET} could be a simple pot, a calibrated multi-turn pot or a switched resistive divider. Powering the AD590 from the 10V reference isolates the AD590 from supply variations while maintaining a reasonable voltage (~7V) across it. Capacitor C_1 is often needed to filter extraneous noise from remote sensors. R_B is determined by the β of the power transistor and the current requirements of the load.

Figure 16 shows how the AD590 can be configured with an 8-bit DAC to produce a digitally controlled set point. This

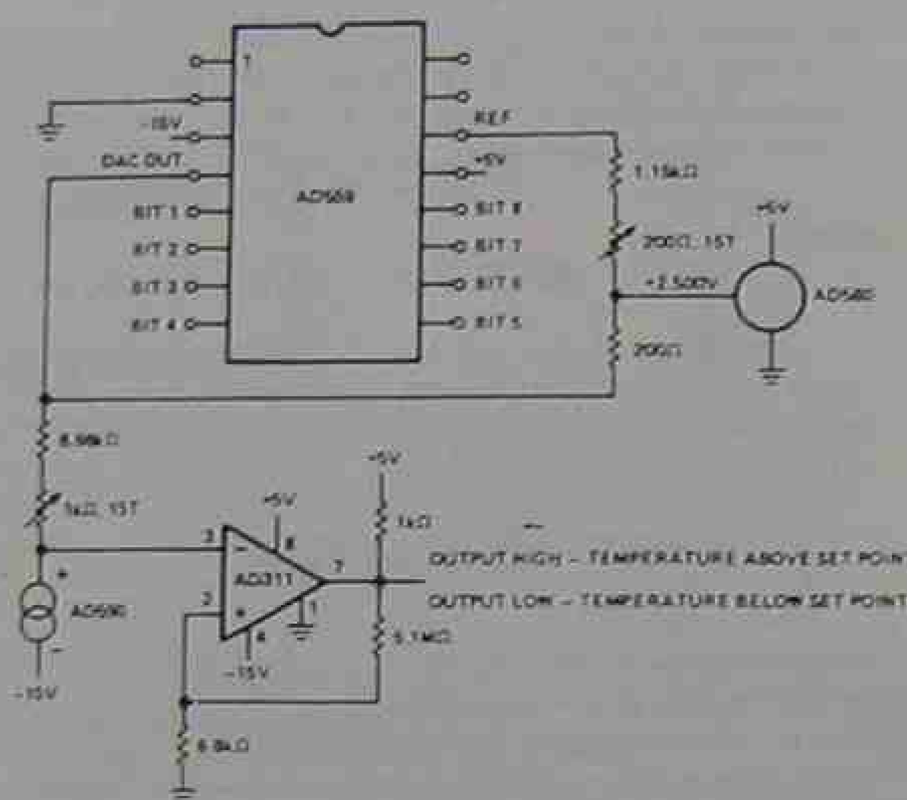


Figure 16. DAC Set Point

particular circuit operates from 0 (all inputs high) to +51°C (all inputs low) in 0.2°C steps. The comparator is shown with 1°C hysteresis which is usually necessary to guard-band for extraneous noise; omitting the 5.1MΩ resistor results in no hysteresis.

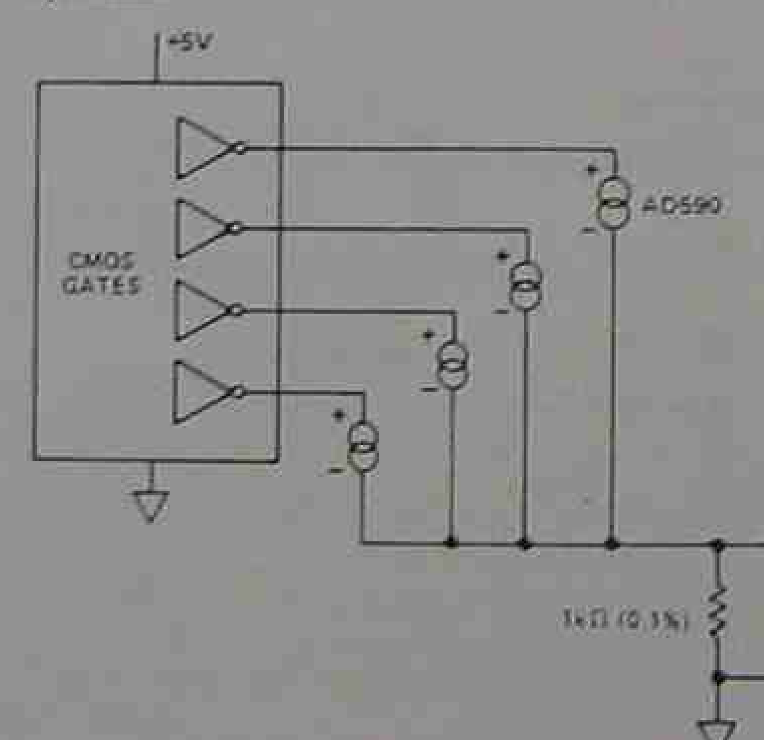


Figure 17. AD590 Driven from CMOS Logic

The voltage compliance and the reverse blocking characteristic of the AD590 allows it to be powered directly from +5V CMOS logic. This permits easy multiplexing, switching or pulsing for minimum internal heat dissipation. In Figure 17 any AD590 connected to a logic high will pass a signal current through the current measuring circuitry while those connected to a logic zero will pass insignificant current. The outputs used to drive the AD590's may be employed for other purposes, but the additional capacitance due to the AD590 should be taken into account.

8

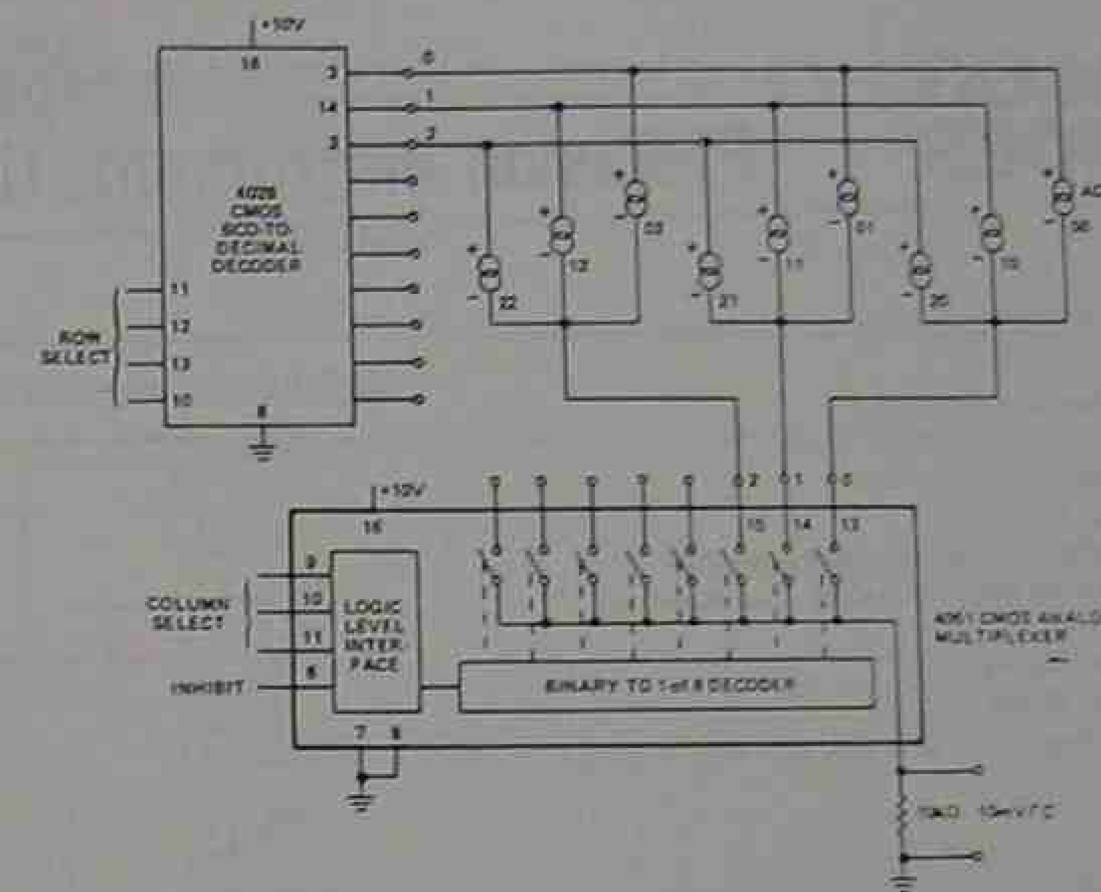


Figure 18. Matrix Multiplexer

CMOS Analog Multiplexers can also be used to switch AD590 current. Due to the AD590's current mode, the resistance of such switches is unimportant as long as 4V is maintained across the transducer. Figure 18 shows a circuit which combines the principal demonstrated in Figure 17 with an 8 channel CMOS Multiplexer. The resulting circuit can select one of eighty sensors over only 18 wires with a 7 bit binary word. The inhibit input on the multiplexer turns all sensors off for minimum dissipation while idling.

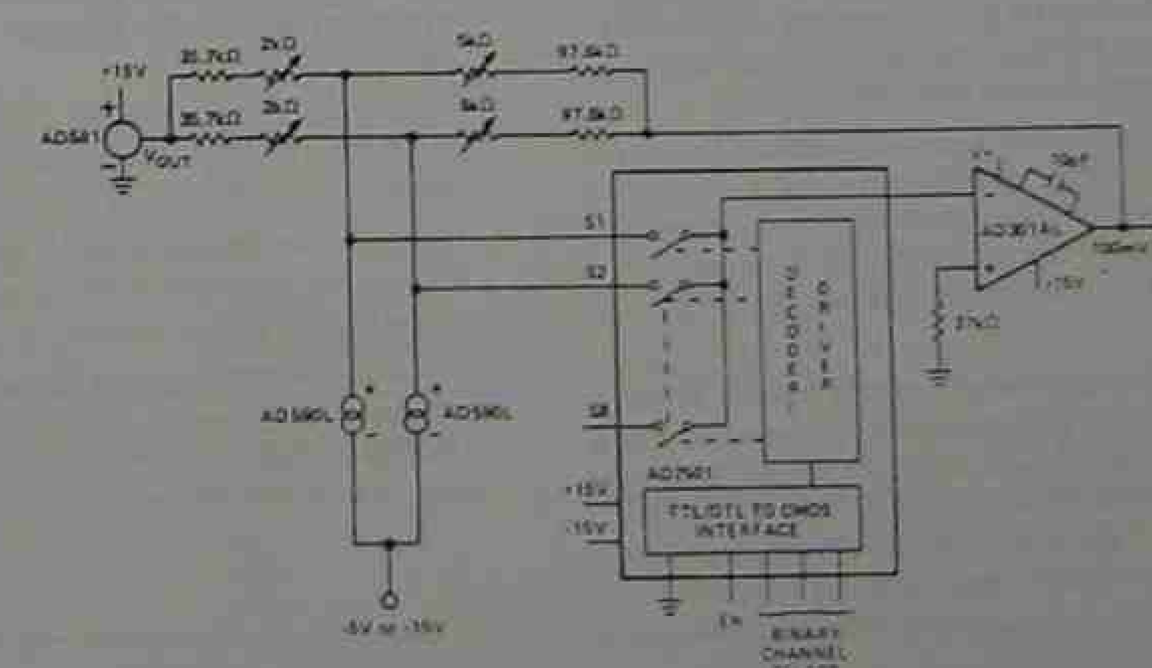


Figure 19. 8-Channel Multiplexer

Figure 19 demonstrates a method of multiplexing the AD590 in the two-trim mode (Figure 7). Additional AD590's and their associated resistors can be added to multiplex up to 8 channels of ±0.5°C absolute accuracy over the temperature range of -55°C to +125°C. The high temperature restriction of +125°C is due to the output range of the op amps; output to +150°C can be achieved by using a +20V supply for the op amp.

ANALOG DEVICES

Integrated Circuit Precision Instrumentation Amplifier

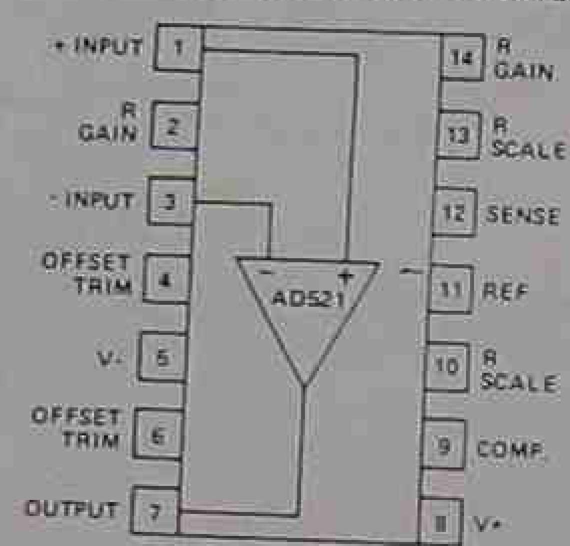
AD521

5

FEATURES

- Programmable Gains from 0.1 to 1000
- Differential Inputs
- High CMRR: 110dB min
- Low Drift: 2µV/°C max (L)
- Complete Input Protection, Power ON and Power OFF
- Functionally Complete with the Addition of Two Resistors
- Internally Compensated
- Gain Bandwidth Product: 40MHz
- Output Current Limited: 25mA
- Very Low Noise: 0.5µV p-p, 0.1Hz to 10Hz, RTI @ G = 1000

AD521 FUNCTIONAL BLOCK DIAGRAM



TO-116

PRODUCT DESCRIPTION

The AD521 is a second generation, low cost, monolithic IC instrumentation amplifier developed by Analog Devices. As a true instrumentation amplifier, the AD521 is a gain block with differential inputs and an accurately programmable input/output gain relationship.

The AD521 IC instrumentation amplifier should not be confused with an operational amplifier, although several manufacturers (including Analog Devices) offer op amps which can be used as building blocks in variable gain instrumentation amplifier circuits. Op amps are general-purpose components which, when used with precision-matched external resistors, can perform the instrumentation amplifier function.

An instrumentation amplifier is a precision differential voltage gain device optimized for operation in a real world environment, and is intended to be used wherever acquisition of a useful signal is difficult. It is characterized by high input impedance, balanced differential inputs, low bias currents and high CMR.

As a complete instrumentation amplifier, the AD521 requires only two resistors to set its gain to any value between 0.1 and 1000. The ratio matching of these resistors does not affect the high CMRR (up to 120dB) or the high input impedance (3 × $10^9\Omega$) of the AD521. Furthermore, unlike most operational amplifier-based instrumentation amplifiers, the inputs are protected against overvoltages up to ±15 volts beyond the supplies.

The AD521 IC instrumentation amplifier is available in four different versions of accuracy and operating temperature range. The economical "J" grade, the low drift "K" grade, and the lower drift, higher linearity "L" grade are specified from 0 to +70°C. The "S" grade guarantees performance to specification over the extended temperature range: -55°C to +125°C.

PRODUCT HIGHLIGHTS

1. The AD521 is a true instrumentation amplifier in integrated circuit form, offering the user performance comparable to many modular instrumentation amplifiers at a fraction of the cost.
2. The AD521 has low guaranteed input offset voltage drift (2µV/°C for L grade) and low noise for precision, high gain applications.
3. The AD521 is functionally complete with the addition of two resistors. Gain can be preset from 0.1 to more than 1000.
4. The AD521 is fully protected for input levels up to 15V beyond the supply voltages and 30V differential at the inputs.
5. Internally compensated for all gains, the AD521 also offers the user the provision for limiting bandwidth.
6. Offset nulling can be achieved with an optional trim pot.
7. The AD521 offers superior dynamic performance with a gain-bandwidth product of 40MHz, full peak response of 100kHz (independent of gain) and a settling time of 5µs to 0.1% of a 10V step.

SPECIFICATIONS

(typical @ V_S = ±15V, R_L = 2kΩ and T_A = 25°C unless otherwise specified)

MODEL	AD521JD	AD521KD	AD521LD	AD521SD
GAIN				
Range (For Specified Operation, Note 1)	1 to 1000	*	*	*
Equation	$G = R_S/R_G$ V/V	*	*	*
Error from Equation	(±0.25–0.004G)%	*	*	*
Nonlinearity (Note 2) 1≤G≤1000	0.2% max	*	0.1% max	*
Gain Temperature Coefficient	±(3 ±0.05G)ppm/°C	*	*	±(15 ±0.4G)ppm/°C
OUTPUT CHARACTERISTICS				
Rated Output	±10V, ±10mA min	*	*	*
Output at Maximum Operating Temperature	±10V @ 5mA min	*	*	*
Impedance	0.1Ω	*	*	*
DYNAMIC RESPONSE				
Small Signal Bandwidth (±3dB)				
G = 1	>2MHz	*	*	*
G = 10	300kHz	*	*	*
G = 100	200kHz	*	*	*
G = 1000	40kHz	*	*	*
Small Signal, ±1.0% Flatness				
G = 1	75kHz	*	*	*
G = 10	26kHz	*	*	*
G = 100	24kHz	*	*	*
G = 1000	6kHz	*	*	*
Full Peak Response (Note 3)	100kHz	*	*	*
Slew Rate, 1≤G≤1000	10V/µs	*	*	*
Settling Time (any 10V step to within 10mV of Final Value)				
G = 1	7µs	*	*	*
G = 10	5µs	*	*	*
G = 100	10µs	*	*	*
G = 1000	35µs	*	*	*
Differential Overload Recovery (±30V Input to within 10mV of Final Value) (Note 4) G = 1000	50µs	*	*	*
Common Mode Step Recovery (30V Input to within 10mV of Final Value) (Note 5) G = 1000	10µs	*	*	*
VOLTAGE OFFSET (may be nulled)				
Input Offset Voltage (V_{OS_1})	3mV max (2mV typ)	1.5mV max (0.5mV typ)	1.0mV max (0.5mV typ)	**
vs. Temperature	15µV/°C max (7µV/°C typ)	5µV/°C max (1.5µV/°C typ)	2µV/°C max	**
vs. Supply	3µV/%	*	*	*
Output Offset Voltage (V_{OS_O})	400mV max (200mV typ)	200mV max (30mV typ)	100mV max	**
vs. Temperature	400µV/°C max (150µV/°C typ)	150µV/°C max (50µV/°C typ)	75µV/°C max	**
vs. Supply (Note 6)	0.005V_{OS_O}/%	*	*	*
INPUT CURRENTS				
Input Bias Current (either input)	80nA max	40nA max	**	**
vs. Temperature	1nA/°C max	500pA/°C max	**	**
vs. Supply	2nA/V	*	*	*
Input Offset Current	20nA max	10nA max	**	**
vs. Temperature	250pA/°C max	125pA/°C max	**	**
INPUT				
Differential Input Impedance (Note 7)	3 × 10^9Ω‖1.8pF	*	*	*
Common Mode Input Impedance (Note 8)	6 × 10^{10}Ω‖3.0pF	*	*	*
Input Voltage Range for Specified Performance (with respect to ground)	±10V	*	*	*
Maximum Voltage without Damage to Unit, Power ON or OFF Differential Mode (Note 9)	30V	*	*	*
Voltage at either input (Note 9)	V_S ±15V	*	*	*
Common Mode Rejection Ratio, DC to 60Hz with 1kΩ source unbalance				
G = 1	70dB min (74dB typ)	74dB min (80dB typ)	**	**
G = 10	90dB min (94dB typ)	94dB min (100dB typ)	**	**
G = 100	100dB min (104dB typ)	104dB min (114dB typ)	**	**
G = 1000	100dB min (110dB typ)	110dB min (120dB typ)	**	**
NOISE				
Voltage RTO (p-p) @ 0.1Hz to 10Hz (Note 10)	$\sqrt{(0.5G)^2 + (225)^2}$ µV	*	*	*
RMS RTO, 10Hz to 10kHz	$\sqrt{(1.2G)^2 + (50)^2}$ µV	*	*	*
Input Current, rms, 10Hz to 10kHz	15pA (rms)	*	*	*
REFERENCE TERMINAL				
Bias Current	3µA	*	*	*
Input Resistance	10MΩ	*	*	*
Voltage Range	±10V	*	*	*
Gain to Output	1	*	*	*
POWER SUPPLY				
Operating Voltage Range	±5V to ±18V	*	*	*
Quiescent Supply Current	5mA max	*	*	*
TEMPERATURE RANGE				
Specified Performance	0 to +70°C	*	*	-55°C to +125°C
Operating	-25°C to +85°C	*	*	-55°C to +125°C
Storage	-65°C to +150°C	*	*	*
PACKAGE OPTION[1]: TO-116 Style (D14A)	AD521JD	AD521KD	AD521LD	AD521SD

NOTES
[1] See Section 19 for package outline information.
* Specifications same as AD521JD.
** Specifications same as AD521KD.
Specifications subject to change without notice.

Applying the AD521

5

NOTES:

1. Gains below 1 and above 1000 are realized by simply adjusting the gain setting resistors. For best results, voltage at either input should be restricted to ±10V for gains equal to or less than 1.

2. Nonlinearity is defined as the ratio of the deviation from the "best straight line" through a full scale output range of ±9 volts. With a combination of high gain and ±10 volt output swing, distortion may increase to as much as 0.3%.

3. Full Peak Response is the frequency below which a typical amplifier will produce full output swing.

4. Differential Overload Recovery is the time it takes the amplifier to recover from a pulsed 30V differential input with 15V of common mode voltage, to within 10mV of final value. The test input is a 30V, 10μs pulse at a 1kHz rate. (When a differential signal of greater than 11V is applied between the inputs, transistor clamps are activated which drop the excess input voltage across internal input resistors. If a continuous overload is maintained, power dissipated in these resistors causes temperature gradients and a corresponding change in offset voltage, as well as added thermal time constant, but will not damage the device.)

5. Common Mode Step Recovery is the time it takes the amplifier to recover from a 30V common mode input with zero volts of differential signal to within 10mV of final value. The test input is 30V, 10μs pulse at a 1kHz rate. (When a common mode signal greater than V_S −0.5V is applied to the inputs, transistor clamps are activated which drop the excessive input voltage across internal input resistors. Power dissipated in these resistors causes temperature gradients and a corresponding change in offset voltage, as well as an added thermal time constant, but will not damage the device.)

6. Output Offset Voltage versus Power Supply Change is a constant 0.005 times the unnulled output offset per percent change in either power supply. If the output offset is nulled, the output offset change versus supply change is substantially reduced.

7. Differential Input Impedance is the impedance between the two inputs.

8. Common Mode Input Impedance is the impedance from *either* input to the power supplies.

9. Maximum Input Voltage (differential or at either input) is 30V when using ±15V supplies. A more general specification is that neither input may exceed either supply (even when V_S = 0) by more than 15V and that the difference between the two inputs must not exceed 30V. (See also Notes 4 and 5.)

10. 0.1Hz to 10Hz Peak-to-Peak Voltage Noise is defined as the maximum peak-to-peak voltage noise observed during 2 of 3 separate 10 second periods with the test circuit of Figure 8.

DESIGN PRINCIPLE

Figure 1 is a simplified schematic of the AD521. A differential input voltage, V_{IN}, appears across R_G causing an imbalance in the currents through Q_1 and Q_2, $\Delta I = V_{IN}/R_G$. That imbalance is forced to flow in R_S because the collector currents of Q_3 and Q_4 are constrained to be equal by their biasing (current mirror). These conditions can only be satisfied if the differential voltage across R_S (and hence the output voltage of the AD521) is equal to $\Delta I \times R_S$. The feedback amplifier, A_{FB} performs that function. Therefore, $V_{OUT} = \frac{V_{IN}}{R_G} \times R_S$ or $\frac{V_{OUT}}{V_{IN}} = \frac{R_S}{R_G}$.

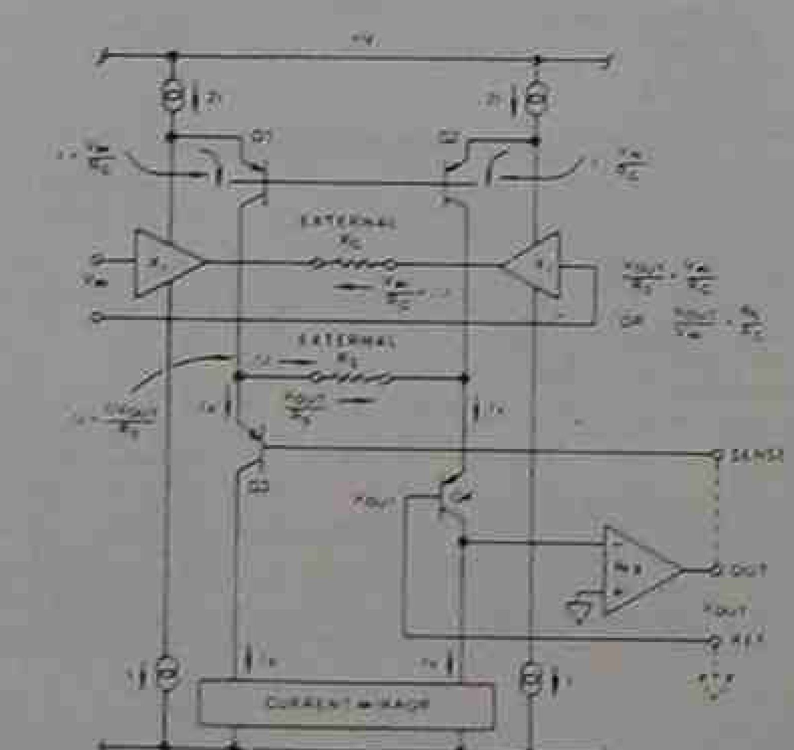


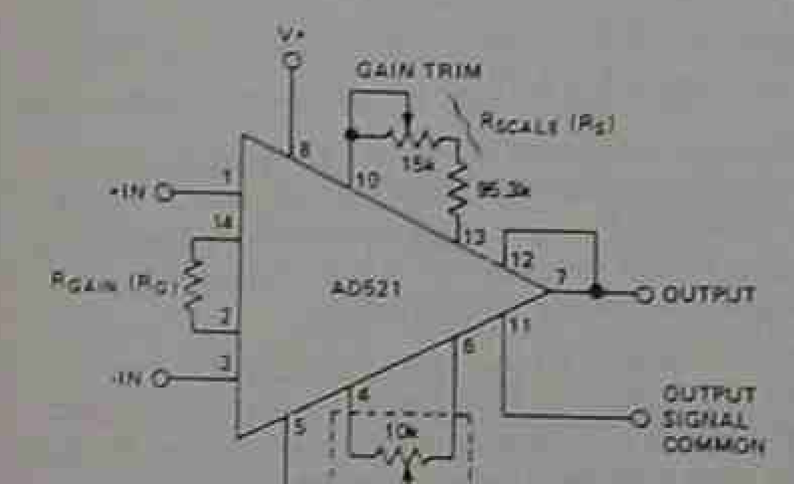
Figure 1. Simplified AD521 Schematic

APPLICATION NOTES FOR THE AD521

These notes ensure the AD521 will achieve the high level of performance necessary for many diversified IA applications.

1. Gains below 1 and above 1000 are realized by adjusting the gain setting resistors as shown in Figure 2 (the resistor, R_S between pins 10 and 13 should remain 100kΩ ±15%, see application note 3). For best results, the input voltage should be restricted to ±10V especially for gain equal to or less than 1.

2. Provide a return path to ground for input bias currents. The AD521 is an instrumentation amplifier, not an isolation amplifier. When using a thermocouple or other "floating" source, this return path may be provided directly to ground or indirectly through a resistor to ground from pins 1 and/or 3, as shown in Figure 3. If the return path is not provided, bias currents will cause the output to saturate. The value of the resistor may be determined by dividing the maximum allowable common mode voltage for the application by the bias current of the instrumentation amplifier.

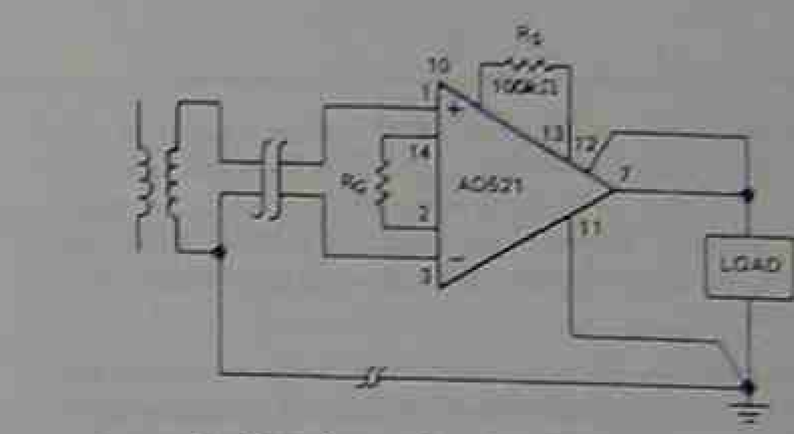


GAIN	VALUE OF R_G
0.1	1MΩ
1	100kΩ
10	10kΩ
100	1kΩ
1000	100Ω

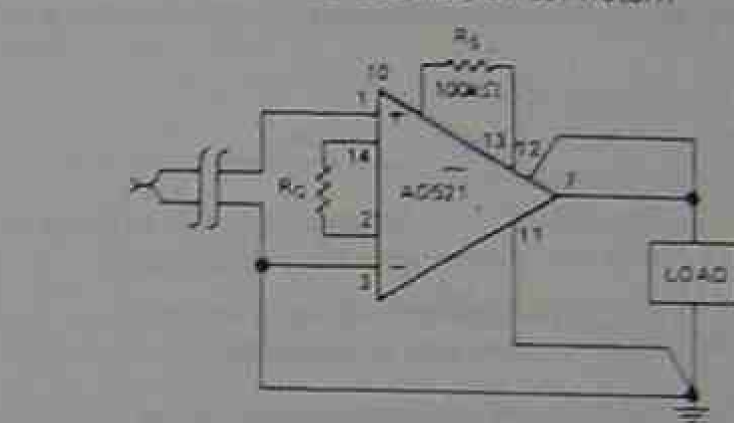
Figure 2. Operating Connections for AD521

3. The resistors between pins 10 and 13, (R_{SCALE}) must equal 100kΩ ±15% (Figure 2). If R_{SCALE} is too low (below 85kΩ) the output swing of the AD521 is reduced. At values below 80kΩ and above 120kΩ the stability of the AD521 may be impaired.

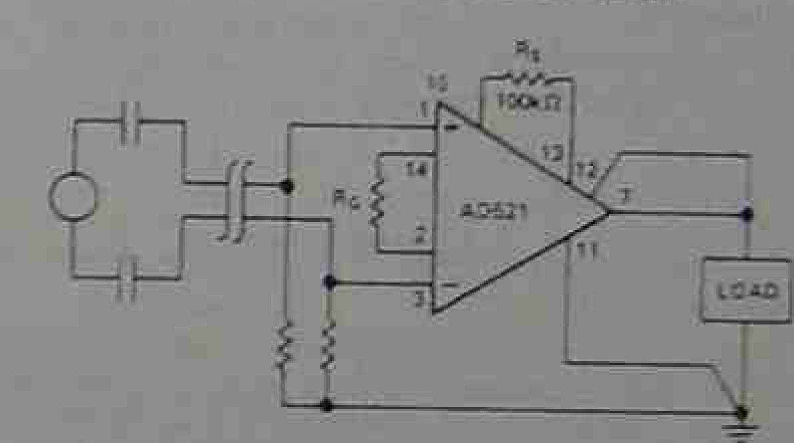
4. Do not exceed the allowable input signal range. The linearity of the AD521 decreases if the inputs are driven within 5 volts of the supply rails, particularly when the device is used at a gain less than 1. To avoid this possibility, attenuate the input signal through a resistive divider network and use the AD521 as a buffer, as shown in Figure 4. The resistor R/2 matches the impedance seen by both AD521 inputs so that the voltage offset caused by bias currents will be minimized.



a). Transformer Coupled, Direct Return

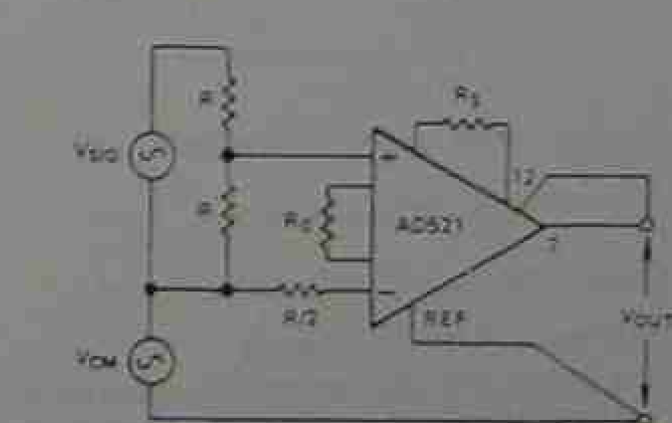


b). Thermocouple, Direct Return



c). AC Coupled, Indirect Return

Figure 3. Ground Returns for "Floating" Transducers



1. INCREASE R_G TO PICK UP GAIN LOST BY R DIVIDER NETWORK
2. INPUT SIGNAL MUST BE REDUCED IN PROPORTION TO POWER SUPPLY VOLTAGE LEVEL

Figure 4. Operating Conditions for $V_{IN} \approx V_S$ = 10V

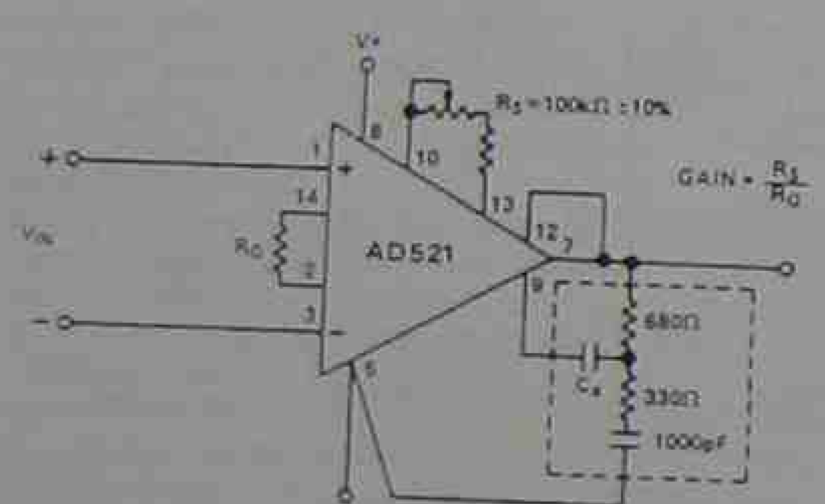
5. Use the compensation pin (pin 9) and the applicable compensation circuit when the amplifier is required to drive a capacitive load. It is worth mentioning that coaxial cables can "invisibly" provide such capacitance since many popular coaxial cables display capacitance in the vicinity of 30pF per foot.

This compensation (bandwidth control) feature permits the user to fit the response of the AD521 to the particular application as illustrated by Figure 5. In cases of extremely high load capacitance the compensation circuit may be changed as follows:

1. Reduce 680Ω to 24Ω
2. Reduce 330Ω to 7.5Ω
3. Increase 1000pF to 0.1µF
4. Set C_X to 1000pF if no compensation was originally used. Otherwise, do not alter the original value.

This allows stable operation for load capacitances up to 3000pF, but limits the slew rate to approximately 0.16V/µs.

6. Signals having frequency components above the Instrumentation Amplifier's output amplifier closed-loop bandwidth will be transmitted from V− to the output with little or no attenuation. Therefore, it is advisable to decouple the V− supply line to the output common or to pin 11.[1]



$$C_x = \frac{1}{100\pi f_t} \text{ when } f_t \text{ is the desired bandwidth.}$$

(f_t in kHz, C_X in µF)

Figure 5. Optional Compensation Circuit

INPUT OFFSET AND OUTPUT OFFSET

When specifying offsets and other errors in an operational amplifier, it is often convenient to refer these errors to the inputs. This enables the user to calculate the maximum error he would see at the output with any gain or circuit configuration. An op amp with 1mV of input offset voltage, for example, would produce 1V of offset at the output in a gain of 1000 configuration.

In the case of an instrumentation amplifier, where the gain is controlled in the amplifier, it is more convenient to separate errors into two categories. Those errors which simply add to the output signal and are unaffected by the gain can be classified as output errors. Those which act as if they are associated with the input signal, such that their effect at the output is proportional to the gain, can be classified as input errors.

As an illustration, a typical AD521 might have a +30mV output offset and a −0.7mV input offset. In a unity gain configuration, the *total* output offset would be +29.3mV or the sum of the two. At a gain of 100, the output offset would be −40mV or: 30mV + 100(−0.7mV) = −40mV.

By separating these errors, one can evaluate the total error independent of the gain settings used, similar to the situation with the input offset specifications on an op amp. In a given gain configuration, both errors can be combined to give a total error *referred to the input* (R.T.I.) or *output* (R.T.O.) by the following formula:

Total Error R.T.I. = input error + (output error/gain)

Total Error R.T.O. = (Gain x input error) + output error

The offset trim adjustment (pins 4 and 6, Figure 2) is associated primarily with the output offset. At any gain it can be used to introduce an output offset equal and opposite to the input offset voltage multiplied by the gain. As a result, the total output offset can be reduced to zero.

As shown in Figure 6, the gain range on the AD521 can be extended considerably by adding an attenuator in the sense terminal feedback path (as well as adjusting the ratio, R_5/R_G). Since the sense terminal is the inverting input to the output amplifier, the additional gain to the output is controlled by R_1 and R_2. This gain factor is $1 + R_2/R_1$.

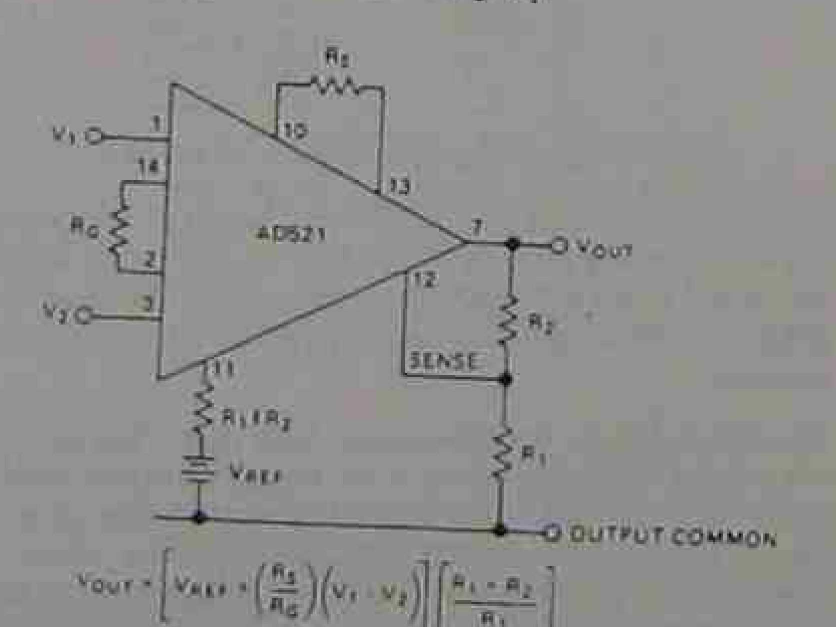


Figure 6. Circuit for utilizing some of the unique features of the AD521. Note that gain changes introduced by changing R1 and R2 will have a minimum effect on output offset if the offset is carefully nulled at the highest gain setting.

[1] For further details, refer to "An I.C. User's Guide to Decoupling, Grounding, and Making Things Go Right for a Change," by A. Paul Brokaw. This application note is available from Analog Devices without charge upon request.

Where offset errors are critical, a resistor equal to the parallel combination of R_1 and R_2 should be placed between pin 11 and V_{REF}. This minimizes the offset errors resulting from the input current flowing in R_1 and R_2 at the sense terminal. Note that gain changes introduced by changing the R_1/R_2 attenuator will have a minimum effect on output offset if the offset is carefully nulled at the highest gain setting.

When a predetermined output offset is desired, V_{REF} can be placed in series with pin 11. This offset is then multiplied by the gain factor $1 + R_2/R_1$ as shown in the equation of Figure 6.

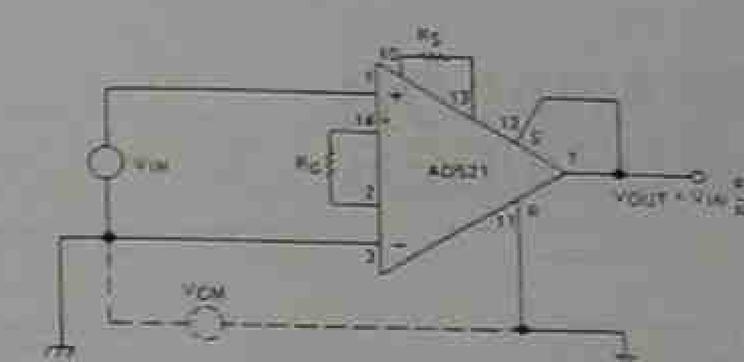


Figure 7. Ground loop elimination. The reference input, Pin 11, allows remote referencing of ground potential. Differences in ground potentials are attenuated by the high CMRR of the AD521.

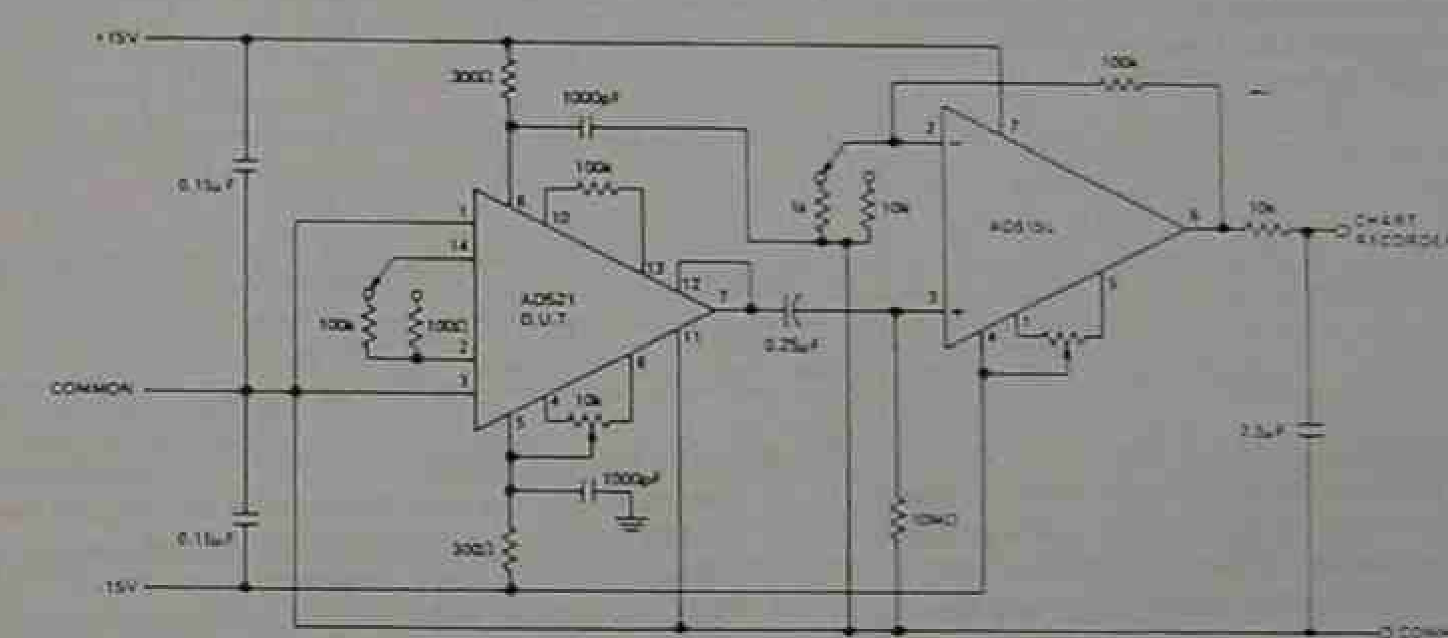


Figure 8. Test circuit for measuring peak to peak noise in the bandwidth 0.1Hz to 10Hz. Typical measurements are found by reading the maximum peak to peak voltage noise of the device under test (D.U.T.) for 3 observation periods of 10 seconds each.

ANALOG DEVICES

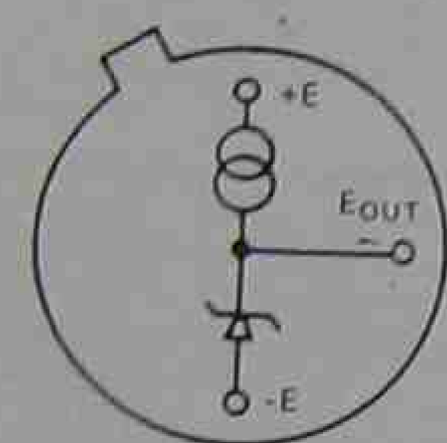
High Precision 2.5 Volt IC Reference

AD580*

FEATURES

- Laser Trimmed to Higher Accuracy: 2.500V ±0.4%, Improved from ±1.0% (AD580M)
- 3-Terminal Device: Voltage In/Voltage Out
- Excellent Temperature Stability: 10ppm/°C (AD580M, U)
- Excellent Long Term Stability: 250μV (25μV/Month)
- Low Quiescent Current: 1.5mA max
- Small, Hermetic IC Package: TO-52 Can

AD580 FUNCTIONAL BLOCK DIAGRAM



TO-52
BOTTOM VIEW

7

PRODUCT DESCRIPTION

The AD580 is an improved three-terminal, low cost, tempera-
ture compensated, bandgap voltage reference which provides
a fixed 2.5V output for inputs between 4.5V and 30V. A
nique combination of advanced circuit design and laser-
rafer-trimmed thin-film resistors provide the AD580 with an
nproved initial tolerance of ±0.4%, a temperature stability of
etter than 10ppm/°C and long term stability of better than
50μV. In addition, the low quiescent current drain of 1.5mA
nax offers a clear advantage over classical zener techniques.

he AD580 is recommended as a stable reference for all 8-,
0- and 12-bit D-to-A converters that require an external refer-
nce. In addition, the wide input range of the AD580 allows
peration with 5 volt logic supplies making the AD580 ideal
r digital panel meter applications or whenever only a single
gic power supply is available.

e AD580J, K, L and M are specified for operation over the
to +70°C temperature range; the AD580S, T and U are speci-
d for operation over the extended temperature range of
5°C to +125°C.

overed by Patent Nos. 3,887,863; RE30,586.

PRODUCT HIGHLIGHTS

1. Laser-trimming the thin-film resistors has reduced the AD580 output error. For example, AD580L output tolerance is now ±10mV, improved from ±50mV.
2. The three-terminal voltage in/voltage out operation of the AD580 provides regulated output voltage without any external components.
3. The AD580 provides a stable 2.5V output voltage for input voltages between 4.5V and 30V. The capability to provide a stable output voltage using a 5-volt input makes the AD580 an ideal choice for systems that contain a single logic power supply.
4. Thin film resistor technology and tightly controlled bipolar processing provide the AD580 with temperature stabilities to 10ppm/°C and long term stability better than 250μV.
5. The low quiescent current drain of the AD580 makes it ideal for CMOS and other low power applications.

SPECIFICATIONS (@ E_{IN} and 25°C)

Model	AD580J Min	AD580J Typ	AD580J Max	AD580K Min	AD580K Typ	AD580K Max	AD580L Min	AD580L Typ	AD580L Max	AD580M Min	AD580M Typ	AD580M Max	Units
OUTPUT VOLTAGE TOLERANCE (Error from Nominal 2.500 Volt Output)			±75			±25			±10			±10	mV
OUTPUT VOLTAGE CHANGE T_{min} to T_{max}			15			7			4.3			1.75	mV
			85			60			25			10	ppm/°C
LINE REGULATION $7V \le V_{IN} \le 30V$		1.5	4		1.5	4			2			2	mV
$4.5V \le V_{IN} \le 7V$		0.3	3		0.3	2			1			1	mV
LOAD REGULATION $\Delta I = 10mA$			10			10			10			10	mV
QUIESCENT CURRENT		1.0	1.5		1.0	1.5		1.0	1.5		1.0	1.5	mA
NOISE (0.1Hz to 10Hz)		60			60			60			60		μV (p-p)
STABILITY Long Term		250			250			250			250		μV
Per Month		25			25			25			25		μV
TEMPERATURE PERFORMANCE Specified	0		+70	0		+70	0		+70	0		+70	°C
Operating	−55		+125	−55		+125	−55		+125	−55		+125	°C
Storage	−65		+175	−65		+175	−65		+175	−65		+175	°C
PACKAGE OPTION[1] - TO-52	AD580JH			AD580KH			AD580LH			AD580MH			

Model	AD580S Min	AD580S Typ	AD580S Max	AD580T Min	AD580T Typ	AD580T Max	AD580U Min	AD580U Typ	AD580U Max	Units
OUTPUT VOLTAGE TOLERANCE (Error from Nominal 2.500 Volt Output)			±25			±10			±10	mV
OUTPUT VOLTAGE CHANGE T_{min} to T_{max}			25			11			4.5	mV
			55			25			10	ppm/°C
LINE REGULATION $7V \le V_{IN} \le 30V$		1.5	6			2			2	mV
$4.5V \le V_{IN} \le 7V$		0.3	3			1			1	mV
LOAD REGULATION $\Delta I = 10mA$			10			10			10	mV
QUIESCENT CURRENT		1.0	1.5		1.0	1.5		1.0	1.5	mA
NOISE (0.1Hz to 10Hz)		60			60			60		μV (p-p)
STABILITY Long Term		250			250			250		μV
Per Month		25			25			25		μV
TEMPERATURE PERFORMANCE Specified	−55		+125	−55		+125	−55		+125	°C
Operating	−55		+150	−55		+150	−55		+150	°C
Storage	−65		+175	−65		+175	−65		+175	°C
ABSOLUTE MAXIMUM RATINGS Input Voltage	40V									
Power Dissipation @ +25°C Ambient Temperature	350mW									
Derate above +25°C	2.8mW/°C									
Lead Temperature (Soldering, 10 sec)	300°C									
Thermal Resistance Junction-to-Case	100°C/W									
Junction-to-Ambient	360°C/W									
PACKAGE OPTION[1] - TO-52	AD580SH			AD580TH			AD580UH			

NOTES
[1]See Section 19 for package outline information.
Specifications subject to change without notice.
Specifications shown in boldface are tested on all production units at final electrical test. Results from those tests are used to calculate outgoing quality levels. All min and max specifications are guaranteed, although only those shown in boldface are tested on all production units.

Applying the AD580

THEORY OF OPERATION

Most precision IC references use complex multichip hybrid designs based on expensive temperature-compensated zener diodes. Others are monolithic with on-chip zener diodes; these often require more than one power supply and, with the zener breakdown occuring near 6.3 volts, will not operate from a low voltage logic supply.

The AD580 family (AD580, AD581, AD584, AD589) uses the "bandgap" concept to produce a stable, low-temperature-coefficient voltage reference suitable for high accuracy data-acquisition components and systems. The device makes use of the underlying physical nature of a silicon transistor base-emitter voltage in the forward-biased operating region. All such transistors have approximately a $-2mV/^{\circ}C$ temperature coefficient, unsuitable for use directly as a low TC reference; however, extrapolation of the temperature characteristic of any one of these devices to absolute zero (with emitter current proportional to absolute temperature) reveals that it will go to a V_{BE} of 1.205 volts 0K, as shown in Figure 1. Thus, if a voltage could be developed with an opposing temperature coefficient to sum with V_{BE} to total 1.205 volts, a zero-TC reference would result and operation from a single, low-voltage supply would be possible. The AD580 circuit provides such a compensating voltage, V_1 in Figure 2, by driving two transistors at different current densities and amplifying the resulting V_{BE} difference (ΔV_{BE} – which now has a positive TC); the sum (V_Z) is then buffered and amplified up to 2.5 volts to provide a usable reference-voltage output. Figure 3 is the schematic diagram of the AD580.

The AD580 operates as a three-terminal reference, that means that no additional components are required for biasing or current setting. The connection diagram, Figure 4 is quite simple.

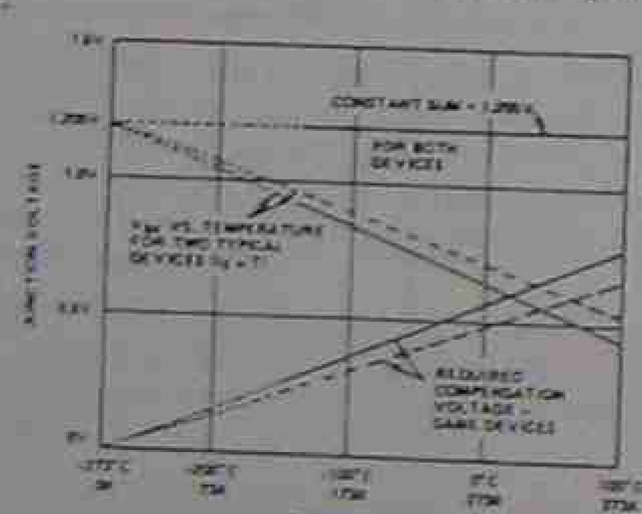


Figure 1. Extrapolated Variation of Base-Emitter Voltage with Temperature ($I_E\alpha T$), and Required Compensation, Shown for Two Different Devices

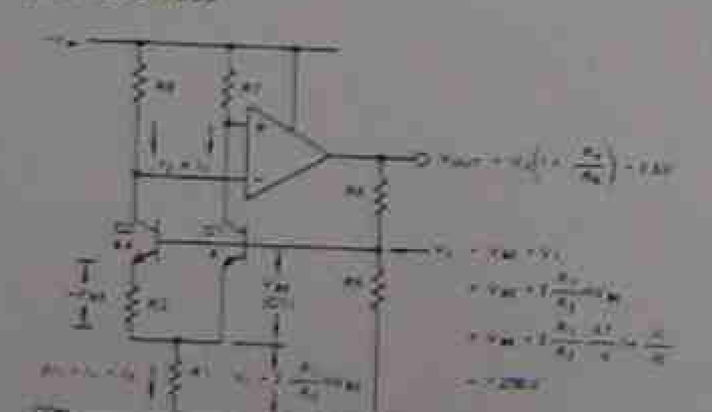


Figure 2. Basic Bandgap-Reference Regulator Circuit

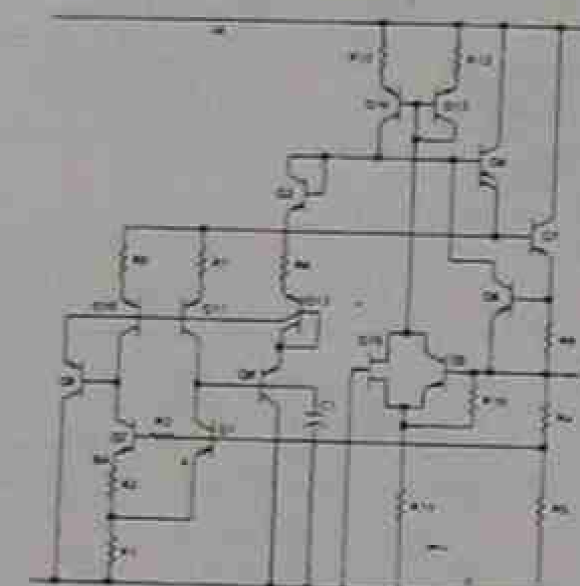


Figure 3. AD580 Schematic Diagram

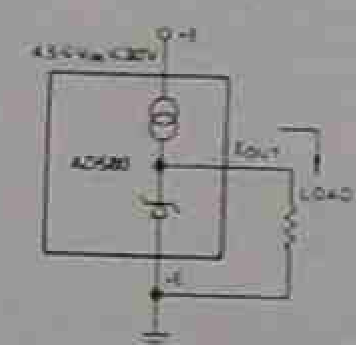


Figure 4. AD580 Connection Diagram

VOLTAGE VARIATION VS. TEMPERATURE

Some confusion exists in the area of defining and specifying reference voltage error over temperature. Historically, references are characterized using a maximum deviation per degree Centigrade; i.e., $10ppm/^{\circ}C$. However, because of the inconsistent nonlinearities in zener references (butterfly or "S" type characteristics), most manufacturers use a maximum limit error band approach to characterize their references. This technique measures the output voltage at 3 to 5 different temperatures and guarantees that the output voltage deviation will fall within the guaranteed error band at these discrete temperatures. This approach, of course, makes no mention or guarantee of performance at any other temperature within the operating temperature range of the device.

The consistent Voltage vs. Temperature performance of a typical AD580 is shown in Figure 5. Note that the characteristic is quasi-parabolic, not the possible "S" type characteristics of classical zener references. This parabolic characteristic permits a maximum output deviation specification over the device's full operating temperature range, rather than just at 3 to 5 discrete temperatures.

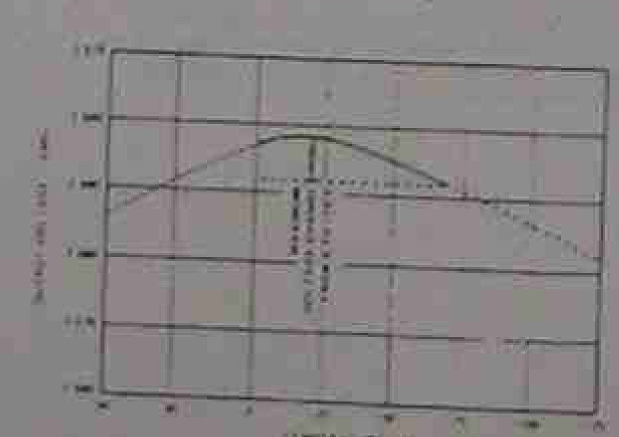


Figure 5. Typical AD580K Output Voltage vs. Temperature

7

VOLTAGE REFERENCES VOL. I, 7-7

The AD580M guarantees a maximum deviation of 1.75mV over the 0 to $+70^{\circ}C$ temperature range. This can be shown to be equivalent to $10ppm/^{\circ}C$ average maximum; i.e. . .

$$\frac{1.75mV\ max}{70^{\circ}C} \times \frac{1}{2.5V} = 10ppm/^{\circ}C\ max\ average$$

The AD580 typically exhibits a variation of 1.5mV over the power supply range of 7 to 30 volts. Figure 6 is a plot of AD580 line rejection versus frequency.

NOISE PERFORMANCE

Figure 7 represents the peak-to-peak noise of the AD580 from 1Hz (3dB point) to a 3dB high end shown on the horizontal axis. Peak-to-peak noise from 1Hz to 1MHz is approximately 600µV.

THE AD580 AS A CURRENT LIMITER

The AD580 represents an excellent alternative to current limiter diodes which require factory-selection to achieve a desired current. This approach often results in temperature coefficients of $1\%/^{\circ}C$. The AD580 approach is not limited to a specially selected factory set current limit; it can be programmed from 1 to 10mA with the insertion of a single external resistor. The approximate temperature coefficient of current limit for the AD580 used in this mode is $0.13\%/^{\circ}C$ for I_{LIM} = 1mA and $0.01\%/^{\circ}C$ for I_{LIM} = 13mA (see Figure 9). Figure 8 displays the high output impedance of the AD580 used as a current limiter for I_{LIM} = 1, 2, 3, 4, 5mA.

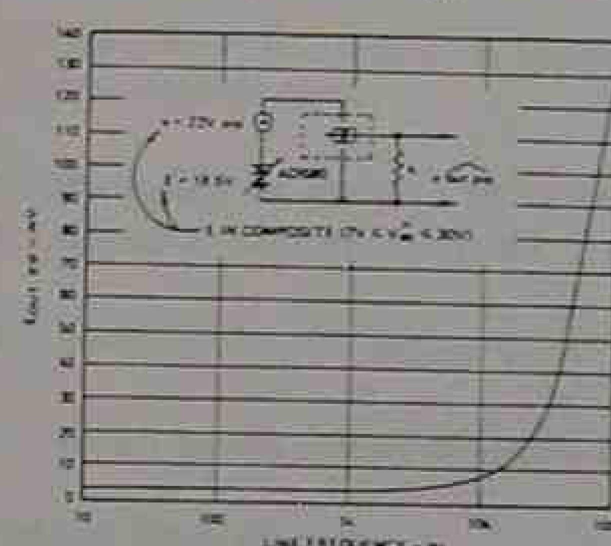


Figure 6. AD580 Line Rejection Plot

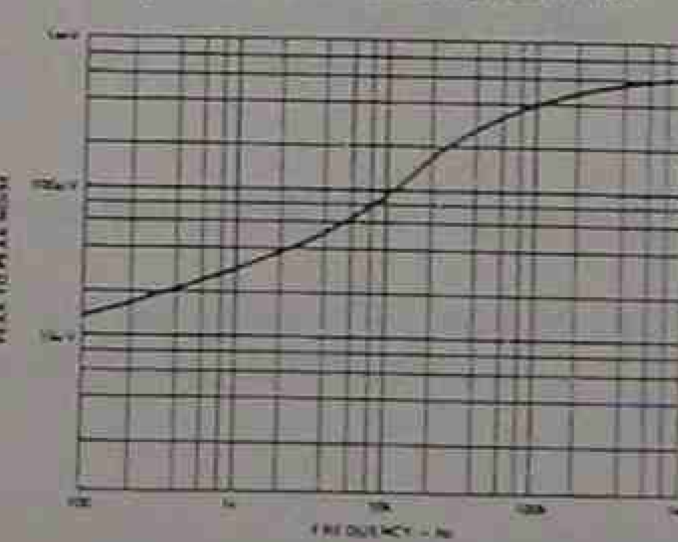


Figure 7. Peak-to-Peak Output Noise vs. Frequency

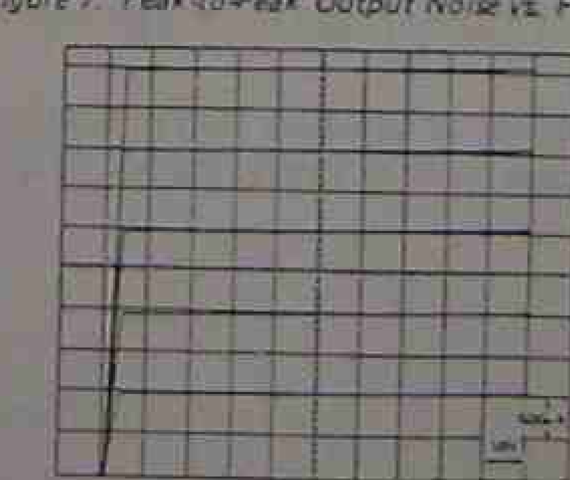


Figure 8. Input Current vs. Input Voltage (Integral Loads)

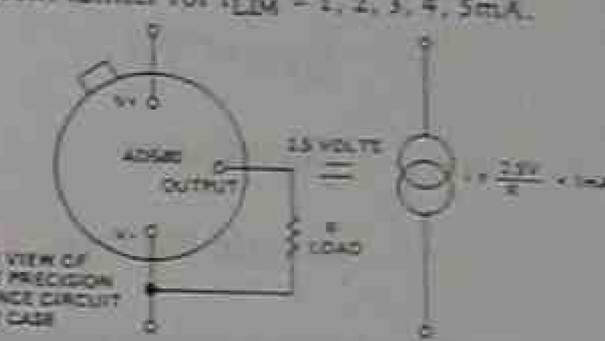


Figure 9. A Two-Component Precision Current Limiter

THE AD580 AS A LOW POWER, LOW VOLTAGE PRECISION REFERENCE FOR DATA CONVERTERS

The AD580 has a number of features that make it ideally suited for use with A/D and D/A data converters used in complex microprocessor-based systems. The calibrated 2.500 volt output minimizes user trim requirements and allows operation from a single low voltage supply. Low power consumption (1mA quiescent current) is commensurate with that of CMOS-type devices, while the low cost and small package complements the decreasing cost and size of the latest converters.

Figure 10 shows the AD580 used as a reference for the AD7542 12-bit CMOS DAC with complete microprocessor interface. The AD580 and the AD7542 are specified to operate from a single 5 volt supply; this eliminates the need to provide a +15 volt power supply for the sole purpose of operating a reference. The AD7542 includes three 4-bit data registers, a 12-bit DAC register, and address decoding logic; it may thus be interfaced directly to a 4-, 8- or 16-bit data bus. Only 8mA of quiescent current from the single +5 volt supply is required to operate the AD7542 which is packaged in a small 16-pin DIP. The AD544 output amplifier is also low power, requiring only 2.5mA quiescent current. Its laser-trimmed offset voltage preserves the ±1/2LSB linearity of the AD7542KN without user trims and it typically settles to ±1/2 LSB in less than 3µs. It will provide the 0 to −2.5 volt output swing from ±5 volt supplies.

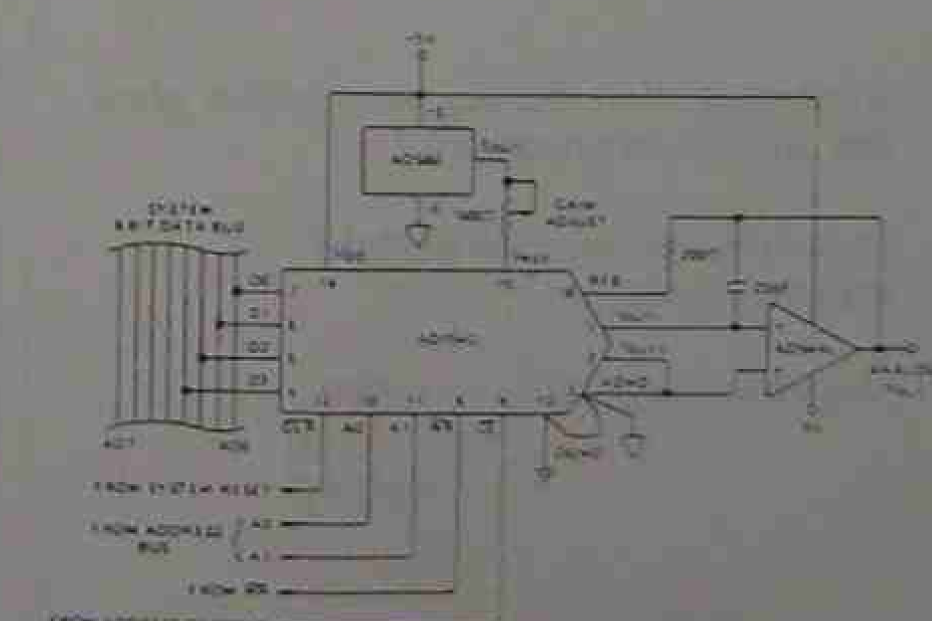


Figure 10. Low Power, Low Voltage Reference for the AD7542 Microprocessor-Compatible 12-Bit DAC

AD590

name the type of device

state the following :-

the output characteristic,

- type of V/I
- impedance
- output figures

temperature sensing range

linearity

power supply range,

- normal supply
- max forward voltage
- max reverse voltage

power consumption, (state conditions)

regulation of output for varying supply voltage, (state conditions)

how remote can the device be placed from the measuring amplifier ?

- explain why

what operation during production gives the AD590 its accuracy ?

state the devices uses in fluid measurement

state the output at; -55°C
+25°C
+150°C

AD521

name the type of device

state the following :-

supply voltages

CMRR

offset voltage drift

gain bandwidth product

output current

range of gain settings

settling time for stepped input (state conditions)

relationship of gain to the two required external resistors

differential input impedance

common mode input impedance

output impedance

what is the effect of not providing a return path for input bias currents?

state the value of R_G for a gain of 0.1

calculate the value of R_G and R_s for the gains of ; 5, 25, 250, & 500.

if the input offset of the IA was 1mV, what is the effect on the output ?

if the output offset voltage is -0.7mV at a gain of 100, what effect is it at a gain of 1000 ?

AD580

name the type of device

state the following :-

type of reference technology

output voltage

output voltage accuracy

- output error in mV

temperature stability

long term stability

- and per month

quiescent current

input voltage range

describe how the temperature stability is achieved

With regard to the complete circuit for accurate temperature measurement using the AD590, AD 521, and AD580,

1. list the areas for consideration in performance error assessment.

2. State the maximum error value for each area.

3. Calculate the total measuring accuracy of the circuit.

Sydney Institute of Technology

School of Electrotechnology

Industrial Sensors

Theory Notes

Area ______________________

Topic ______________________

Session No.

For further information on this module, or this subject
contact Jim Hafford, (02) 217 3620, Bld. K. Ultimo, S.I.T.

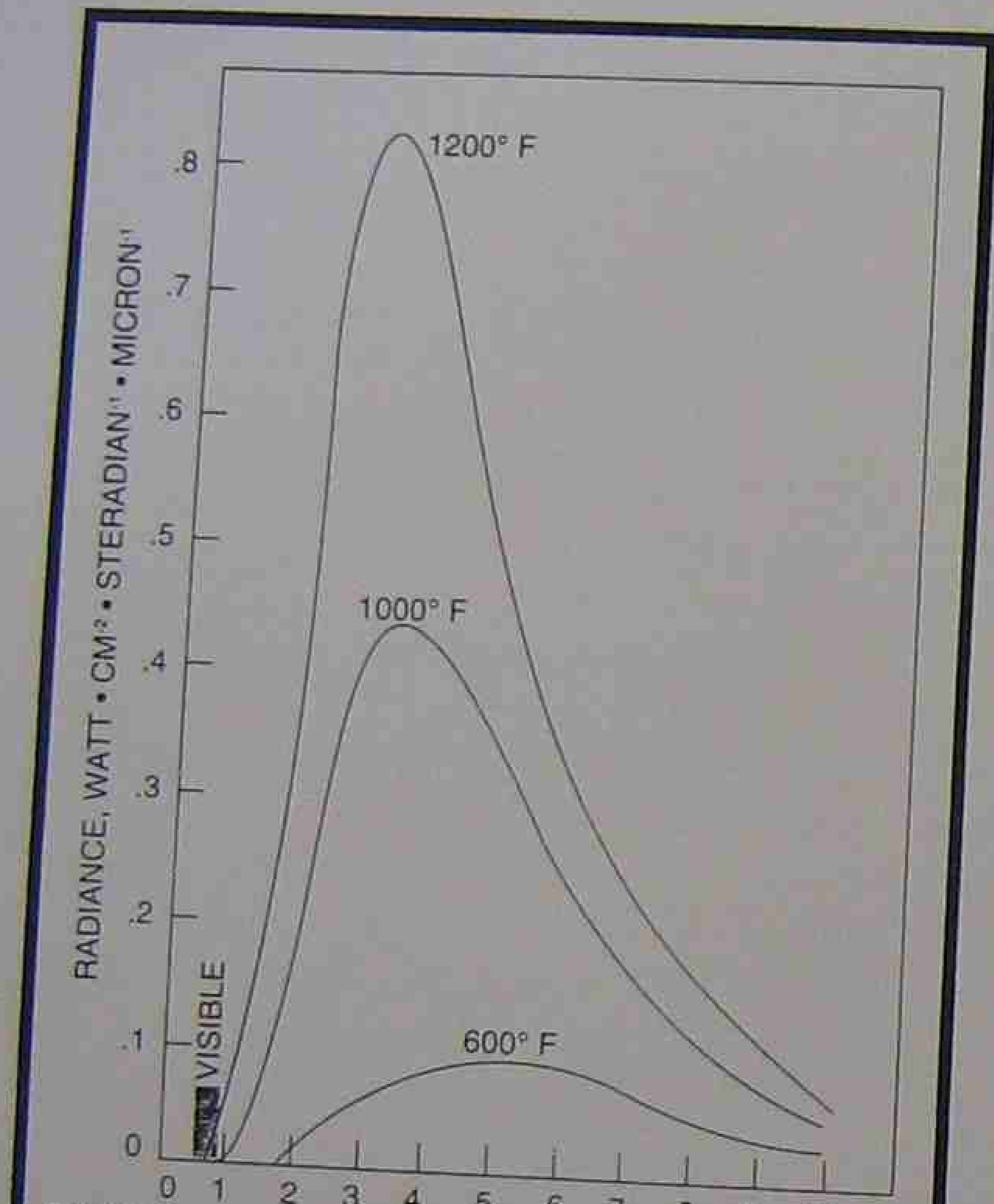


FIGURE 1
Blackbody Radiation Characteristics

GENERAL

Infrared thermometers measure the temperature of an object without requiring physical contact. The ability to accomplish this is based on the fact that every object emits radiant energy and the intensity of this radiation is a function of its temperature.

The following sections represent a qualitative presentation of the fundamentals of radiation physics upon which infrared thermometry or radiation pyrometry are based. Several of the many ways of applying these fundamentals to the practical methods of temperature measurement will be discussed.

RADIANT EMISSION WITH TEMPERATURE

Everyone observes that a sufficiently hot object will emit light or visible radiation. A light bulb filament, a smoldering ember and a billet of "red hot" steel are all obvious examples of this phenomenon. Furthermore, it is readily observed that the hotter the object the brighter and whiter its color and one can estimate the temperature of an object in this way. Experienced workers in the steel industry do this regularly.

Not as widely recognized is the fact that each of these incandescent objects is emitting a tremendous amount of "invisible" infrared radiation. For example, a steel billet at 1500°F radiates 100,000 times more energy in the infrared than it does in the visible. The intensity of this infrared radiation is a function of the billet's temperature.

The general relationship between the intensity of radiation as a function of wavelength and temperature of a perfect emitter is shown in Fig. 1. Observe that only a tiny fraction of the energy radiated is visible. Below about 1000°F the intensity of visible radiation is so small that we can't see it. However, there is still copious emission of infrared. Note that the radiant intensity at every wavelength increases with increasing temperature and the determination of the radiant intensity at any wavelength can serve to establish the emitter's temperature.

NATURE OF RADIATION

The difference between infrared radiation and visible radiation is the wavelength of the electromagnetic wave. Red light has a longer wavelength than blue light and infrared radiation has longer wavelengths than both. In all other respects these radiations behave similarly. All can be considered to be composed of elementary packets of energy called photons. All photons travel in straight lines at the "speed of light". They all can be reflected by appropriate mirrors and their paths can be bent and focused by the proper refractive elements or lenses.

All photons will dissipate their energy as heat on being absorbed by an appropriate absorber. The only fundamental difference between a blue photon, a red photon or a 2 micron infrared photon is one of its wavelength and the amount of energy it carries. The energy of a photon is inversely proportional to its wavelength.

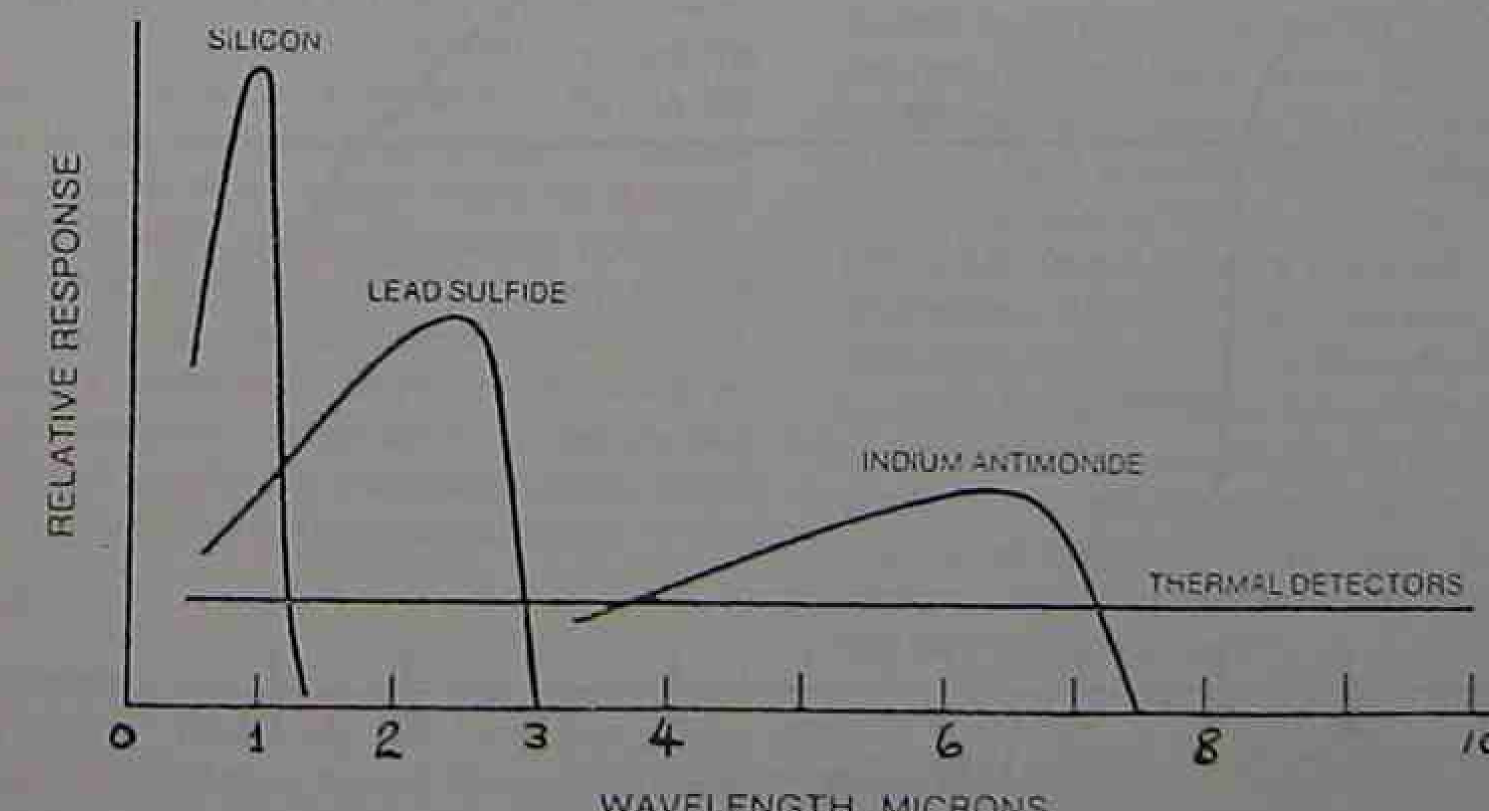


FIG. 2 — SPECTRAL RESPONSE CHARACTERISTICS OF SEVERAL INFRARED DETECTORS

ELEMENTS OF AN INFRARED THERMOMETER

A simple analysis of the eye, one form of radiation thermometer, clearly reveals the basic components used in any practical infrared thermometer. The eye contains a lens which focuses the photon flux from the emitter onto the retina or radiation detector of the human system. The retina is stimulated by the incident radiation and produces a signal that is transmitted to the brain. The brain serves as the indicator or recorder which measures the radiant intensity of the emitter and, if properly calibrated by experience, relates this radiant intensity to temperature.

The same basic elements comprise an industrial infrared thermometer. These include the collecting optics, the radiation detector and some form of indicator. It is the remarkable capabilities of available detectors that result in the apparently magical capabilities of present day infrared thermometers.

RADIATION DETECTORS

Radiation detectors take many forms but all serve the same purpose of converting an incident photon flux into an electrical signal. The two main types are the thermal detector and the quantum detector.[1] The thermal detector absorbs the incident flux, and the power dissipated increases its temperature to change some measurable physical property (for example, its resistance). This type of detector generally has a completely black receiving surface so that it is sensitive to all wavelengths. Depending as it does on its own temperature rise it has an inherently slow response.

[1] The response characteristics of several infrared detectors are shown in figure 2.

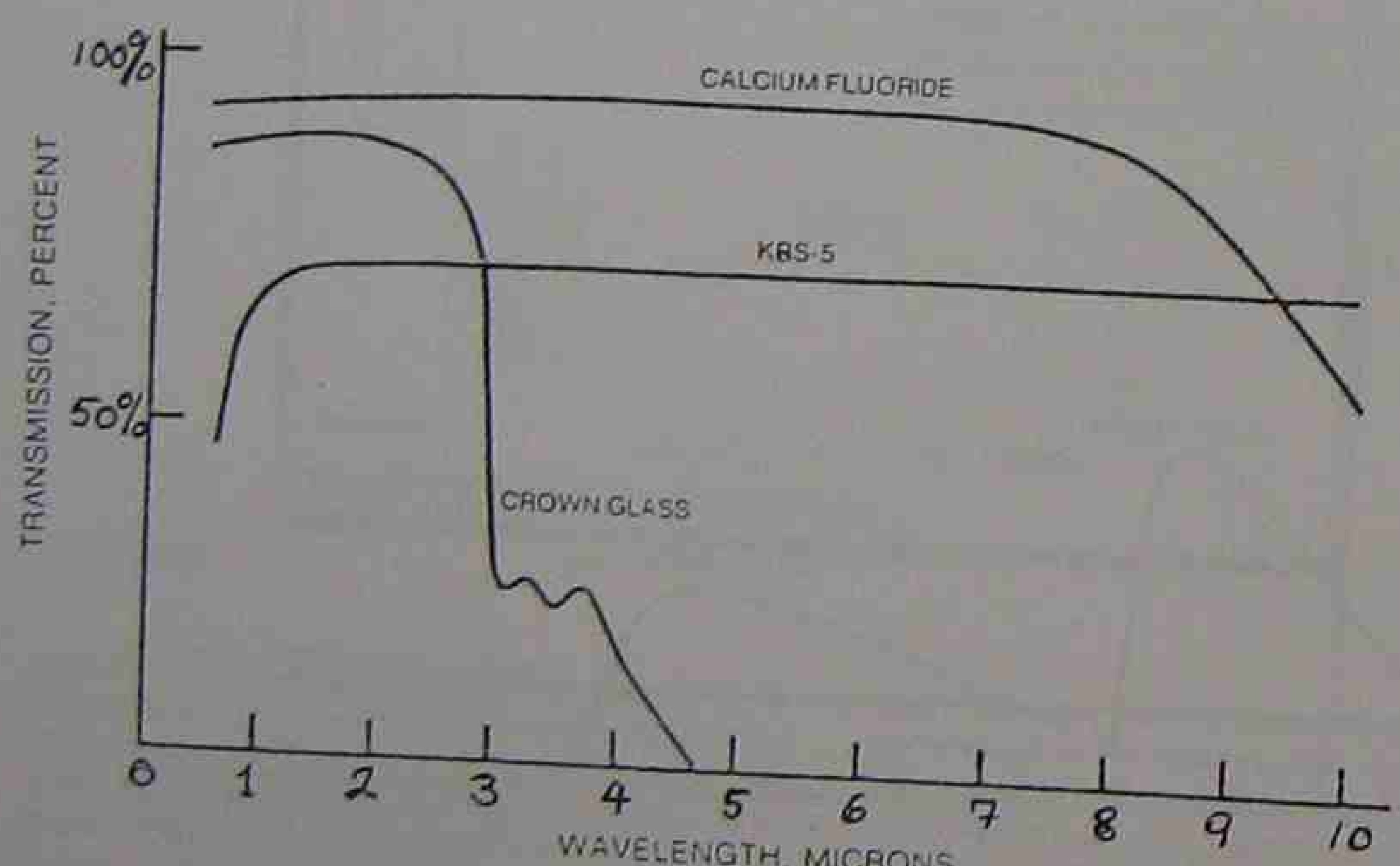


FIG. 3 — TRANSMISSION CHARACTERISTICS OF SEVERAL INFRARED OPTICAL MATERIALS

The quantum detector senses radiation in a different way. One form of quantum detector, and the type generally employed consists of a semiconductor crystal. The incident photon interacts with a bound electron within the crystal lattice. The photon's energy, if sufficiently large is transferred to the electron to free it from its immobile state permitting it to move through the crystal. During the time it is free it can produce a signal voltage in the detector. After a short interval it will return to its bound state. These intervals are generally far shorter than the thermal time constants of thermal detectors.

The quantum detector is a photon counter which is equally sensitive to all photons having the minimum energy necessary to free a bound electron. Each detector of this type will exhibit a fairly uniform response to all photons up to a particular wavelength. Photons beyond this wavelength will not have enough energy to liberate electrons to produce a signal.

The great practical advantage of these detectors lies in their ability to produce electrical signals which faithfully measure the incident photon flux completely without human attendance. This of course permits a method of continuous temperature measurement and control without contact. Where the eye is limited to temperature measurements above 1000°F, present day infrared thermometers extend the measurement range down to zero degrees.

OPTICAL ELEMENTS

The collecting optics of the radiation thermometer are chosen to be compatible with the spectral response of the detector employed. Mirrors are suitable for use over wide spectral regions. Lenses are restricted to those regions where the materials employed maintain good transmission properties. Certain design characteristics strongly favor the use of lenses for most practical systems. Fig. 3 shows the spectral transmission properties of several infrared lens materials. These same materials are also employed as windows in those applications where the target is situated in a sealed chamber.

OUTPUT

The radiation thermometer provides an electrical voltage output which can be used for simple temperature indication or any of the many forms of closed loop temperature control.[2]

CHOICE OF SPECTRAL REGION

At first glance it would appear that the radiation thermometer should utilize the entire spectrum or at least a broad enough portion of the spectrum to capture most of the radiant emission of the target in its particular temperature range. There are several reasons why this is not generally advantageous.

RADIANT EMISSION VS. WAVELENGTH

One reason relates to the rate at which the radiant emission increases with temperature. An inspection of Fig. 1 will show that the radiant emission at 2 microns increases far more rapidly with temperature than it does at (say) 6 microns. The rate of change of radiant emission with temperature is always greater at shorter wavelengths. It is clear that the greater this rate of change the more precise the temperature measurement and the tighter the temperature control. On the other hand this can't be carried to extremes because at a given short wavelength there is a lower limit to the temperature that can be measured. For example, the eye becomes useless below about 1000°F. For these reasons alone we can understand the general rule that the spectral range of the appropriate infrared thermometer shifts to longer wavelengths as the process temperature decreases.

EMITTANCE, REFLECTANCE AND TRANSMITTANCE

Another important reason for the use of different spectral regions relates to the specific emission characteristics of particular target materials. The curves of Fig. 1 show the emission characteristics of the ideal emitter or "blackbody". No material can emit more strongly than a blackbody at a given temperature. Many materials, however, can and do emit less than a blackbody at the same temperature in various portions of the spectrum. The ratio of the radiant emittance at wavelength λ of a material to that of a blackbody at the same temperature is called the spectral emittance (ε_λ). The value of ε_λ for the substance can range between 0 and 1 and this value may vary with wavelength.

[2] The detector in some infrared thermometers can provide voltages high enough to drive meters and recorders directly. Other infrared thermometers, particularly those covering the lower temperature ranges, require built-in amplifiers to provide proper output levels.

The emittance of a substance depends on its detailed interaction with radiation. A stream of radiation incident on the surface of a substance can suffer one of three fates. A portion may be reflected. Another portion may be transmitted through the substance. The remainder will be absorbed and degraded to heat. The sum of the fraction reflected (r), the fraction transmitted (t) and the fraction absorbed (a) will be equal to the total amount incident on the substance. Furthermore, the emittance (ϵ) of a substance is identical to the absorptance (a) and we can write

$$\epsilon = a = 1 - t - r$$

For the blackbody the transmittance and reflectance are zero and the emittance is unity. For any opaque substance the transmittance is zero and

$$\epsilon = 1 - r$$

An example of this case is oxidized steel in the visible and near infrared where the transmittance is 0, the reflectance is 0.20 and the emittance is 0.80. A good example of a material whose emittance characteristics change radically with wavelength is glass. Fig. 4 shows the overall transmission of several specimens of soda-lime-glass. The reflectance of the glass is about 0.03 or less through most of the spectral region shown. At wavelengths below about 2.6 microns the glass is very highly transparent and the emittance is essentially zero. Beyond 2.6 microns the glass become increasingly opaque. From this it is seen that beyond 4 microns glass is completely opaque and the emittance is above 0.98.

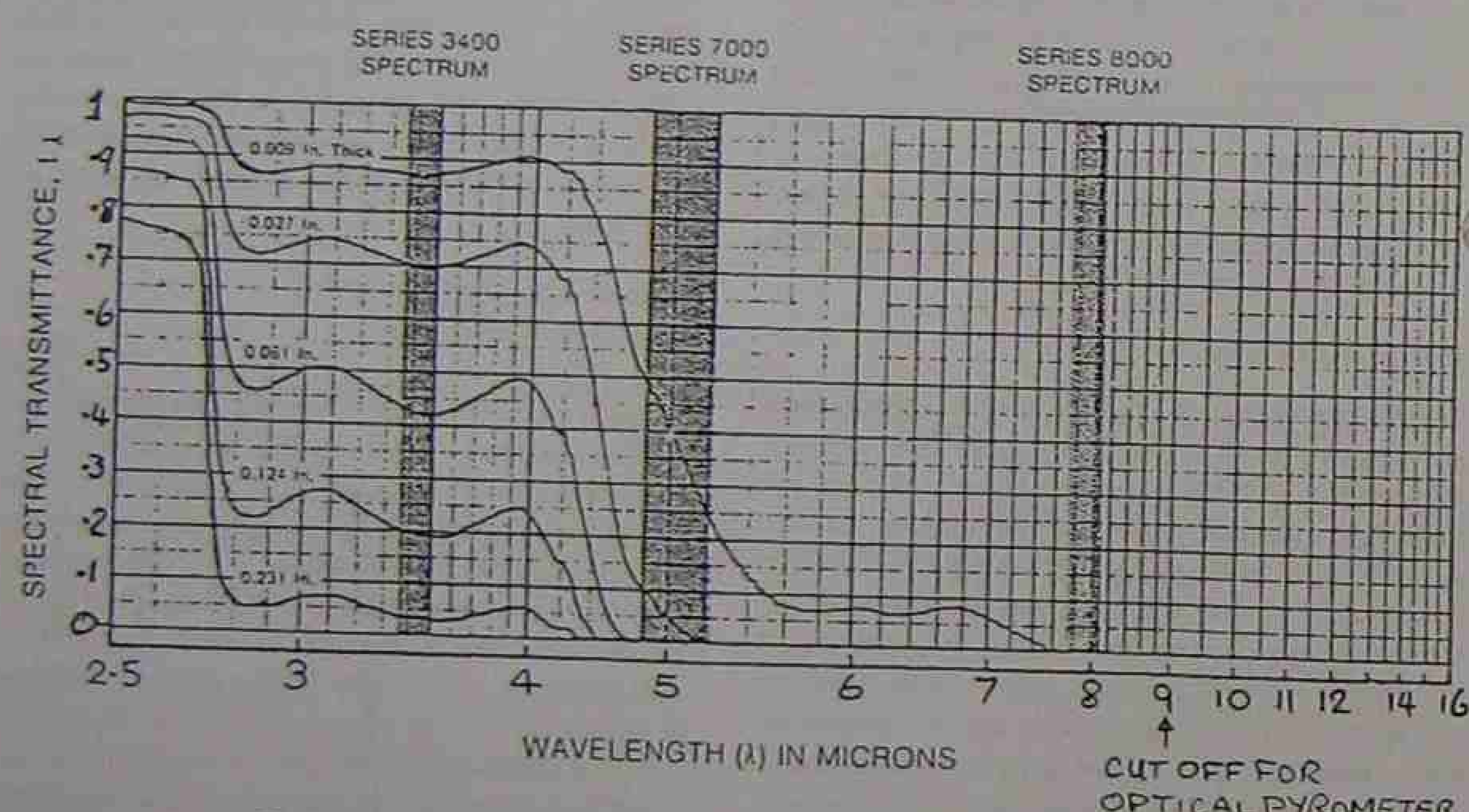


FIG. 4 — EFFECT OF THICKNESS ON SPECTRAL TRANSMITTANCE CURVES FOR SODA-LIME-SILICA GLASS

This example of glass clearly illustrates how the detailed characteristics of the material can dictate the choice of the spectral region of measurement. For example, consider the problem of measuring and controlling the temperature of this glass sheet during manufacture at a point where its temperature is 600°F. The rule that suggests a short wavelength infrared thermometer, because of the high temperature, obviously fails. To use the region around 1 micron would be useless because the emittance is close to 0. Furthermore, since the glass is highly transparent the radiation thermometer will "see through" the glass and can give false indications because of a hot wall behind the glass. One can recognize that glass can be used as an effective "window" with a short wavelength radiation thermometer. By employing the spectral region between 3 and 4 microns the internal temperature of the glass can be effectively measured and controlled. By operating out at 5 or more microns the surface temperature of the glass is measured. Each of these cases represents a practical application of infrared thermometry.

ATMOSPHERIC TRANSMISSION

A third important consideration affecting the choice of spectral region is that of the transmission of the atmosphere between the target substance and the radiation thermometer. The normal atmosphere always contains a small but definite amount of carbon dioxide and a variable amount of water vapor. Carbon dioxide strongly absorbs radiation between 4.2 and 4.4 microns and the water vapor absorbs strongly between 5.6 and 8.0 microns and also somewhat in the region 2.6 to 2.9 microns (see Fig. 5). It is obvious that these spectral regions should be avoided, particularly in the region of the water bands. If this is not done the temperature calibration will vary with path length and also with humidity. If the air temeprature is comparable to or higher than the target temperature the improperly designed infrared thermometer could provide temperature measurements strongly influenced by air temperatures.

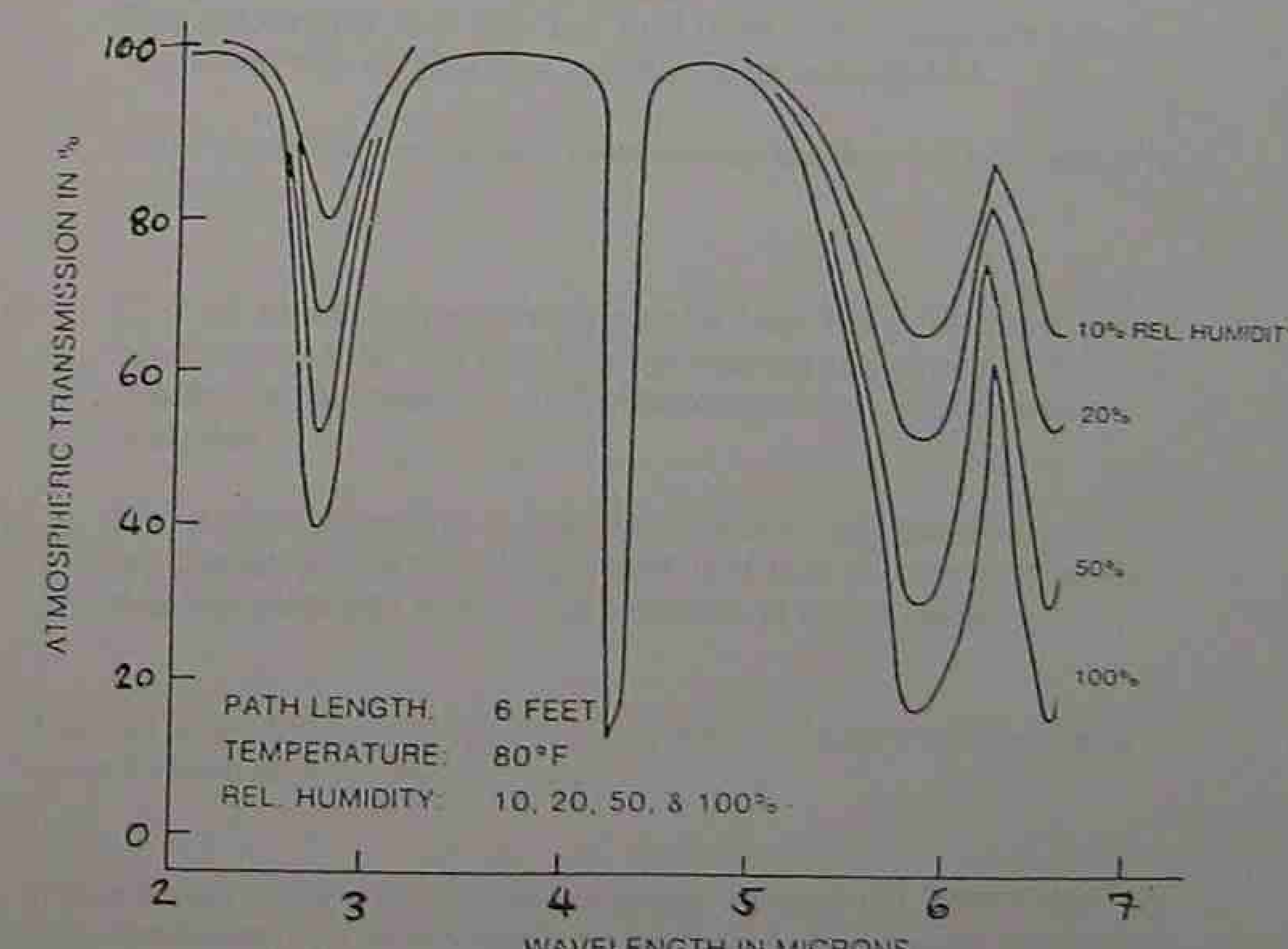


FIG. 5 — TRANSMISSION FOR ATMOSPHERE AT 80°F, AND SEVERAL RELATIVE HUMIDITIES

PRACTICAL APPLICATIONS

Infrared thermometers are currently used in a wide range of laboratory and industrial temperature control applications. A few low temperature examples include extrusion, lamination and drying of plastics, paper and rubber — curing of resins, adhesives and paints — cold rolling and forming of metals.

Some high temperature examples include forming, tempering and annealing of glass — smelting, casting, rolling, forging and heat treating of metals — calcining and firing of ceramics and cement.

In short, the infrared thermometer can be used in almost any application in the range 0 to 6500°F where its unique capabilities can turn a seemingly impossible measurement and control problem into a practical working process. Many processes now controlled manually can be converted into continuous, automated systems.

Most probably your specific application has not been covered in this discussion. The staff at IRCON, INC. is available to consider your particular problem and will promptly provide recommendations for its solution. It is asked only that your request be accompanied with as detailed information as possible including such information as target material and dimensions, surface condition, temperature range, working distance, ambient conditions and a simple sketch of the actual installation. Naturally, there is no obligation for this service.

Procedure for using an Optical Pyrometer

1. **Focus the internal reference lamp onto your eye with the eyepiece adjuster.**

2. **Focus the object whose temperature is being measured, using the 'objective' lens adjustment**

3. **Familiarise yourself with the controls. Change the reference lamp brightness above and below that of the object to accustomise your eye. (NOTE:- it is less tiring to keep both your eyes open while taking readings).**

4 a) **For each reading, match the reference lamp brightness to the object brightness first by beginning with reference lamp too cold, and then too hot.**

   b) **The mean of these two readings is taken as the actual temperature for that operation.**
   It is important that the 'up' and 'down' readings are taken consecutively so that the mean can be established.

   c) **Approach the match in a continuous motion without stops or reversals.**

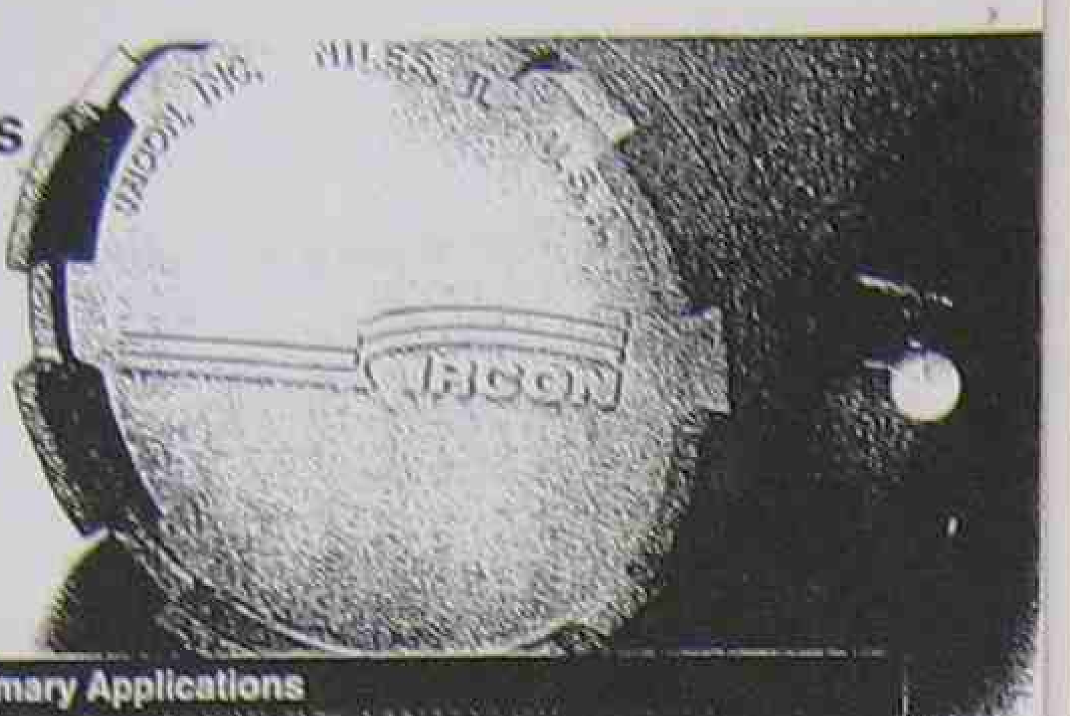
   d) **If at all possible have someone else take the pyrometer scale readings so that you can concentrate on the process of obtaining the readings and obtaining the most consistent matches.**

   e) **If the pyrometer has a 'standardisation' adjustment, (emissivity), it is very important that this be correctly set to suit the material whose temperature is being taken.**

5. **Take at least four readings (i.e. four pairs of observations), and record them in a table.**

6. **When calibrating a pyrometer, against a standard 'blackbody' or 'standard-lamp', it is advisable to include in the results the date, equipment type and equipment readings, that may be relevant when looking for 'long-term' trends of the pyrometers stability and accuracy.**

Virtually Unlimited Applications



Series	Temp. Span	Wavelength	Primary Applications
43	120 to 400°F 200 to 600°F 300 to 1000°F 50 to 200°C 100 to 400°C 150 to 500°C	3.43 microns	Ideally suited for thin polyethylene film applications such as extrusion coating, blown film cooling, blow molding, thermo-forming, film laminating and printing.
44	0 to 200°F 0 to 500°F 0 to 1000°F 0 to 100°C 0 to 250°C 0 to 600°C	8 to 14 microns	Ideally suited to all types of very low temperature applications such as print drying, food, wood, paper and textile processing, vacuum forming and infrared heating. *For low temperature food applications from –50 to 200°F (–50 to 100°C) consult factory for availability.
46	500 to 1000°F 600 to 1400°F 300 to 800°C 300 to 1300°C	2.0 to 2.6 microns	This series is best choice for thick plastics, rubber, textiles and metal applications. It is ideal for general applications involving medium temperatures and can look through glass and quartz windows down to 500°F.
47	200 to 1000°F 500 to 1500°F 500 to 2500°F 100 to 600°C 300 to 800°C 300 to 1300°C	4.8 to 5.2 microns	Measures glass surface temperature in such operations as forming, bending, tempering annealing and sealing. Also suitable for infrared heating.
48	0 to 600°F 500 to 1500°F 500 to 2500°F 0 to 300°C 300 to 800°C 300 to 1300°C	7.5 to 8.5 microns	Ideally suited for thin films of plastics such as polyesters, fluorocarbons, synthetic textiles, oils, paper, wood and other organic materials. Also for processes such as extruding, stretching, heat setting, PET bottle molding, thermoforming, glass tempering, adhesive curing, blow molding, laminating and coating.

Why Infrared Thermometers?

In many temperature measurement and control applications, infrared thermometers are the only solution because they don't require contact for the measurement. Where the product is small, fragile, moving, or in a vacuum or other controlled atmosphere, there really is no other solution.

In many other applications, infrared thermometers are the most intelligent solution. Since there is no contact, they don't add or remove heat or disturb the process in any way. They measure the product temperature and not the oven, furnace or surrounding environment. They measure continuously with fast response times. These characteristics often result in substantial energy savings, consistent higher quality finished products and improved control of temperature dependent product variables.

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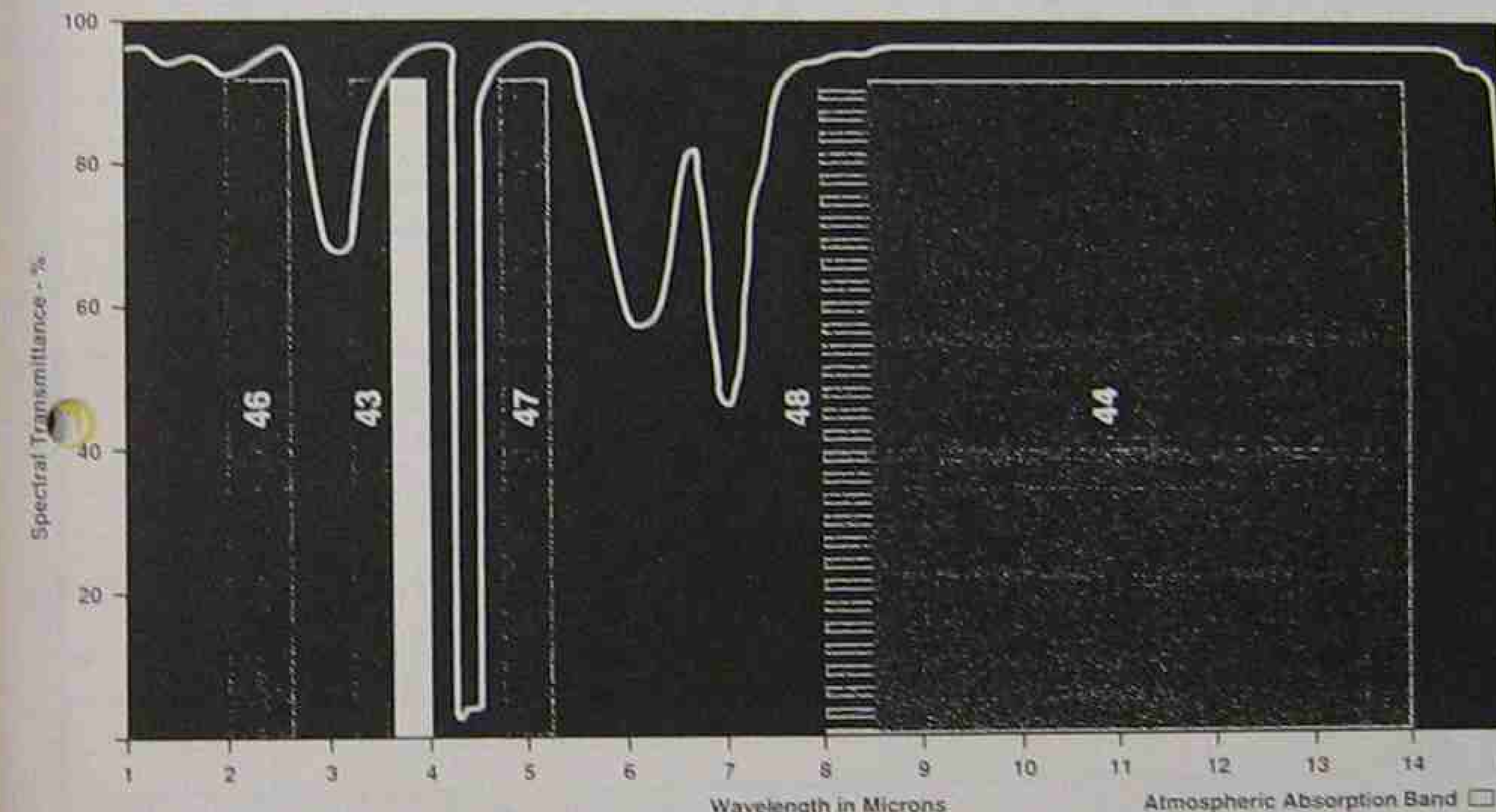
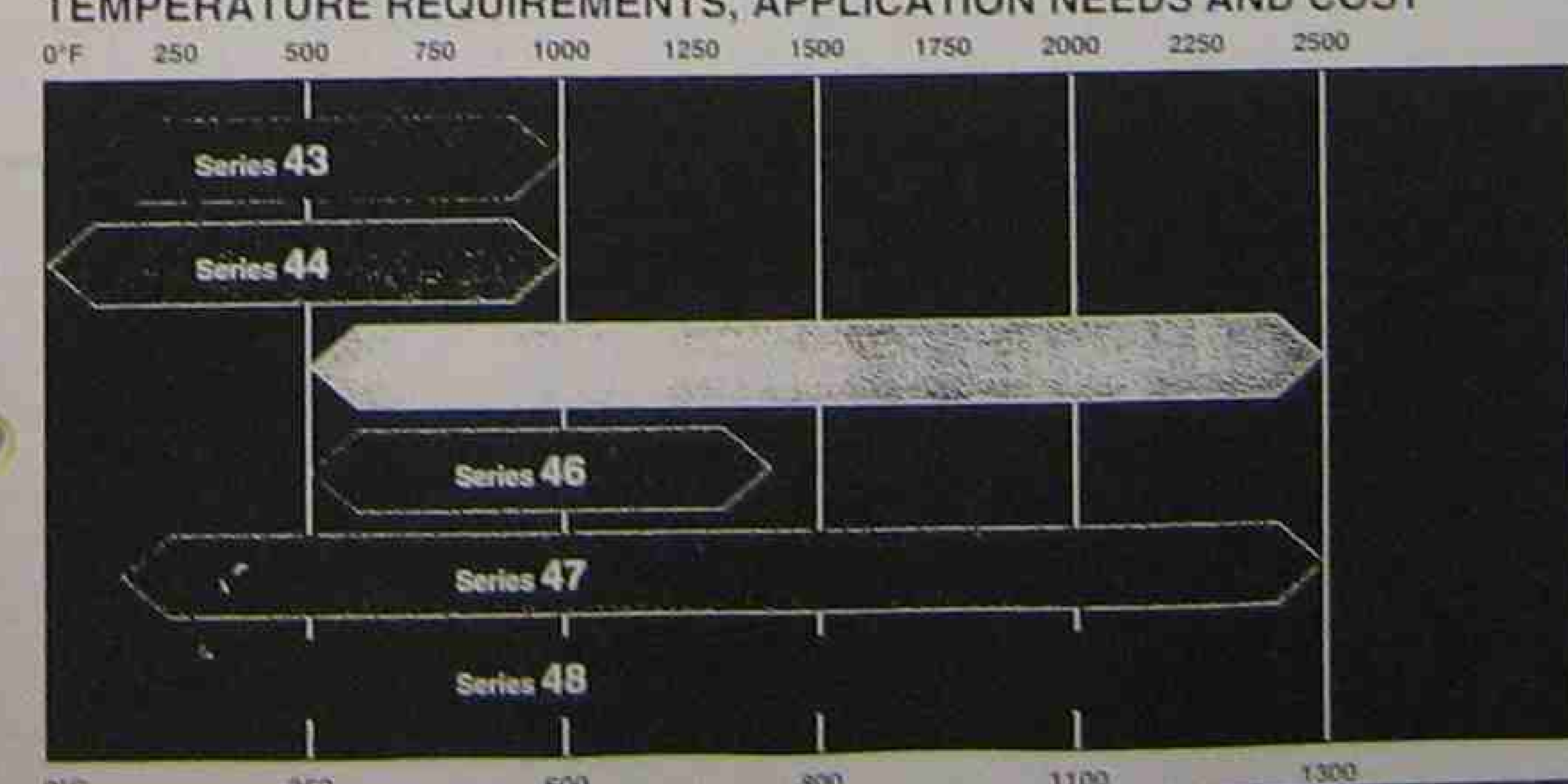


YOUR CHOICE OF SIX SENSOR SERIES TO MATCH YOUR TEMPERATURE REQUIREMENTS, APPLICATION NEEDS AND COST



APPLICATION WAVELENGTH SELECTION GUIDE

TYPICAL APPLICATIONS	WAVELENGTH (µm) 0.65	0.9	1.0	.7-1.08 RATIO	1.65	2.0	3.43	3.9	5.0	7.9	8-14
ALUMINUM			•		•	•					
ASPHALT										•	•
AUTOMOTIVE		•	•	•	•	•	•			•	•
APPLIANCES						•	•			•	•
AMMUNITION						•	•				•
BATTERIES						•					•
CEMENT	•	•	•	•		•		•		•	•
CONSTRUCTION MATERIALS							•			•	•
PHARMACEUTICAL											•
FIBERGLASS	•		•	•		•	•		•		•
FOOD PROCESSING						•	•				•
FOUNDRY	•	•	•	•		•					
GLASS - MELTING	•	•	•	•							
GLASS - FLAT									•		
GLASS BOTTLES/ CONTAINERS		•							•	•	
HEAT TREATING		•		•	•	•					
INDUCTION HEATING		•		•	•	•					
KILNS	•	•		•		•		•	•		•
METAL WORKING		•		•	•	•					
MINING				•							
NON-FERROUS METALS			•		•	•					
OVENS		•	•	•	•	•	•	•	•	•	•
PAPER							•			•	•
PLASTICS							•			•	•
RUBBER							•			•	•
SEMICONDUCTORS		•	•	•		•		•	•		
STEEL	•		•	•	•	•		•			
TEXTILES						•	•			•	•
UTILITIES				•							•

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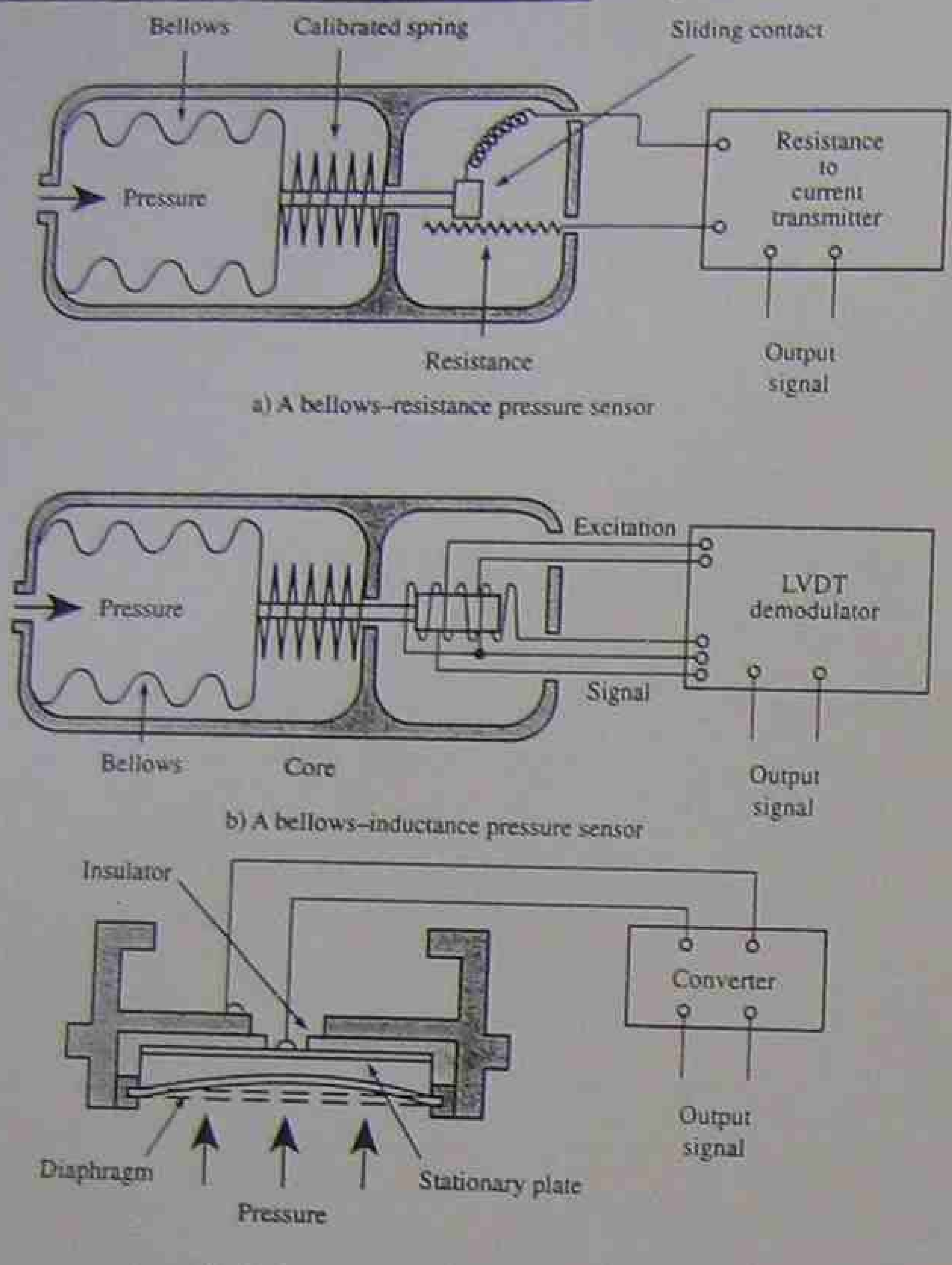
PRESSURE SENSORS.



a) A bellows–resistance pressure sensor

b) A bellows–inductance pressure sensor

c) A diaphragm–capacitance pressure sensor

Figure 10.12 Examples of deflection-type pressure sensors.

A variable resistance pressure sensor is illustrated in Figure 10.12a. The calibrated spring is displaced by an amount proportional to the pressure in the bellows. The sliding contact causes a change in the resistance between the two leads connected to the transmitter. The transmitter, in turn, produces an electrical signal based on the resistance value.

A variable inductance pressure sensor is illustrated in Figure 10.12b. The LVDT displacement transducer and demodulator produces a linear dc voltage signal proportional to the displacement of the core from a central null position. As the core moves in one direction from null, a positive voltage is produced. Movement in the other direction produces a negative voltage. A major advantage of the LVDT is the fact that it does not touch the internal bore of the transformer. This eliminates problems of mechanical wear, errors due to friction, and electrical noise due to a rubbing action.

A variable capacitance pressure transducer is illustrated in Figure 10.12c. The diaphragm and the stationary plate form the two plates of the capacitor. The displacement of the diaphragm reduces the distance between the two plates, thereby increasing the capacitance. The signal conditioner produces an electrical signal based on the capacitance value of the primary element.

Linear Variable Differential Transformers

A linear variable differential transformer (LVDT) is a transformer that consists of a primary winding and two identical secondary windings positioned symmetrically on both sides of the primary. The primary and the secondaries are wound on a hollow plastic or ceramic tube, into which a ferrous cylinder (core) can be placed. The core may be threaded to screw into the tube, or it may be smooth, to slide in freely. The position of the core with respect to the two secondary windings determines the output of the device. Therefore, LVDTs are used to measure physical displacement, both linear and rotary.

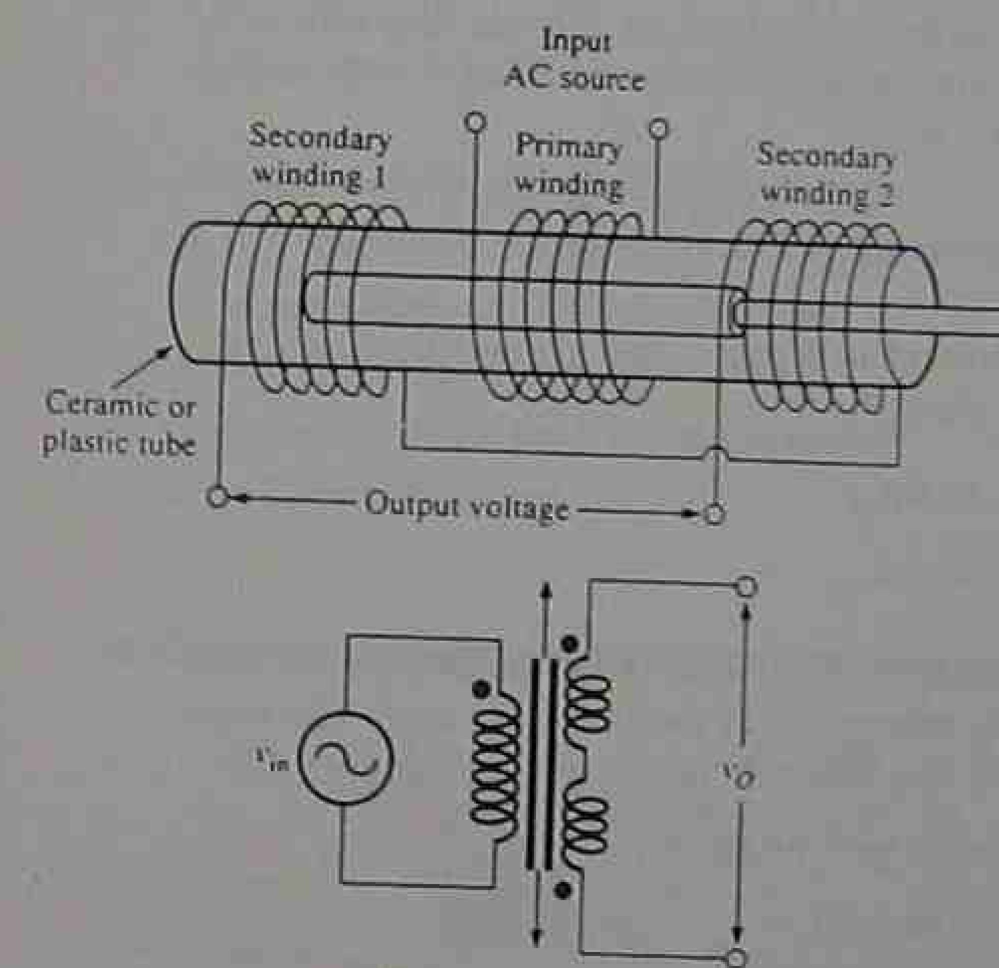


FIGURE 7–35
Structure of an LVDT (From Berlin and Getz, *Principles of Electronic Instrumentation and Measurement*. Merrill, 1988.)

The secondaries of an LVDT are usually connected in a series opposing arrangement, as shown in Figure 7–35, so if the voltage outputs from both secondaries are equal, the net output from the LVDT is zero. The factor that affects the secondary output is the amount of magnetic flux that is coupled from the primary to the secondary by the core. If the core is positioned so that both of the secondaries are equally coupled, the secondaries will produce identical outputs, 180° out of phase, resulting in an output of 0 V. If, however, the core moves to a position that results in more magnetic flux coupled through one of the secondaries, that particular secondary winding will produce a larger output than the other. This difference will result in a measurable output from the LVDT. Since the secondaries produce outputs that are 180° out of phase, the dominating secondary determines the phase of the LVDT output. The amount of displacement of the core determines the amplitude of the output. Oftentimes this output is rectified in such a way that the DC voltage indicates the amount of displacement, and the polarity of this DC voltage indicates the direction of the displacement.

10.2 FLOW RATE MEASUREMENT

Sensing Methods

The flow rate of liquids and gases is an important variable in industrial processes. The measurement of the flow rate indicates how much fluid is used or distributed in a process. Flow rate is frequently used as a controlled variable to help maintain the economy and efficiency of a given process.

The average flow rate is usually expressed in terms of the volume of liquid transferred in 1 s or 1 min.

$$\text{Average flow rate} = q_{avg} = \frac{\text{change in volume}}{\text{change in time}} = \frac{\Delta V}{\Delta t}$$

The instantaneous flow rate is determined by the limit of the average flow rate as Δt is reduced to zero. In mathematics, this limit is called the *derivative* of V with respect to t and is represented by the symbol dV/dt.

$$\text{Instantaneous flow rate} = q = \lim_{\Delta t \to 0} \frac{\Delta V}{\Delta t} = \frac{dV}{dt}$$

The flow rate in a pipe can also be expressed in terms of the average fluid velocity, v_{avg}, and the cross-sectional area of the pipe, A.

$$q_{avg} = A v_{avg}$$

The SI unit of flow rate is cubic meter/second.

Sometimes it is preferable to express the flow rate in terms of the mass of fluid transferred per unit time. This is usually referred to as the *mass flow rate*. The mass flow rate (W) is obtained by multiplying the flow rate (q) by the fluid density (ρ).

$$\text{Mass flow rate} = W = \rho q$$

The SI unit of mass flow rate is kilogram/second.

Vortex Shedding Flow Meters

A *vortex shedding flow meter* uses an unstreamlined obstruction in the flow stream to cause pulsations in the flow. The pulsations are produced when vortices (or eddies) are alternately formed and then shed on one side of the obstruction and then on the other side of the obstruction. The resulting pulsations are sensed by a piezoelectric crystal. The frequency of the pulses is directly proportional to the fluid velocity, thus forming the basis of a volumetric flow meter. Figure 10.8 illustrates a vortex shedding flow meter.

The frequency (f) of the vortex shedding is proportional to the average fluid velocity (v_{avg}) and inversely proportional to the width of the obstruction (w). The expression fw/v_{avg} is called the *Strouhal number*. The Strouhal number is constant over many ranges of Reynolds number. The relationship between the mass flow rate (W) and the vortex frequency (f) is given by the following equation:

$$W = \frac{\rho w A f}{St} \tag{10.3}$$

where A = cross-sectional area of the pipe, meter^2
f = frequency of the vortex shedding, hertz
W = mass flow rate, kilogram/second
w = width of the obstruction, meter
ρ = density of the fluid, kilogram/meter^3
St = Strouhal number

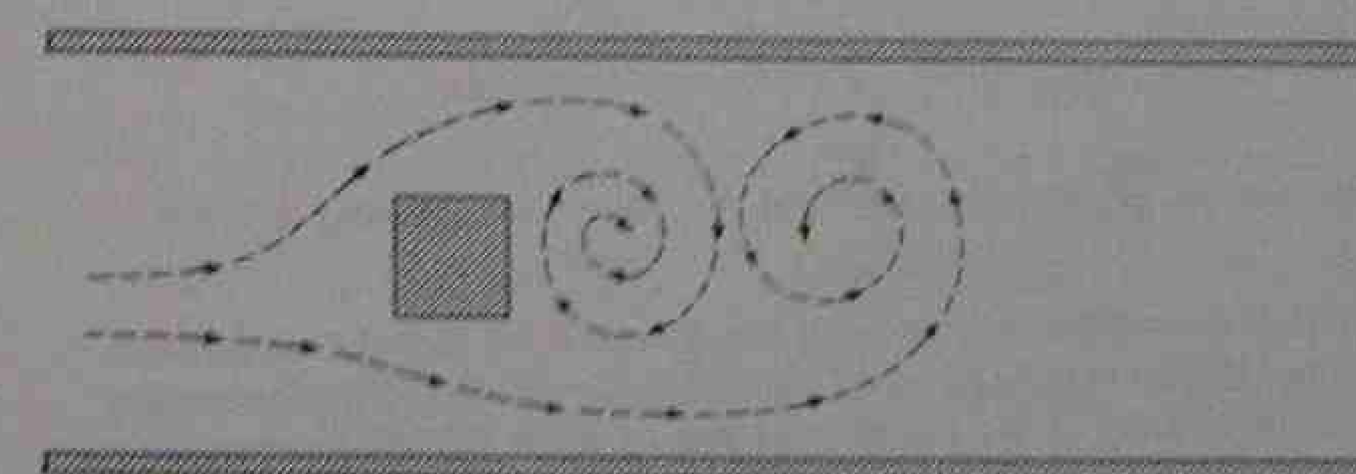


Figure 10.8 In a vortex flow meter, vortices are alternately formed and shed on one side of an obstruction and then on the other side. The resulting pulsations are sensed by a piezoelectrical crystal, and the frequency of the pulses is directly proportional to the volumetric flow rate.

Desirable features of the vortex shedding meter include a linear digital output signal, good accuracy over a wide range of flow, no moving or wearing parts, and low installed cost. Less desirable features include decreasing rangeability with increasing viscosity, and practical considerations limit the size to a diameter range of 1 to 8 in.

Turbine Flow Meters

A *turbine flow meter* is illustrated in Figure 10.7. A small permanent magnet is embedded in one of the turbine blades. The magnetic sensing coil generates a pulse each time the magnet passes by. The number of pulses is related to the volume of liquid passing through the meter by the following equation: $V = KN$, where V is the total volume of liquid, K the volume of liquid per pulse, and N the number of pulses. The average flow rate, q_{avg}, is equal to the total volume V divided by the time interval Δt.

$$q_{avg} = \frac{V}{\Delta t} = K\frac{N}{\Delta t}$$

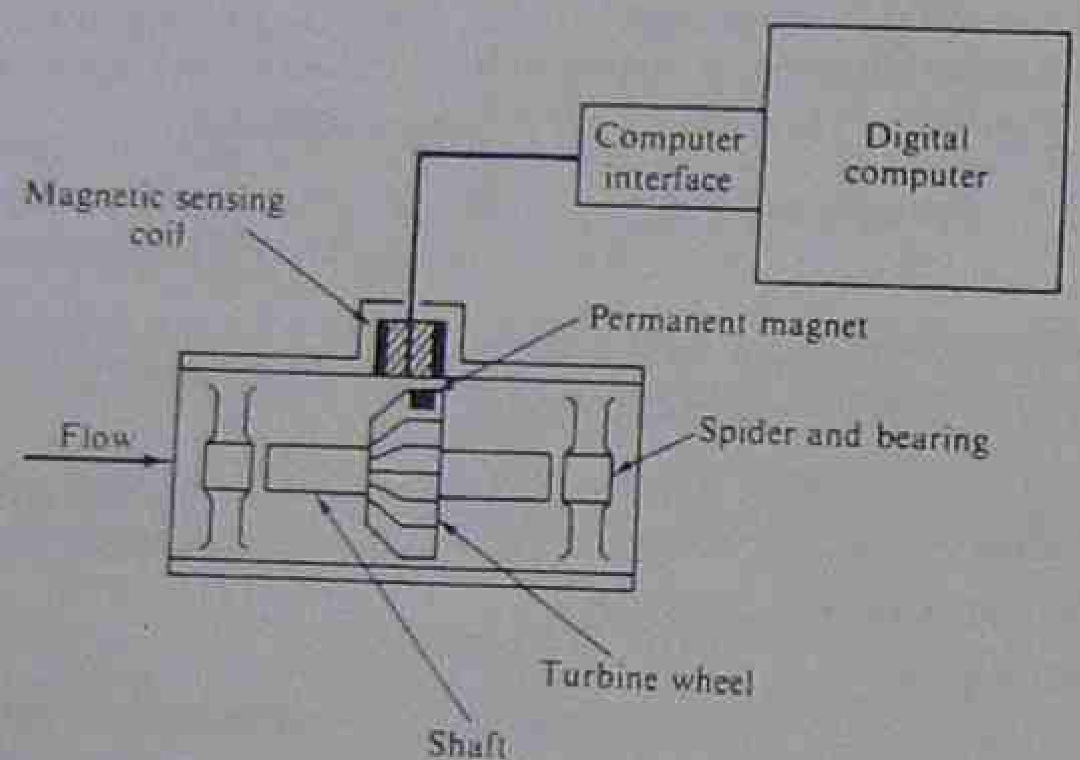


Figure 10.7 A turbine flow meter produces an accurate linear, digital flow signal. Turbine meters are used in the petrochemical and other industries for a broad range of applications.

But $N/\Delta t$ is the number of pulses per unit time (i.e., the pulse frequency f). Thus

$$q = Kf \tag{10.2}$$

The pulse output of the turbine flow meter is ideally suited for digital counting and control techniques. Digital blending control systems make use of turbine flow meters to provide accurate control of the blending of two or more liquids. Turbine flow meters are also used to provide flow rate measurements for input to a digital computer, as shown in Figure 10.7.

Example 10.2

A turbine flow meter has a K value of 12.2 cm^3 per pulse. Determine the volume of liquid transferred for each of these pulse counts: (a) 220; (b) 1200; (c) 470. Also determine the flow rate, if each of the pulse counts above occurs during a period of 140 s.

Solution

$$V = KN$$

a. For 220 pulses in 140 s,

b. For 1200 pulses in 140 s,

c. For 470 pulses in 140 s,

Magnetic Flow Meters

The *magnetic flow meter* has no moving parts and offers no obstructions to the flowing liquid. It operates on the principle that a voltage is induced in a conductor moving in a magnetic field. A magnetic flow meter is illustrated in Figure 10.9. The saddle-shaped coils placed around the flow tube produce a magnetic field at right angles to the direction of flow. The flowing fluid is the conductor, and the flow of the fluid provides the movement of the conductor. The induced voltage is perpendicular to both the magnetic field and the direction of motion of the conductor. Two electrodes are used to detect the induced voltage, which is directly proportional to the liquid flow rate. The magnetic flow transmitter converts the induced ac voltage into a dc electric current signal suitable for use by an electronic controller.

Desirable features of magnetic flow meters include no obstruction to the fluid flow, no moving parts, low electric power requirements, excellent for slurries, and very low flow capabilities. Less desirable features include the fact that the fluid must have a minimum electrical conductivity, the large size and high cost of a magnetic flow meter, and the fact that periodic zero flow checks are required.

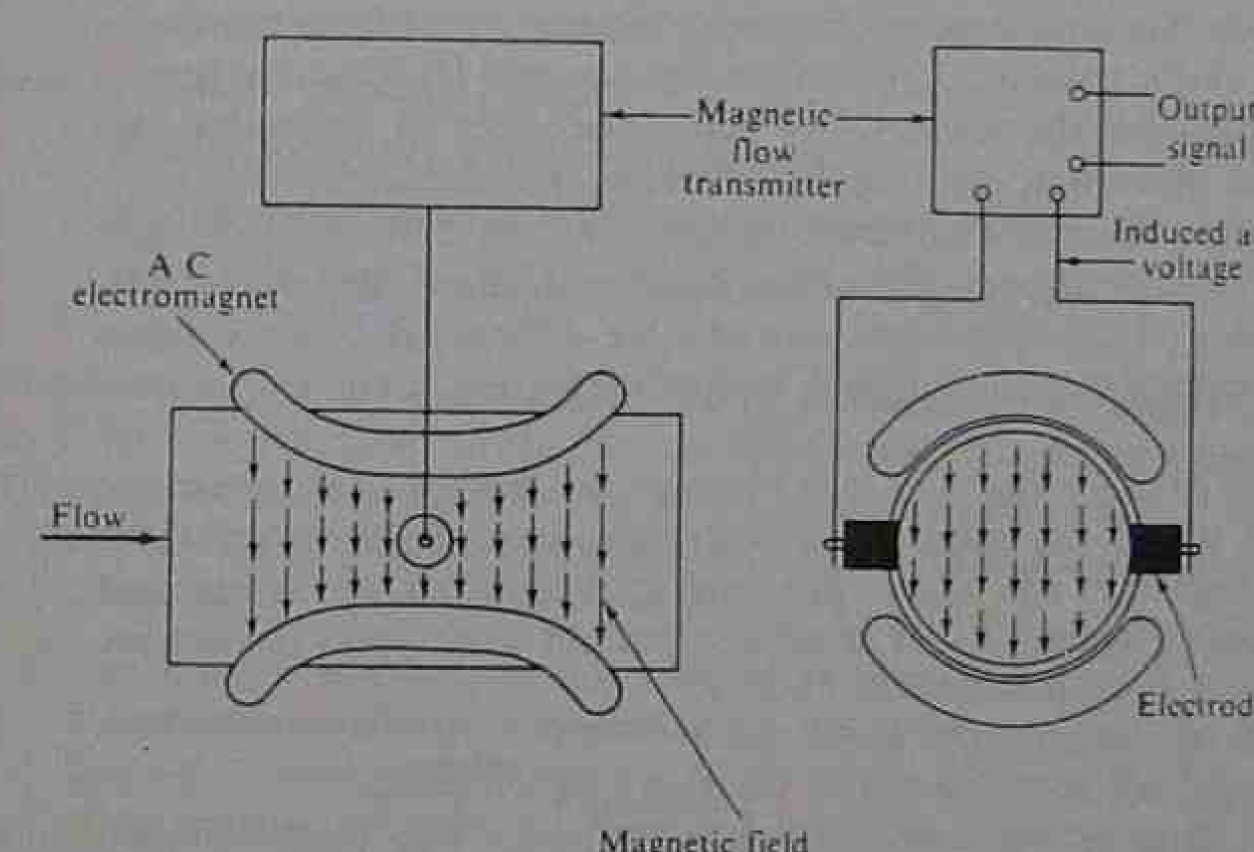


Figure 10.9 A magnetic flow meter has a completely unobstructed flow path, a decided advantage for slurries and food products.

Differential Pressure Flow Meters

Differential pressure flow meters operate on the principle that a restriction placed in a flow line produces a pressure drop proportional to the flow rate squared. A differential pressure transmitter is used to measure the pressure drop (h) produced by the restriction. The flow rate (q) is proportional to the square root of the measured pressure drop.

$$q = K\sqrt{h} \tag{10.1}$$

The restriction most often used for flow measurement is the orifice plate—a plate with a small hole, which is illustrated in Figure 10.6a. The orifice is installed in the flow line in such a way that all the flowing fluid must pass through the small hole (see Figure 10.6b).

Special passages transfer the fluid pressure on each side of the orifice to opposite sides of the diaphragm unit in a differential pressure transmitter. The diaphragm arrangement converts the pressure difference across the orifice into a force on one end of a force beam. A force transducer on the other end of the beam produces an exact counterbalancing force. A displacement detector senses any motion resulting from an imbalance of the forces on the force arm. The amplifier converts this displacement signal into an adjustment of the current input to the force transducer that restores the balanced condition. The counterbalancing force produced by the force transducer is proportional to both the pressure drop and the input current (I). Thus the current (I) is directly proportional to the pressure drop across the orifice (h). This same electric current is used as the output signal of the differential pressure transducer.

In Figure 10.6 the orifice is the primary element, and the differential pressure transmitter is the secondary element. The orifice converts the flow rate into a differential pressure signal, and the transmitter converts the differential pressure signal into a proportional electric current signal. A typical calibration curve is illustrated in Figure 10.6c.

Desirable features of the orifice flow meter include the fact that it is simple and easy to fabricate, has no moving parts, that a single differential pressure transmitter can be used without regard to pipe size or flow rate, and that it is a widely accepted standard. A less desirable feature is that an orifice does not work well with slurries.

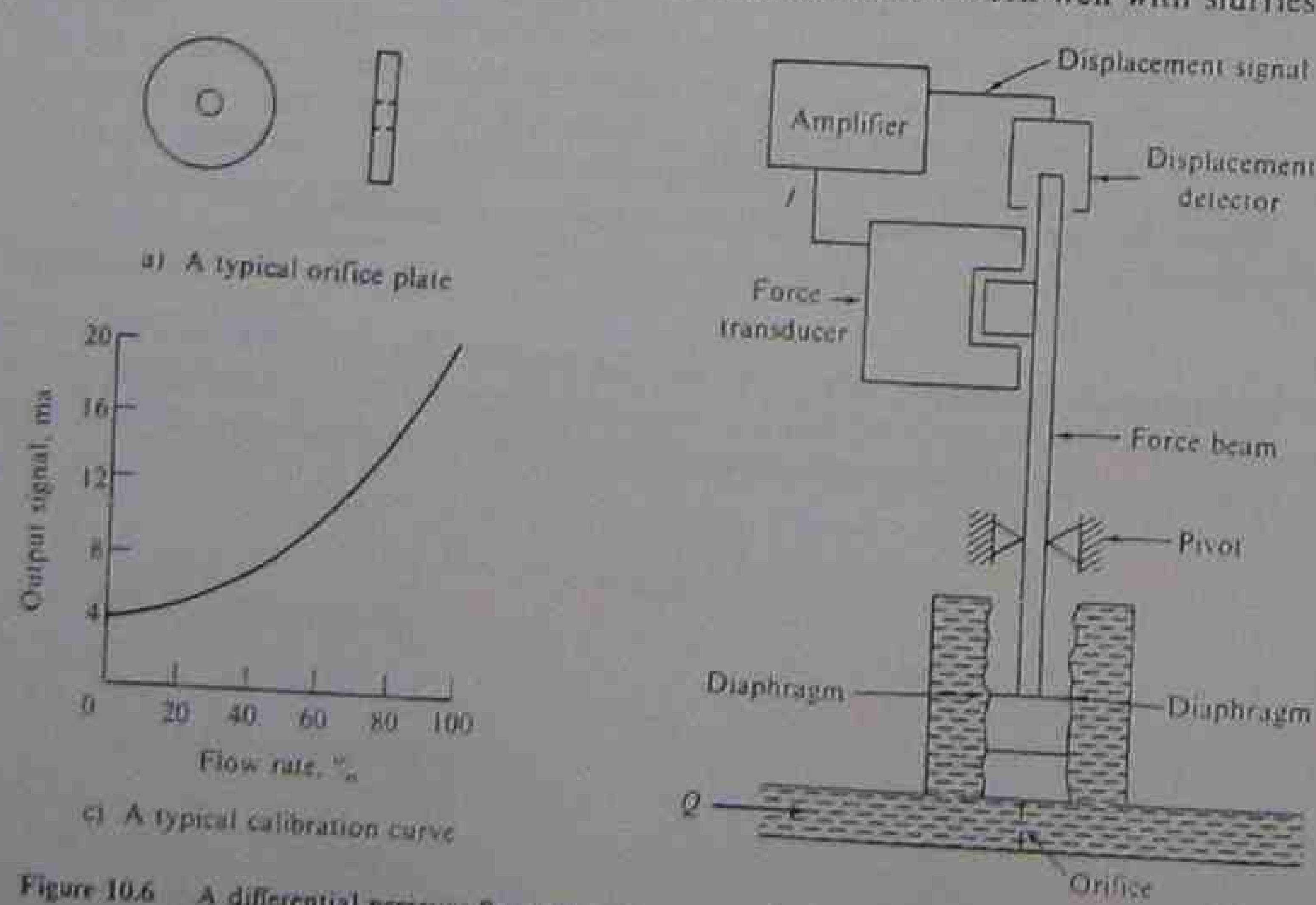


Figure 10.6 A differential pressure flow transmitter is the most widely used method of flow measurement.

10.4 LIQUID LEVEL MEASUREMENT

Sensing Methods

The measurement of the level or weight of material stored in a vessel is frequently encountered in industrial processes. Liquid level measurement may be accomplished directly by following the liquid surface, or indirectly by measuring some variable related to the liquid level. The direct methods include sight glasses and various floats with external indicators. Although simple and reliable, direct methods are not easily modified to provide a control signal. Consequently, indirect methods provide most level control signals.

Many indirect methods employ some means of measuring the static pressure at some point in the liquid. These methods are based on the fact that the static pressure is proportional to the liquid density times the height of liquid above the point of measurement.

$$p = \rho g h \tag{10.4}$$

where p = static pressure, pascal
ρ = liquid density, kilogram/cubic meter
h = height of liquid above the measurement point, meter
g = 9.81 meter/second^2 (acceleration due to gravity)

Thus any static pressure measurement can be calibrated as a liquid level measurement. If the vessel is closed at the top, the differential pressure between the bottom and the top of the vessel must be used as the level measurement.

The following are examples of some of the other indirect methods used to measure liquid level.

1. The displacement float method is based on the fact that the buoyant force on a stationary float is proportional to the liquid level around the float.
2. The capacitance probe method is based on the fact that the capacitance between a stationary probe and the vessel wall depends on the liquid level around the probe.
3. The gamma-ray system is based on the fact that the number of gamma rays that penetrate a layer of liquid depends on the thickness of the layer.

Static Pressure Level Sensors

Static pressure level sensors use the static pressure at some point in the liquid as a measure of the level. They are based on the fact that the static pressure is proportional to the height of the liquid above the point of measurement. The relationship is given by the equation

$$p = \rho g h \tag{10.4}$$

where p = static pressure, pascal

ρ = liquid density, kilogram/cubic meter

h = height of liquid above the measurement point, meter

$g = 9.81\ \text{m/s}^2$

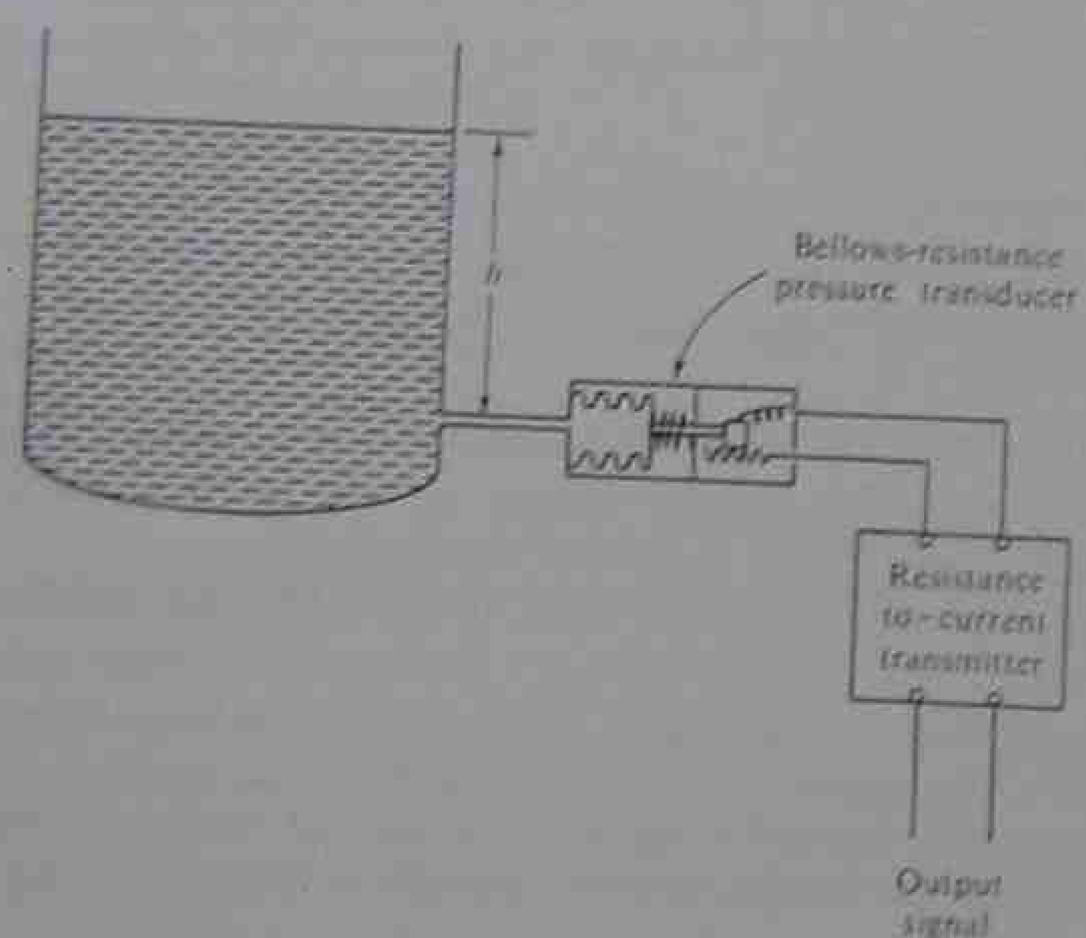


Figure 10.14 A static pressure sensor uses the pressure near the bottom of the tank as a measure of the liquid level.

If the top of the tank is open to atmospheric pressure, an ordinary pressure gage may be used to measure the pressure at some point in the liquid. A variety of methods is used to measure the static pressure. One method is illustrated in Figure 10.14, where a bellows resistance pressure sensor and transmitter is used to measure the level. The output of the transmitter is a 4- to 20-mA current signal corresponding to a level range from 0 to 100%.

Displacement Float Level Sensors

A displacement float level sensor is illustrated in Figure 10.13. The float applies a downward force on the force beam equal to the weight of the float minus the buoyant force of the liquid around the float. The force on the beam is given by the following equation:

$$f = Mg - \rho g A h \tag{10.5}$$

where f = net force, newton

M = mass of the float, kilogram

$g = 9.81\ \text{m/s}^2$ (gravity)

ρ = liquid density, kilogram/cubic meter

A = horizontal cross-sectional area of the float, square meter

h = length of the float below the liquid surface, meter

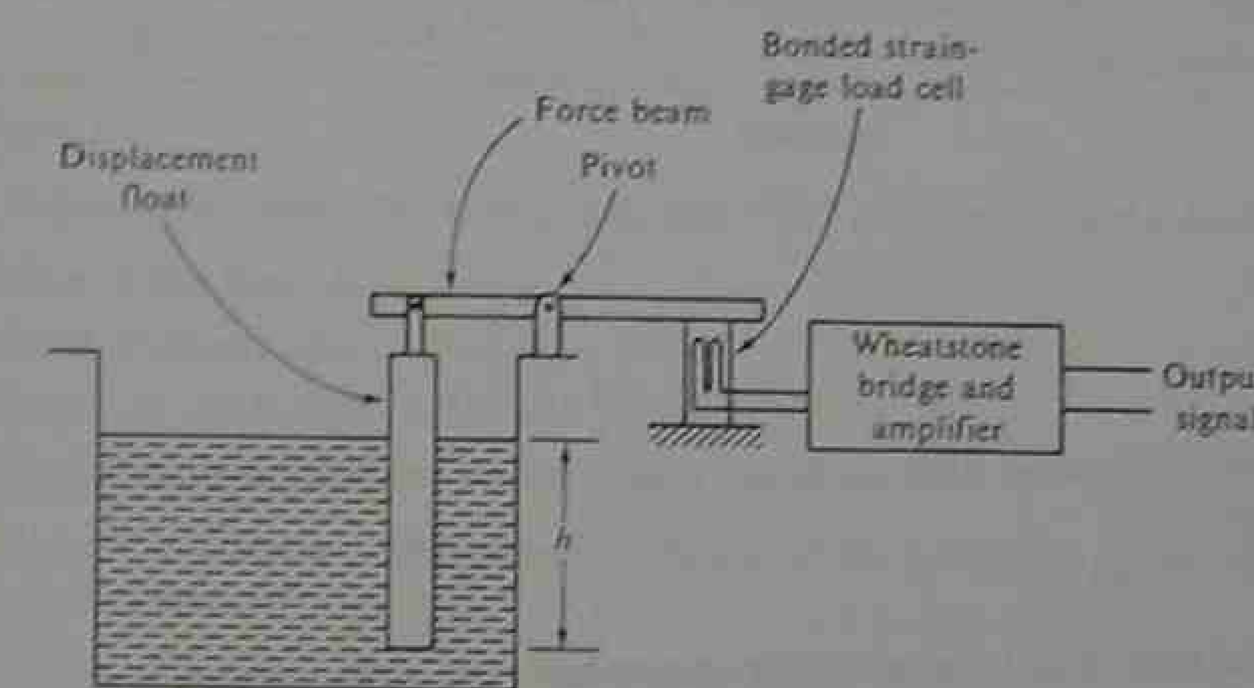


Figure 10.13 A float and a force or displacement sensor use the buoyant force on the float as a measure of the liquid level.

Equation (10.5) shows that the force, f, bears a linear relationship to the liquid level.

The load cell applies a balancing force on the force beam that is proportional to f and, consequently, bears a linear relationship with the liquid level. The load cell is a strain gage force transducer that varies its resistance in proportion to the applied force. A signal conditioner converts the load cell resistance into a usable electrical signal.

STATIC PRESSURE LEVEL SENSOR – IN A CLOSED TANK

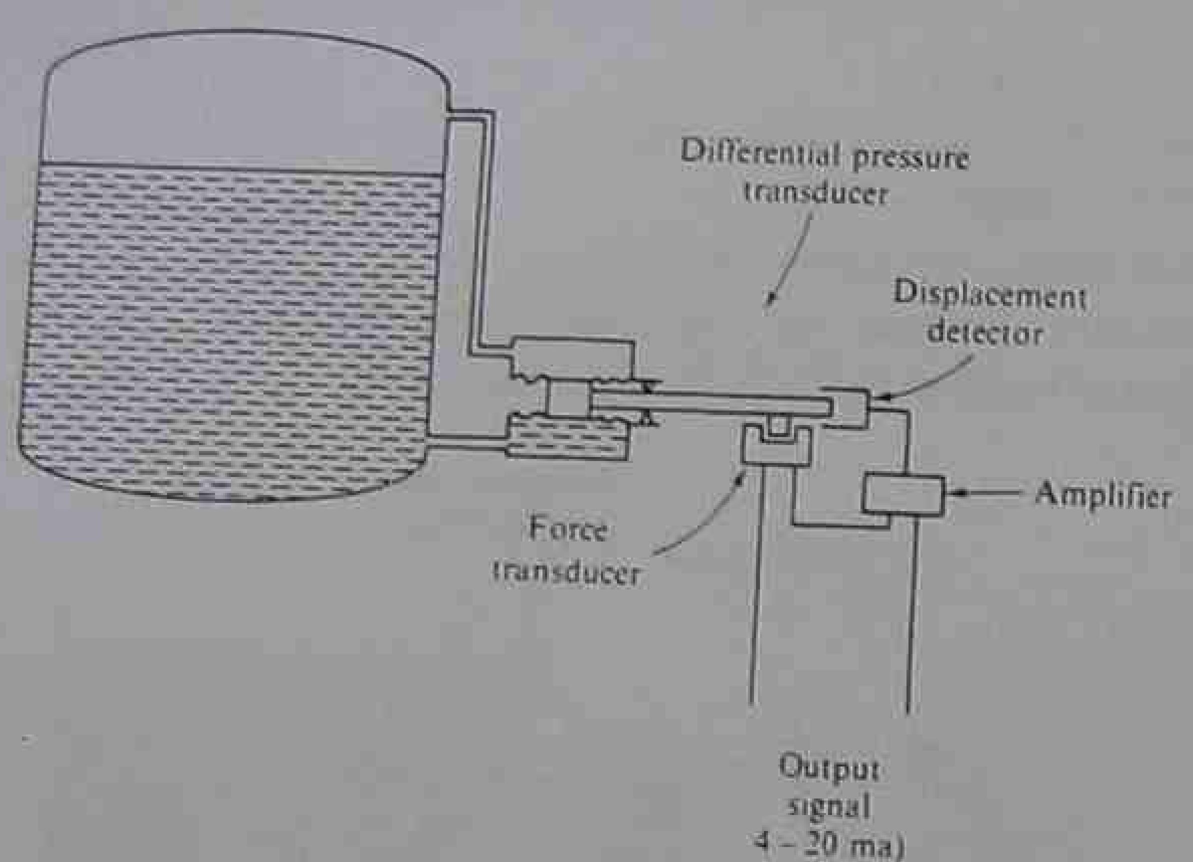


Figure 10.15 A differential pressure sensor can be used to measure liquid level in a closed tank that is under high pressure.

If the top of the tank is not vented to the atmosphere, the static pressure is increased by the pressure in the tank at the liquid surface. The height of the liquid above the point of measurement is proportional to the difference between the static pressure and the pressure at the top of the tank. A differential pressure measurement is required. Figure 10.15 illustrates the use of a differential pressure transducer to measure the level in a closed tank.

Example 10.3

A bellows pressure element similar to Figure 10.12b has the following values.

Effective area of bellows = 12.9 cm^2
Spring rate of the spring = 80 N/cm
Spring rate of the bellows = 6 N/cm

What is the pressure range of the sensor if the motion of the bellows is limited to 1.5 cm?

Solution

The total spring rate is + = N/cm. The force required to deflect the spring a distance of 1.5 cm is (cm) × (N/cm) = N. The pressure required to produce this force is (N)/(cm^2) = N/cm^2. The range is 0 to N/cm^2, or 0 to kN/m^2.

Example 10.4

The displacement float level sensor in Figure 10.13 has the following data.

Mass of the float, $M = 2.0$ kg
Cross-sectional area of float, $A = 20\ cm^2$
Length of the float, $L = 2.5$ m
Liquid in the vessel, kerosene

Determine the minimum and maximum values of the force, f, applied to the force beam by the float.

Solution

From Table 3 in Appendix A, the density (ρ) of kerosene is 800 kg/m^3. The force, f, is given by Equation (10.5).

$$f = Mg - \rho g A h$$
$$= (\ )(\ ) - (\ )(\ )\left(\frac{}{}\right)h$$
$$= \quad - \quad h \text{ N}$$

The minimum force occurs when $h = L$:

$$f_{min} = \quad - (\ )(\ )$$
$$= - \quad \text{N}$$

The maximum force occurs when $h = 0$:

$$f_{max} = \quad - (\ )(\ )$$
$$= \quad \text{N}$$

The force applied by the float on the force beam ranges from 19.62 N when the vessel is empty to −19.63 N when the vessel is full.

Example 10.5

Water is to be stored in an open vessel. The static pressure measurement point is located 2 m below the top of the tank. A pressure transducer is selected to measure the liquid level. Determine the range of the pressure transducer for 100% output when the tank is full.

Solution

Equation (10.4) gives the desired relationship. The density of water is obtained from Table 3 in Appendix A.

$$\rho = \qquad \text{kg/m}^3$$

$$p = \rho g h$$

$$p_{max} = (\qquad)(\qquad)(\qquad)$$

$$= \qquad \text{Pa}$$

or

$$p_{max} = \qquad \text{kPa}$$

10.7 A differential pressure flow meter is used as the sensor in a liquid flow control system. A technician obtained the following data in a series of calibration tests. Each test consisted of measuring the time required to fill a 0.5-gal container. The tests were conducted at controller settings of 25, 50, 75, and 100%.

Controller Setting (%)	Time to Fill a 0.5-gal Container (min)
25	6.63
50	4.78
75	3.95
100	3.45

Determine the average flow rate in gal/min for each controller setting by dividing 0.5 gal by the time in minutes. Then construct a calibration curve from your calculated results.

10.8 The flow control system in Exercise 10.7 is described by the equation

$$q = K\sqrt{sp}$$

where q = flow rate, gallon/minute

K = constant

sp = controller setting, percent

Calculate the value of K for each of the four calibration test results in Exercise 10.7. Find the average of the four K values and use this value to express the relationship between the flow rate, q, and the square root of the controller setting, sp.

10.9 A turbine flow meter has a K value of 34.1 cm³ per pulse. Determine the volume of liquid transferred for each of the following pulse counts.

a. 8200 b. 32,060 c. 680

10.10 Determine the flow rate for each pulse count in Exercise 10.9 if the pulse counts occur during a 210-s time interval.

10.11 A bellows pressure element similar to Figure 10.12a has the following values.

Effective area of the bellows = 21 cm²

Spring rate of the spring = 200 N/cm

Spring rate of the bellows = 10 N/cm

What is the pressure range of the transducer if the motion of the bellows is limited to 2 cm?

10.12 A bellows pressure element similar to Figure 10.12b has the following values.

Effective area of the bellows = 4 cm²

Spring rate of the spring = 400 N/cm

Spring rate of the bellows = 25 N/cm

What stroke of the bellows is required for a range of 0 to 50 N/cm²?

10.13 A displacement float level transducer has the following data.

Mass of the float, M = 6.5 kg

Area of the float, A = 50 cm²

Length of the float, L = 4.0 m

Liquid in the vessel, water

Determine the minimum and maximum values of the force, F, applied to the force beam by the float.

10.14 Gasoline is to be stored in a vented vessel. The static-pressure measurement point is 8 m below the top of the vessel. A pressure sensor is to be used as the level sensor. Determine the range of the pressure sensor for 100% output when the tank is filled.

TABLE-3

Properties of Liquids

Liquid	Density kilogram/meter³	Absolute Viscosity pascal second	Thermal Conductivity watt/meter kelvin	Heat Capacity joule/kg kelvin
Ethyl alcohol	800	0.0013	0.18	2,300
Gasoline	740	0.0005	0.14	2,100
Glycerine	1,260	0.83	0.29	2,400
Kerosene	800	0.0024	0.15	2,010
Mercury	13,600	0.0015	8.0	140
Oil	880	0.160	0.15	2,180
Turpentine	870	0.0015	0.13	1,720
Water	1,000	0.001	0.6	4,190

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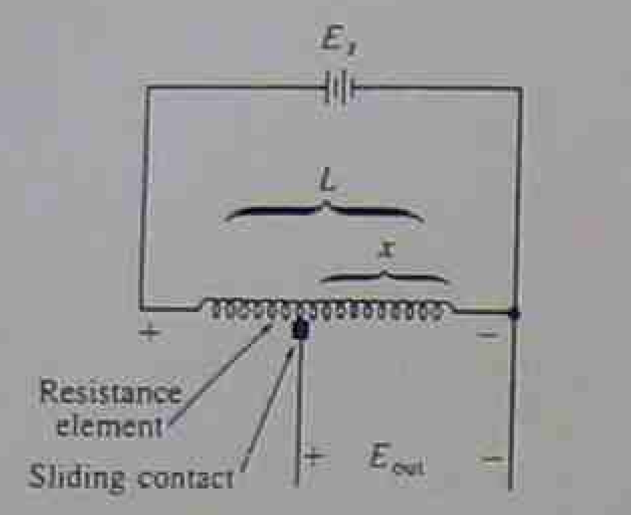
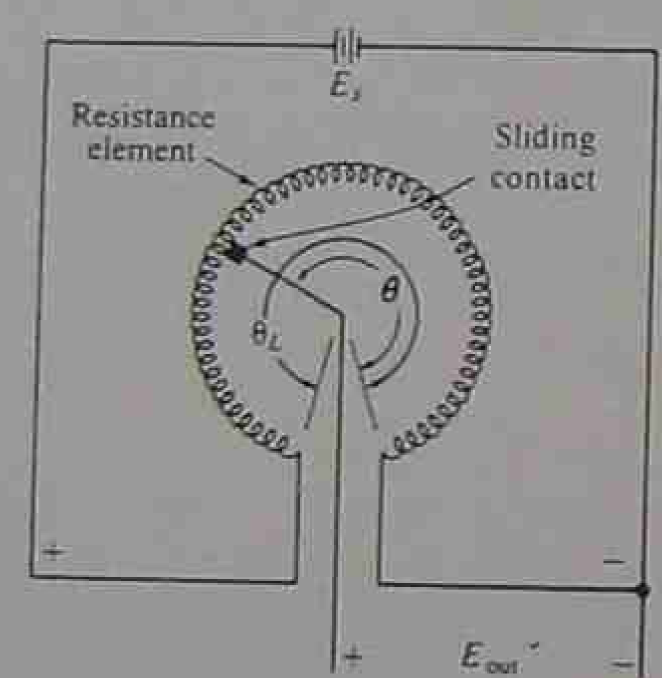
Potentiometers

A *potentiometer* consists of a resistance element with a sliding contact that can be moved from one end to the other. Potentiometers are used to measure both linear and angular displacement, as illustrated in Figure 9.3. The resistance element produces a uniform drop in the applied voltage, E_s, along its length. As a result, the voltage of the sliding contact is directly proportional to its distance from the reference end.

Figure 9.3 Two types of potentiometric displacement sensors: (a) linear; (b) angular. In both types, E_{out} is a measure of the position of the sliding contact.



a) A linear displacement potentiometer



b) An angular displacement potentiometer

$$\text{Linear potentiometer:} \quad E_{out} = \left(\frac{x}{L}\right)E_s$$

$$\text{Angular potentiometer:} \quad E_{out} = \left(\frac{\theta}{\theta_L}\right)E_s$$

When the resistive element is wirewound, the resolution of the potentiometer is determined by the voltage step between adjacent loops in the element. If there are N turns in the element, the voltage step between successive turns is $E_T = E_S/N$, where E_S is the full-scale voltage. Expressed as a percentage of the full-scale output, the percentage resolution is given by the following relationship.

$$\text{Resolution } (\%) = \frac{100E_T}{E_S} = \frac{100(E_s/N)}{E_S}$$

or

$$\text{Resolution } (\%) = \frac{100}{N} \qquad \textbf{(9.1)}$$

Potentiometers are subject to an error whenever a current passes through the lead wire connected to the sliding contact. This error is called a *loading error* because it is caused by the load resistor connected between the sliding contact and the reference point. A potentiometer with a load resistor is illustrated in Figure 9.4. If R_p is the resistance of the potentiometer and a is the proportionate position of the sliding contact, then aR_p is the resistance of the portion of the potentiometer between the sliding contact and the reference point. The load resistor, R_L, is connected in parallel with resistance aR_p. The equivalent resistance of this parallel combination is $(R_L)(aR_p)/(R_L + aR_p)$.

The resistance of the remaining portion of the potentiometer is equal to $(1 - a)R_p$, and the equivalent total resistance is the sum of the last two values.

$$R_{EQ} = (1 - a)R_p + \frac{aR_LR_p}{R_L + aR_p}$$

The output voltage, E_{out}, may be obtained by voltage division as follows:

$$E_{out} = \left(\frac{aR_LR_p/(R_L + aR_p)}{(1 - a)R_p + aR_LR_p/(R_L + aR_p)}\right)E_s$$

$$E_{out} = \left(\frac{a}{1 + ar - a^2r}\right)E_S$$

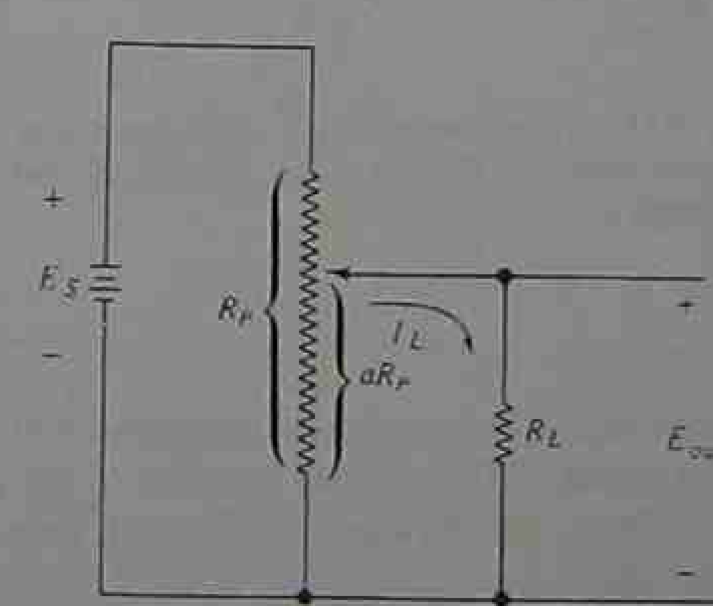
where

$$r = \frac{R_p}{R_L}$$

The loading error is the difference between the loaded output voltage (E_{out}) and the unloaded output voltage (aE_S).

$$\begin{aligned}\text{Loading error} &= aE_S - E_{out}\\ &= aE_S - \left(\frac{a}{1 + ar - a^2r}\right)E_S\\ &= \left(\frac{a^2r(1 - a)}{1 + ar - a^2r}\right)E_S \quad \text{volts}\end{aligned}$$

Figure 9.4 A loading error is produced in a potentiometer when a load resistor is connected between the sliding contact and the reference terminal.



The loading is usually expressed as a percentage of the full-scale range, E_S.

$$\text{Loading error } (\%) = 100\left[\frac{a^2r(1 - a)}{1 + ar(1 - a)}\right] \qquad \textbf{(9.2)}$$

Resolvers

A *resolver* is a rotary transformer that produces an output signal that is a function of the rotor position. Figure 9.10 shows the position of the coils in a resolver. The two rotor coils are placed 90° apart. The two stator coils are also placed 90° apart. Either pair of coils can be used as the primary with the other pair forming the secondary. The following equations define the secondary voltages in terms of the primary voltages when the rotor coils are used as the primary.

$$E_1 = K(E_3 \cos \theta - E_4 \sin \theta) \quad (9.5)$$
$$E_2 = K(E_4 \cos \theta + E_3 \sin \theta) \quad (9.6)$$

When a resolver is used as a sensor, one of the rotor windings is shorted as shown in Figure 9.11. If E_4 is the shorted coil, Equations (9.5) and (9.6) simplify to the following.

$$E_1 = KE_3 \cos \theta \quad (9.7)$$
$$E_2 = KE_3 \sin \theta \quad (9.8)$$

Equations (9.7) and (9.8) define the output of the resolver shown in Figure 9.11. This output is reasonably linear for values of θ over a range of $\pm 35°$. The excitation voltage, E_3, is a sinusoidal voltage that can be represented as follows:

$$E_3 = A \sin \omega t$$

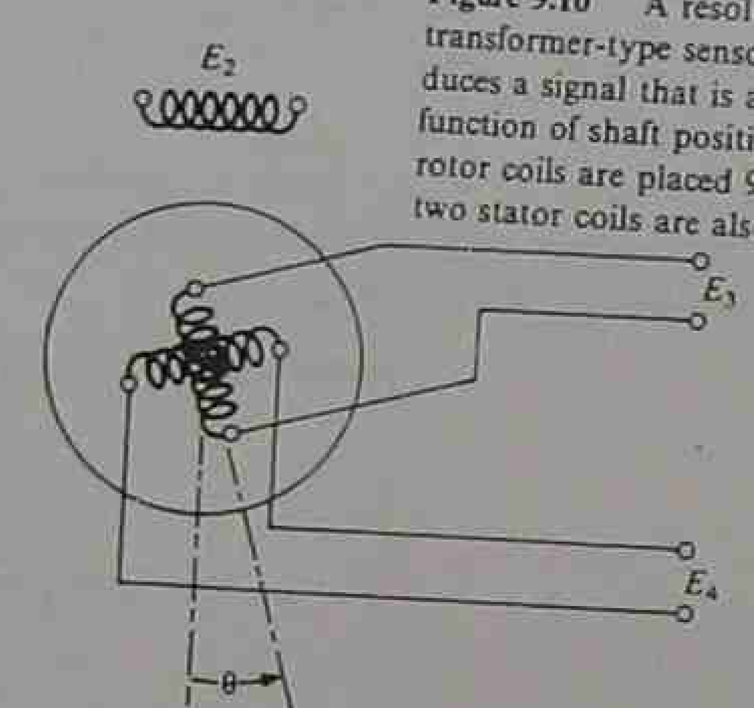


Figure 9.10 A resolver is a transformer-type sensor that produces a signal that is a trigonometric function of shaft position, θ. The two rotor coils are placed 90° apart. The two stator coils are also 90° apart.

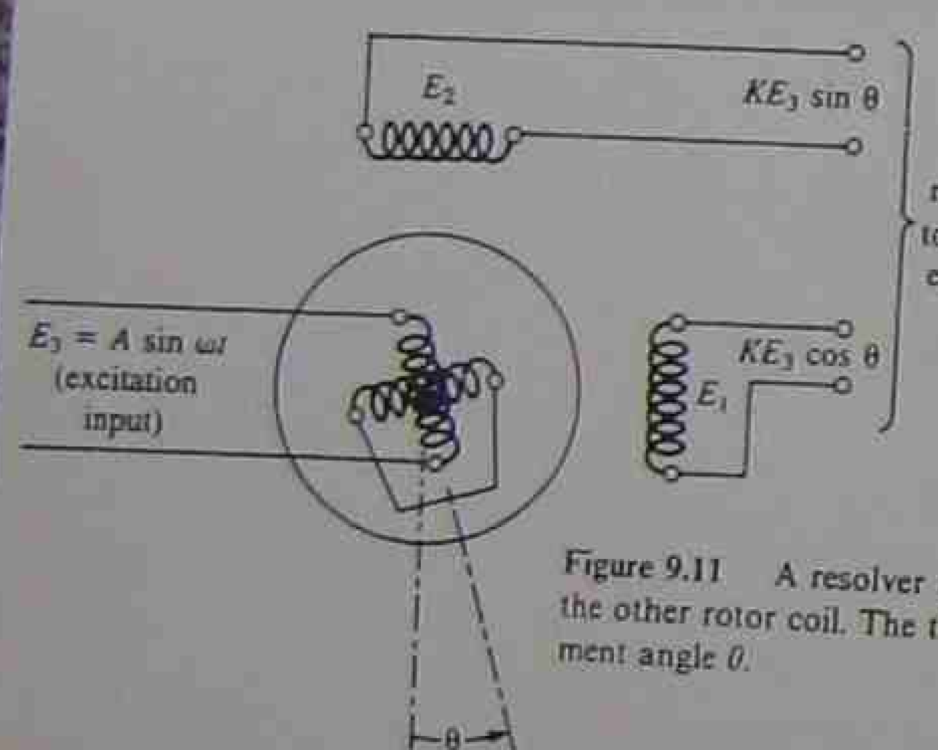


Figure 9.11 A resolver used as a sensor has one of its rotor coils shorted. An ac voltage is applied to the other rotor coil. The two output voltages, E_1 and E_2, are trigonometric functions of the displacement angle θ.

Output voltage, E_1, is essentially a sinusoidal voltage whose amplitude varies according to the cosine of the angular position of the rotor. Output voltage, E_2, is also a sinusoidal voltage, but its amplitude varies according to the sine of the angular position of the rotor.

The resolver position sensor requires a signal conditioning circuit that can convert the two voltages, E_1 and E_2, into a usable signal representing the position of the rotor, θ. If a digital signal is required, the signal conditioner must also convert the signal from an analog form to a digital form. We will refer to a signal conditioner that performs both functions as a *resolver-to-digital converter.*

One problem with a resolver is the necessity of brushes and slip rings to bring the excitation voltage to coil E_3 on the rotor. The brushes are subject to wear and must be protected from the dirty environment encountered in industry. The brushless resolver has been developed to solve this problem. A brushless resolver uses a transformer to couple the excitation voltage to the rotor coil, eliminating the need for brushes and slip rings.

Optical Encoders

An *encoder* is a device that provides a digital output in response to a linear or angular displacement. The resolver and digital converter discussed in the preceding section is an angular encoder. In this section we describe another type of encoder, the optical encoder.

An optical encoder has four main parts: a light source, a code disk, a light detector, and a signal conditioner. This section deals with the first three parts. Position encoders can be classified into two types: incremental encoders and absolute encoders.

An *incremental encoder* produces equally spaced pulses from one or more concentric tracks on the code disk. The pulses are produced when a beam of light passes through accurately placed holes in the code disk. Each track has its own light beam; thus an encoder with three tracks will have three light sources and three light sensors. Figure 9.12 illustrates an optical encoder with three tracks. Each track has a series of equally spaced holes in an otherwise opaque disk. The inside track has only one hole,

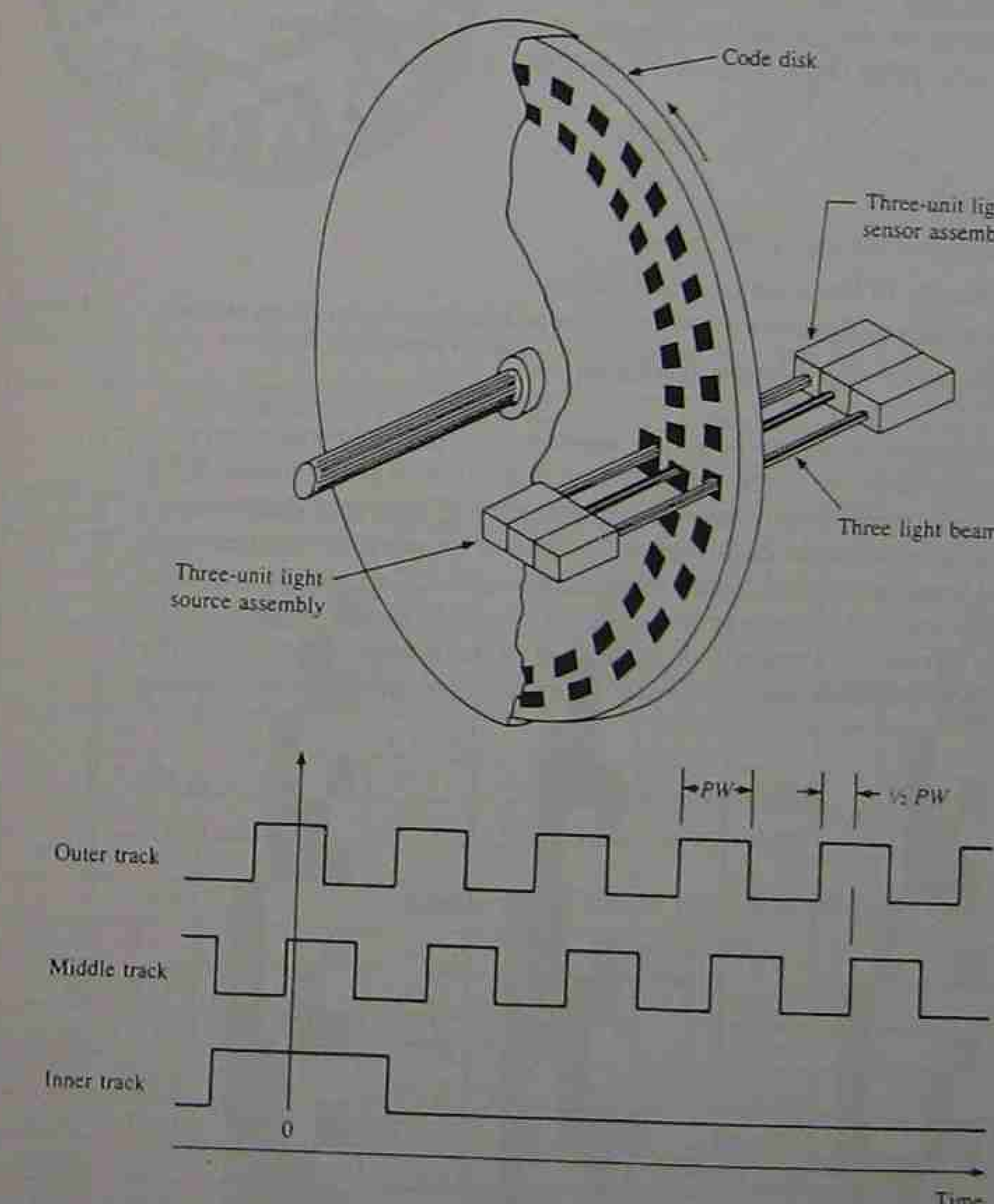


Figure 9.12 The incremental encoder has three tracks. The inner track provides a reference signal to locate the home position. The middle track provides information about the direction of rotation. In one direction, the middle track lags the outside track; in the other direction, the middle track leads the outside track.

which is used to locate the "home" position on the code disk. The other two tracks have a series of equally spaced holes that go completely around the code disk. The holes in the middle track are offset from the holes in the outside track by one-half the width of a hole. The purpose of the offset is to provide directional information. The diagram of the track pulses in Figure 9.12 was made with the disk rotating in the counterclockwise direction. Notice that the pulses from the outer track lead the pulses from the inner track by one-half the pulse width. If the direction is reversed to a clockwise direction, the pulses from the middle track will lead the pulses from the outer track by the same one-half of the pulse width.

The primary functions of the signal conditioner for an incremental encoder are to determine the direction of rotation and count pulses to determine the angular displacement of the code disk. The pulse count is a digital signal, so an analog-to-digital converter is not required for an encoder.

An angular, incremental encoder can be used to measure a linear distance by coupling the encoder shaft to a tracking wheel as shown in Figure 9.13. The wheel rolls along the surface to be measured, and the signal conditioner counts the pulses. The total displacement that can be measured in this manner is limited only by the capacity of the counter in the signal conditioner. The incremental encoder simply rotates as many times as the application requires. The measured displacement is obtained from the total pulse count as given by the following equation:

$$x = \frac{\pi d N_T}{N_R} \qquad (9.9)$$

where x = measured displacement, meter
d = diameter of the tracking wheel, meter
N_T = total pulse count
N_R = number of pulses in one revolution

An *absolute encoder* produces a binary number that uniquely identifies each position on the code disk. Absolute encoders may have from 6 to 20 tracks. Each track produces one bit of the binary number according to the code that is established by the hole pattern in the code disk. Figure 9.14 shows an absolute encoder with seven tracks that form the natural binary representation of 128 unique positions on the code disk. The number of unique positions on the code disk is related to the number of bits in the binary number (which is equal to the number of tracks on the code disk). This also establishes the resolution of the encoder according to the following equations:

$$\text{Number of positions} = 2^N \qquad (9.10)$$

$$\text{Resolution} = 1 \text{ part in } 2^N \qquad (9.11)$$

where N = number of tracks = number of bits in the number

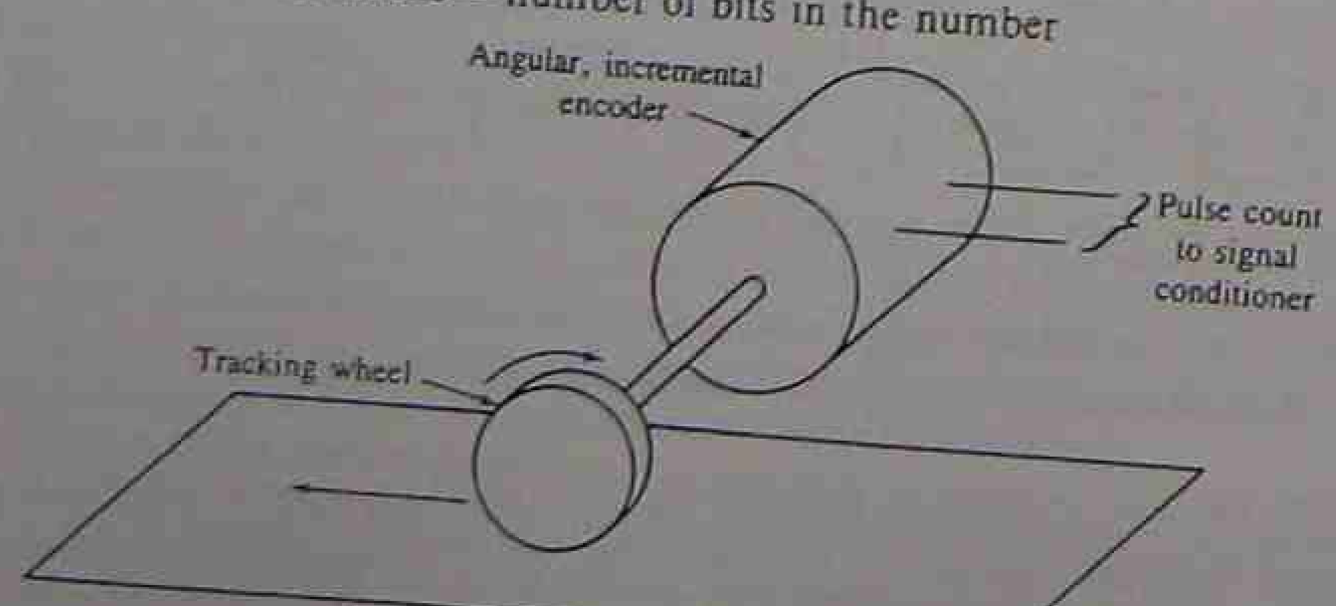


Figure 9.13 An incremental encoder coupled to a tracking wheel is used to measure linear displacement.

There are a number of binary codes that could be used in an encoder. The three most popular codes are the natural binary code, the Gray code, and the BCD code. Figure 9.15 shows the pattern for these three codes for the numbers from 0 to 10. The Gray code is a popular code for counters because only one bit changes each time the count increases by one.

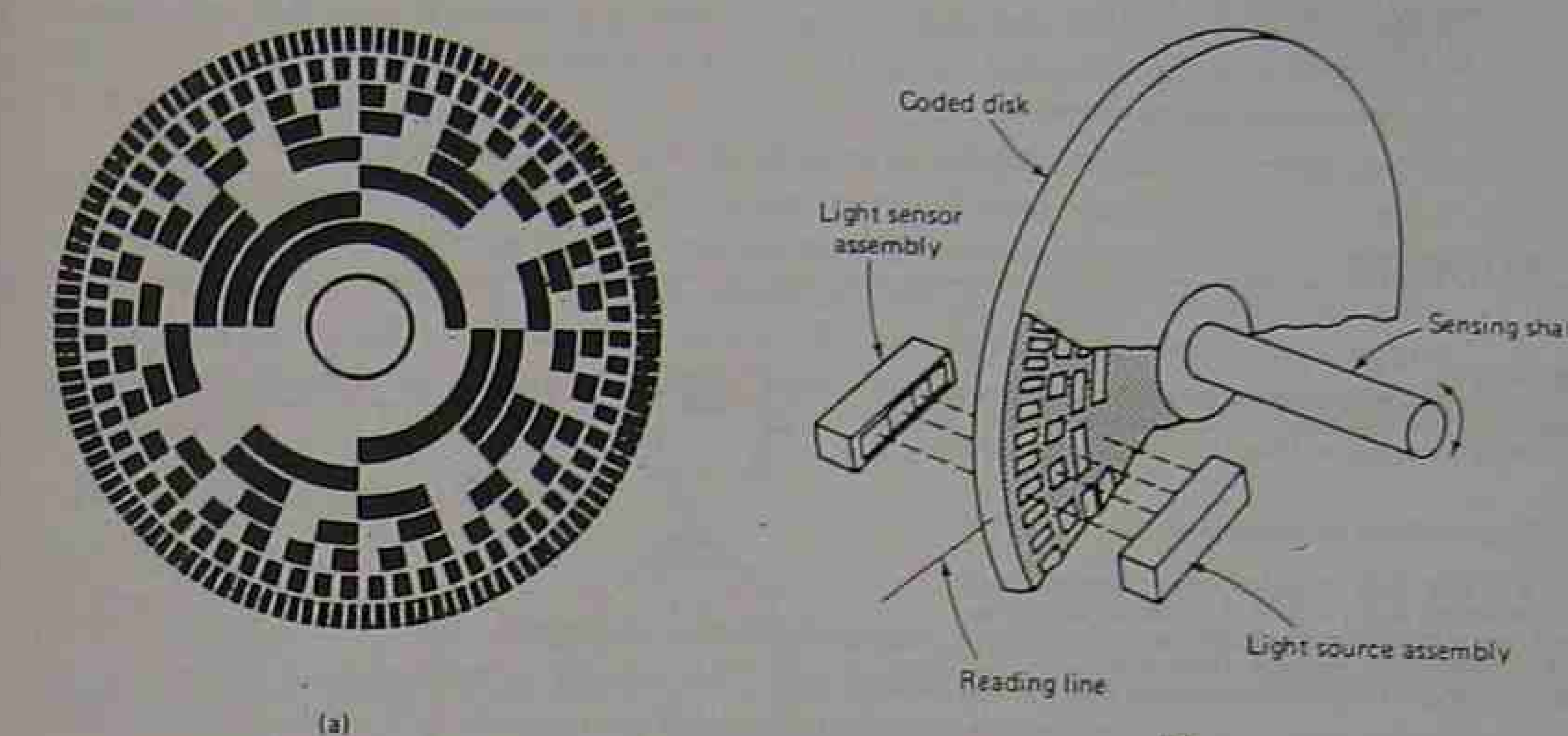
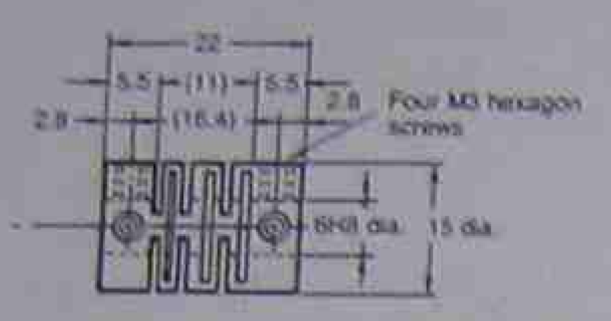


Figure 9.14 Absolute optical encoder: (a) typical code disk; (b) encoder elements. [From H. Norton, *Sensor and Analyzer Handbook* (Englewood Cliffs, N.J.: Prentice-Hall, Inc., 1982), Fig. 1-44, p. 107.]

Arabic number	(Natural) Binary: Digital number 8 4 2 1	(Natural) Binary: Code pattern $2^3\ 2^2\ 2^1\ 2^0$	Gray (Binary): Digital number	Gray (Binary): Code pattern $G_3\ G_2\ G_1\ G_0$	BCD: Digital number Tens 8 4 2 1	BCD: Digital number Units 8 4 2 1	BCD: Code pattern Tens 2^0	BCD: Code pattern Units $2^3\ 2^2\ 2^1\ 2^0$
0	0000		0000		0000	0000		
1	0001		0001			0001		
2	0010		0011			0010		
3	0011		0010			0011		
4	0100		0110			0100		
5	0101		0111			0101		
6	0110		0101			0110		
7	0111		0100			0111		
8	1000		1100			1000		
9	1001		1101		0000	1001		
10(A)	1010		1111		0001	0000		
11(B)	1011		1110			0001		
12(C)	1100		1010			0010		
13(D)	1101		1011			0011		
14(E)	1110		1001			0100		
15(F)	1111		1000			0101		

Figure 9.15 Digital code structure for absolute encoders.

Coupling E69-C06B (included)



Material: Glass-reinforced polyacetal resin (GC-25)

Precautions

Power application
The rotary encoder may output wrong pulses for 1 second on power application. Start operating the equipment connected to the encoder at least 1 second after power has been applied to the encoder.

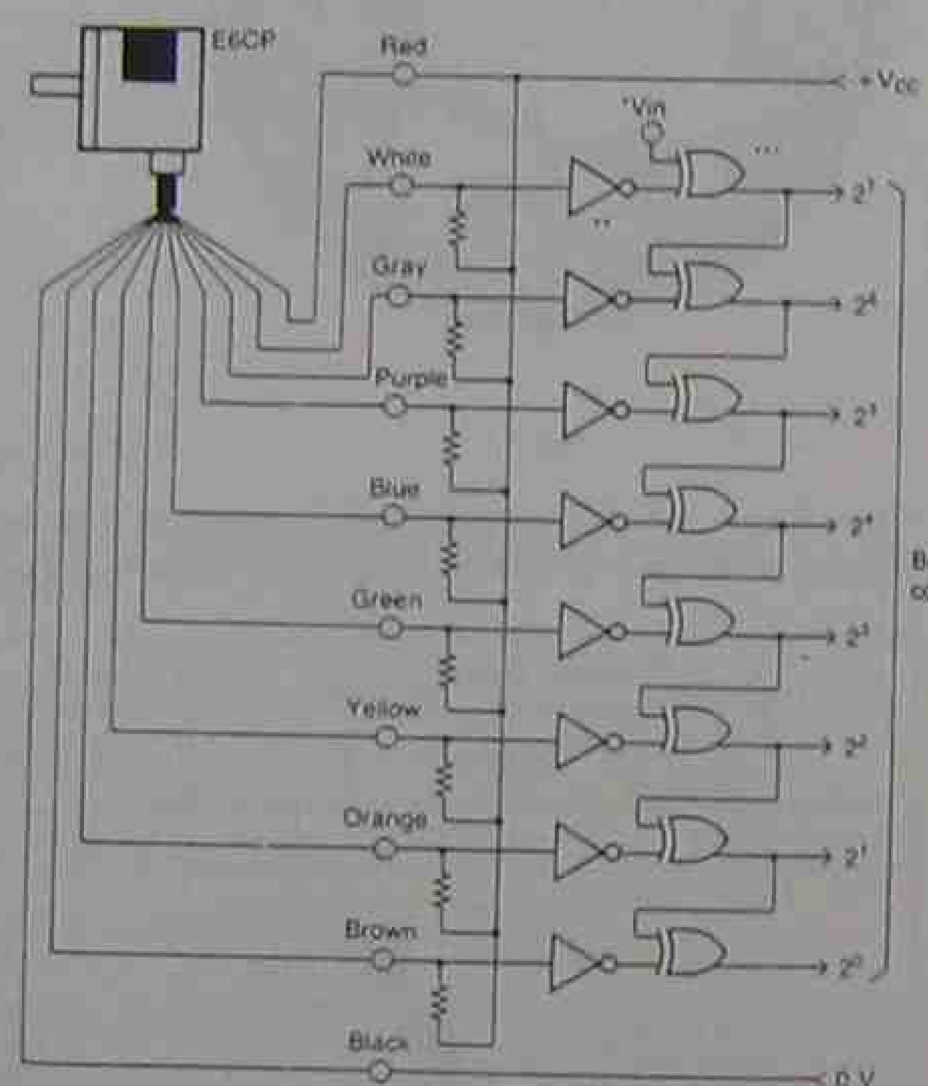
Reference (Gray-to-binary converter circuit)
Binary code
Binary code is a basic code for digital signal processing and consists of numerals 0 and 1 only. It is, however, difficult to change two or more digits simultaneously when a number represented by binary code changes. Consequently, the reading timing is very delicate, which may occasionally cause a read error.

Gray code
As shown in the table below, only one digit changes when a number represented by Gray code changes. Gray code therefore features that a read error hardly occurs and is employed in many rotary encoders (absolute) and electronic balances.

Output codes

Decimal	Binary	Gray
	2^3 2^2 2^1 2^0	
0	0 0 0 0	0 0 0 0
1	0 0 0 1	0 0 0 1
2	0 0 1 0	0 0 1 1
3	0 0 1 1	0 0 1 0
4	0 1 0 0	0 1 1 0
5	0 1 0 1	0 1 1 1
6	0 1 1 0	0 1 0 1
7	0 1 1 1	0 1 0 0
8	1 0 0 0	1 1 0 0
9	1 0 0 1	1 1 0 1
10	1 0 1 0	1 1 1 1
11	1 0 1 1	1 1 1 0
12	1 1 0 0	1 0 1 0
13	1 1 0 1	1 0 1 1
14	1 1 1 0	1 0 0 1
15	1 1 1 1	1 0 0 0

Use the circuit on the right to current Gray code into binary code.



* Gray code can be converted into positive logic binary code when the Vin terminal is connected to 0 V.
** Inverter
*** Exclusive OR

VELOCITY MEASUREMENT

Sensing Methods

Velocity is the rate of change of displacement or distance. It is measured in units of length per unit time. Velocity is a vector quantity that has both magnitude (speed) and direction. A change in velocity may constitute a change in speed, a change in direction, or both.

Angular velocity is the rate of change of angular displacement. It is measured in terms of radians per unit time or revolutions per unit time. Angular velocity measurement is more common in control systems than linear velocity measurement. When linear velocity is measured, it is often converted into an angular velocity and measured with an angular velocity transducer. Three methods of measuring angular velocity are considered in this section: dc tachometers, ac tachometers, and optical tachometers.

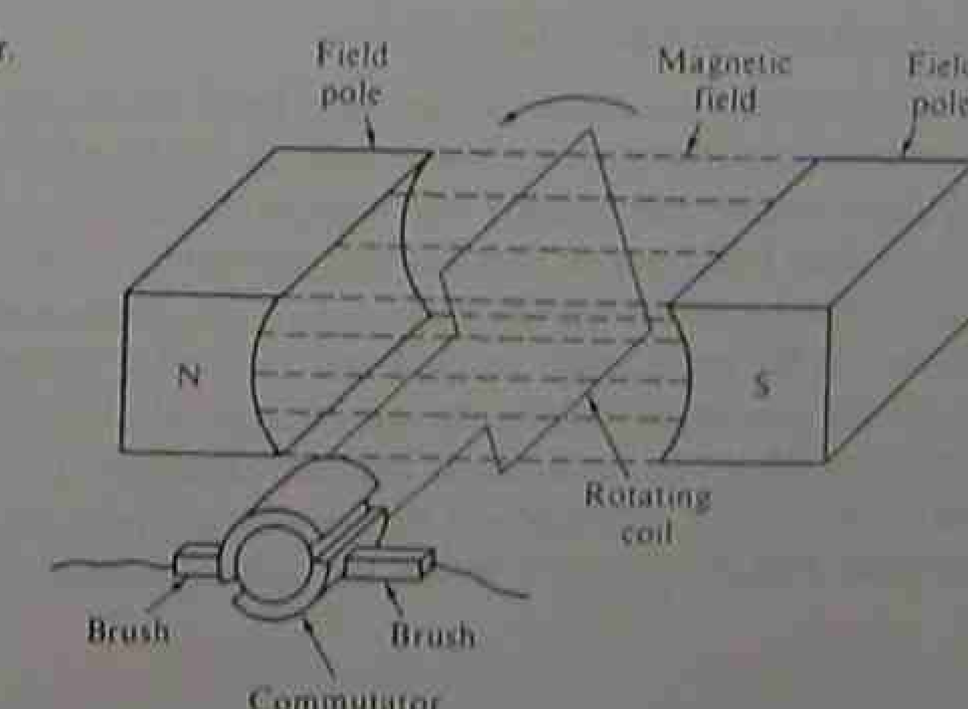
DC Tachometers

A *tachometer* is an electric generator used to measure angular velocity. A *brush-type dc tachometer* is illustrated in Figure 9.17. The coil is mounted on a metal cylinder called the *armature*. The armature is free to rotate in the magnetic field produced by the two permanent-magnet field poles. The two ends of the coil are connected to opposite halves of a segmented connection ring called the *commutator*. There are two segments on the commutator for each coil on the armature (only one is shown in Figure 9.17). For example, an armature with 11 coils would have a commutator with 22 segments.

The two carbon brushes connect the lead wires to the commutator segments. The brushes and commutator act as a reversing switch that reverses the coil connection once for each 180° rotation of the armature. This switching action converts the ac voltage induced in the rotating coil into a dc voltage. In other words, the commutator and brush constitute an ac-to-dc converter.

The tachometer produces a dc voltage that is directly proportional to the angular velocity of the armature. This voltage is based on the following fact: A voltage is induced in a conductor when it moves through a transverse magnetic field. If the

Figure 9.17 Tachometer generator.



conductor, magnetic field, and velocity are mutually perpendicular, the induced voltage (E_L) is equal to the length of the conductor (L) times the flux density (B) times the velocity of the conductor (V).

$$E_L = LBV$$

In a tachometer, the velocity is perpendicular to the magnetic field only twice during each rotation. The velocity (V) in the above equation is replaced by ($V \sin \theta$), which is the component perpendicular to the magnetic field. This means that a sinusoidal voltage is induced in each coil. However, when several coils are spaced evenly around the armature, the rectified voltage is very nearly equal to $E_L = LBV$.

The velocity of the conductors may be expressed in terms of the average radius (R), and the angular velocity (ω) in radians per second, or the angular velocity (S) in revolutions per minute.

$$V = R\omega = R2\pi S/60$$

$$E_L = 2\pi RBLS/60$$

Finally, the tachometer has N conductors of length L connected in series. The total voltage is the sum of the identical voltages induced in each conductor:

$$E = NE_L = \left(\frac{2\pi RBNL}{60}\right)S$$

The term enclosed in parentheses is called the EMF constant of the tachometer. It is designated by K_E and has units of volts per revolution per minute, or volts/rpm. Equations (9.12) and (9.13) define the electromotive force (EMF) constant and the output voltage for a dc tachometer.

$$E = K_E S = \frac{30K_E\omega}{\pi} \qquad (9.12)$$

$$K_E = \frac{2\pi RBNL}{60} \qquad (9.13)$$

where E = tachometer output, volt
K_E = EMF constant, volt/rpm
S = angular velocity, revolution/minute
ω = angular velocity, radian/second
R = average radius, meter
B = flux density of the magnetic field, weber/square meter
N = effective number of conductors
L = length of each conductor, meter

A harsh industrial environment can be very hard on brush-type tachometers. Particulate contaminants can cause excessive wear in the brushes. Gaseous contaminants build up films on the commutator which cause inaccuracies. A sealed enclosure results in excessive heat buildup and thermal drift problems. A *brushless dc tachometer* solves these problems by reversing the positions of the permanent magnet and the coil. The armature is the permanent magnet and the coil is stationary. The brushes and commutator are not required because there are no electrical connections necessary to the armature. However, additional circuitry is required to sense the position of the armature and provide appropriate solid-state switching to produce a dc output. The solid-state switching circuit serves the same function as the brushes and commutator.

AC Tachometers

An ac tachometer is a three-phase electric generator with a three-phase rectifier on its output. The ac tachometer works well at high speeds, but the output becomes nonlinear at low speeds, due to the voltage drop across the rectifiers (about 0.7 V). For this reason, ac tachometers are usually limited to speed ranges of 100 to 1, compared with 1000 to 1 for dc tachometers. The ac tachometer has no brushes and has the same ability to withstand a contaminated environment as the brushless dc generator.

Optical Tachometers

An incremental encoder connected to a rotating shaft produces a sequence of pulses from which a digital velocity signal can be easily obtained. The major signal conditioning requirement is a timed counter. For example, assume that an incremental encoder has 1000 holes in the outside track, and the counter produces a new total every 10 ms. A shaft speed of 600 rpm (10 revolutions per second) will produce $10 \times 1000 = 10{,}000$ pulses per second. The counter will count $0.01 \times 10{,}000 = 100$ pulses during a 10-ms interval. Thus a count of 100 corresponds to an angular velocity of 600 rpm. Equations (9.14) and (9.15) define the relationship between the shaft speed and the timed count for an optical tachometer.

$$S = \frac{60C}{NT_c} \qquad (9.14)$$

$$C = \frac{SNT_c}{60} \qquad (9.15)$$

where S = shaft speed, revolution/minute
N = number of pulses per shaft revolution
C = total count during time interval T_c
T_c = counter time interval, second

When a speed measurement is obtained from an absolute encoder, the track with the greatest number of holes (least significant digit) is used in the same manner as an incremental encoder. Optical encoders can handle very large dynamic ranges with extremely high accuracy and excellent long-term stability.

Example 9.1

The potentiometer in Figure 9.4 has a resistance of 10,000 Ω and a total of 1000 turns. Determine the resolution of the potentiometer and the loading error caused by a 10,000-Ω load resistor when $a = 0.5$.

Solution

The resolution is given by Equation (9.1).

$$\text{Resolution} = 100/N = 100/1000 = 0.1\%$$

The loading error is given by Equation (9.2).

$$\text{Loading error} = 100\left[\frac{a^2r(1-a)}{1+ar(1-a)}\right]$$
$$= 100\left[\frac{\quad}{\quad}\right]$$
$$=$$

Example 9.3

An incremental encoder is used with a tracking wheel to measure linear displacement as shown in Figure 9.13. The tracking wheel diameter is 5.91 cm, and the code disk has 180 holes in the outside track and 180 holes in the middle track. Determine the linear displacement per pulse and the displacement measured by each of the following total pulse counts.

a. $N_T = 700$
b. $N_T = 2220$

Solution

The linear displacement per pulse can be determined by dividing Equation (9.9) by N_T.

$$\text{Displacement per pulse} = \frac{\pi d}{N_R}$$
$$= \frac{\pi(\quad)}{\quad}$$
$$=$$
$$=$$

Total displacement = meter

a. $N_T = 700$

$x =$

b. $N_T = 2220$

$x =$

Example 9.4

An absolute encoder is to be used for measurements that require a resolution of at least 1 minute of arc. Determine the number of bits required to meet the specified resolution.

Solution

First determine the number of minutes in a full circle.

$$N =$$

Next find the smallest power of 2 that is larger than 21,600.

The encoder must have 15 bits to have a resolution of at least 1 minute of arc.

Example 9.5

A dc tachometer has the following specifications:

$$R = 0.03 \text{ m}$$
$$B = 0.2 \text{ Wb/m}^2$$
$$N = 220$$
$$L = 0.15 \text{ m}$$

Determine K_E and the output voltage at each of the following speeds:

$$S = 1000, 2500, \text{ and } 3250 \text{ rpm}$$

Solution

$$K_E = \frac{2\pi RBNL}{60}$$

$$= \frac{2\pi(\qquad\qquad)}{}$$

$$=$$

For $S = 1000$ rpm,

$$E =$$

For $S = 2500$ rpm,

$$E =$$

For $S = 3250$ rpm,

$$E =$$

Example 9.6

An incremental encoder has 2000 pulses per shaft revolution.

a. Determine the count produced by a shaft speed of 1200 rpm if the timer count interval is 5 ms.
b. Determine the speed that produced a count of 224 for a timer count interval of 5 ms.

Solution

a. Equation (9.15) applies:

$$C = \frac{\qquad}{\qquad} =$$

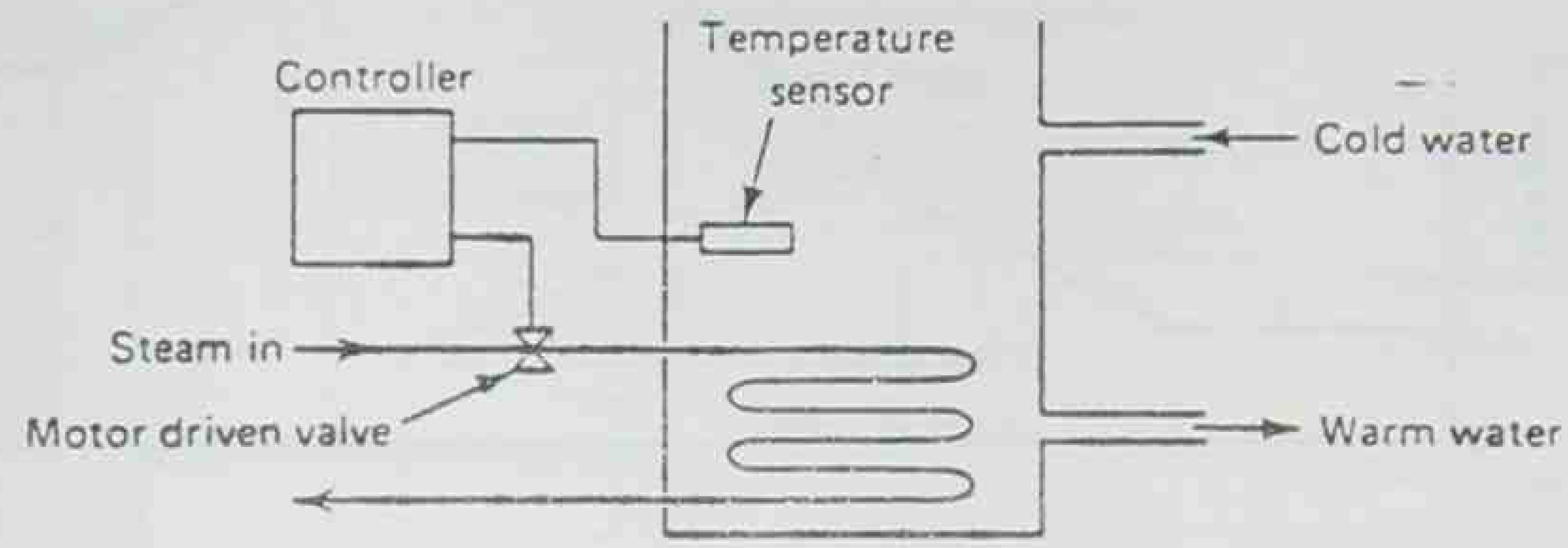
b. Equation (9.14) applies:

$$S = \frac{\qquad}{\qquad} =$$

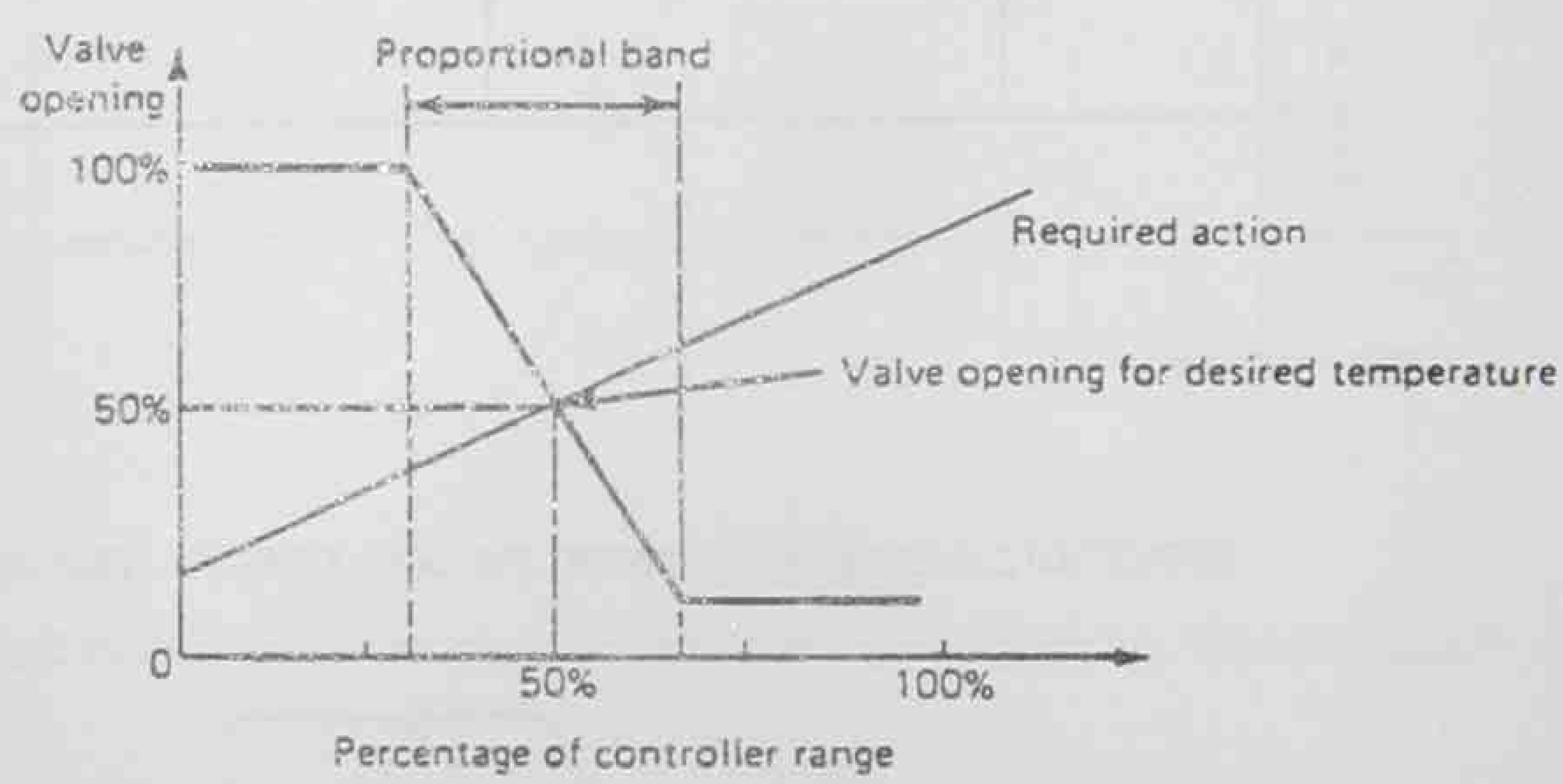
Proportional control

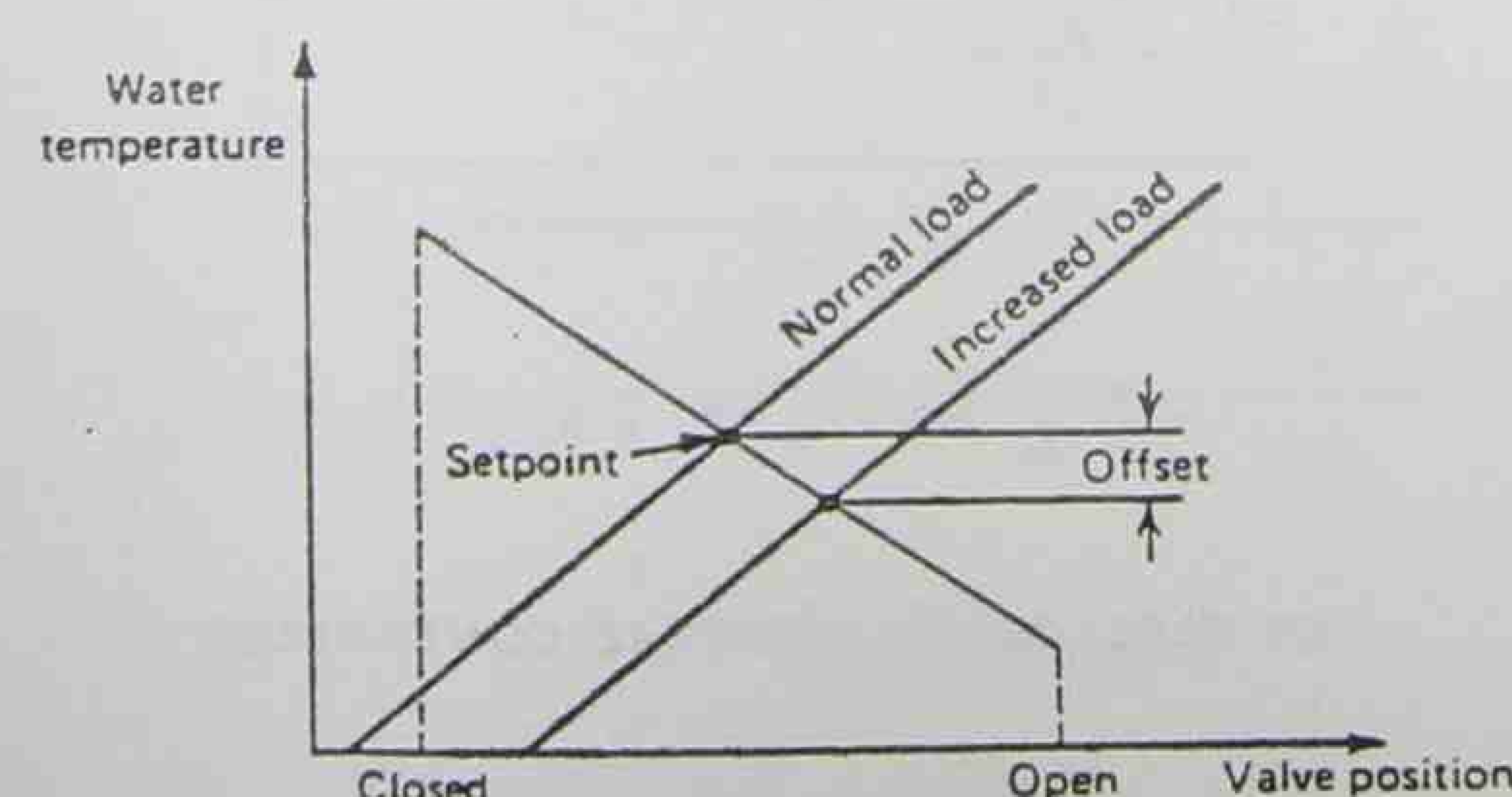
PROPORTIONAL control produces a change in the controller output that is proportional to the error signal. There is a fixed linear relationship between the controlled variable and the position of the final control element.

Proportional control is used on processes with a small capacitance and fast moving load changes where the gain can be made large enough to reduce the offset to an acceptable level. This implies a process with a capacitance which is too small to permit the use of two position or floating control.



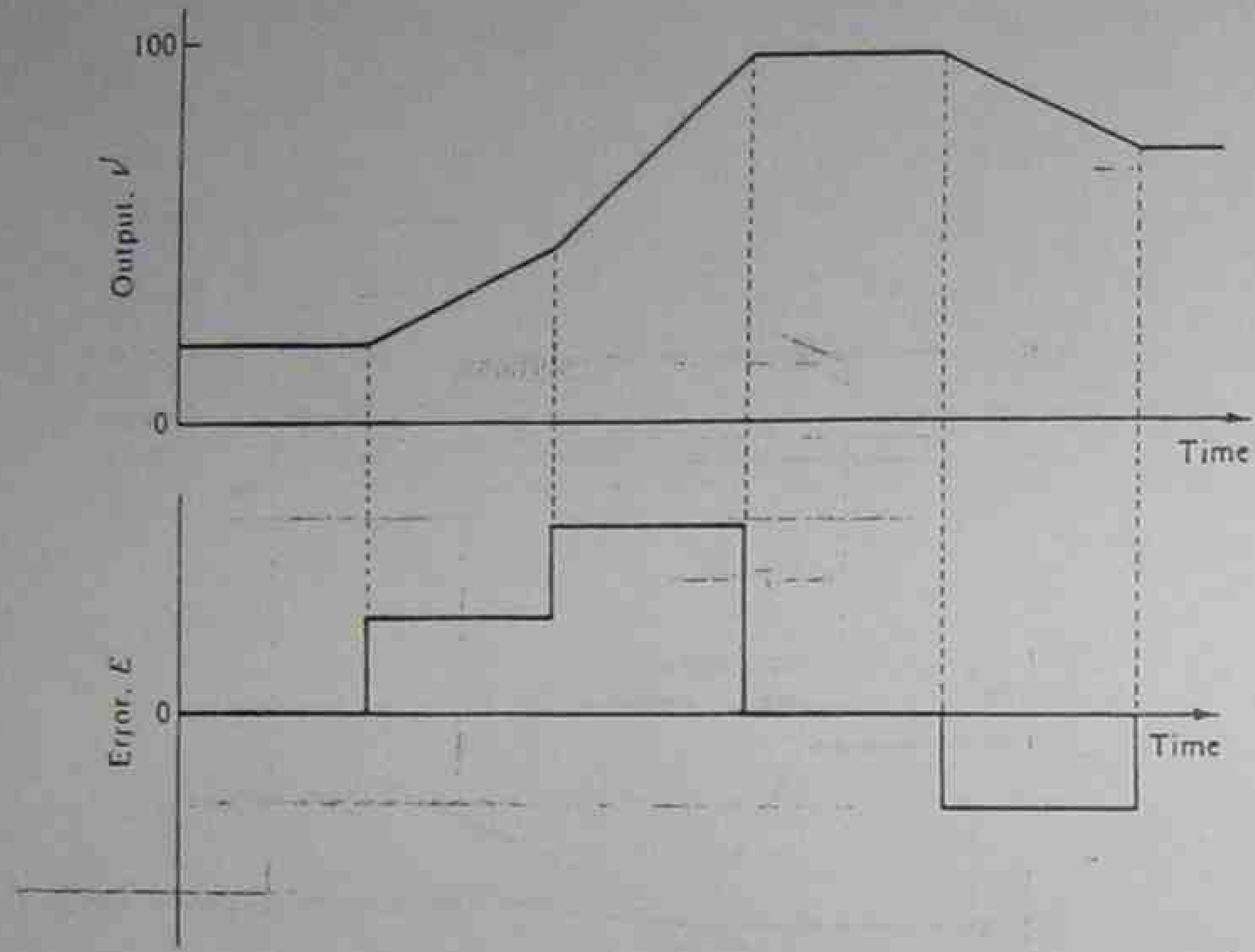
(a)



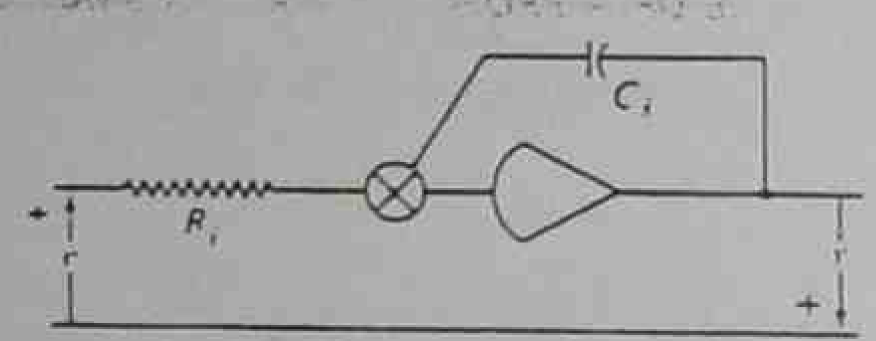


(b)

INTEGRAL control (reset control) is continuous and the output of the control element changes at a rate proportional to the magnitude and duration of the error signal. Integral control is rarely used by itself, but is usually used in conjunction with proportional control, where it eliminates offset.



INTEGRAL-MODE RESPONSE TO AN ERROR SIGNAL

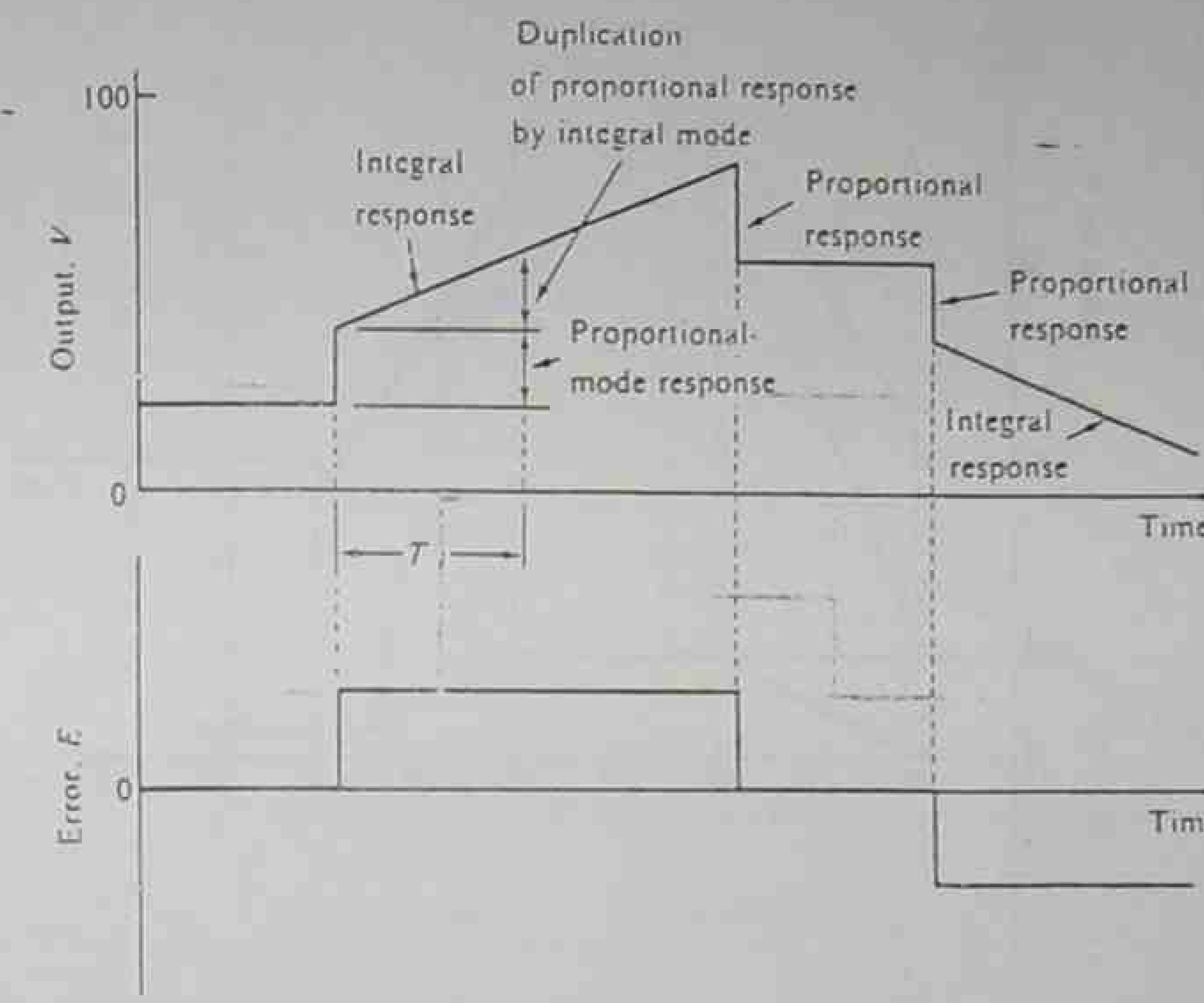


$$r = \frac{1}{T_i}\int_o^t e\,dt + r_o ;\ T_i = R_i C_i$$

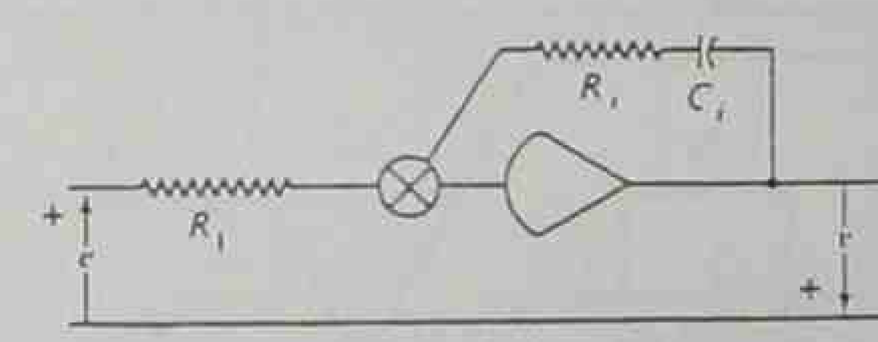
AN ELECTRONIC INTEGRAL CONTROLLER

PROPORTIONAL plus INTEGRAL (PI) mode provides an automatic reset action which eliminates the proportional offset. PI control is used on processes with large load changes when the proportional mode alone is not capable of reducing the offset to an acceptable level.

The integral mode provides the rest action which eliminates the proportional offset.



STEP RESPONSE OF A PROPORTIONAL-PLUS-INTEGRAL CONTROLLER

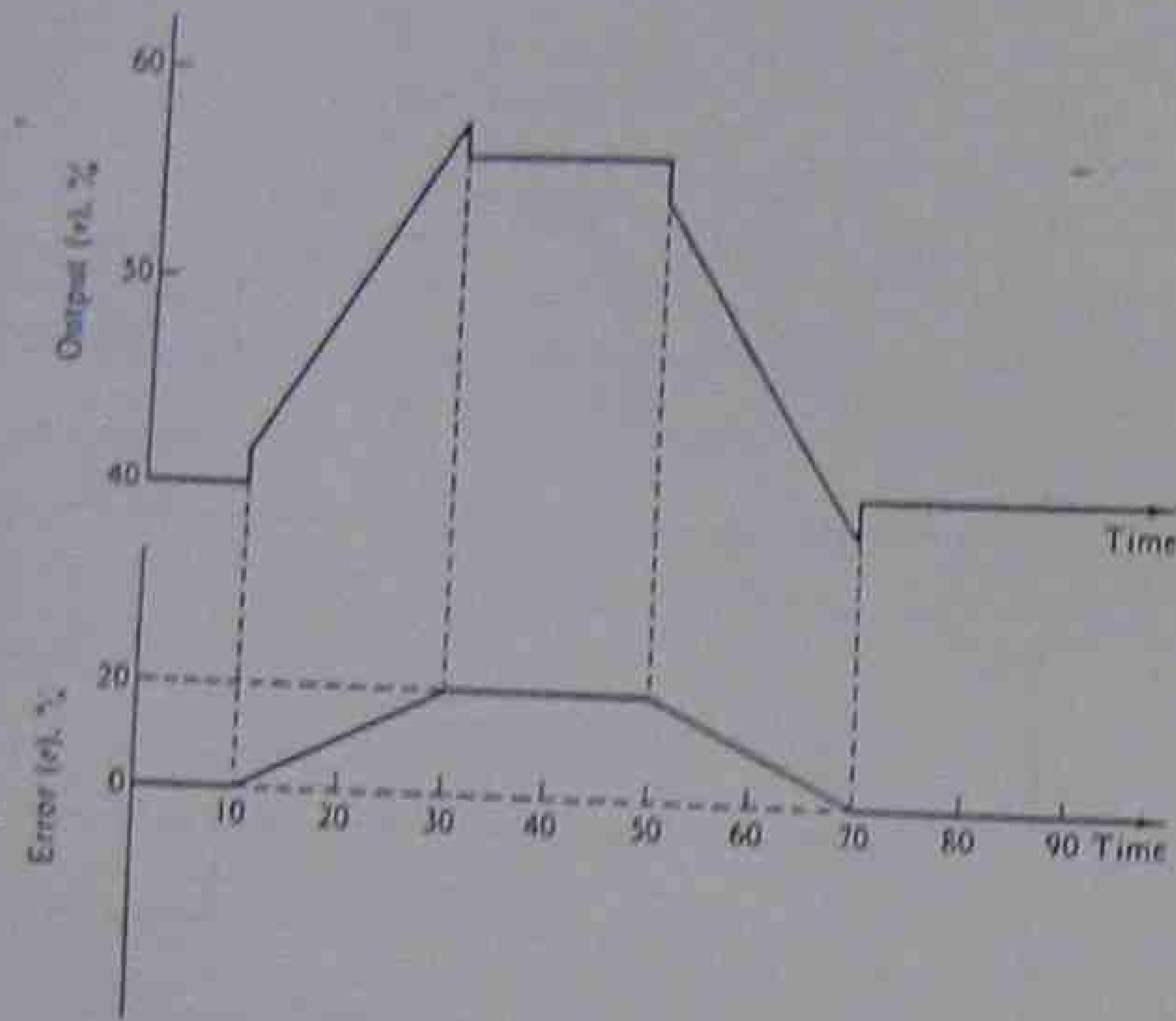


$$r = Ke + \frac{K}{T_i}\int_o^t e\,dt + r_o$$

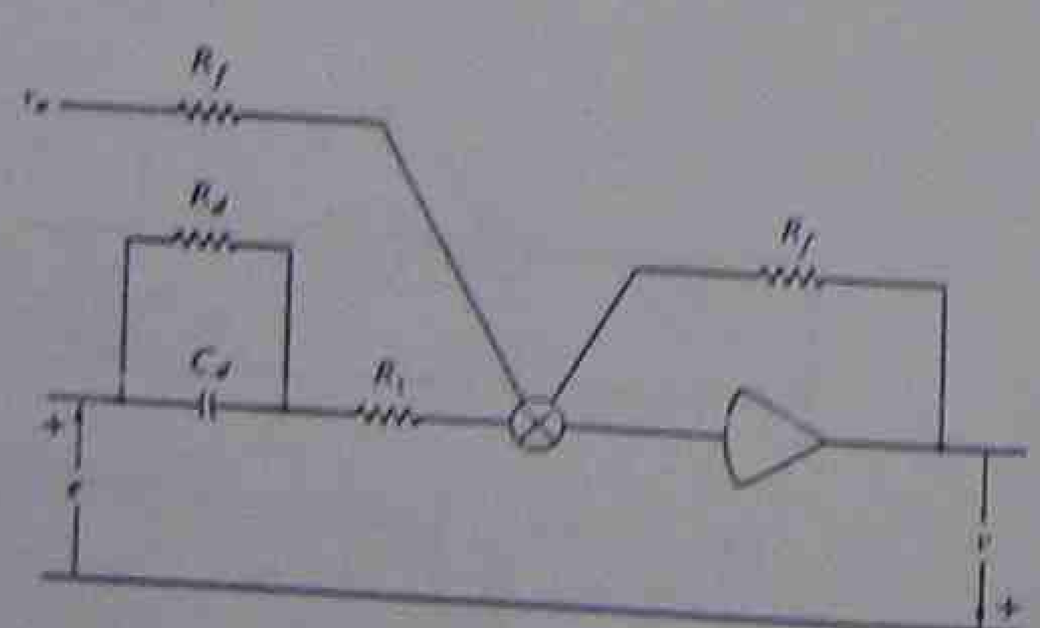
$$T_i = R_i C_i ;\ K = R_i / R_1$$

AN ELECTRONIC PROPORTIONAL-PLUS-INTEGRAL CONTROLLER

PROPORTIONAL plus DERIVATIVE (PD) control provides a change in the controller output which is proportional to the error signal and the derivative mode provides an additional change in the controller output which is proportional to the rate of change of error signal. The derivative mode anticipates the future value of the error signal and changes the controller output accordingly.



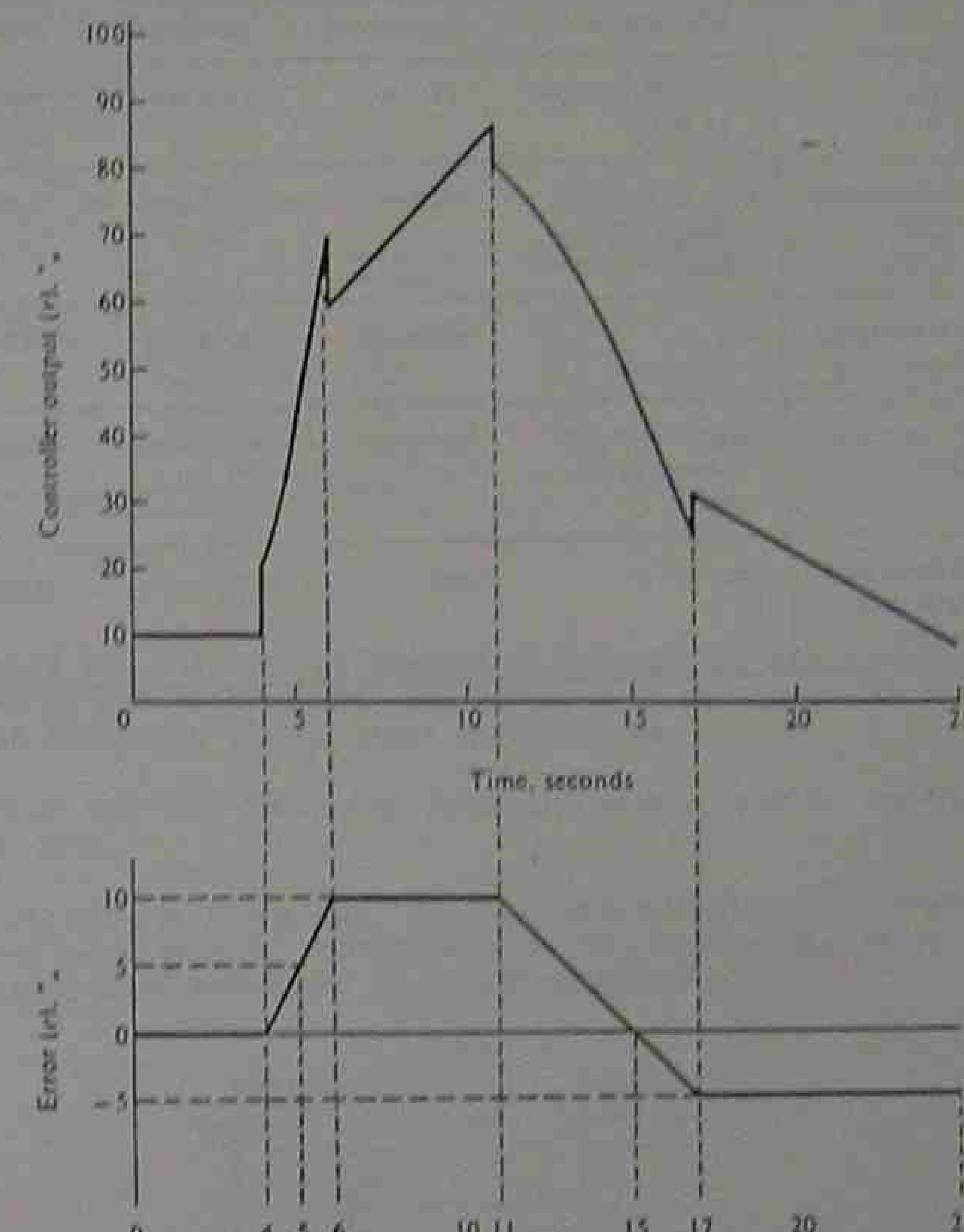
ERROR AND CONTROLLER OUTPUT GRAPHS

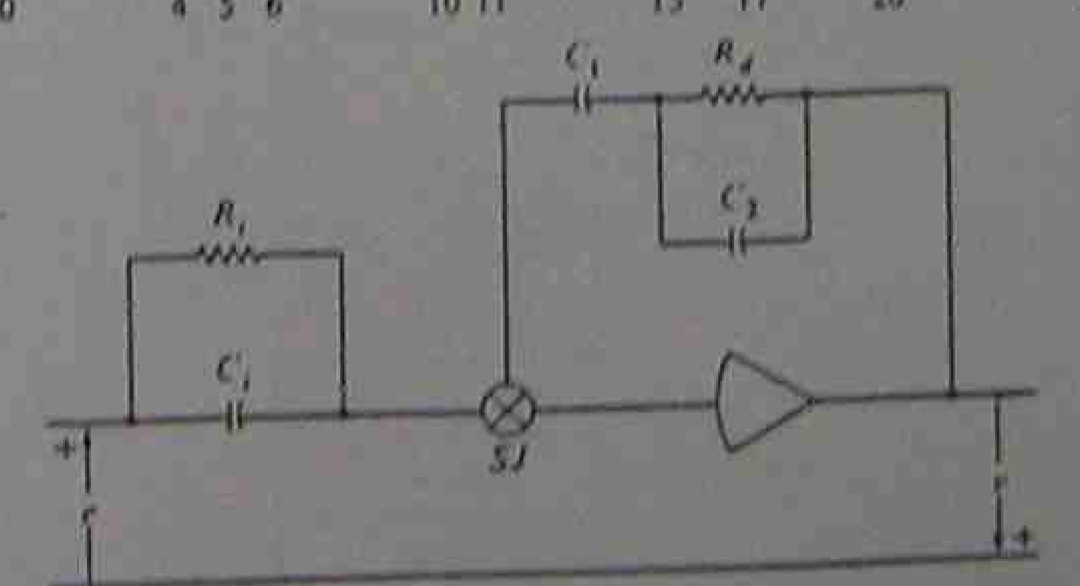


AN ELECTRONIC PROPORTIONAL-PLUS-DERIVATIVE CONTROLLER

PROPORTIONAL plus INTEGRAL plus DERIVATIVE (PID) is referred to as a three mode controller. The PID control mode is used on processes with sudden, large load changes when one or two mode control is not capable of keeping the error within acceptable limits.

The derivative mode produces an anticipatory action which reduces the maximum error produced by sudden load changes. The integral mode provides a reset action which eliminates the proportional offset.





AN ELECTRONIC THREE-MODE CONTROLLER WITH INTEGRAL INPUT AND DERIVATIVE OUTPUT

CONTROL MODE SUMMARY

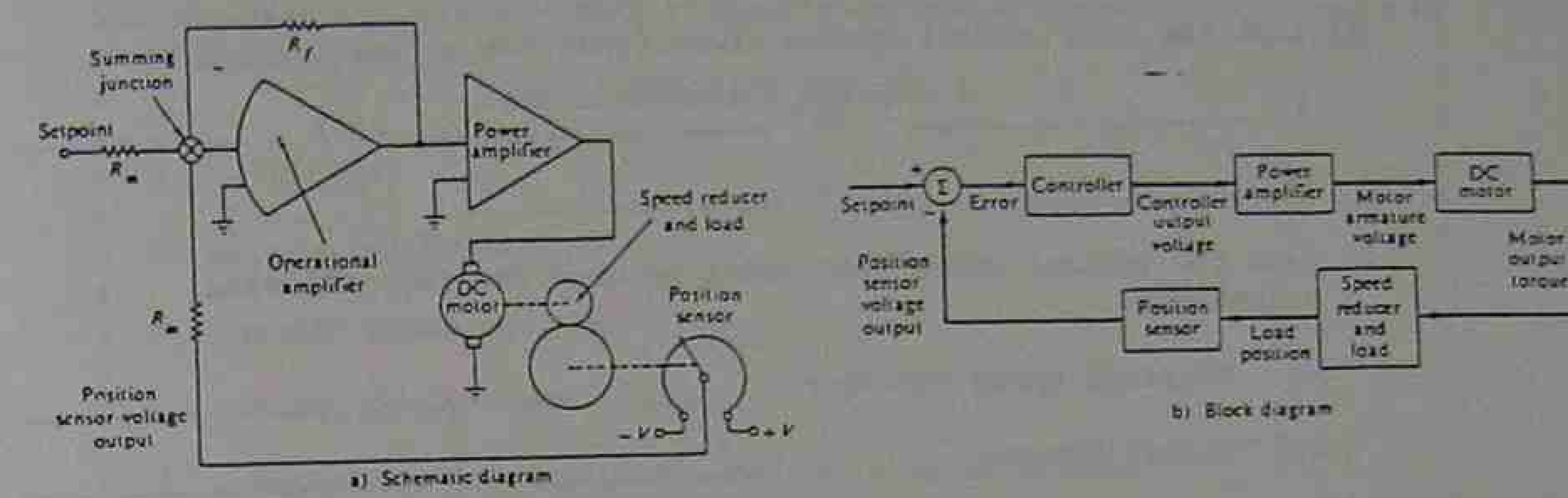
Control mode	Process reaction delay (minimum)	Transfer lag (maximum)	Dead time (maximum)	Size of load disturbance (maximum)	Speed of load disturbance (maximum)
On-Off	Long only (cannot be short)	Very short	Very short	Small	Slow
Proportional only	Long or moderate (cannot be too short)	Moderate	Moderate	Small	Slow
Proportional plus integral	Any	Moderate	Moderate	Any	Slow
Proportional plus derivative	Long or moderate (cannot be too short)	Moderate	Moderate	Small	Any
Proportional plus integral plus derivative	Any	Any	Any	Any	Any

CONTROL SIGNALS

The signals in a control system can be divided into two general categories, Analog or Digital.

ANALOG CONTROL

Analog signals vary in a continuous manner and may take on any value between some minimum and maximum limits. Analog control systems do not use digital signals within the system.



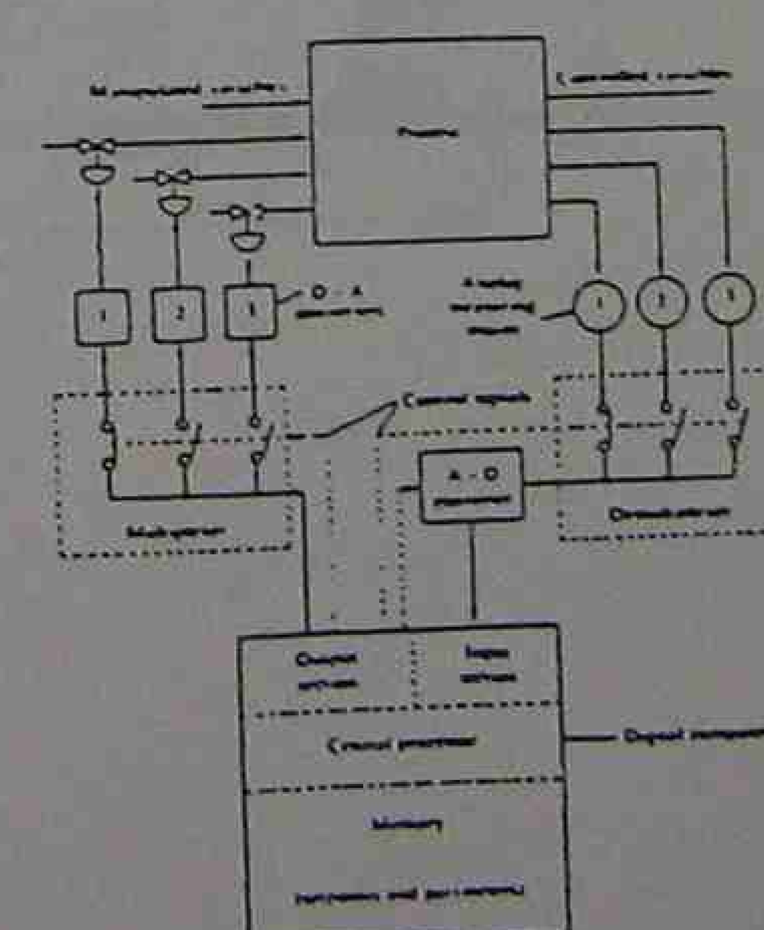
a) Schematic diagram b) Block diagram

DIGITAL CONTROL

Digital signals can only be represented in a form that has fixed values within a maximum set of possible values. The values will depend on the number of bits to be used to represent the original analog value.

Digital systems can use computers, microprocessors, pla's, rom logic or hard wired controllers.

In any digital system there must be interface circuits to convert the analog input signals to a digital representation for processing, while the digital output may need to be converted to an analog output.



ADVANCED CERTIFICATE IN APPLIED INDUSTRIAL ELECTRONICS

YEAR 2 : INDUSTRIAL CONTROL (6016K) by I Eggleton, Mar.1991

THEORY ASSIGNMENT : WEEK 2

Question 1

a) List the five control system classifications by application :-

i) ________ ii) ________ iii) ________

iv) ________ v) ________

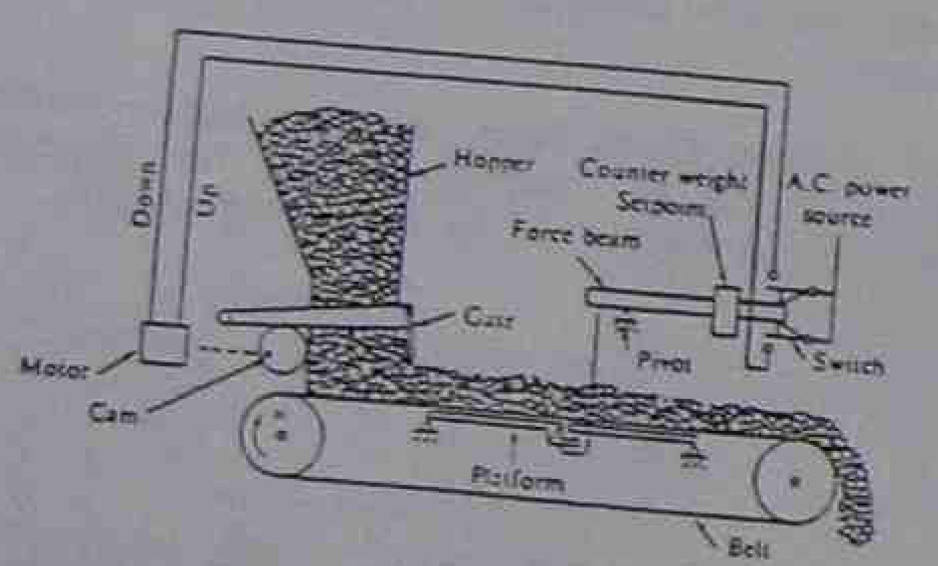
b) List two control modes that could be used for the following situations.

i) Constant speed conveyer : ________ ________

ii) Billet furnace : ________ ________

Question 2

For the process control example below, indicate the following :- measuring means, controlled variable, manipulated variable, measured variable, setpoint, final control element.



NEW SOUTH WALES DEPARTMENT of TECHNICAL & FURTHER EDUCATION
SCHOOL OF APPLIED ELECTRICITY
DIVISION OF INDUSTRIAL ELECTRONICS

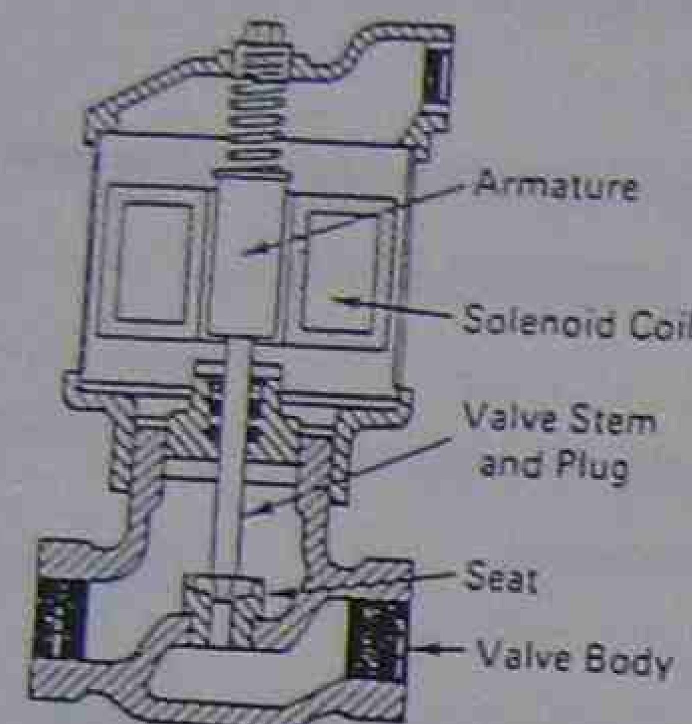
COURSE : ADVANCED CERTIFICATE of INDUSTRIAL ELECTRONICS (6016)
STAGE : 2 (ELECTIVE)
SUBJECT : *INDUSTRIAL CONTROL* (6016K)
WEEK : 3
TOPIC : FINAL CORRECTING I

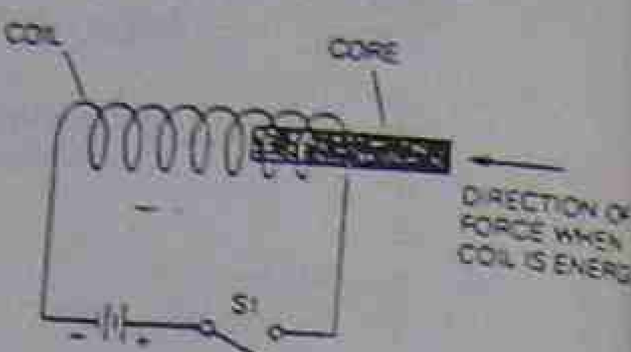
FINAL CORRECTING DEVICES I

1. Solenoids.
2. Relays and Contactors.
3. Contact phases of operation.
4. Contact protection.
5. Relay contacts in digital and counting circuits.
6. Protecting interface circuitry.
7. Types of relays.
8. Relay data sheets.
9. Bipolar switching transistors.
10. DC output modules.
11. AC output modules.
12. Power FET's.

SOLENOIDS

The most commonly used electrical actuator used in industry is the solenoid The solenoid is an electromagnetic device producing a straight line mechanica force. The motion produced can be used with levers, valves and locks, in al cases the operation of the solenoid is inherently a two position device.

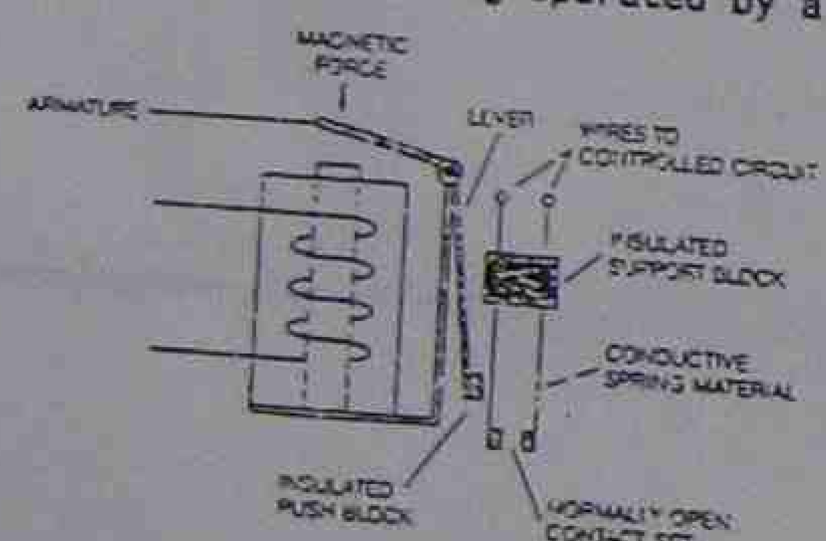




RELAYS AND CONTACTORS

When electric current is the manipulated variable in a control system, the final correcting device is often a relay or contactor.

The relay uses the mechanical motion of a lever that is connected to a set of one or more electrical contacts being operated by a solenoid.



Simplified Relay

The only difference between a relay and contactor is the current carrying and interrupting capability of the contacts. Contactors are capable of handling large currents whereas relays are used for low current operations.

DC RELAYS

The dc relay consists of an electromagnetic coil with a mechanical lever that opens or closes one or more sets of electrical contacts when a supply voltage is applied to the coil. Since the supply voltage is dc (constant) the electromagnetic field is also constant and the armature (core) of the coil will remain energised while the supply voltage is connected.

Relays and contactors provide a differential gap for on/off control because of the hysteresis effect inherent in their operation. If the armature of a relay is to energise, the coil current must reach a specified minimum value, this is known as the "pick up current".

Once the relay has energised, the coil current must be reduced below a specified maximum current "drop out current" to allow the armature to return to its normal position.

The value of current above the drop out current is known as the "hold in current".

The hysteresis effect is therefore provided by the upper value of current, (pick up current) and the lower value of current, (drop out current). These differences are due to the magnetic circuit when de-energised and the inertia of the armature.

A.C RELAYS

If a dc relay has its coil connected to an ac supply voltage the relay would continually pull in and drop out, (chatter). This happens when the ac voltage passes through zero, the magnetic field will also be zero. The iron core must be laminated and a method to prevent chattering must be adopted.

There are three methods of modifying the construction of the basic dc relay to enable its operation on an ac supply voltage.

- *Large armature mass being held in position by its own inertia.*
- *Using a second winding that is out of phase with the other.*
- *Using a shading band to generate an out of phase field.*

RELAY CONTACTS

The relay or contactor will usually have one or more sets of contacts.

The types of relay contacts available are :-

. *Change over (C.O)*
. *Multiple*
. *Normally open (N.O)*
. *Normally closed (N.C)*

In turn these contacts can be :- *Break before make* or
Make before break.

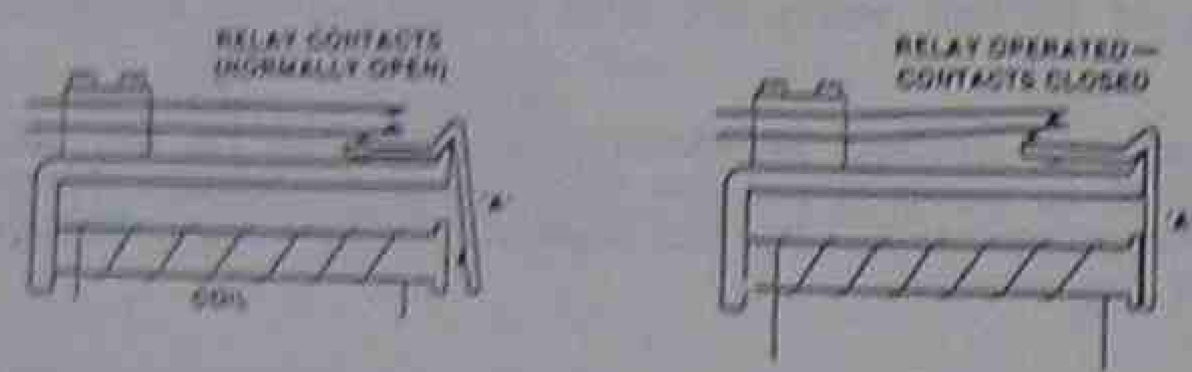


Figure 1. A relay with normally open contacts shown unoperated at left and operated at right. Note the over travel of the contact leaves. The armature is stopped by the core here.

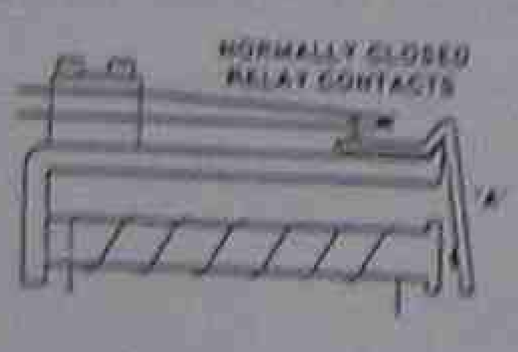


Figure 2. A relay with normally closed contacts. Note that the contact leaves are pre-loaded and the plunger on the armature operates the upper contact leaf.

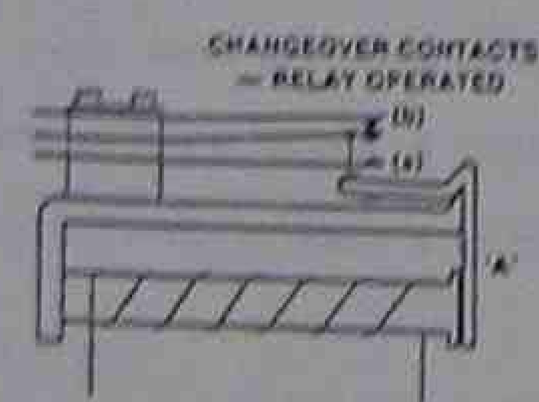


Figure 3. Relay with changeover contacts — one set of normally open contacts and one set of normally closed contacts. The armature is shown here in the operated position.

CONTACT PHASES OF OPERATION

There are three phases of operation that a relay contact must go through in the control of current flow.

. *Making the load.*
. *Carrying the load.*
. *Breaking the load.*

The conditions that exist during each of the phases will depend on the type of load to be controlled.

Resistive or Inductive, and the type of current used by the load, ie. *AC or DC.*

CONTACT MAKING AND CARRYING OF A.C RESISTIVE LOADS

When a contact makes for an inductive load eg. motor, the resulting inrush of start current can be up to 600% of the load running current.

The inrush current for a tungsten lamp, including the change of resistance due to the temperature, is about 10 times the normal current while electromagnetic loads such as relays and contactors the inrush current is up to 5 times greater.

Transformers that retain some remnant magnetism after switch off and are then switched on can draw an inrush current of 1000% for several cycles, if the switching occurs when the supply voltage is at the zero cross over point and the increasing supply voltage is of the same polarity as the remnant magnetism the resulting inrush may be as high as 3000%. In this case, zero voltage switching is not beneficial.

CONTACT BREAKING OF A.C RESISTIVE AND INDUCTIVE LOADS

When alternating current is being controlled, and arc formed by the breaking of the contacts will be extinguished when the a.c cycle passes through zero.

CONTACT BREAKING OF D.C RESISTIVE LOADS

When a dc resistive current is interrupted there will be heavy arcing that will persist for some period of time. The arcing will cause metal transfer from one contact to the other and is increased for an increase in voltage.

The contact rating for dc current is therefore significantly reduced from the normally rated ac current carrying value, it is also advisable to use suitable quenching or suppression to protect the contacts.

CONTACT BREAKING OF D.C INDUCTIVE LOADS

When a dc inductive load is broken (also relay or coil) the collapsing field will induce reverse emf that will be much greater than the supply voltage and this will cause arching across the contacts or could damage relay or coil electronic interface circuitry.

Contacts and coils must be protected using suitable quenching or suppression.

The example below indicates how the allowable release current changes for resistive and inductive loads. "NOM" is the nominal dc rating not the ac rating which is higher.

CONTACT PROTECTION FOR DC SWITCHING

The worst case switching conditions occur when an inductive load i[s] interrupted because of the high voltage being generated at the contacts.

There are three main types of contact protection :-

- *Extinguishing method*
- *Absorption method*
- *Multiple contact method*

EXTINGUISHING METHOD

This method uses a magnetic field or compressed air to stretch and break the arc, this method may incorporate arc shields or arc chutes. The contactor has the facility for the fitting of magnetic blowout coils.

ABSORPTION METHOD

The absorption method is commonly used with a resistor/capacitor combination across the load. Polyester capacitors may be used with peak voltages up to five times higher than the line voltage. Another form of absorption uses a varistor across the load or contacts.

MULTIPLE CONTACTS

In this method several contacts are connected in series, this will divide the voltage surge and reduce the over voltage for each contact.

CONTACT PROTECTION FOR AC SWITCHING

Contact life in inductive ac circuits can be greatly extended by connecting a resistor/capacitor combination across the load for low voltage (<48v) circuits and across the contacts for higher voltages.

The time constant of the resistor/capacitor combination should be approximately equal to the time constant of the load and the impedance of the load is much less than the protection impedance.

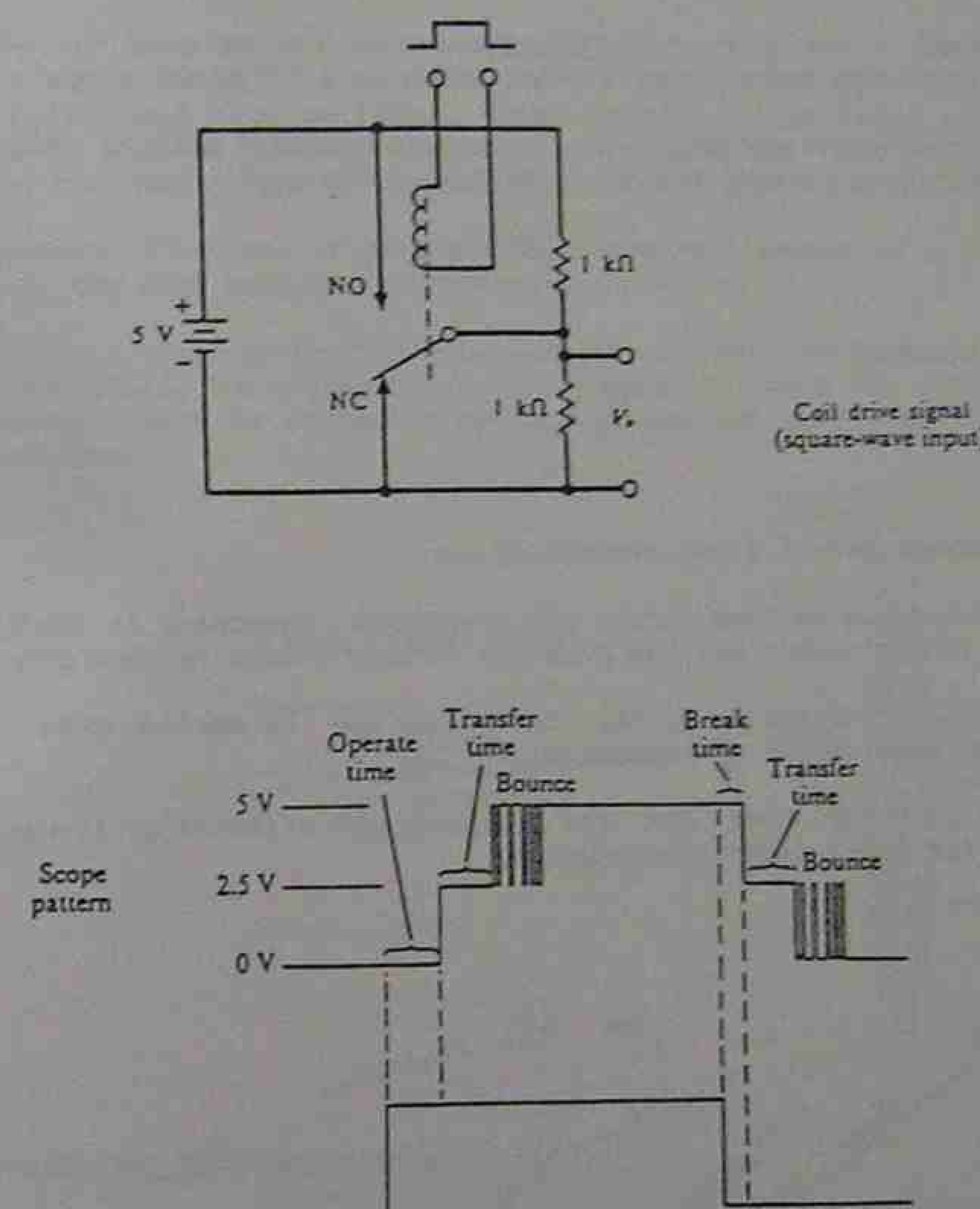
Improved protection can be obtained by using a diode with a resistor/capacitor. The diode should have a PIV of 800 volts, the capacitor should be approximately 1uf per ampere switched with a 400VW rating, the resistor should be approximately 100k per 2 Watts of load.

RELAY CONTACTS IN DIGITAL AND COUNTING CIRCUITS

When a relay contact is used to control a digital or counting circuit the output required should be a clean square wave shape.

Most relays require several milliseconds to complete the transition from one contact state to the other, this is known as "contact bounce" and can seriously distort the switched signal and severely limit switching speeds.

This effect is reduced by special relays that have "wetted" contacts, that is, consist of a small amount of mercury within the contact region.



PROTECTING INTERFACE CIRCUITRY

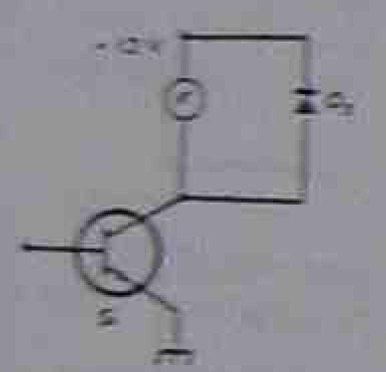
The circuitry that interfaces an inductive load such as a relay or coil to system via electronic components could be damaged by the back emf produced b the collapsing field when the inductive load is switched off.

The magnitude of the back emf depends on the load and rate of change o current at switch off, this voltage will be many times the value of the supp voltage.

For example, a 12 volt relay coil would produce a back emf of at least 30 volts, while a 24 volt telephone relay could produce a back emf up to 300 volts.

Although a resistor or resistor/capacitor can be used for di/dt protection the preferred method uses a diode known as a "flywheel diode".

When electronic circuitry is involved, protection must be added, this involve connecting a reverse biased diode across the coil.

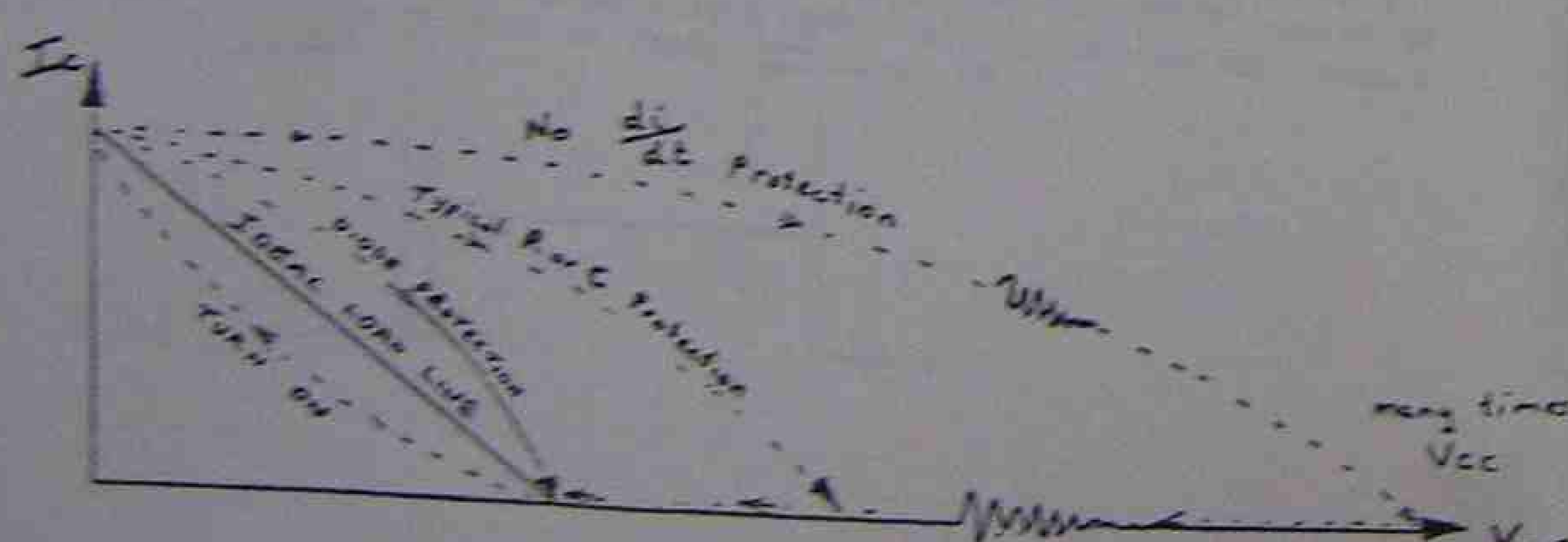


The advantages of diode protection are :-

. *The maximum voltage across the electronic component is limited to the supply voltage plus the forward biased voltage drop.*

. *The current rating of the transistor can be smaller than with other types of protection.*

. *The pick up time for the relay is not effected as it would be for other types of protection.*



TYPES OF RELAYS

There are a wide range of relays available, some of these are :-

. Differential	. Electrostatic	. Electrostrictive
. High speed	. High voltage	. Hot wire
. Impulse	. Linear expansion	. Low level
. Magnetic latching	. Magnetostrictive	. Polarised
. Mechanical latching	. Mercury plunger	. PCB
. Phase sequence	. Solid state	. Reed

A general description of these relays can be found in an article by Collyn Rivers, ETI Feb. 84

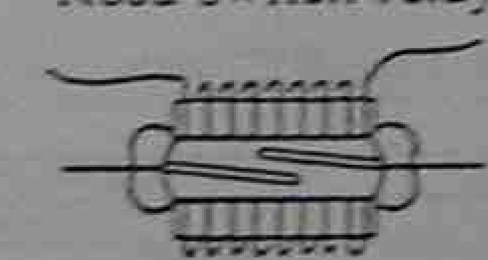
REED RELAYS

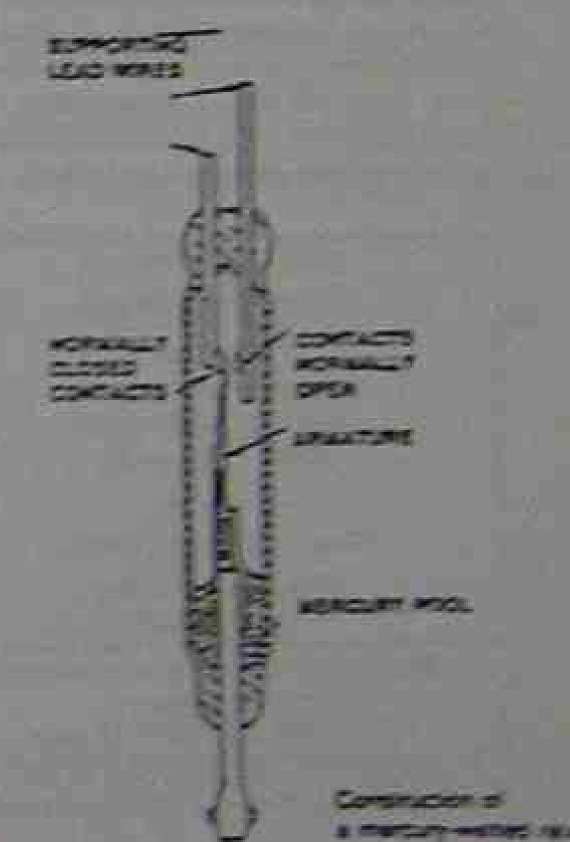
Reed relays use reed switches, these consist of at least two metal reeds sealed in a glass tube filled with inert gas. The reed relay operates by placing the reed switch in the proximity of a magnetic field.

The magnetic field may originate from a permanent magnet or a coil, if a coil is used, the reed switch is placed within the coil.

Since only a small spring force is used to separate the contacts when the coil is de-energised, the arcing on contact closure can weld the contacts together. The contacts must be protected by suppression and only used for low current applications.

Reed switch relay





Construction of a mercury-wetted relay

ADVANTAGES AND DISADVANTAGES OF RELAYS

Advantages :- . Very good electrical isolation.
. More than one output available.
. Wide range of contact configurations.

Disadvantages :- . Mechanical wear and tear on moving parts.
. Contact wear.
. May need costly maintenance.

RELAY DATA SHEETS

Information that must be determined from relay data sheets when the particul[...] characteristics are determined.

The relay characteristics include :-

. Coil type :- ac/dc, voltage, current and resistance.
. Pick up voltage. eg. % of nominal value.
. Operating / release time.
. Temperature range of use.
. Contact arrangement :- NO, NC, DP, SP.
. Contact current rating :- ac, or dc for inductive/resistive load.

TYPE H GENERAL PURPOSE RELAYS

Description — The Bulletin 700 Type H line of general purpose relays includes the Type HA tube base relay, Type HB square base-relay, Type HC miniature-square base relay, Type HT single range timing relay and Type HTM-multi-range-timing relay. A variety of options and accessories are available for each of these relays.

Type HA and HB relays — Type HA and HB relays are rated at 10 amperes and available with AC or DC coils. Type HA relays are available in 2 or 3 pole versions with tube base plug-in terminations. Type HB relays are available in 2 or 3 pole versions with square base, plug-in quick-connect solder terminations.

SPECIFICATIONS
(Types HA and HB)

CONTACTS
Arrangement:
DPDT (2 Form C) or
3PDT (3 Form C)

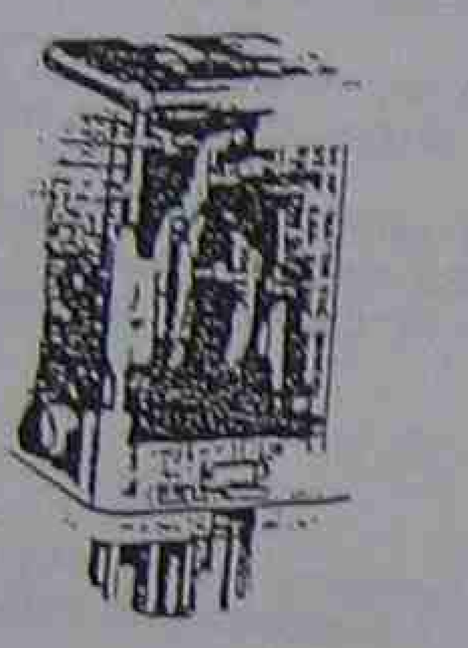
Material:
Silver Cadmium Oxide

Pilot Duty Rating:
NEMA B300

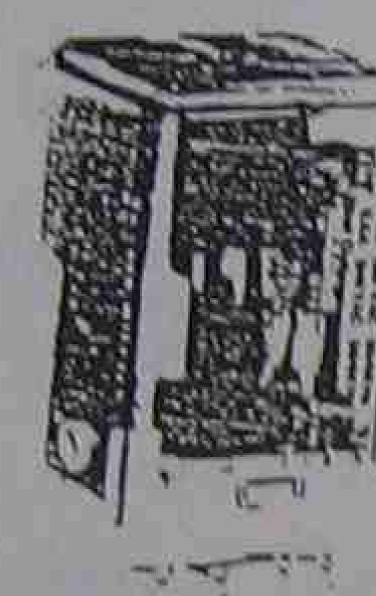
Dielectric Withstand Voltage:
(tested for 1 minute)
1500 VAC rms pole to pole
1500 VAC rms contact to coil
1500 VAC rms contact to frame

COILS
Duty Cycle:
continuous

Pickup Voltages At
Operating Temperature:
AC — 85% of nominal voltage at (60 Hz)
AC — 80% of nominal voltage at (50 Hz)
DC — 80% of nominal voltage



Type HA
Tube Base Relay



Type HB
Square Base Relay

Contact Ratings For Types HA and HB

Volts	AC Amperes					DC Amperes	
	Inductive			Resistive			
	Make	Break	Continuous	Make, Break & Continuous	HP	Volts	Make, Break & Continuous
120	30	3	10	10	1/3	24 VDC	10
240	15	1.5	10□	10□	1/2		

□ 3 pole devices have a 20 ampere maximum total current rating for all three poles

Coil Data For Types HA and HB

DC Coils				AC Coils (50/60 Hz)				
Nominal Voltage (Volts)	Nominal Resistance (Ohms)	Nominal Current (mA)	Nominal Power (Watts)	Nominal Voltage (Volts)	Nominal Resistance (Ohms)	Nominal Current (mA) 50 Hz	Nominal Current (mA) 60 Hz	Nominal Sealed VA*
6	17.1	351	2.0	6	4.26	500	417	3.0 VA at 50 Hz
12	73.7	163		12	17.1	250	208	
24	322	76		24	73.7	125	104	
48	1140	42		120	2,030	25	21	2.5 VA at 60 Hz
110	5870	18		240	8,140	13	10	

* Approximate inrush VA equals 1.5 times sealed VA

BIPOLAR SWITCHING TRANSISTORS

The ideal switch has the following characteristics :-

. *Infinite off state resistance.*
. *Zero on state resistance.*
. *Zero switching time (on - off) & (off - on).*
. *Low actuation level.*

TRANSISTOR CHARACTERISTICS

BV_{ebo} :- The reverse breakdown voltage between the base and emitter with the collector open circuit. This can be increased by the addition of a forward biased series diode in the base emitter circuit.

CURRENT RATING :- The continuous dc current and the peak current drawn by the load must determined.

POWER REQUIREMENTS :- The power dissipation must be determined for the on-off dc and transient conditions.

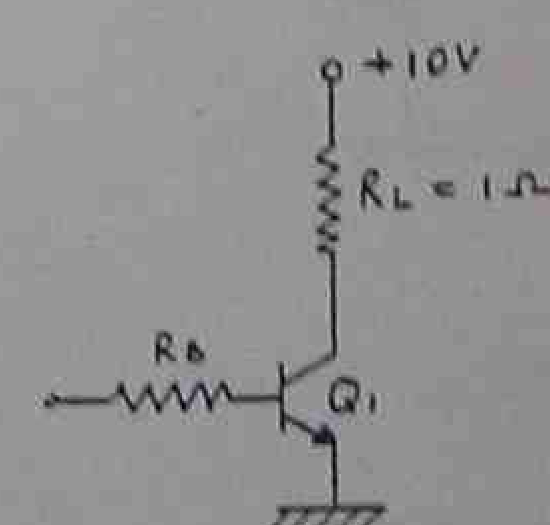
The *ON - OFF* conditions have a power dissipation of :-

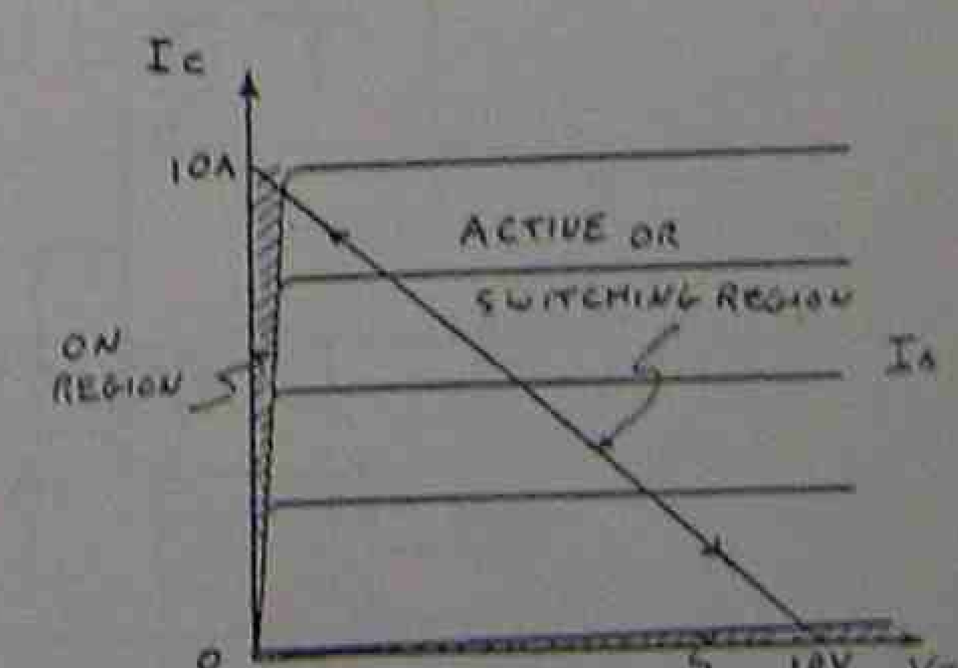
. *POWER OFF = VCC x I_{ce}* . *POWER ON = VCE sat x Ic max*

The on-off (dc) power dissipation is directly related to the biasing of the transistor, that is, the transistor must be driven into saturation for the on state to reduce Vce (sat) and completely cut off when in the on state.

The "transient" power is dissipated when the transistor is either switching from (on-off) or (off-on) and is directly related to the switching time of the transistor.

The maximum transient power dissipation occurs at the mid point of the load line.



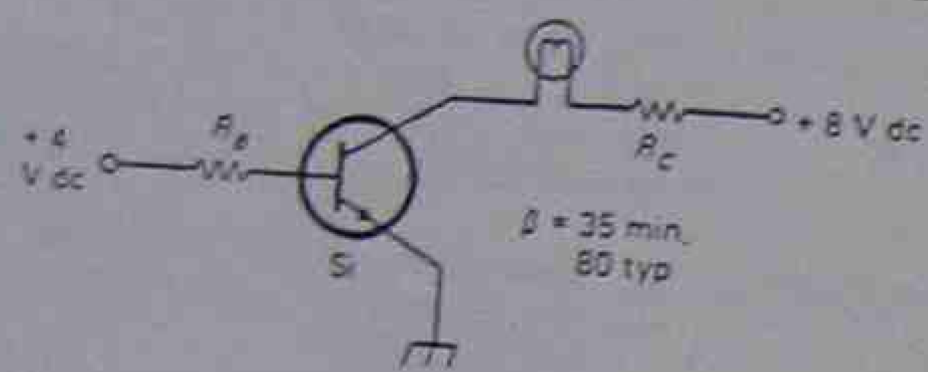


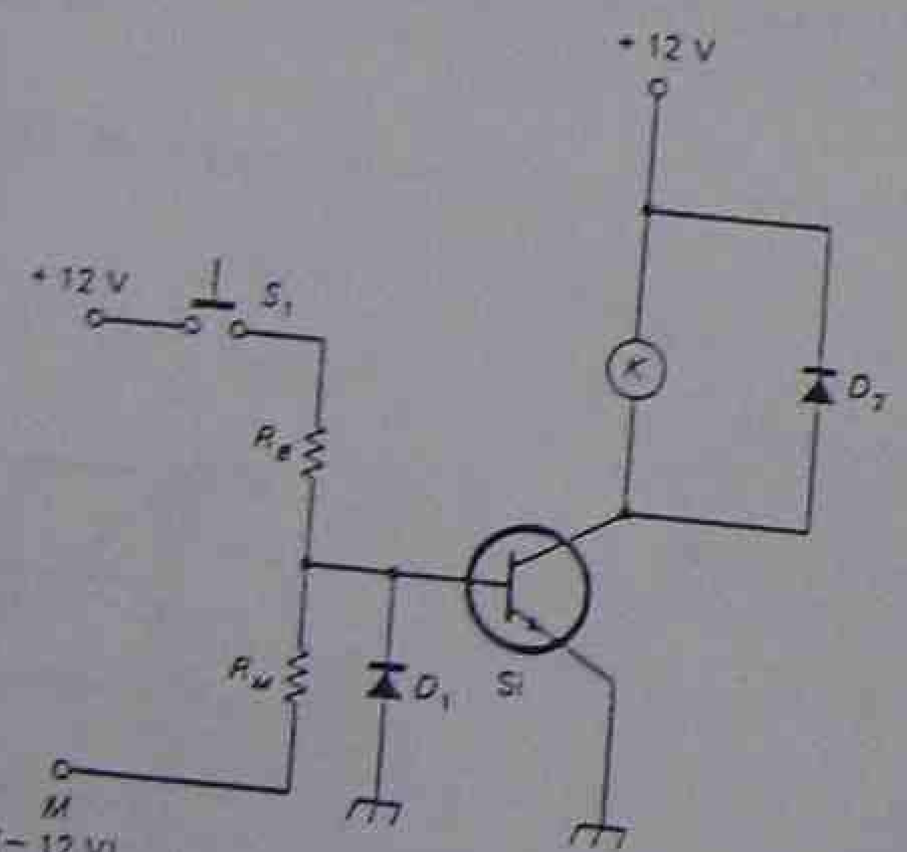
DESIGN OF TRANSISTOR DRIVE CIRCUITS

The transistor needs to be overdriven to minimise Vce, this is usual achieved by setting the "on" base current to three times (3x) the minimum ba current necessary to start saturation.

If the available base current is insufficient to drive the transistor in saturation a darlington configuration must be used as a current amplifier.

If Germanium transistors are used, the "off" power dissipation would be hig this is overcome by driving the base voltage slightly more negative than t emitter.



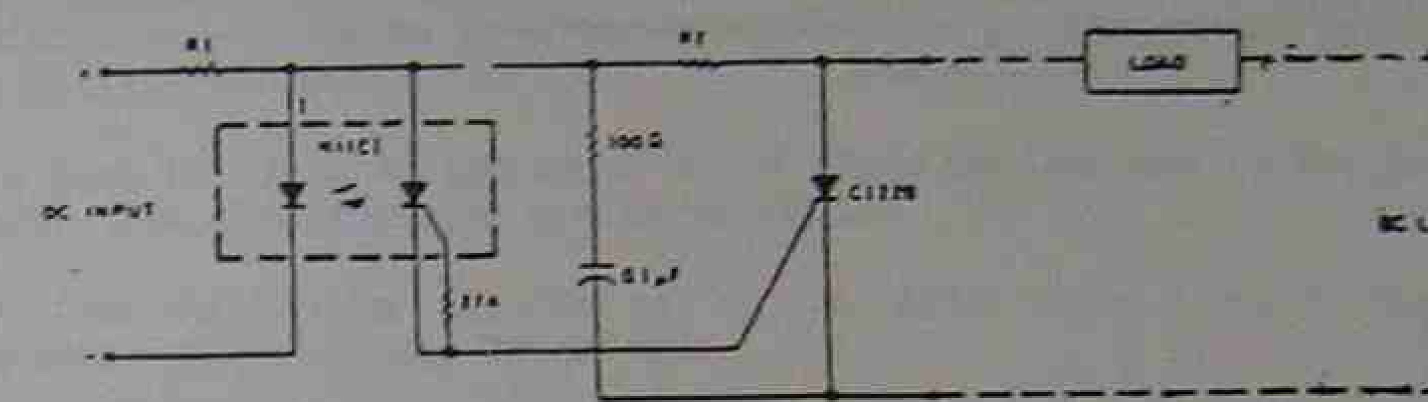


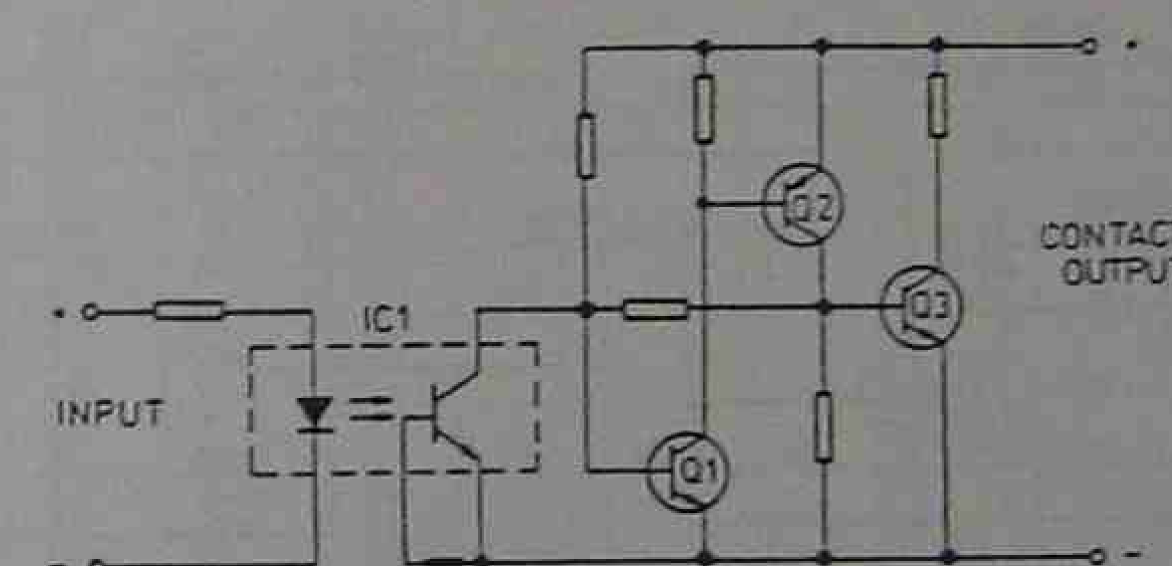
DC OUTPUT MODULES

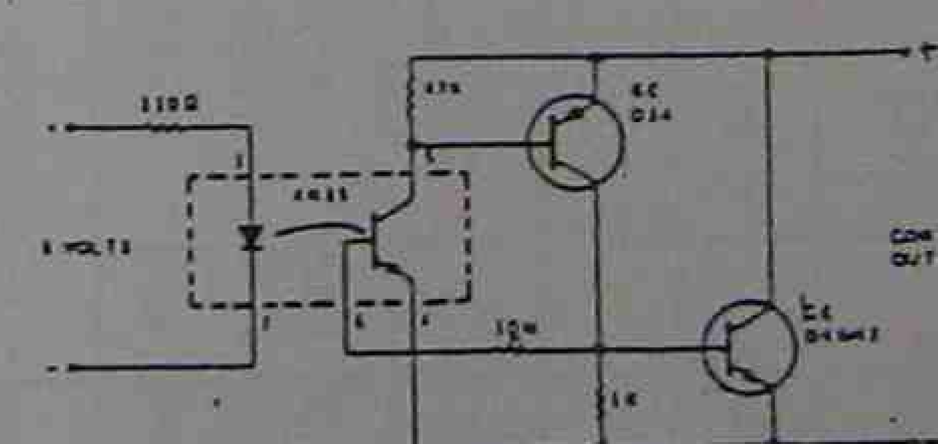
Modules using discrete circuitry can be commercially obtained or constructed for a specific control system.

Examples of discrete dc output modules are shown below :-

. DC output module to be used with dc a commutating motor.
. Normally closed dc output module.
. Normally open dc output module.



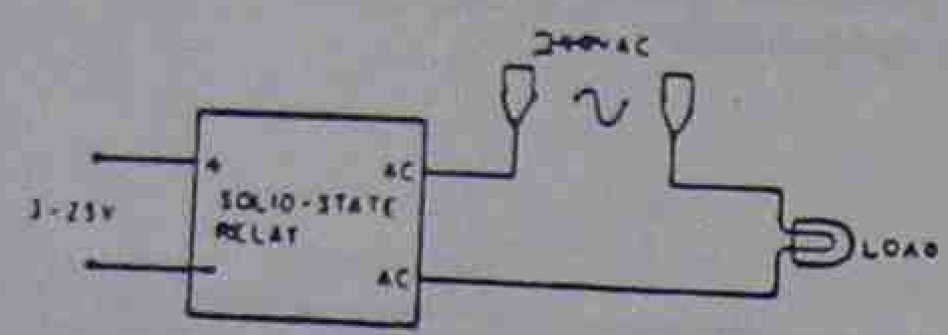


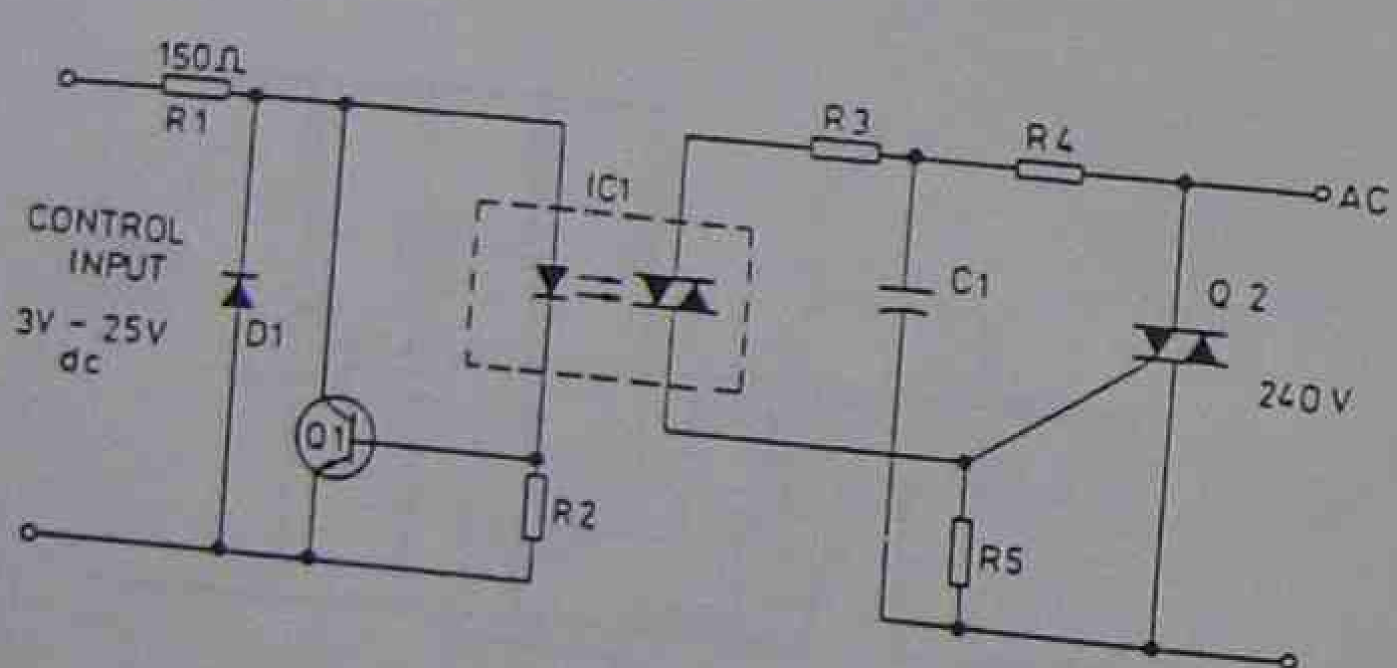


AC OUTPUT MODULES

Modules using discrete circuitry can be commercially obtained or constructed for a specific control system.

An example of a discrete component solid state relay for ac loads is shown below.





POWER FET's

ADVANTAGES OF USING FET's

. High input impedance :- allows direct interfacing to CMOS and TTL.
. Nanosecond switching times.
. No thermal runaway.
. Low on-state voltage.
. Simple to use.

DISADVANTAGES

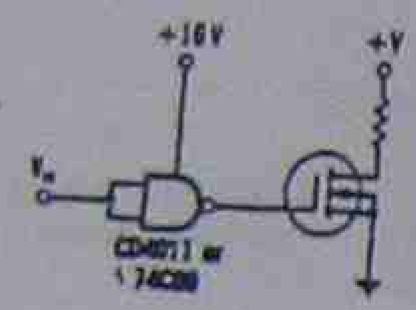
When compared to bipolar transistors the on state voltage drop of these mosfet's can be up to 2 volts and exhibit a greater power loss when used in low voltage circuits.

Care must be taken in the handling of these devices due to the possibility of damage from electrostatic discharge (ESD).

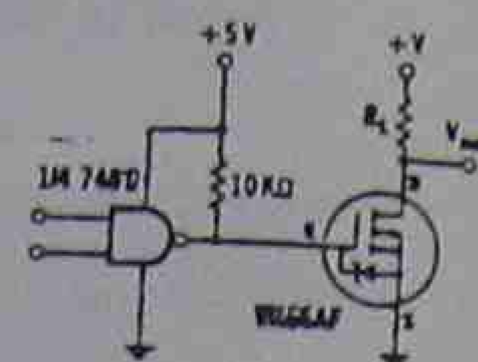
For the following sample of a MOS data sheet, note the following characteristics :- Vds, Id and Rds.

V_{DS} / V	$R_{DS(on)}$ / Ohm	TO 220 BUZ..	I_D/A	TO 218 BUZ..	I_D/A	TO 3 BUZ..	I_D/A	TO 238 BUZ..	I_D/A
50	0.10 0.12 0.10 0.04 0.06 0.03	71 71A 71L 11 11A	13.0 13.0 12.0 30.0 25.0	348 347	39.0 42.0	14 15	39.0 45.0	17 18	32.0 37.0
60	0.04	1152	30.0						
100	0.2 0.25 0.20 0.085 0.6	72 72A 20 21	10.0 9.0 12.0 21.0	349	32.0	23 25 24	10.0 21.0 32.0	27	26.0
200	0.40 0.60 0.20 0.40 0.12	73 73A 31 32	7.0 5.8 12.5 9.5	350	22.0	34 35 36	14.0 9.9 22.0	37 38	13.0 18.0
400	1.80 2.50 1.00 1.50 0.40 0.50	76 76A 60 60B	3.0 2.6 5.5 4.5	351 326	11.5 9.5	63 64	5.9 11.5	67	9.6
500	3.00 4.00 1.50 2.00 0.50 0.60 0.60 0.80 0.80	74 74A 41A 42	2.4 2.0 4.5 4.0	330 331	9.5 8.0	45B 45 45A	10.0 9.6 8.3	48 48A	7.5 6.8
600	5.0 2.0 2.5 0.9	77 90 90A	1.9 4.0 3.5			94	7.8		
800	8.0 3.0 4.0 2.0 2.0 1.5	78 80A 80	1.5 3.0 2.6	307 308 356 355	3.0 2.6 5.3 6.0	83A 83 84A 84	3.4 2.9 6.5 5.7	88A 88	5.0 4.3
1000	5.0 6.0 8.0 2.0 2.6	50A 50C 50B	2.5 2.3 2.0	310 311 357 358	2.5 2.3 5.0 4.5	53A 54 54A	2.6 5.1 4.5	58 55A	5.8 4.9

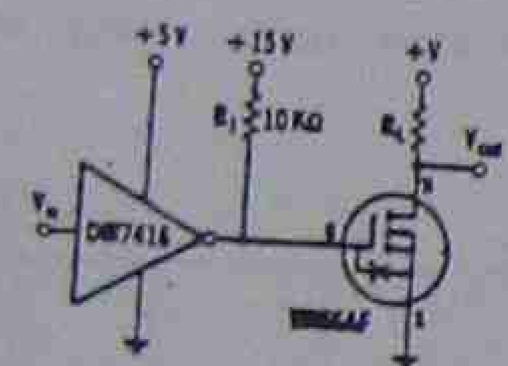
Mosfet's prove extremely versatile in their applications and since th inputs are voltage driven, with virtually zero current, their use with digi circuits are ideal.



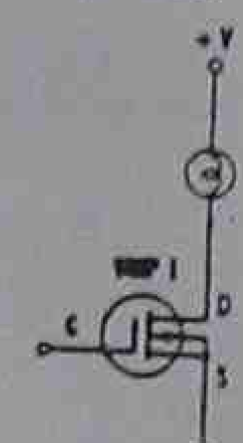
Driving VMOS from CMOS



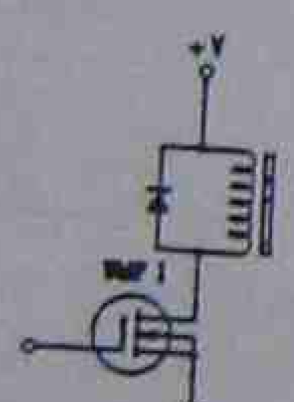
Driving VMOS from TTL



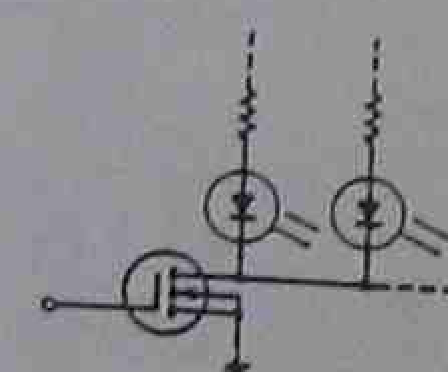
Driving VMOS from o/c TTL



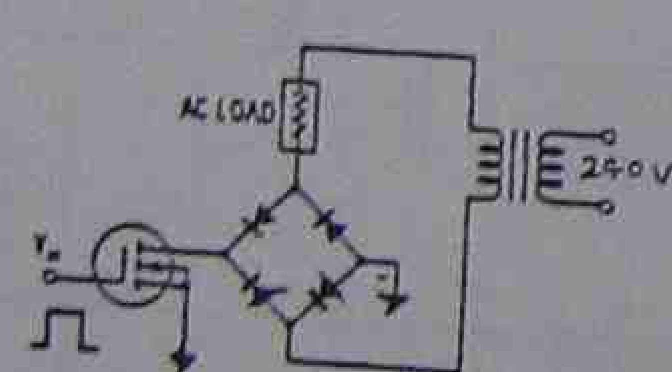
Lamp driver



Relay driver



Led driver



A.C load control using VMOS

ADVANCED CERTIFICATE IN APPLIED INDUSTRIAL ELECTRONICS

YEAR 2 : INDUSTRIAL CONTROL (6016K) by I Eggleton, Mar.1991

THEORY ASSIGNMENT : WEEK 3

Question 1

Sketch a circuit using an npn transistor to interface a 12 volt, 300 Ohm dc relay to a circuit having an output ranging from 0 to 2 volts.

Question 2

For the AC output module below describe the function of :-

. D1 : ____________________

. Q1 & R2 : ____________________

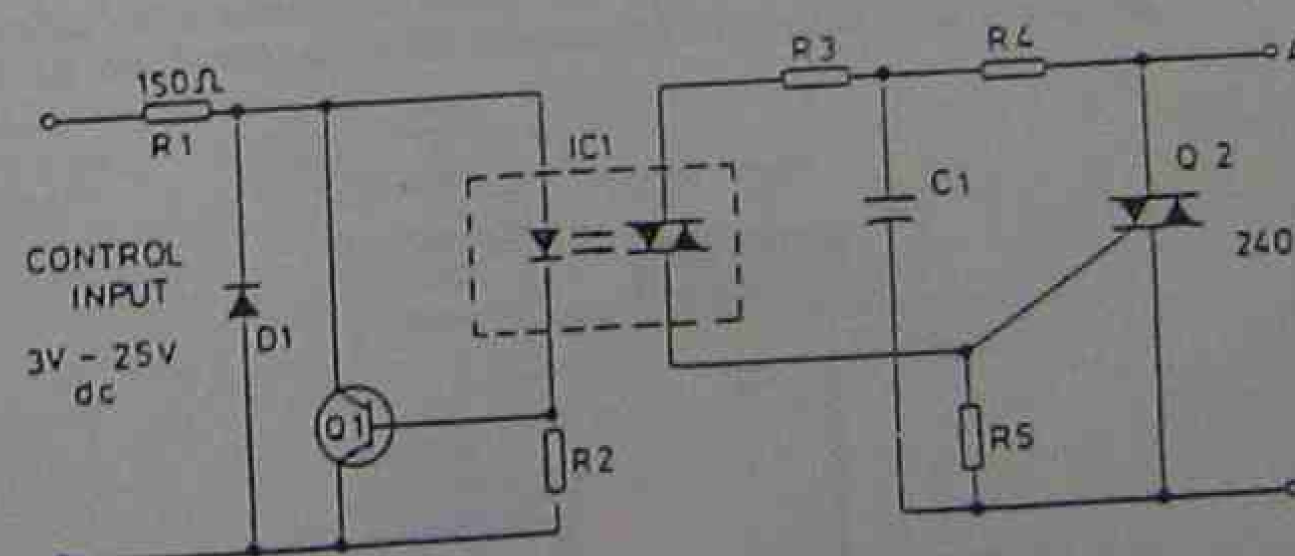
. IC1 : ____________________

. R3 : ____________________

. Q2 : ____________________

. R4 & C1 : ____________________

. R1 : ____________________



SCHOOL OF APPLIED ELECTRICITY
DIVISION OF INDUSTRIAL ELECTRONICS

COURSE : ADVANCED CERTIFICATE of INDUSTRIAL ELECTRONICS (6016)
STAGE : 2 (ELECTIVE)
SUBJECT : *INDUSTRIAL CONTROL* (6016K)
WEEK : 4
TOPIC : FINAL CORRECTING DEVICES II

FINAL CORRECTING DEVICES II

1. Valve characteristics.
2. Single phase motors.
3. Three phase motors.
4. DC machines.
5. AC servo motors.
6. DC servo motors.
7. Stepping motors.
8. Solid state relays.
9. SSR specifications.
10. SSR advantages & disadvantages.

MOTOR DRIVEN VALVES

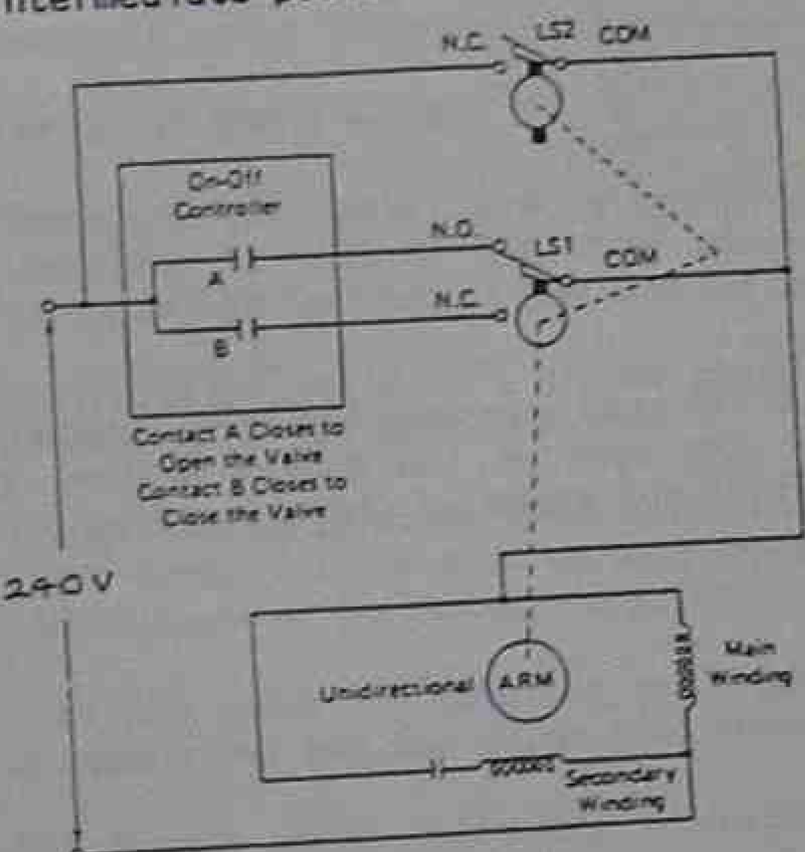
When a large valve is to control high fluid pressure, a motor must be connected to a mechanical linkage to operate the opening and closing.

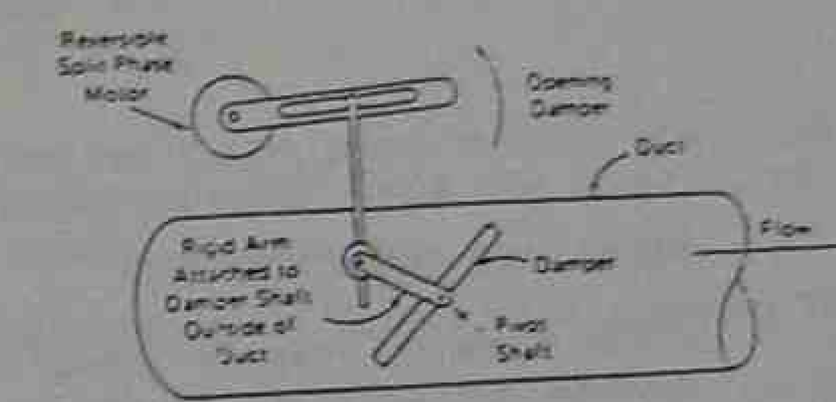
Most two position valves of this type are operated by a unidirectional split phase motor. The motor is geared down to provide a slow moving output shaft speed with high torque.

The out shaft rotates from 0° to 180° for the linkage to open the vale, the shaft must be rotated from 180° to 360° for the vale to be closed.

The travel time for these motors are up to 4 minutes and since the valve, moves slowly this type of control falls between on-off and proportional control, ie. floating control.

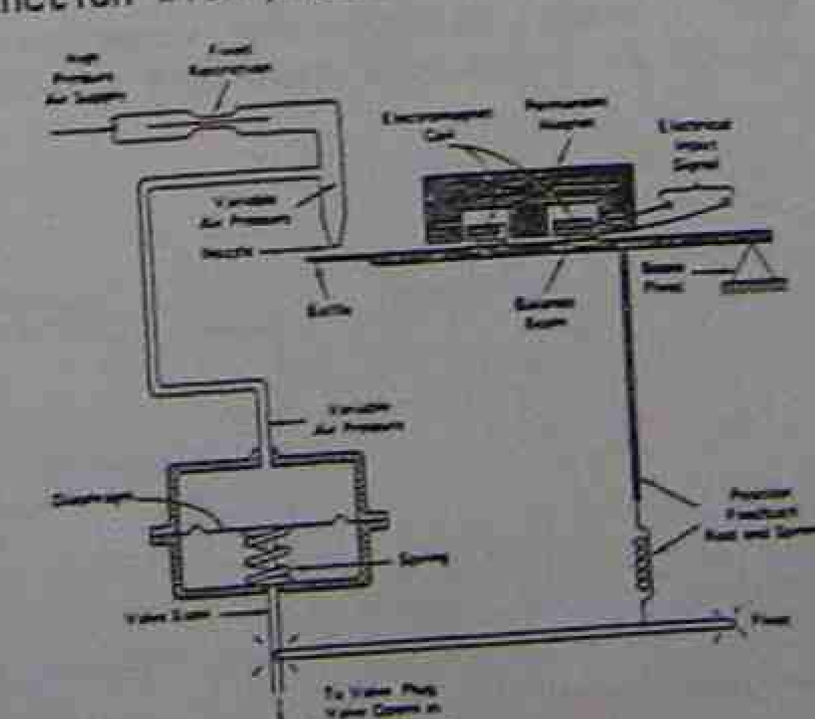
For proportional control the positioning of the control vale will be in an intermediate position.





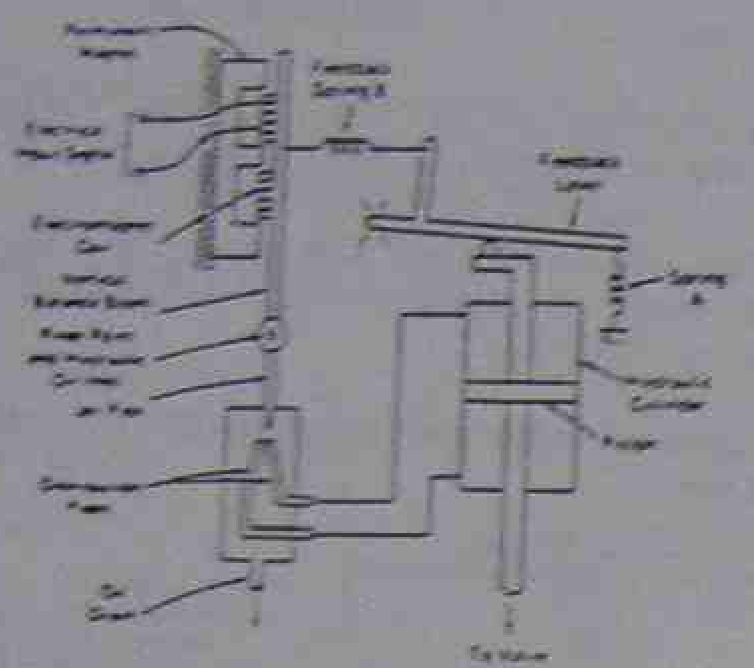
ELECTROPNEUMATIC VALVES

For large vales the electric motor is not practical therefore the valve must be operated in conjunction with pneumatic or hydraulic pressure.



ELECTROHYDRAULIC VALVES

The electrohydraulic system is used when the valve is massive, difficult to hold in position or may become stuck in any position.



VALVE FLOW CHARACTERISTICS

Fluid flow is exactly linear but in real systems the flow characteristic depends not only on the valve but the rest of the piping system. Designers overcome this problem by using varying the shape of the valve plug.

SINGLE PHASE MOTORS

INTRODUCTION

There are two basic forms of construction for single phase motors, one almost identical to that of the three phase induction motor while the other is of a form similar to that of the d.c series motor.

The rotating field in a single phase induction motor occurs by simulating the effects of a two phase motor.

The single phase motor must have two currents at an appropriate phase angle to each other. This can be achieved by having two windings of different inductances or by adding a capacitor in series with one of the windings.

Once the motor is rotating at a suitable speed,one of the windings can be disconnected and the motor will continue to rotate.
The single phase motor has a vibration of twice the line frequency and this makes it noisy in operation.

THE SPLIT PHASE MOTOR

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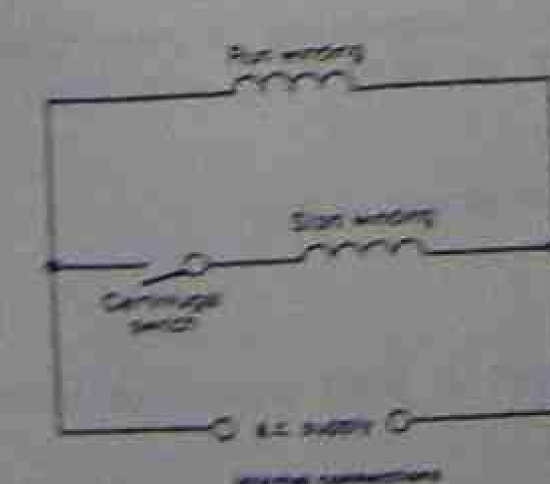
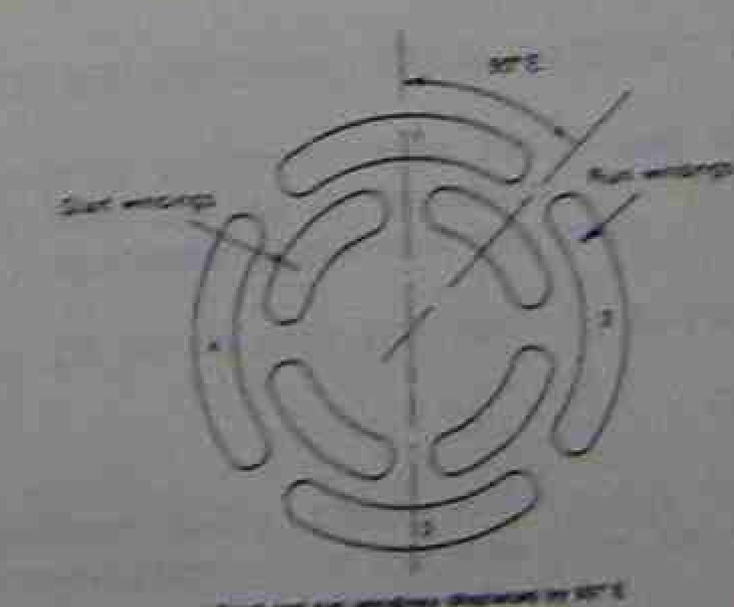
THE SPLIT PHASE MOTOR

The standard split phase motor has windings (start / run) connected across the supply during the start, but when the motor is up to speed only the run winding is used.

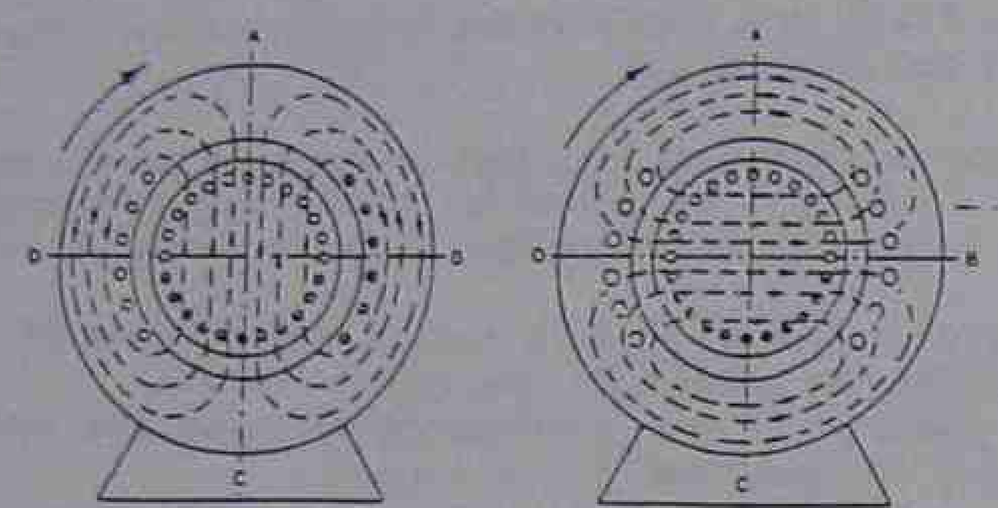




The startor flux rotates at a speed determined by the supply frequency and the number of poles in a winding. The direction of rotation is in the same direction as that of the magnetic field.

$$n = \frac{120 \times f}{p}$$

n = speed (rpm)
f = supply frequency
p = number of poles



USES

Split phase motors have only moderate starting torque so they are limited to such typical uses as washing machines, blowers,buffing machines,grinders and machine tools.

THE CAPACITOR START MOTOR

To increase the phase angle and produce improved characteristics a capacitor is connected in series with the start winding.

Reversal of rotation is achieved by reversing connections of either (but not both) winding. This is the same method as for the split phase motor.

USES

General purpose heavy duty applications requiring high rotor torque, such as starting refrigerators and air compressors.

THE CAPACITOR START CAPACITOR RUN MOTOR

This motor has both windings permanently connected across the supply, with a capacitor connected in series with each of them. The windings are referred to as the main and auxiliary windings.

Reversal of rotation is similar to split phase motor.

USES

Heavy duty loads where quiet running and a good starting torque is required such as wall mounted air conditioners.

PERMANENTLY SPLIT PHASE MOTOR

The permanently split phase motor also both windings connected permanently across the supply, with a capacitor in series with one of them.

USES

Light applications with a low starting torque as fans, blowers and dampers for regulating air flow.

SHADED POLE MOTOR

The shaded pole motor has a cage rotor with salient poles in the stator. On the side of each pole a shading ring is embedded which produces a flux that opposes the main flux.

USES

Since the speed can be varied within a limited range by a series resistor or inductor this suitable for torque fans and blowers.

THE SERIES MOTOR

The series motor is known as the universal motor because it can effectively operate on d.c and a.c supplies.

Like a d.c series motor it has a variable speed characteristic up to 15000 rpm.

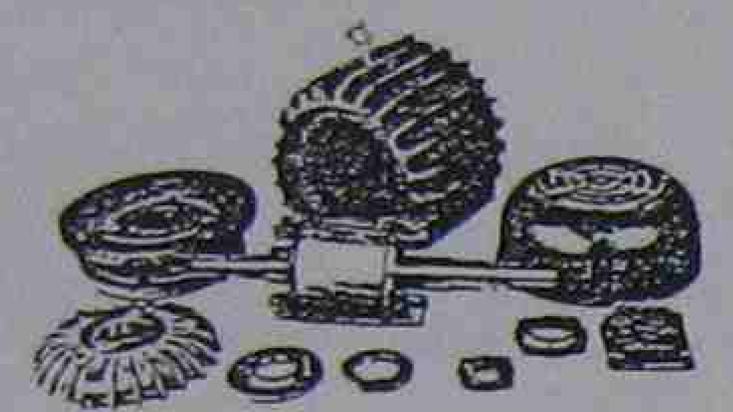
For large motors the load should not be removed when the motor is running at speed but in small motors the internal losses such as friction and windage will limit the speed to a safe value.

USES

Portable appliances such as saws and drills, sewing machines, food mixers and vacuum cleaners.

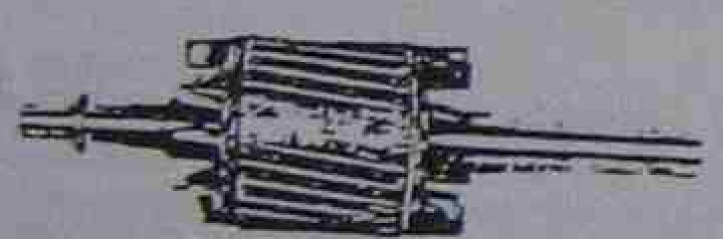
THREE PHASE INDUCTION MOTORS

Three phase induction motors consist of a laminated stator with three identical windings. The laminated rotor generally has single turn conductors within the slots and shorts circuited at each end. The motor derives its name from the fact that the currents flowing in the rotor are induced and drawn directly from the supply.

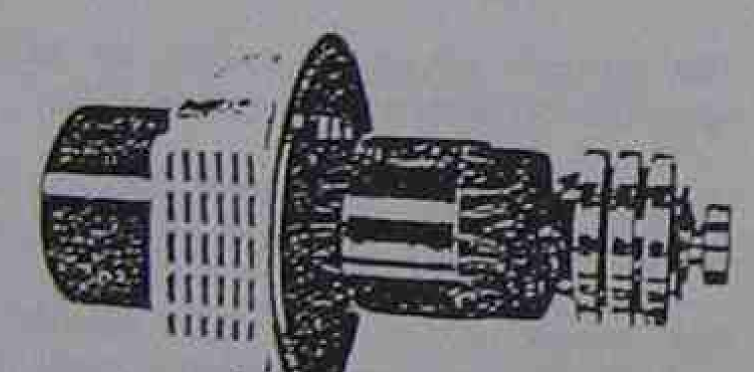


Rotors with bars in slots rather than windings are known as squirrel cage rotors and have a lower starting torque than the wound rotor.

The wound rotor has each of the three phase windings connected in star and terminated at three slip rings. The slip rings are connected to the external circuit using brushes, these motors are known as slip ring motors.

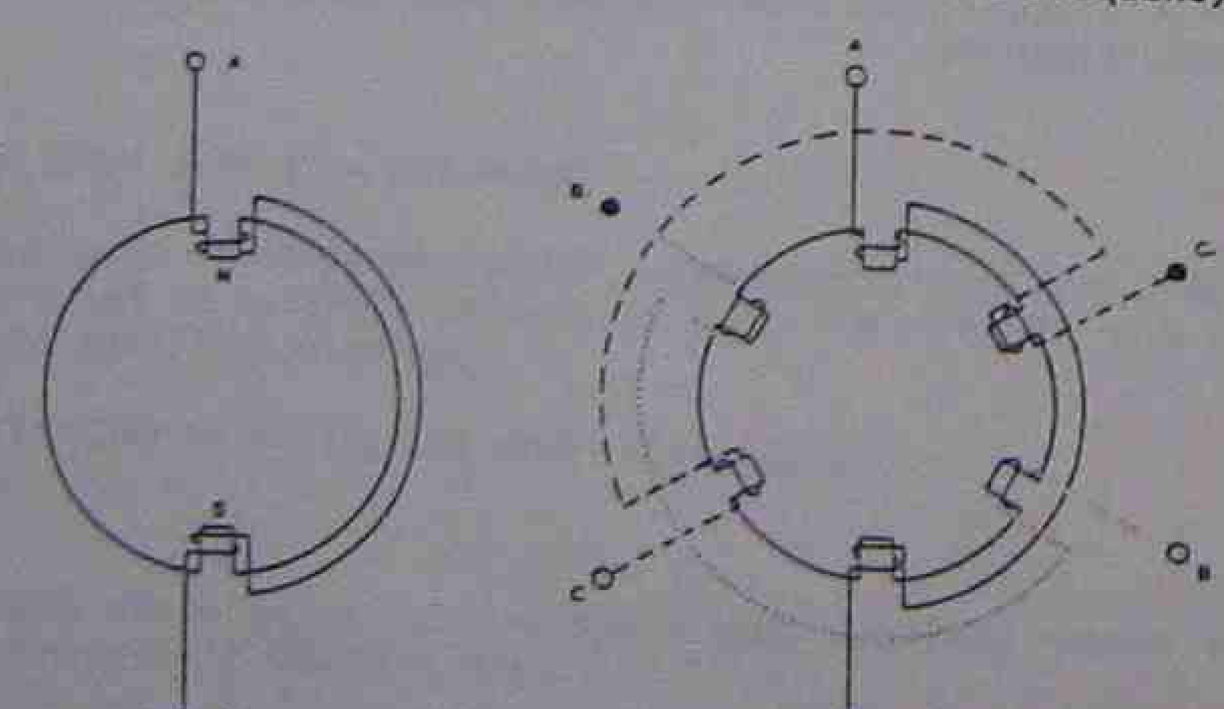


Squirrel-cage rotor for an induction motor



Wound rotor for an induction motor

The rotating magnetic field is established by three phase supply and the spe
is therefore related to the number of poles and the line frequency.



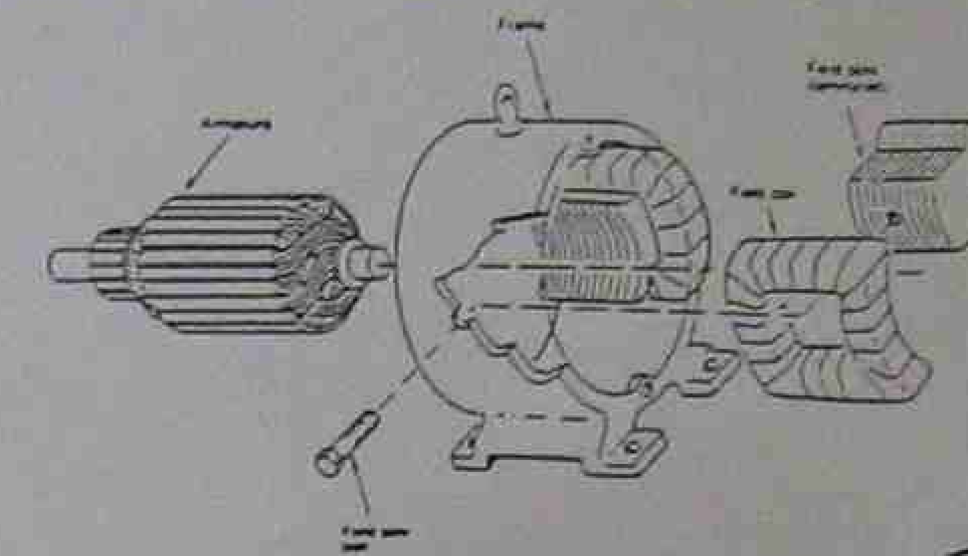
(a) One phase

(b) Three phases

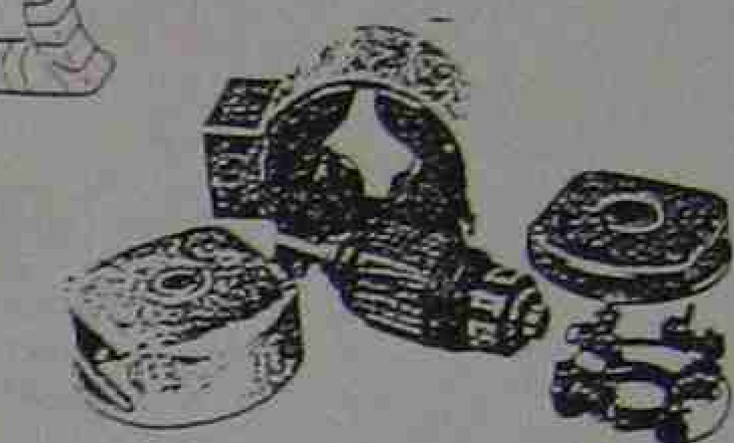
Polarities and connections in a two-pole, three-phase motor

DIRECT CURRENT MACHINES

A d.c machine can be a d.c generator or d.c motor,the construction of either is similar, that is, a d .c motor can be used as a generator and a d . c generator can be used as a motor.

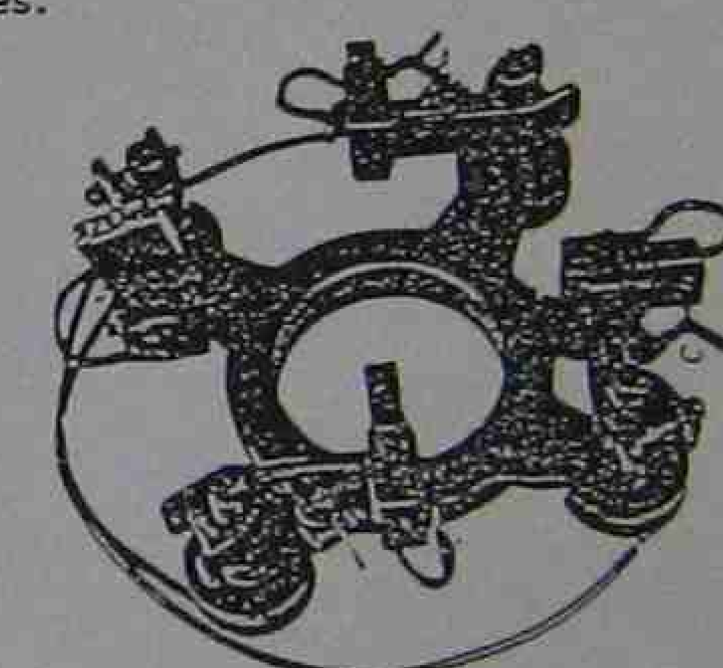


Component parts of a d.c. machine



Dismantled d.c. machine

The armature is electrically connected to the outside world by the use of a commutator and brushes.



Brush gear assembly of a d.c. machine

The commutator by its nature of operation will rectify the A.C voltage as the armature rotates.

D.C motors can use permanent magnets for their fields, these are usually limited to toys using batteries such trains or radio controlled planes.

For larger motors such as those used in furnace drives and steel mills, Alnico magnets are used, and speed control is usually achieved by varying the voltage applied to the armature.

The wound field version of the separately excited motor has no size limitations, speed control is usually by means of controlling field current.