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[1]	−55		+150	−55		+150	°C
Storage Temperature Range[1]	−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[2]			±0.1			±0.1	°C
Long Term Drift[3]			±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[4] (Reverse Voltage = 10V)		10			10		pA
PACKAGE OPTIONS[5]							
"H" Package: TO-52		AD590LH			AD590MH		
"F" Package: Flat Pack (F2A)		AD590LF			AD590MF		

CIRCUIT DESCRIPTION[2]

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.

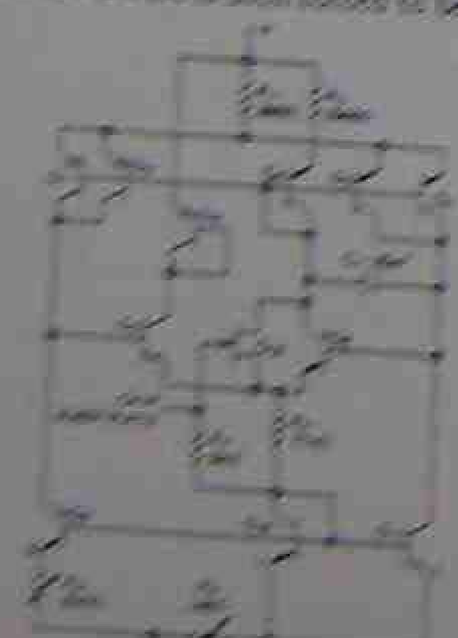

Figure 1. Schematic Diagram

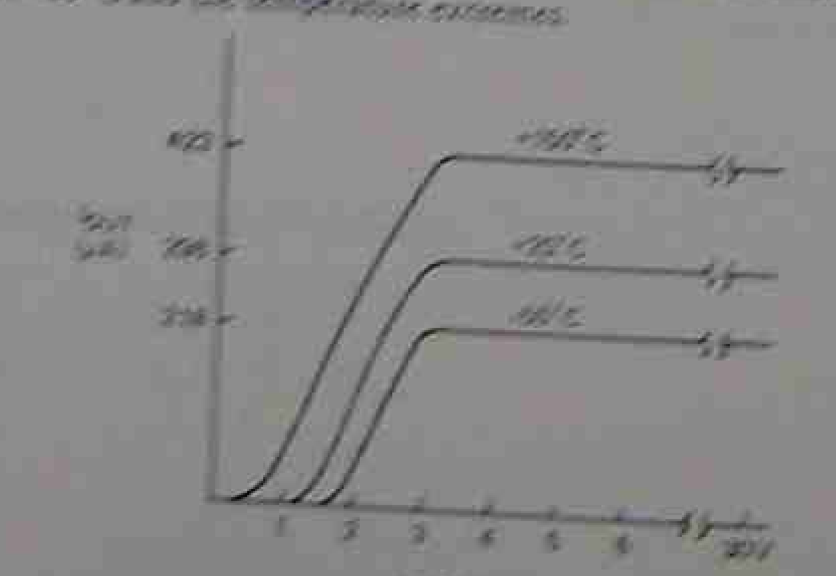

Figure 2. V–I Plot

[2] 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.

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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)[1] 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.

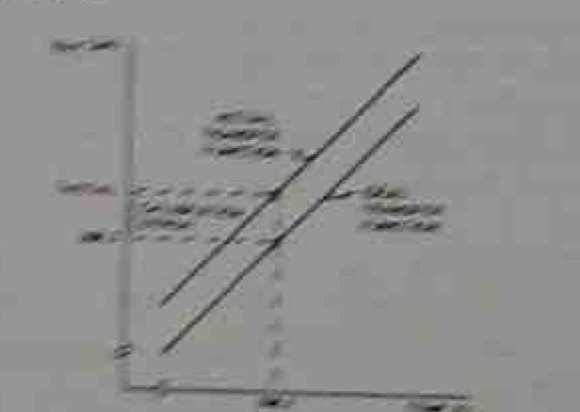

Figure 3. Calibration Error vs. Temperature

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

[1] 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.

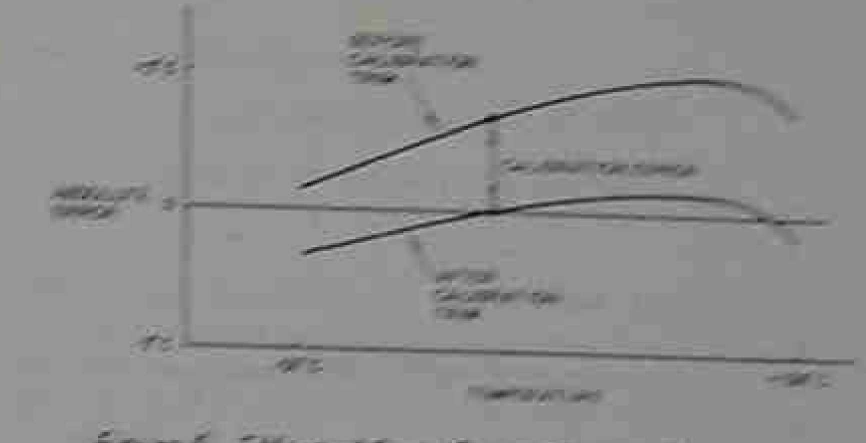

Figure 5. Effect of Scale Factor Trim on Accuracy

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.

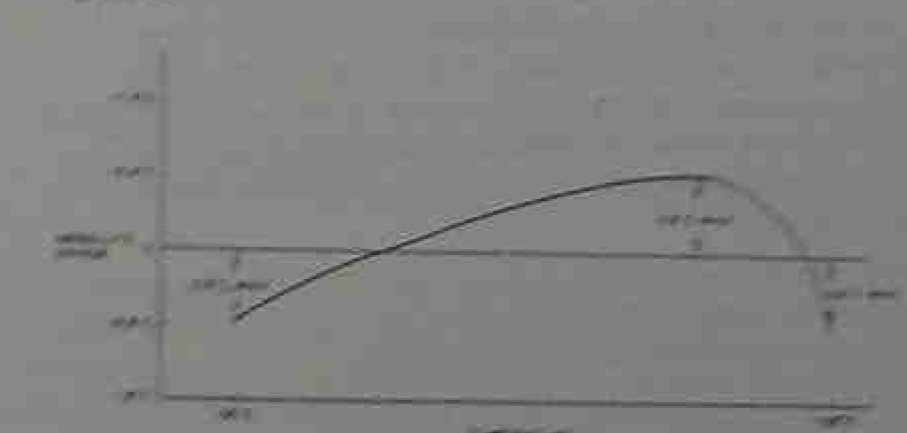

Figure 6. Nonlinearity

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.

Understanding the AD590 Specifications

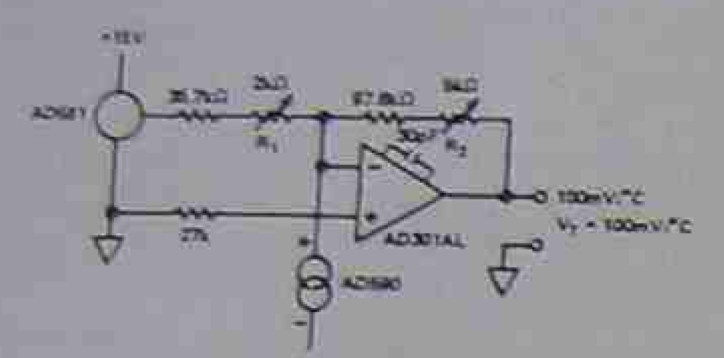

Figure 7A. Two Temperature Trim

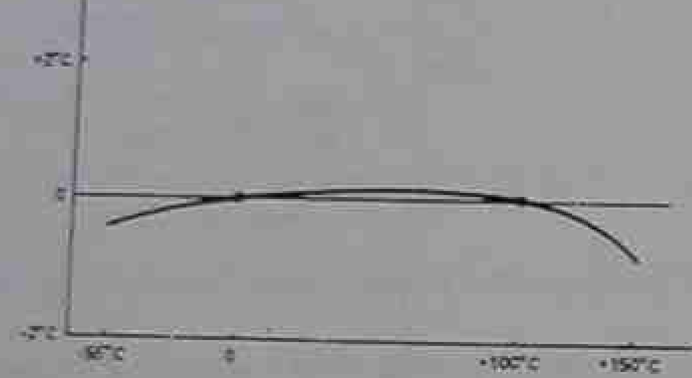

Figure 7B. Typical Two-Trim Accuracy

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.

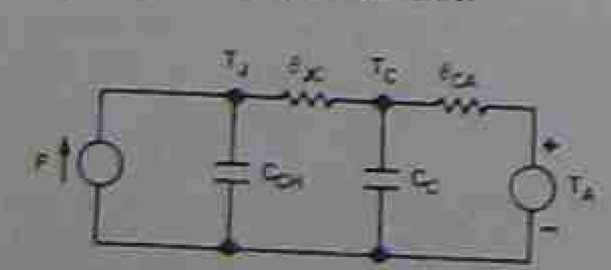

Figure 8. Thermal Circuit Model

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.

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MEDIUM	$\theta_{JC} + \theta_{CA}$ (°C/watt)		τ (sec) (Note 3)	
	H	F	H	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.

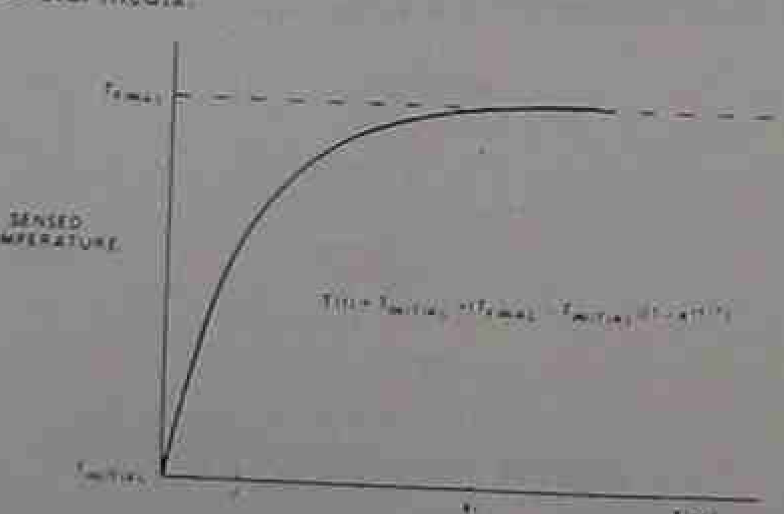

Figure 9. Time Response Curve

GENERAL APPLICATIONS

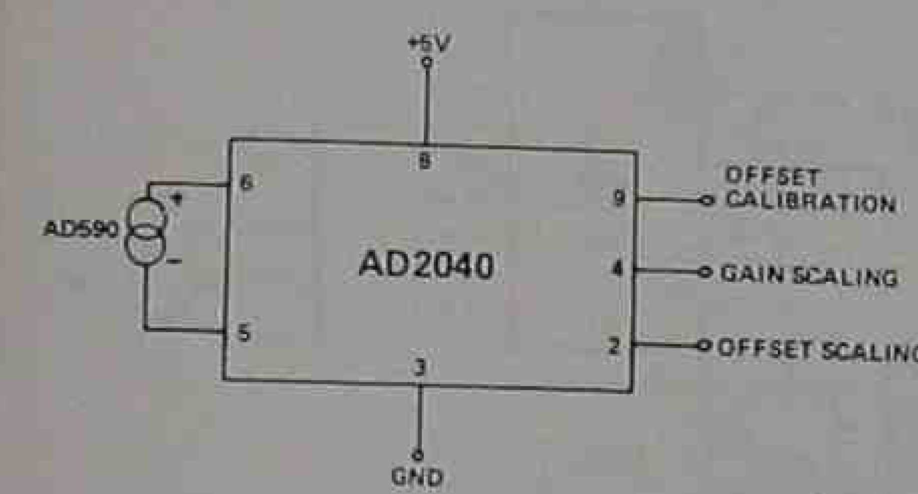

Figure 10. Variable Scale Display

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.

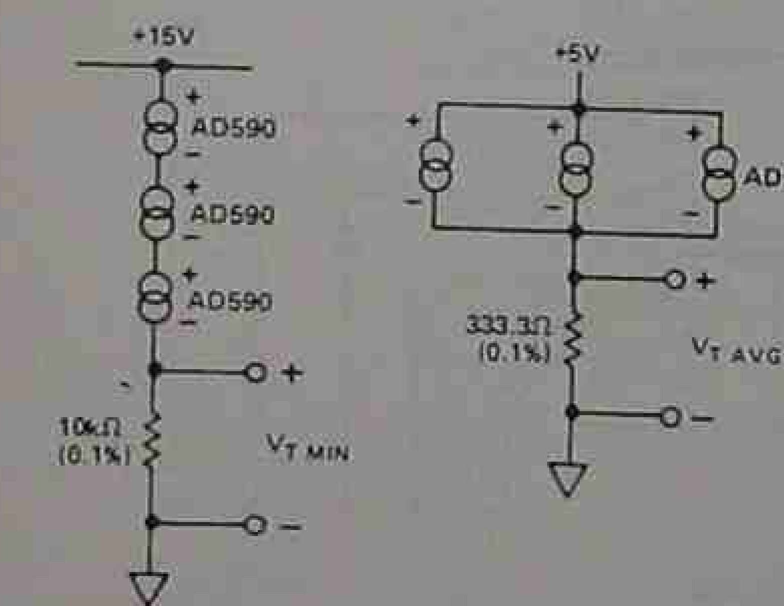

Figure 11. Series & Parallel Connection

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.

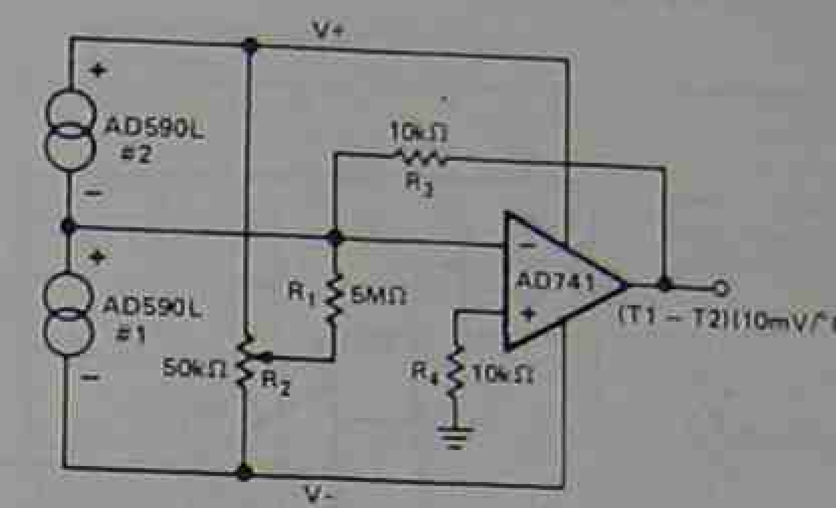

Figure 12. Differential Measurements

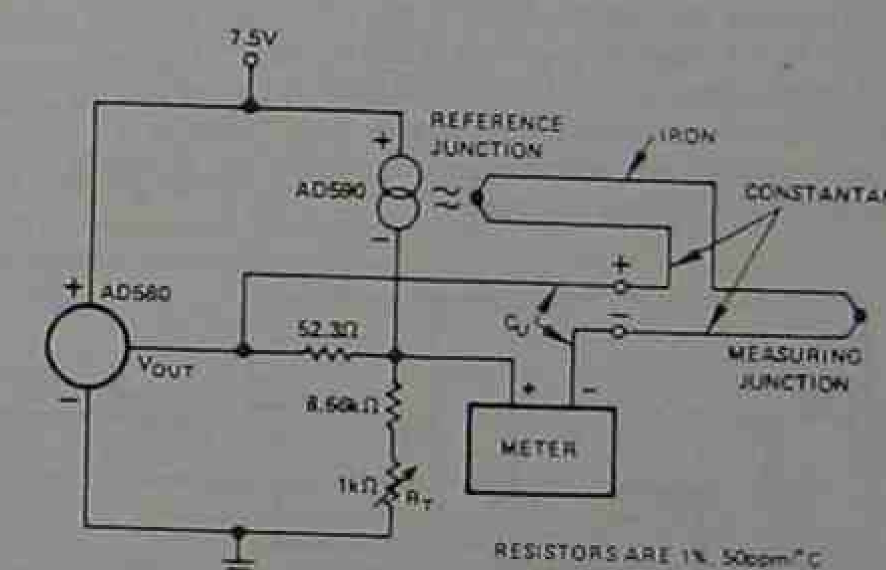

Figure 13. Cold Junction Compensation Circuit for Type J Thermocouple

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.

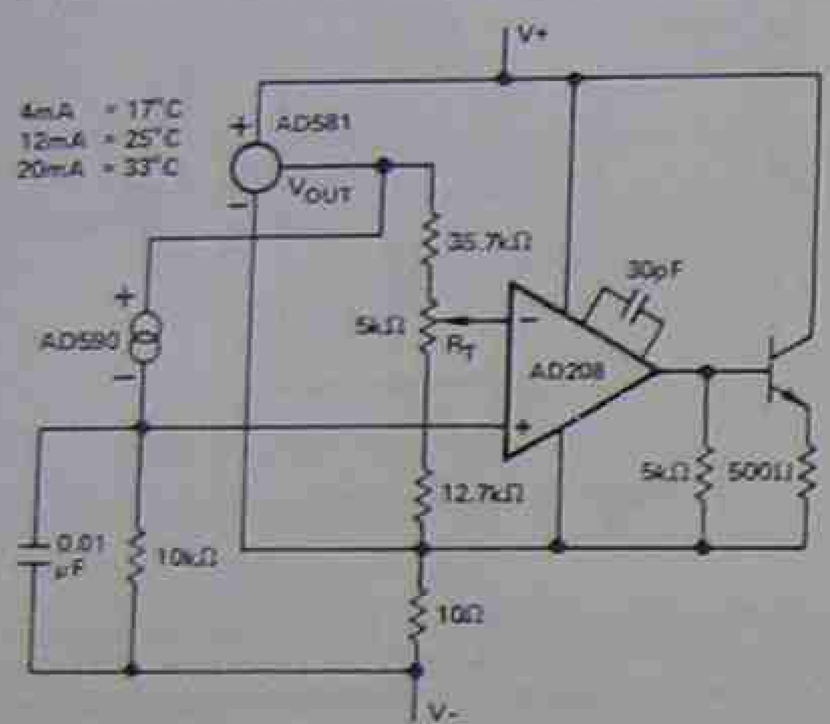

Figure 14. 4 to 20mA Current Transmitter

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.

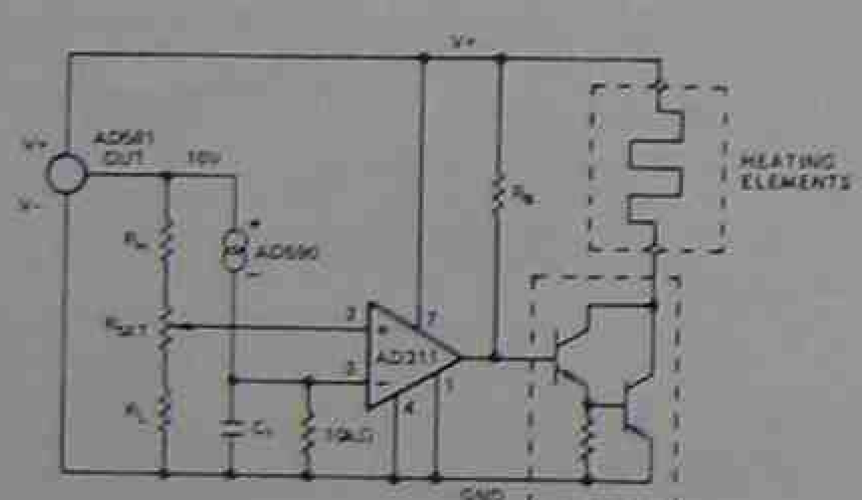

Figure 15. Simple Temperature Control Circuit

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

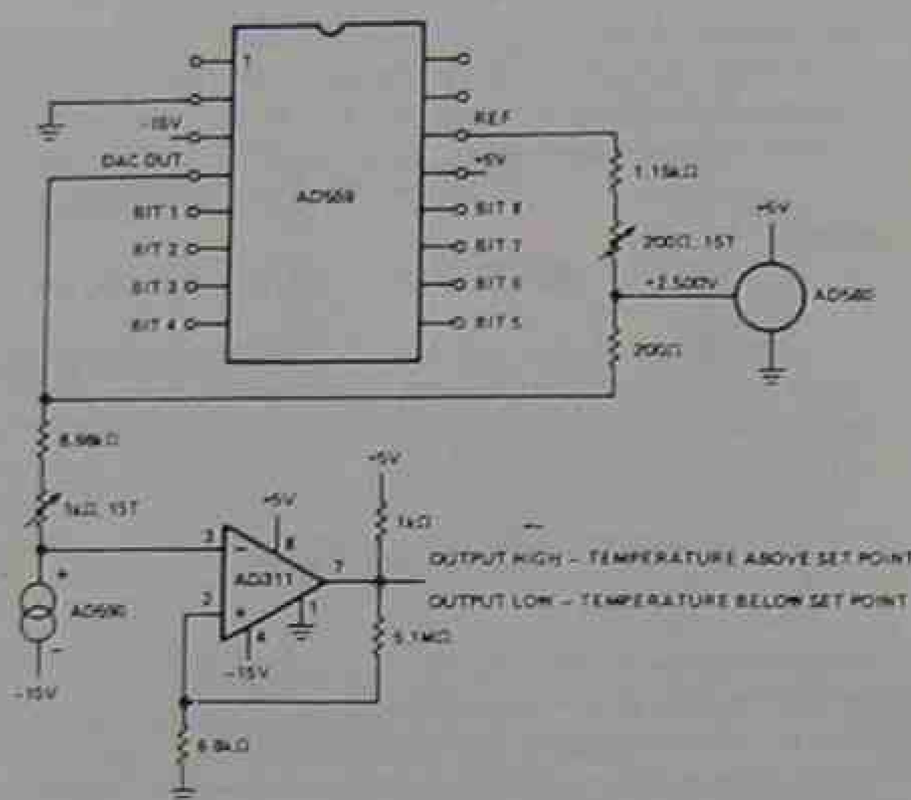

Figure 16. DAC Set Point

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.

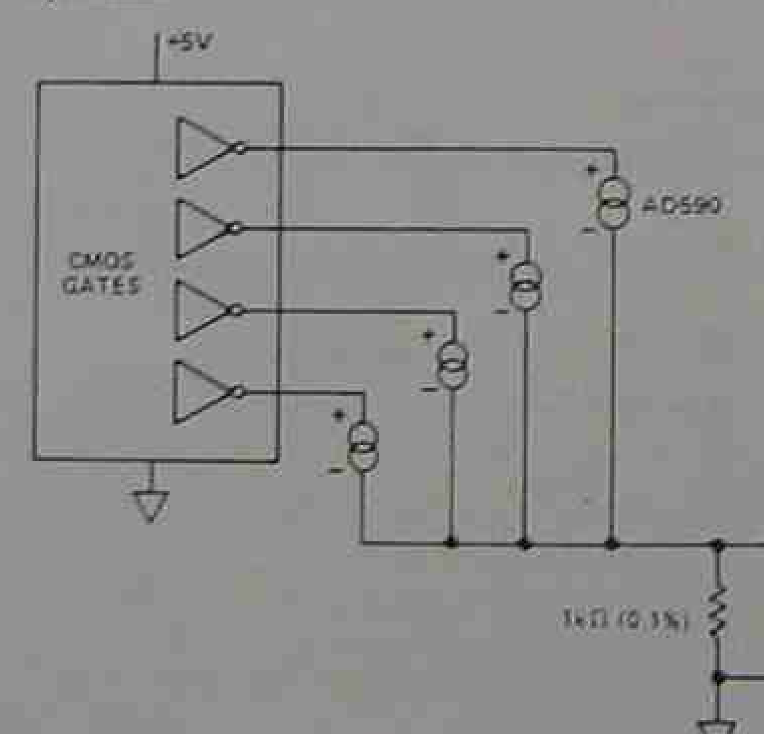

Figure 17. AD590 Driven from CMOS Logic

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.

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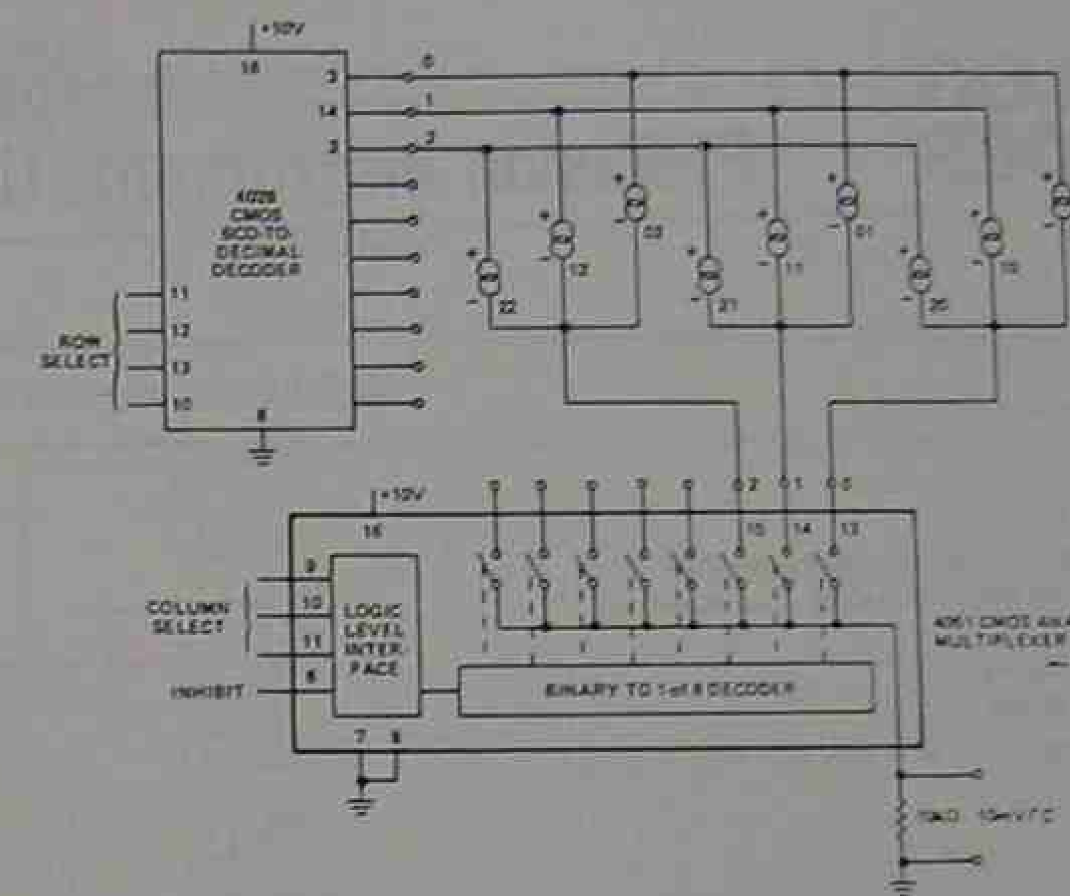


Figure 18. Matrix Multiplexer

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.

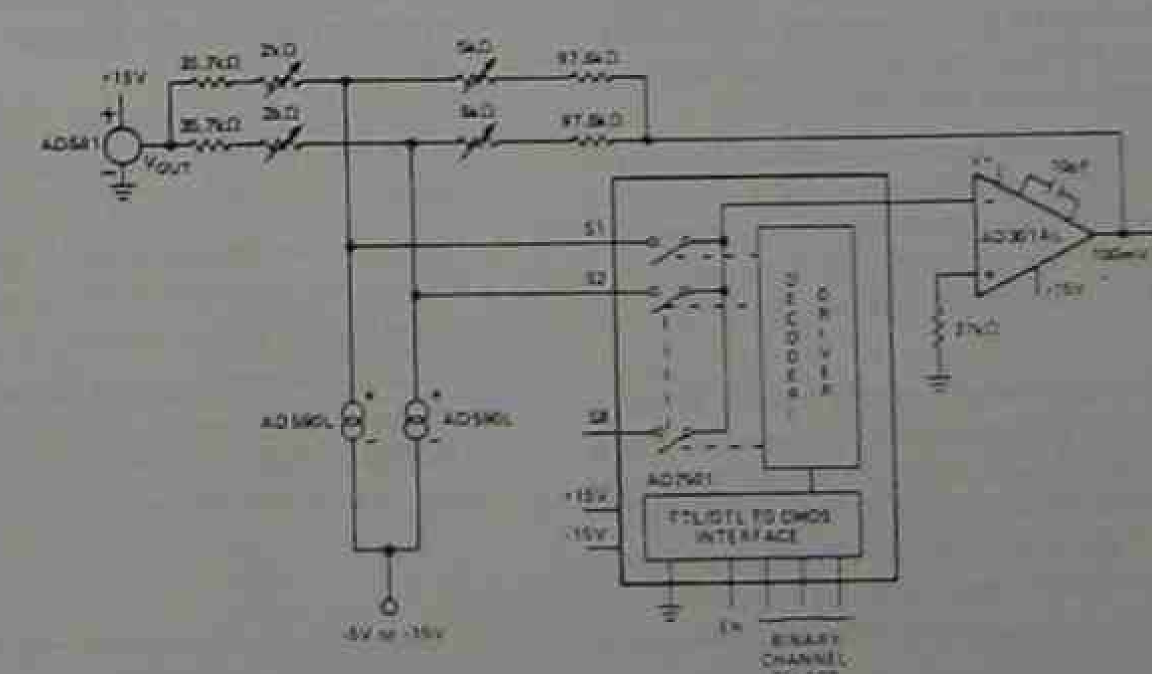


Figure 19. 8-Channel Multiplexer

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

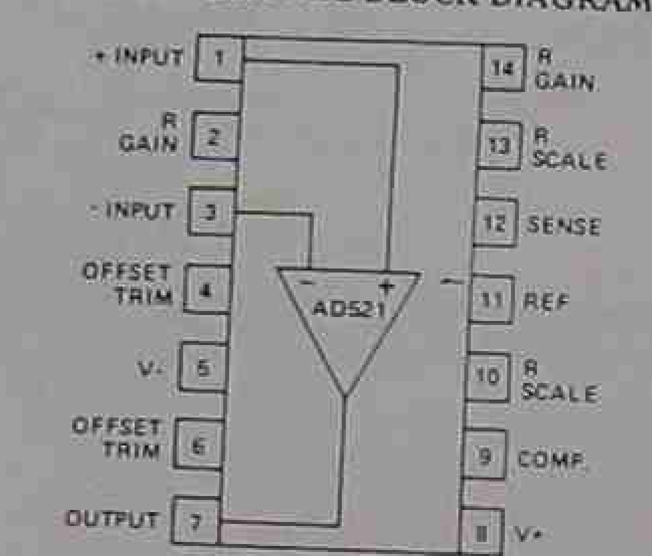
Integrated Circuit Precision Instrumentation Amplifier

AD521

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

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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 ($3 \times 10^9\Omega$) 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_I})	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 $\frac{V_{OUT}}{V_{IN}} = \frac{R_S}{R_G}$.

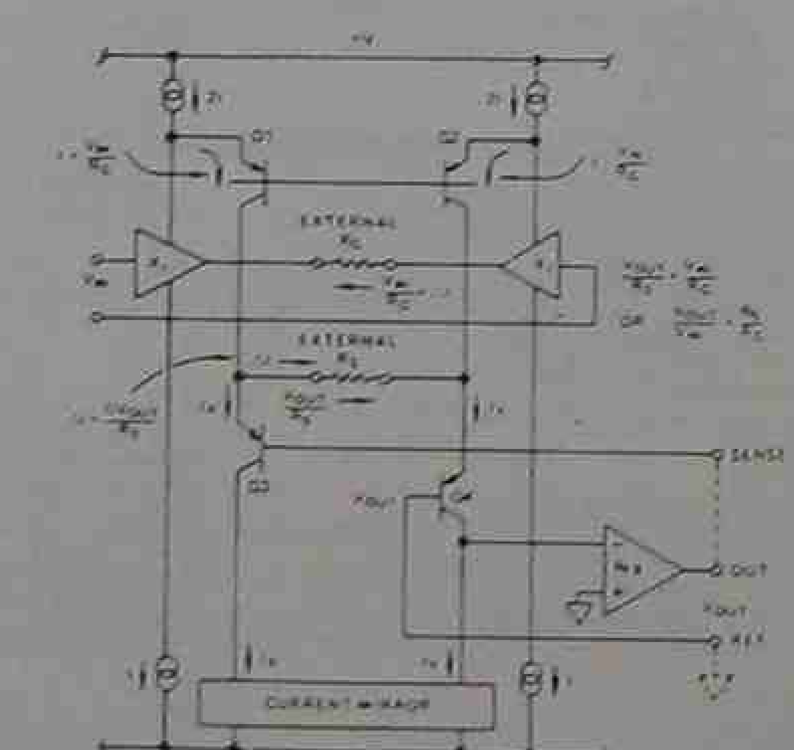


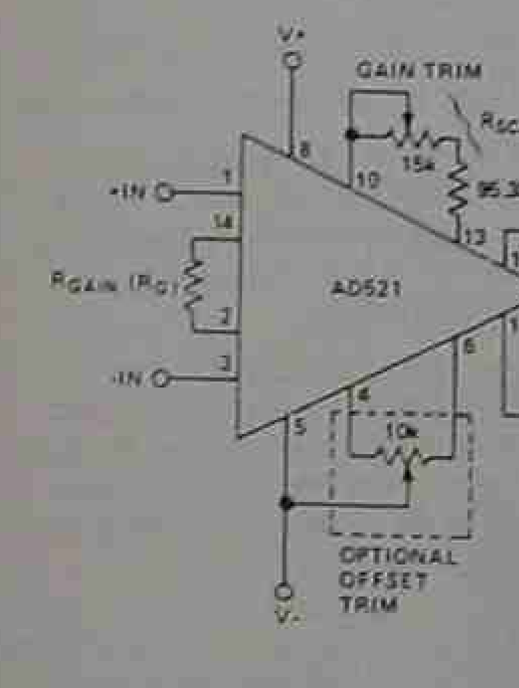
Figure 1. Simplified AD521 Schematic

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.

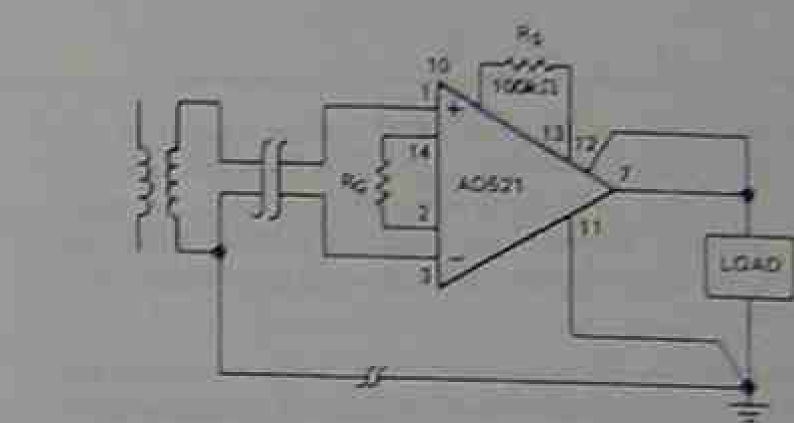


GAIN	VALUE OF R_G
0.1	1MΩ
1	100kΩ
10	10kΩ
100	1kΩ
1000	100Ω

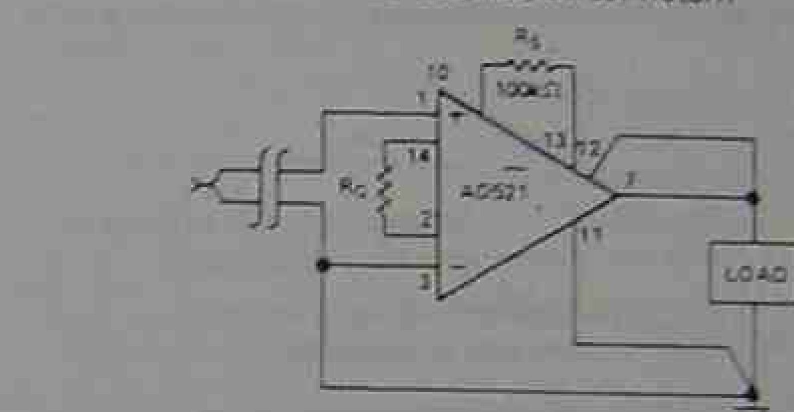
Figure 2. Operating Connections for AD521

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.

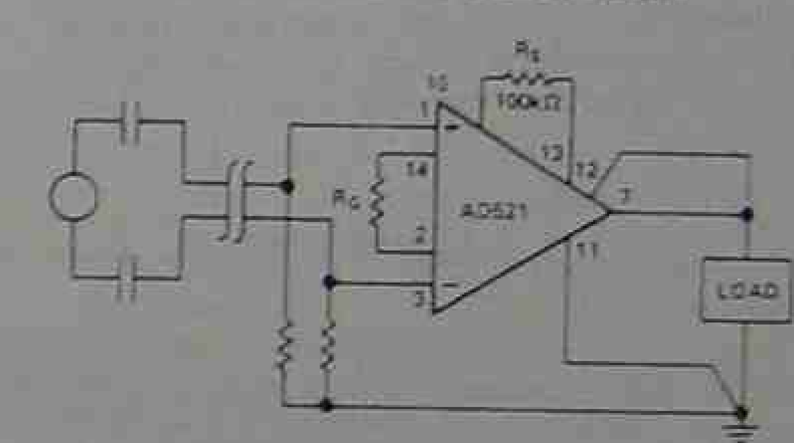
4. Do not exceed the allowable input signal range. The linearity of the AD521 decreases if the inputs are driven within 5 volts of the supply rails, particularly when the device is used at a gain less than 1. To avoid this possibility, attenuate the input signal through a resistive divider network and 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

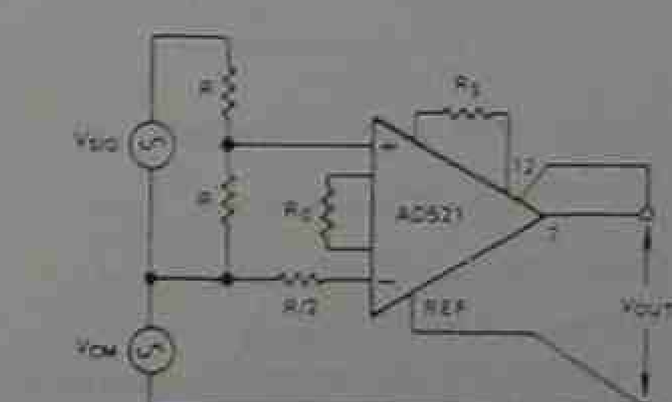


b). Thermocouple, Direct Return



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

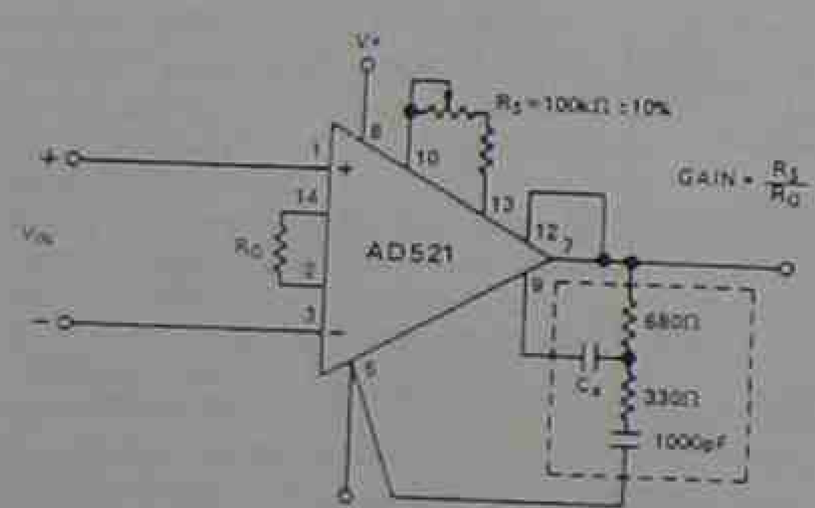
5. Use the compensation pin (pin 9) and the applicable compensation circuit when the amplifier is required to drive a capacitive load. It is worth mentioning that coaxial cables can "invisibly" provide such capacitance since many popular coaxial cables display capacitance in the vicinity of 30pF per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



$C_x = \frac{1}{100\pi f_t}$ when f_t 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.

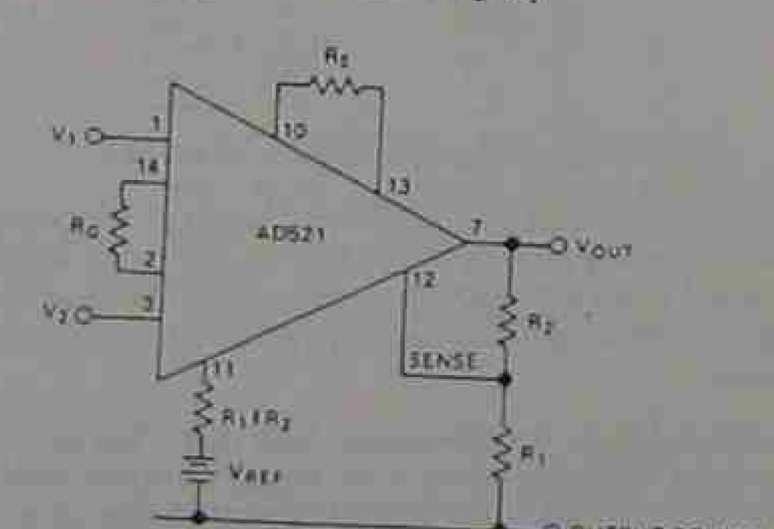
By separating these errors, one can evaluate the total error independent of the gain settings used, similar to the situation with the input offset specifications on an op amp. In a given gain configuration, both errors can be combined to give a total error *referred to the input* (R.T.I.) or *output* (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_S/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 $1 + R_2/R_1$.



$$V_{OUT} = \left[V_{REF} + \left(\frac{R_S}{R_G}\right)(V_1 - V_2)\right]\left[\frac{R_1 + R_2}{R_1}\right]$$

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.

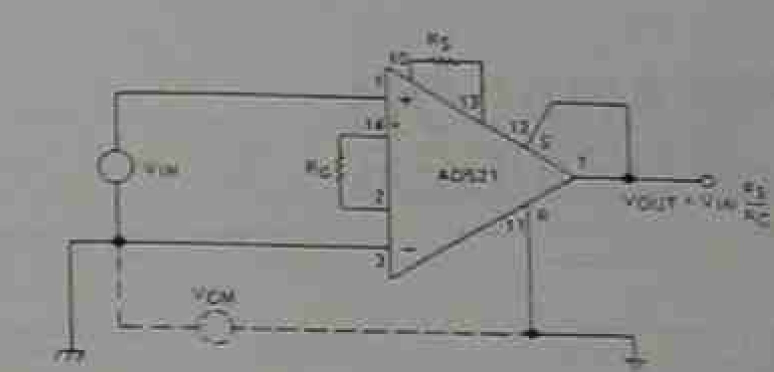


Figure 7. Ground loop elimination. The reference input, Pin 11, allows remote referencing of ground potential. Differences in ground potentials are attenuated by the high CMRR of the AD521.

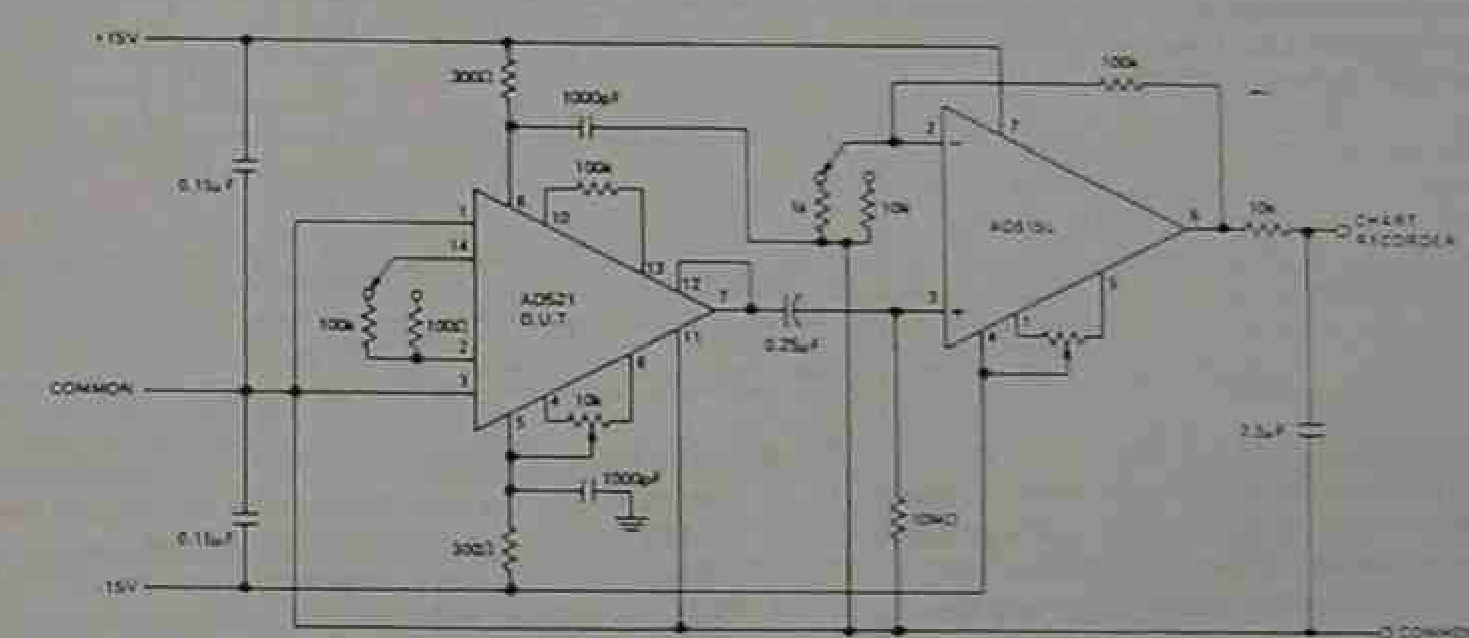


Figure 8. Test circuit for measuring peak to peak noise in the bandwidth 0.1Hz to 10Hz. Typical measurements are found by reading the maximum peak to peak voltage noise of the device under test (D.U.T.) for 3 observation periods of 10 seconds each.

ANALOG DEVICES

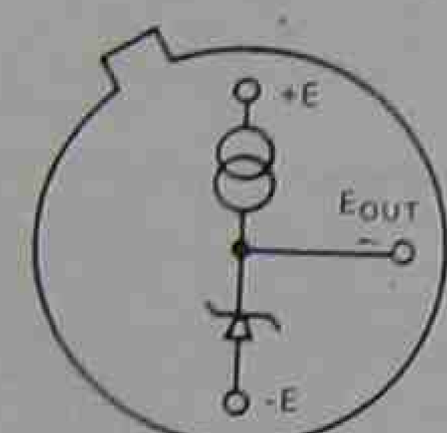
High Precision 2.5 Volt IC Reference

AD580*

7

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

PRODUCT DESCRIPTION

The AD580 is an improved three-terminal, low cost, tempera-
ure 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-
afer-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
ax 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			AD580K			AD580L			AD580M			
	Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	Min	Typ	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 85			7 60			4.3 25			1.75 10	mV ppm/°C
LINE REGULATION $7V \le V_{IN} \le 30V$ $4.5V \le V_{IN} \le 7V$		1.5 0.3	4 3		1.5 0.3	4 2			3 1			2 1	mV mV
LOAD REGULATION ΔI = 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 Per Month		250 25			250 25			250 25			250 25		µV µV
TEMPERATURE PERFORMANCE Specified Operating Storage	0 −55 −65		+70 +125 +175	0 −55 −65		+70 +125 +175	0 −55 −65		+70 +125 +175	0 −55 −65		+70 +125 +175	°C °C °C
PACKAGE OPTION[1] – TO-52	AD580JH			AD580KH			AD580LH			AD580MH			

Model	AD580S			AD580T			AD580U			
	Min	Typ	Max	Min	Typ	Max	Min	Typ	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 55			11 25			4.5 10	mV ppm/°C
LINE REGULATION $7V \le V_{IN} \le 30V$ $4.5V \le V_{IN} \le 7V$		1.5 0.3	6 3			2 1			2 1	mV mV
LOAD REGULATION ΔI = 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 Per Month		250 25			250 25			250 25		µV µV
TEMPERATURE PERFORMANCE Specified Operating Storage	−55 −55 −65		+125 +150 +175	−55 −55 −65		+125 +150 +175	−55 −55 −65		+125 +150 +175	°C °C °C
ABSOLUTE MAXIMUM RATINGS Input Voltage Power Dissipation @ +25°C Ambient Temperature Derate above +25°C Lead Temperature (Soldering, 10 sec) Thermal Resistance Junction-to-Case Junction-to-Ambient	40V 350mW 2.4mW/°C 300°C 100°C/W 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/°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.

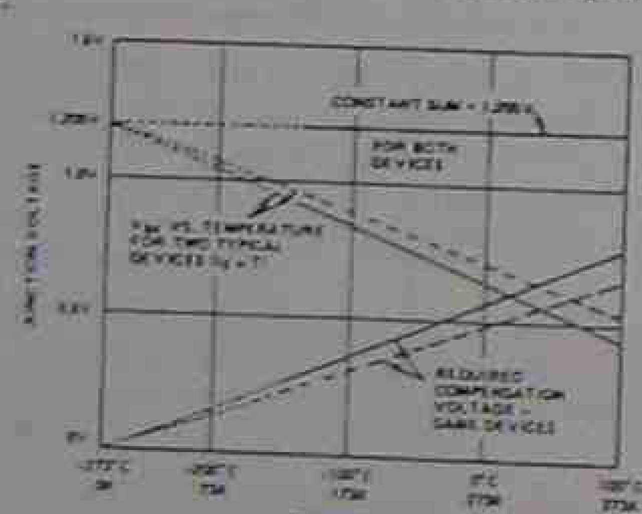


Figure 1. Extrapolated Variation of Base-Emitter Voltage with Temperature ($I_E \alpha T$), and Required Compensation, Shown for Two Different Devices

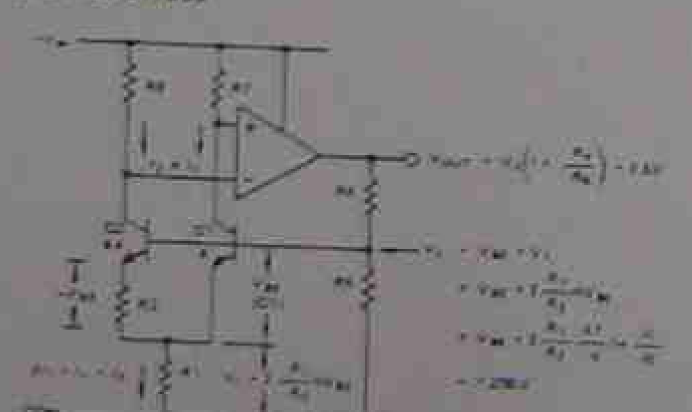


Figure 2. Basic Bandgap-Reference Regulator Circuit

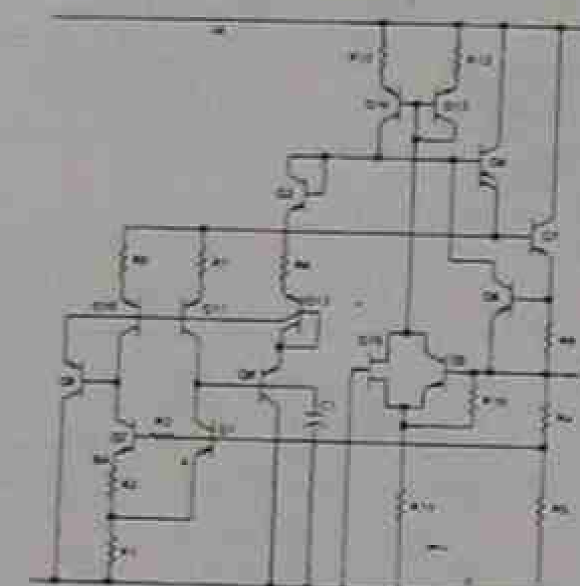


Figure 3. AD580 Schematic Diagram

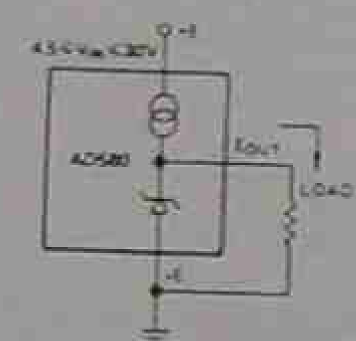


Figure 4. AD580 Connection Diagram

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/°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.

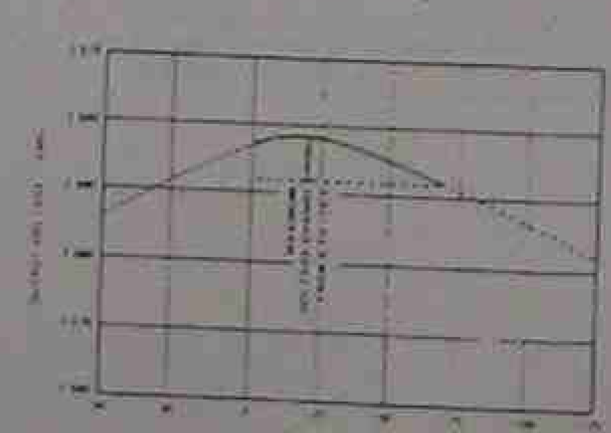


Figure 5. Typical AD580K Output Voltage vs. Temperature

7

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

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

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%/°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%/°C for I_{LIM} = 1mA and 0.01%/°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.

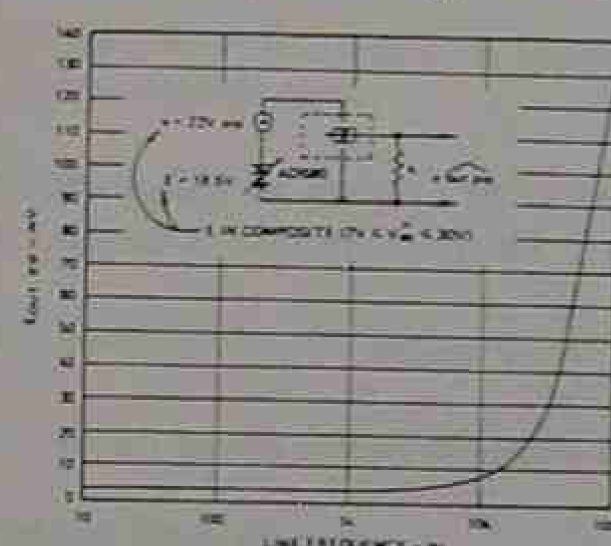


Figure 6. AD580 Line Rejection Plot

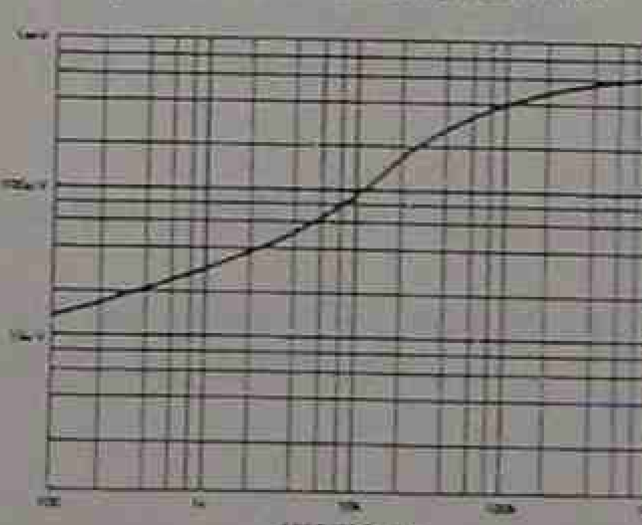


Figure 7. Peak-to-Peak Output Noise vs. Frequency

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

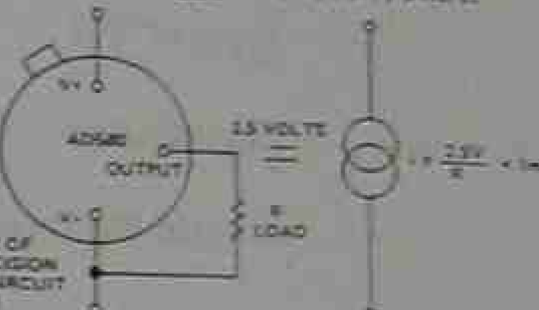


Figure 9. A Two-Component-Precision Current Limiter

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.

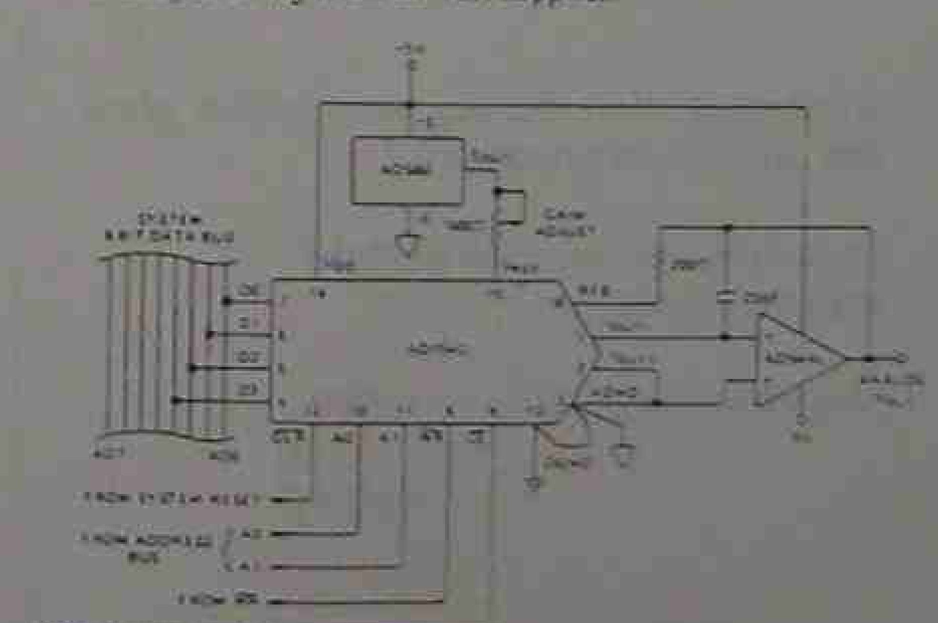


Figure 10. Low Power, Low Voltage Reference for the AD7542 Microprocessor-Compatible 12-Bit DAC

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

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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.

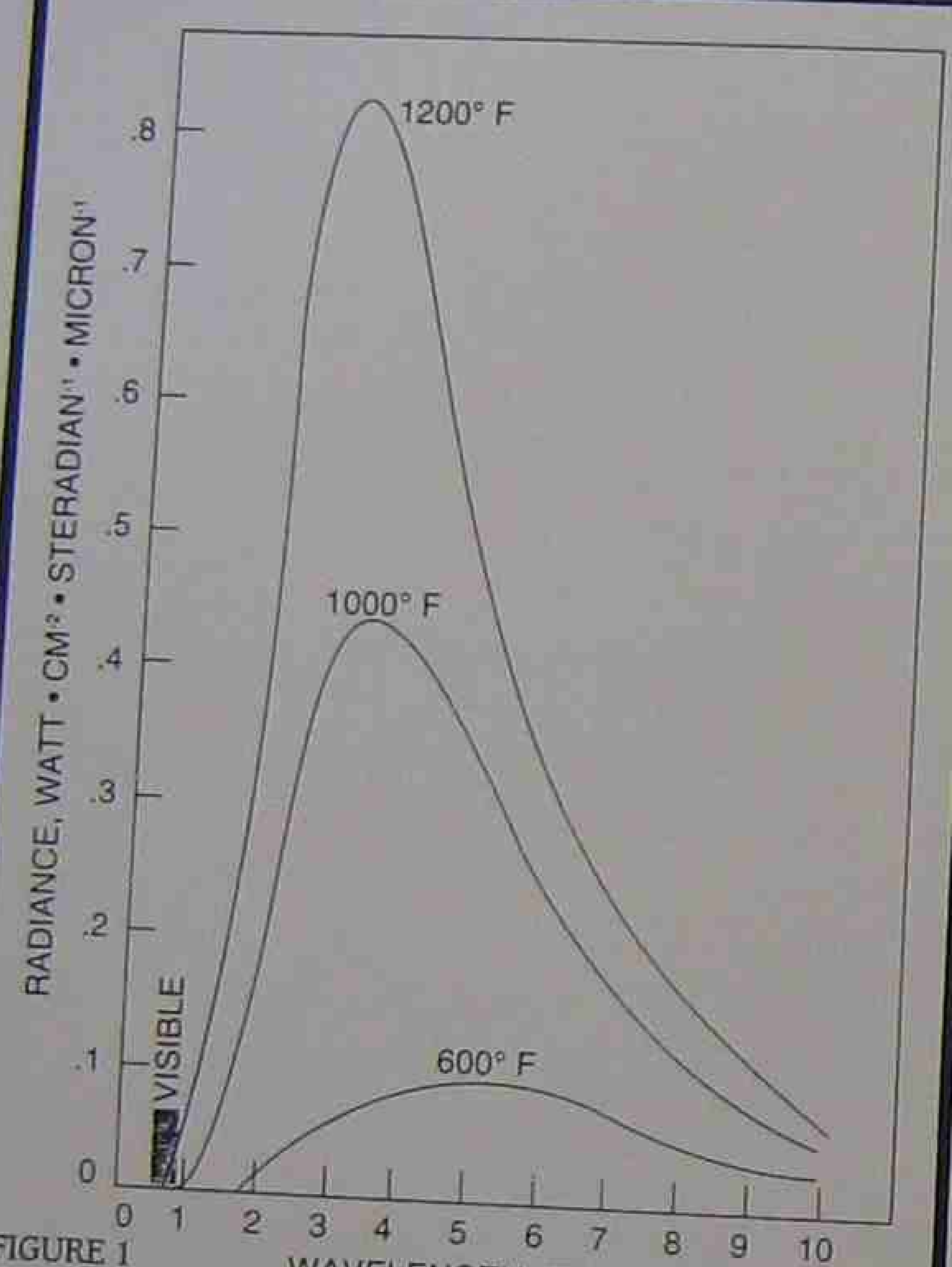


FIGURE 1
Blackbody Radiation Characteristics

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.

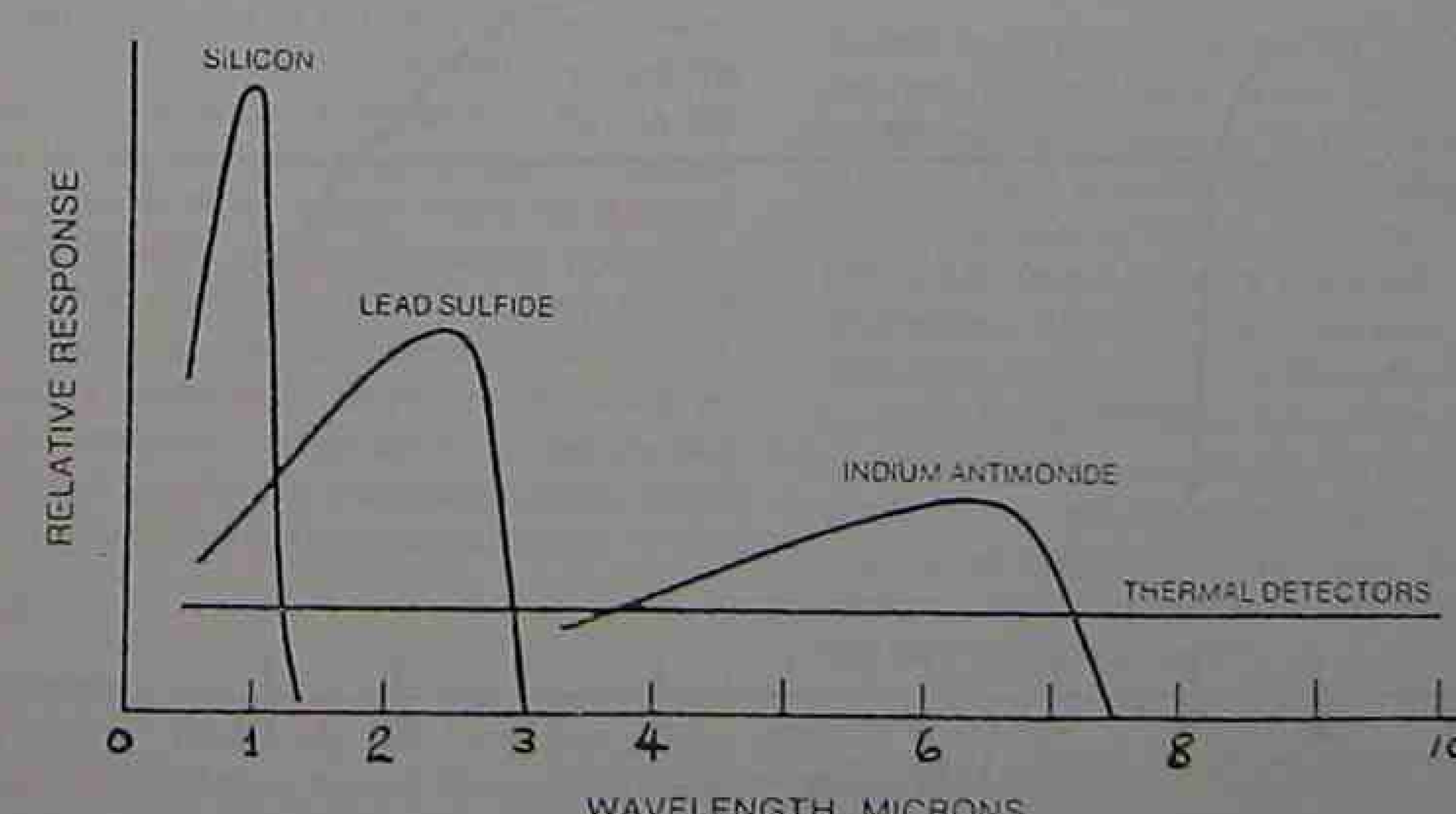


FIG. 2 — SPECTRAL RESPONSE CHARACTERISTICS OF SEVERAL INFRARED DETECTORS

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.

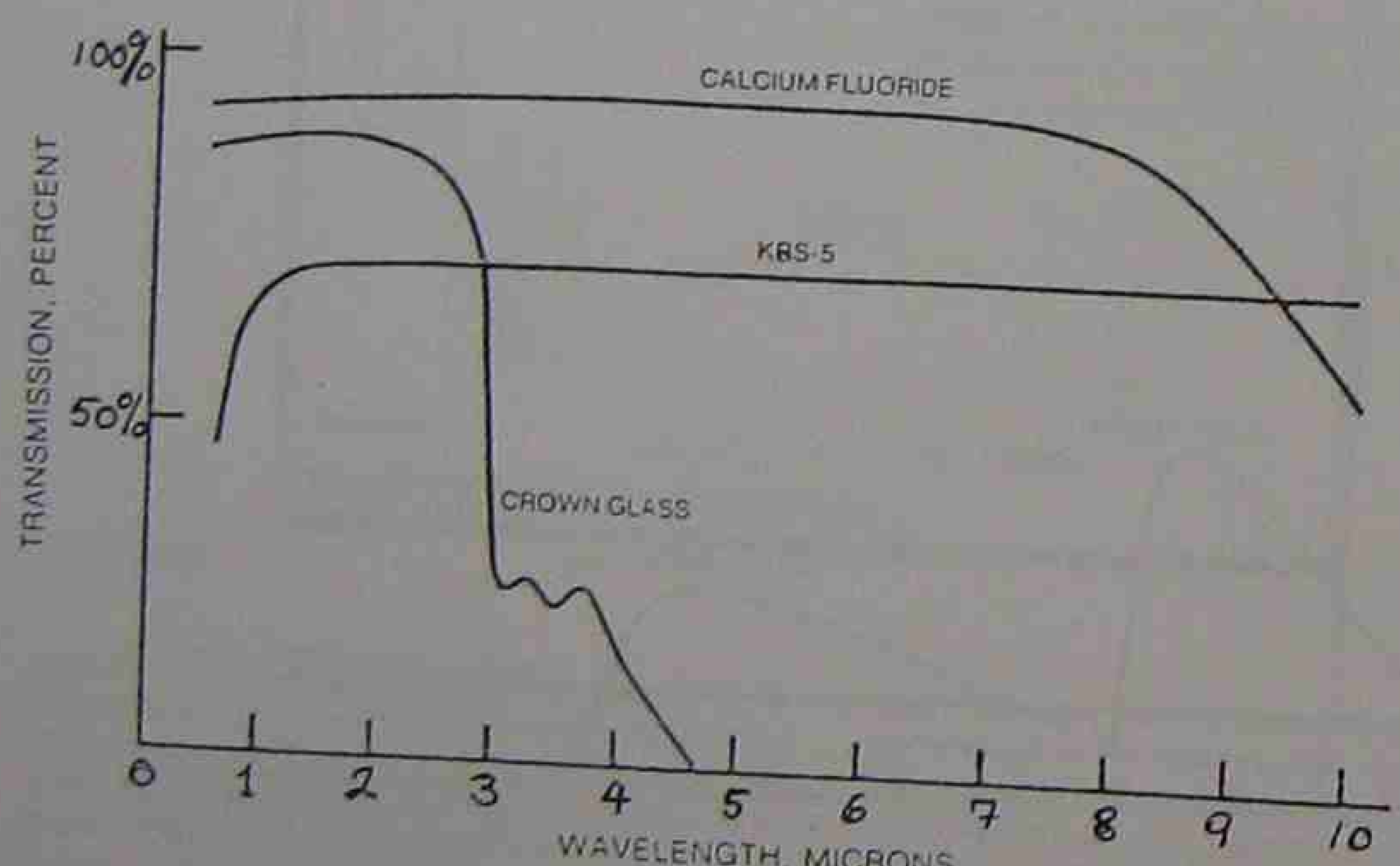


FIG. 3 — TRANSMISSION CHARACTERISTICS OF SEVERAL INFRARED OPTICAL MATERIALS

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.

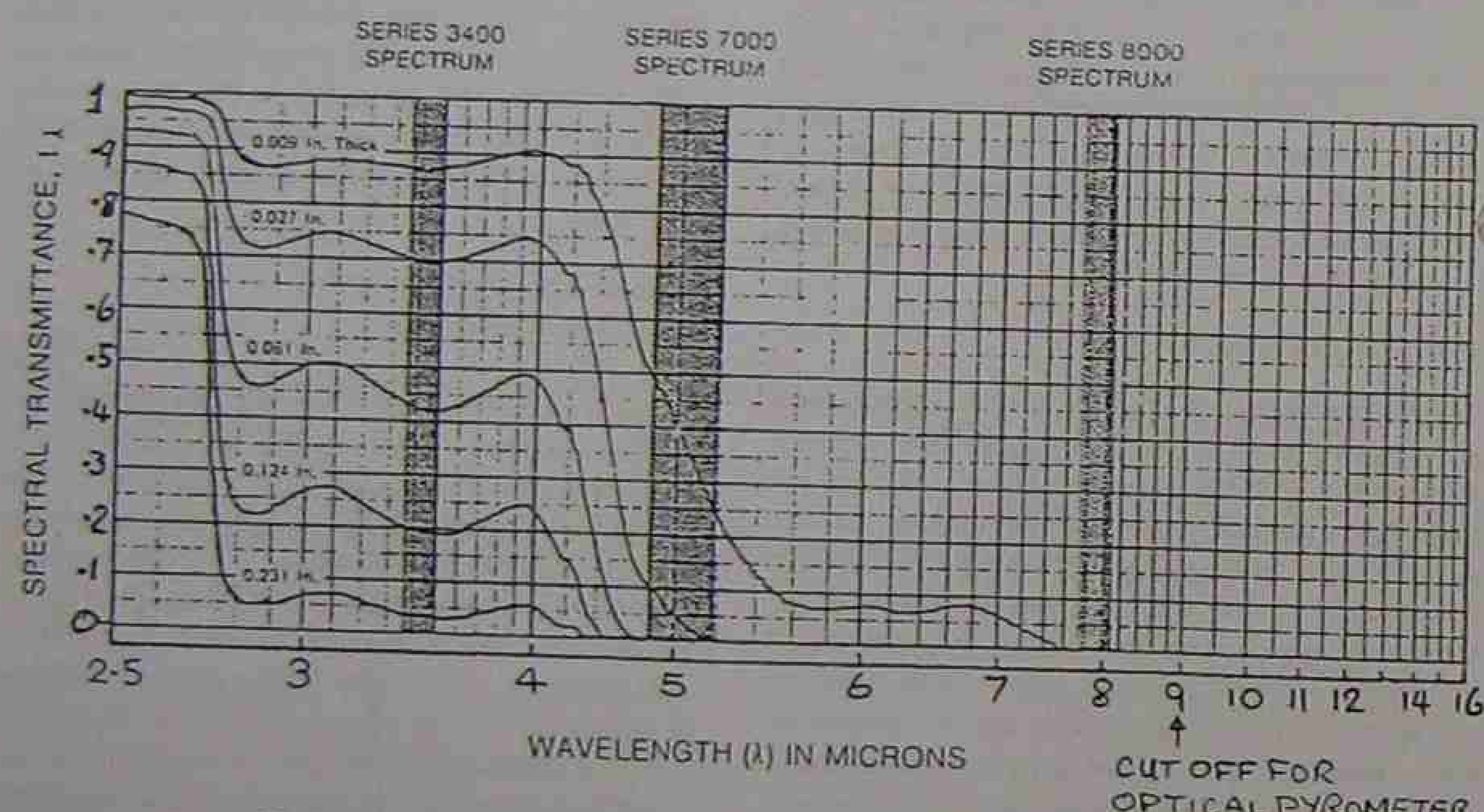


FIG. 4 — EFFECT OF THICKNESS ON SPECTRAL TRANSMITTANCE CURVES FOR SODA-LIME-SILICA GLASS

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.

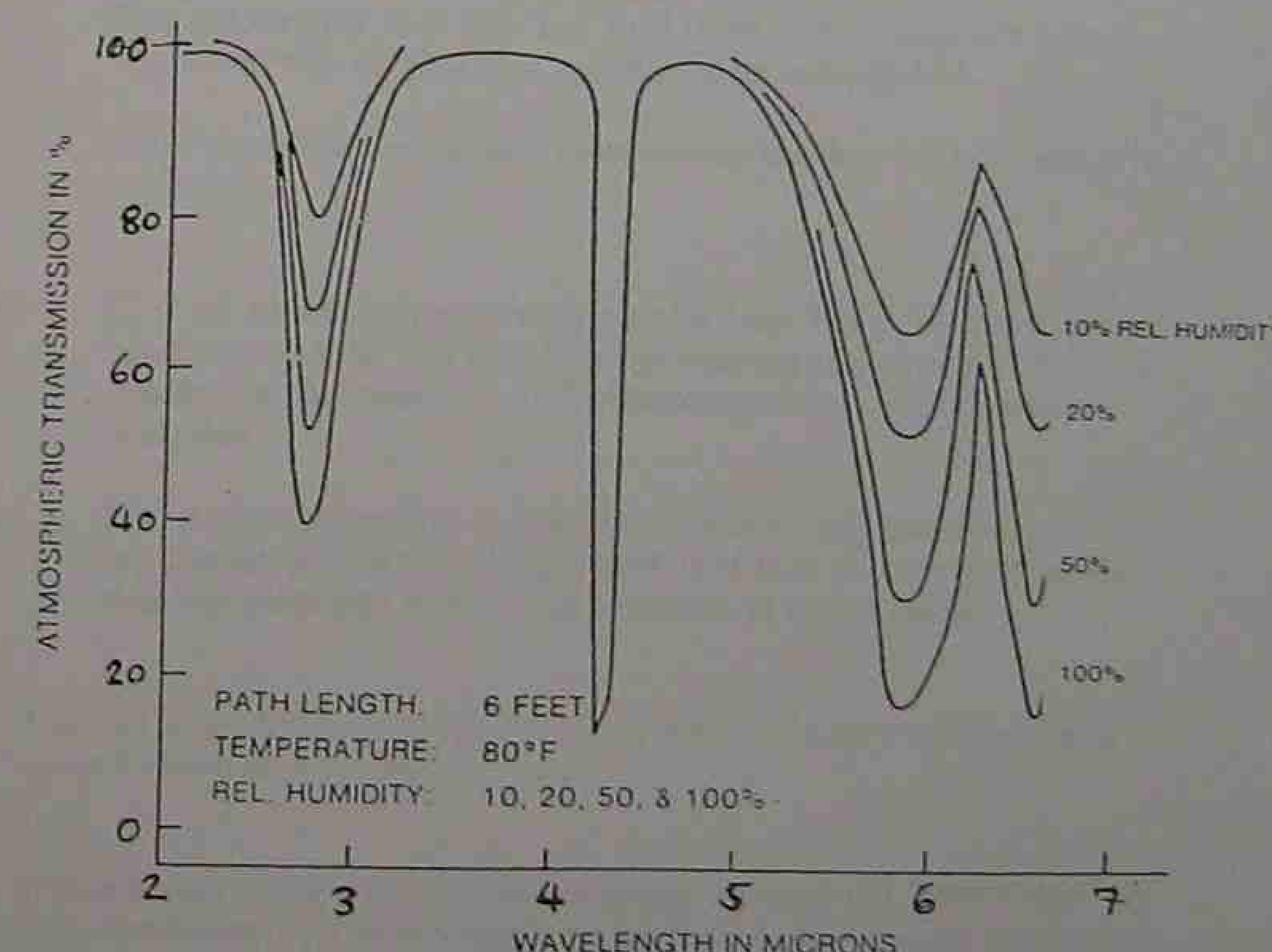


FIG. 5 — TRANSMISSION FOR ATMOSPHERE AT 80°F, AND SEVERAL RELATIVE HUMIDITIES

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.

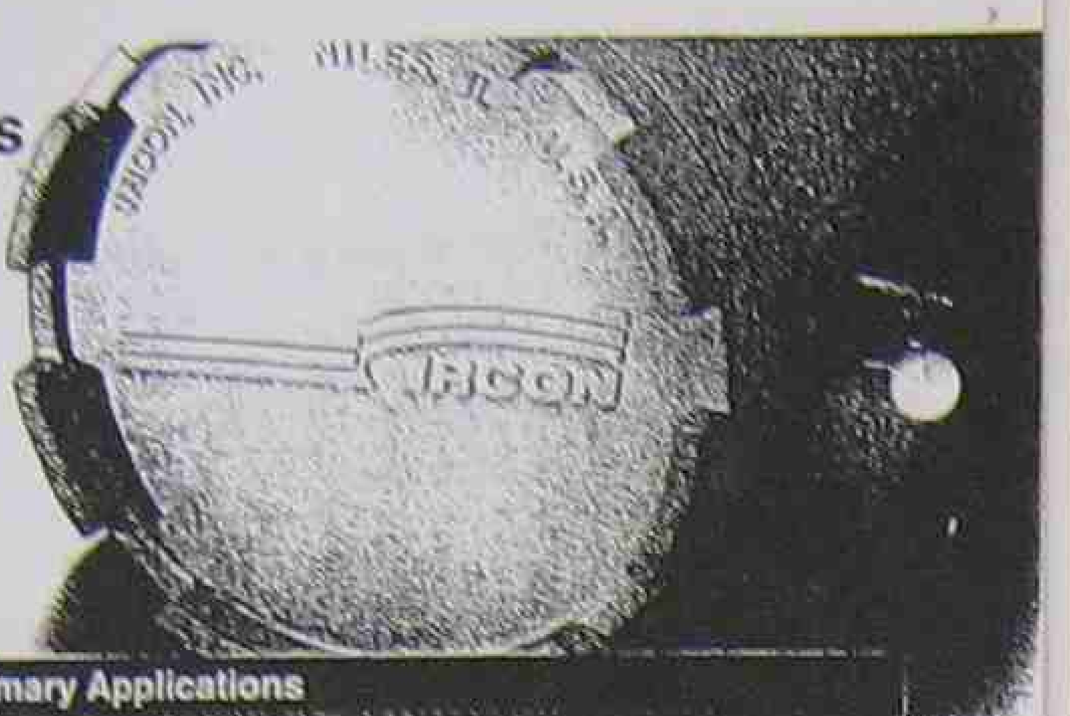
d) If at all possible have someone else take the pyrometer scale readings so that you can concentrate on the process of obtaining the readings and obtaining the most consistent matches.

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



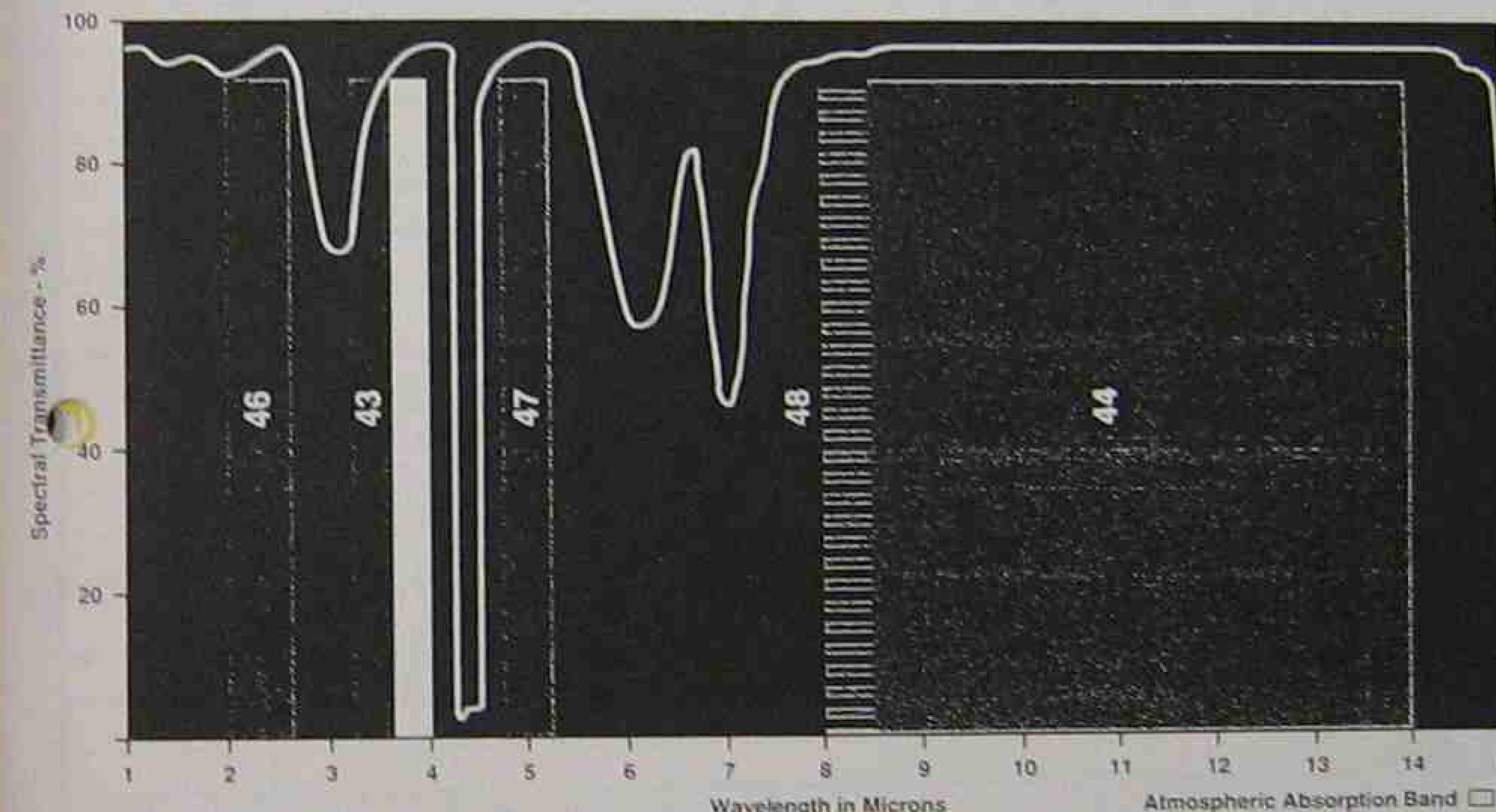
Series	Temp. Span	Wavelength	Primary Applications
43	120 to 400°F 200 to 600°F 300 to 1000°F 50 to 200°C 100 to 400°C 150 to 500°C	3.43 microns	Ideally suited for thin polyethylene film applications such as extrusion coating, blown film cooling, blow molding, thermo-forming, film laminating and printing.
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?

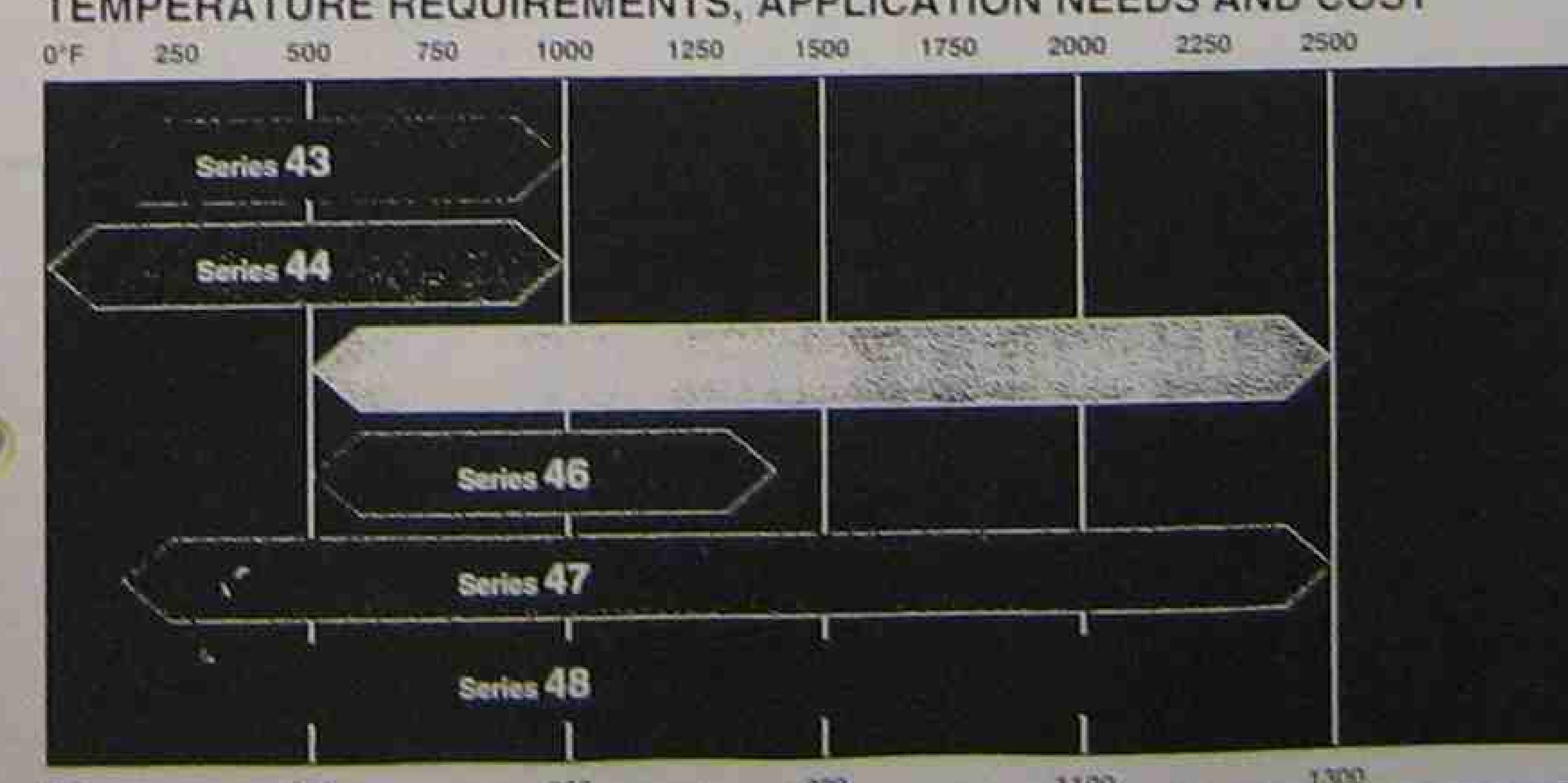
In many temperature measurement and control applications, infrared thermometers are the only solution because they don't require contact for the measurement. Where the product is small, fragile, moving, or in a vacuum or other controlled atmosphere, there really is no other solution.

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.

YOUR CHOICE OF SIX SPECTRAL RANGES



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



APPLICATION WAVELENGTH SELECTION GUIDE

TYPICAL APPLICATIONS	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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Theory Notes

Area

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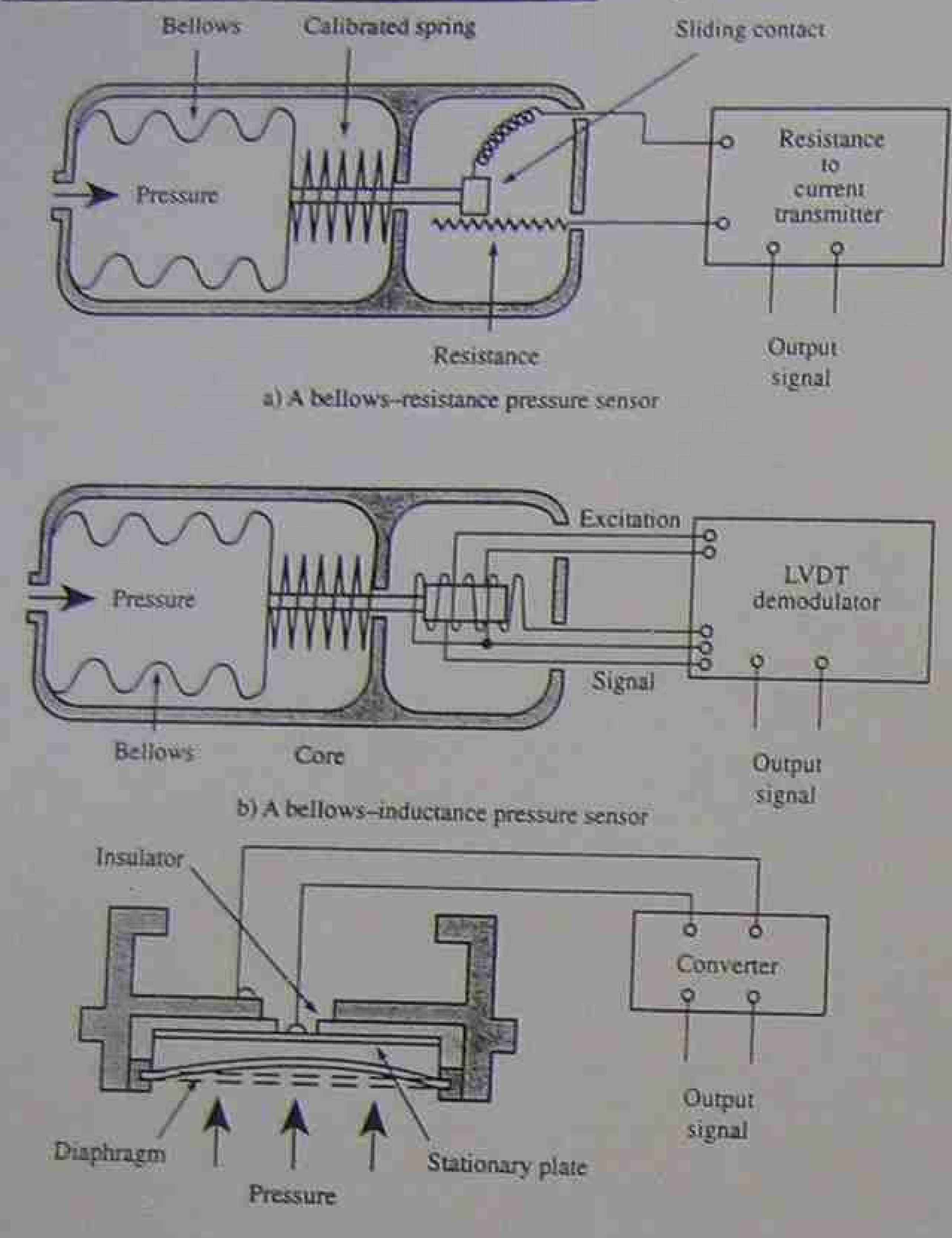
Pressure Sensors.



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.

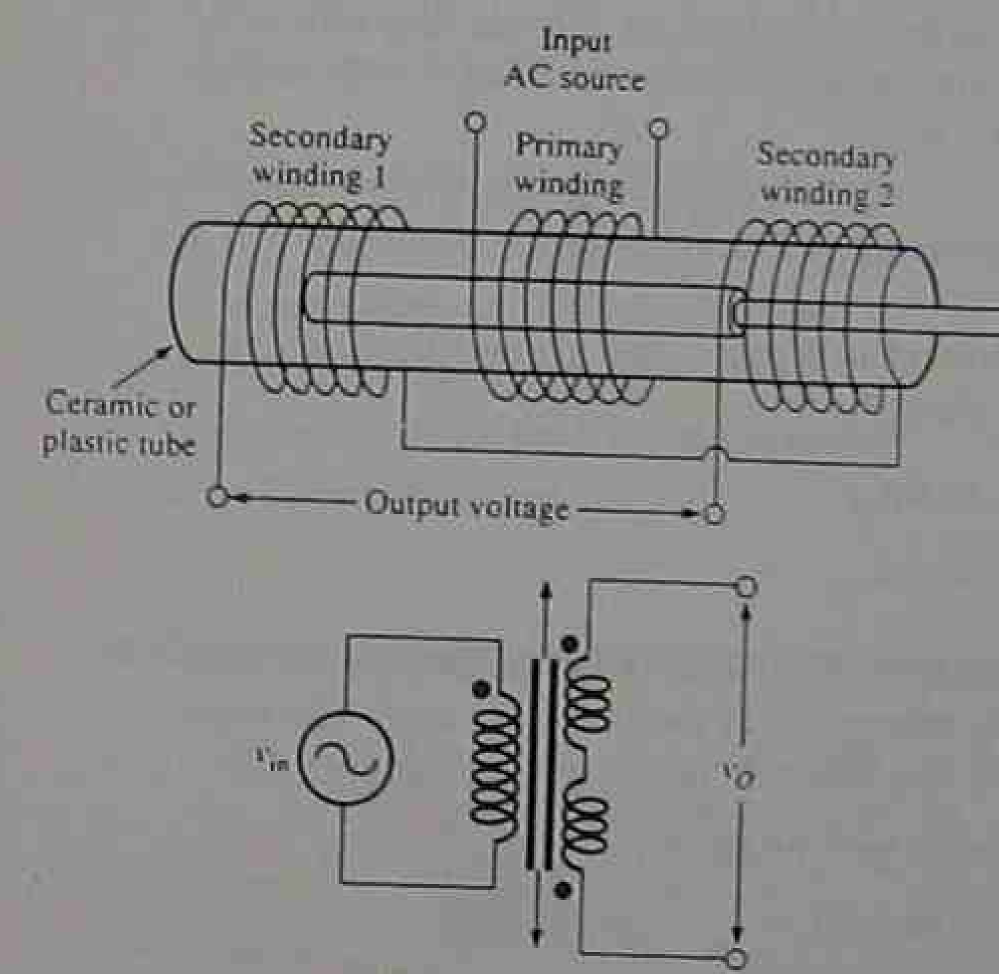


FIGURE 7–35
Structure of an LVDT (From Berlin and Getz, *Principles of Electronic Instrumentation and Measurement*. Merrill, 1988.)

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.

$$q_{avg} = Av_{avg}$$

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

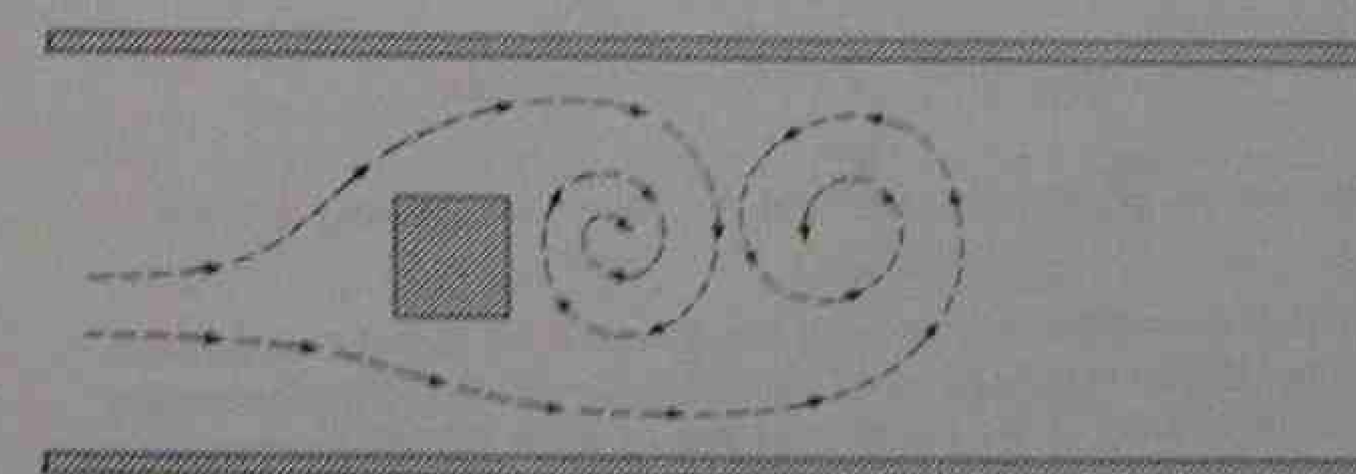


Figure 10.8 In a vortex flow meter, vortices are alternately formed and shed on one side of an obstruction and then on the other side. The resulting pulsations are sensed by a piezoelectrical crystal, and the frequency of the pulses is directly proportional to the volumetric flow rate.

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.

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

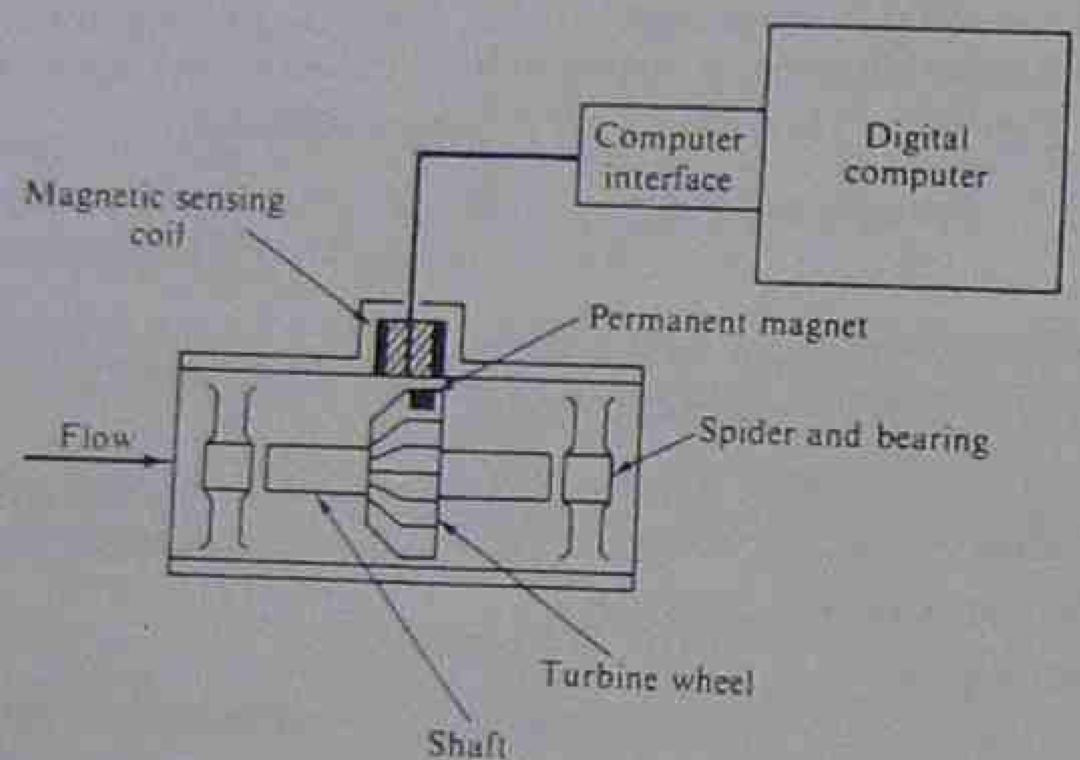


Figure 10.7 A turbine flow meter produces an accurate linear, digital flow signal. Turbine meters are used in the petrochemical and other industries for a broad range of applications.

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.

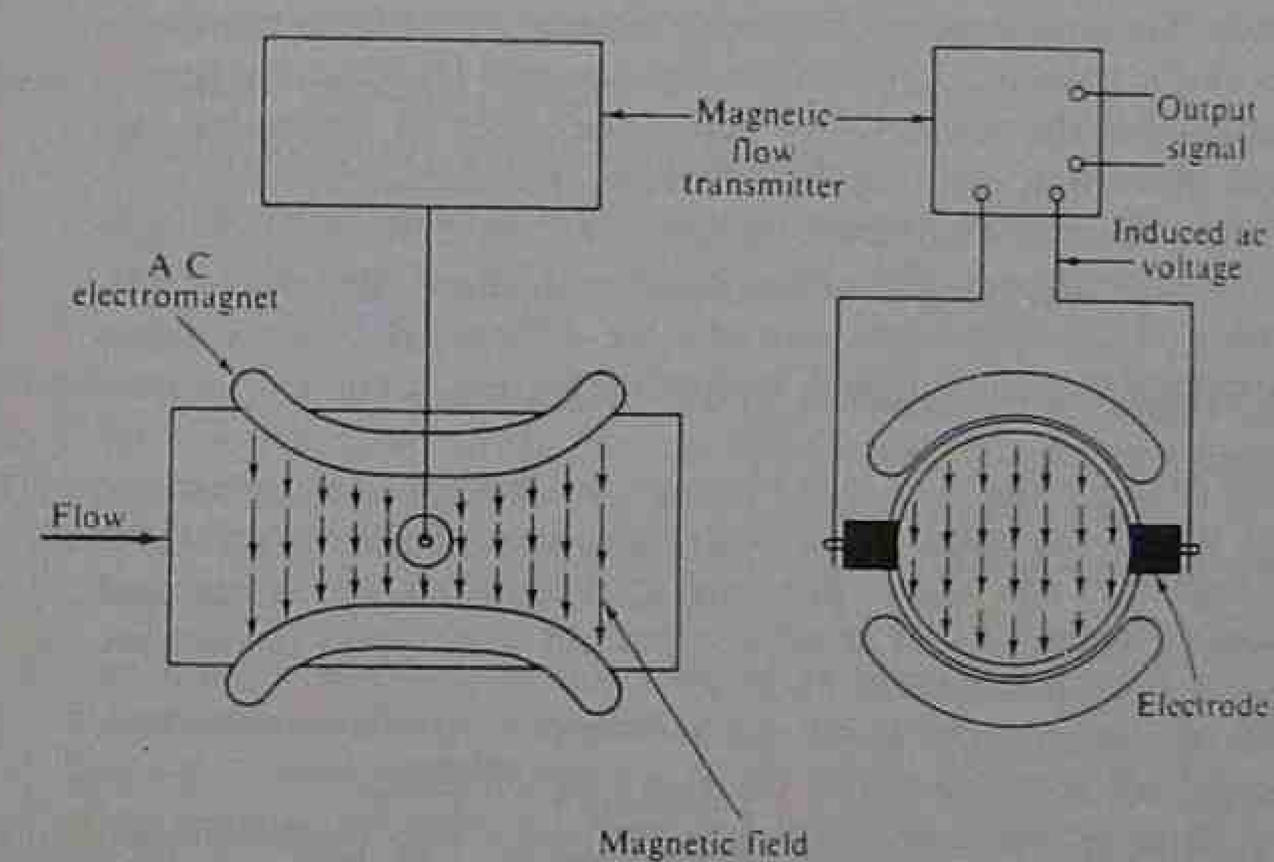


Figure 10.9 A magnetic flow meter has a completely unobstructed flow path, a decided advantage for slurries and food products.

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.

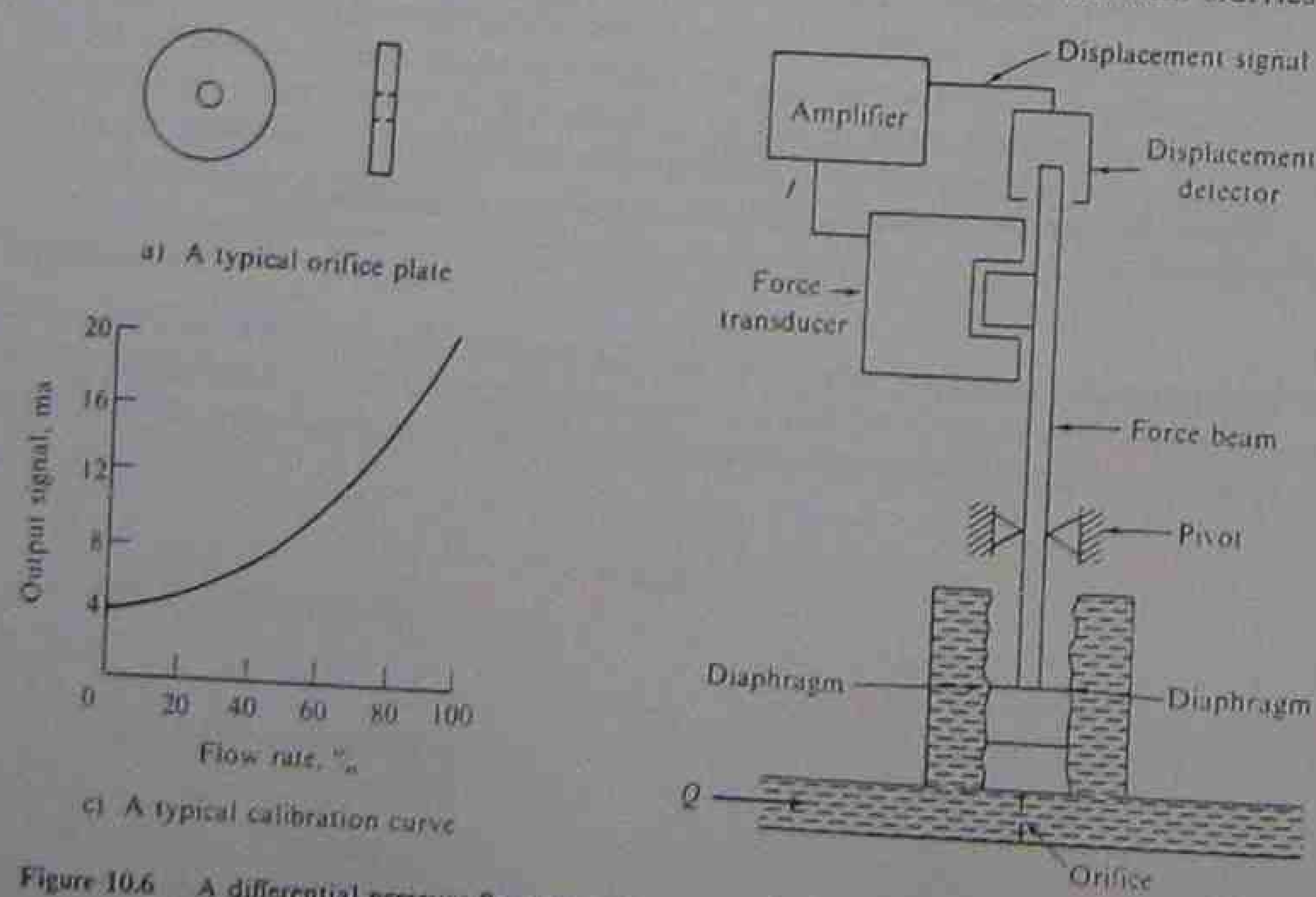
$$q = K\sqrt{h} \tag{10.1}$$

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.



a) A typical orifice plate

c) A typical calibration curve

b) A typical differential pressure flow transducer

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

$$p = \rho g h \qquad (10.4)$$

where p = static pressure, pascal
ρ = liquid density, kilogram/cubic meter
h = height of liquid above the measurement point, meter
$g = 9.81\ m/s^2$

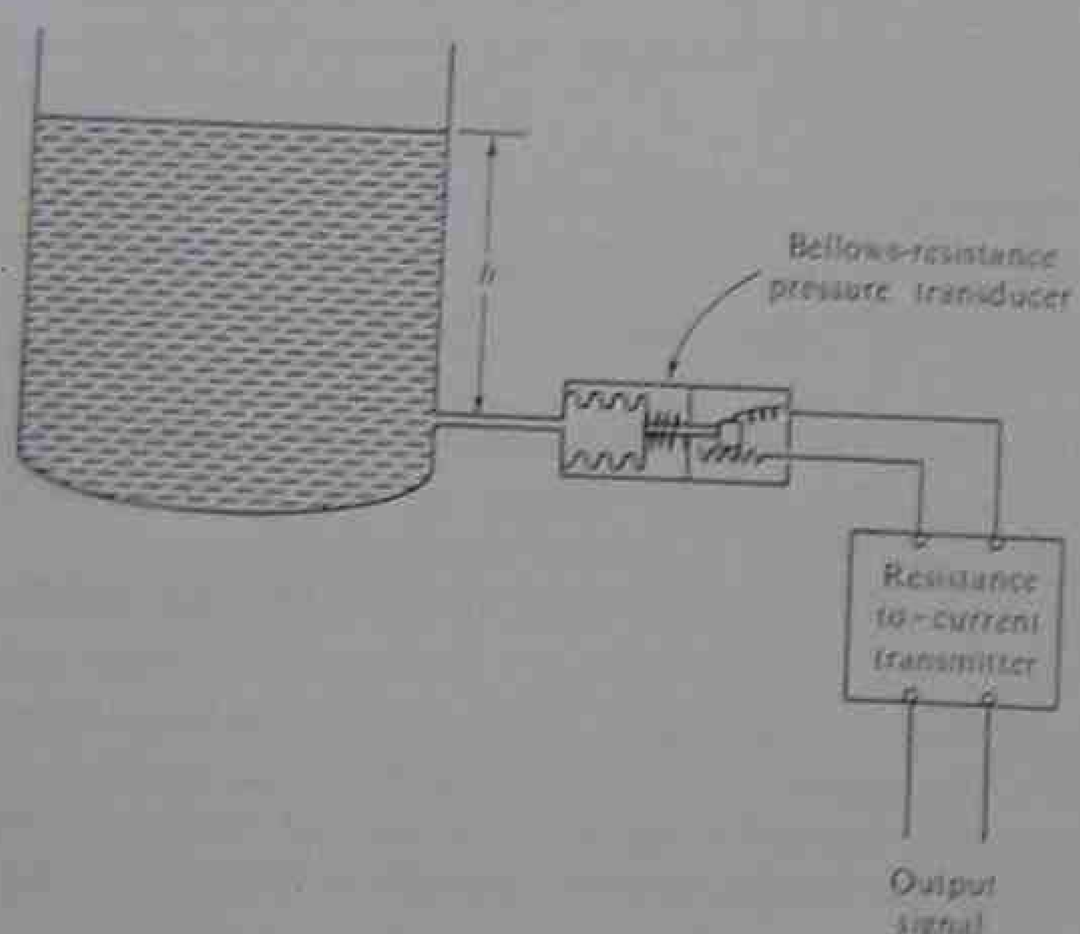


Figure 10.14 A static pressure sensor uses the pressure near the bottom of the tank as a measure of the liquid level.

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:

$$f = Mg - \rho g A h \qquad (10.5)$$

where f = net force, newton
M = mass of the float, kilogram
$g = 9.81\ m/s^2$ (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

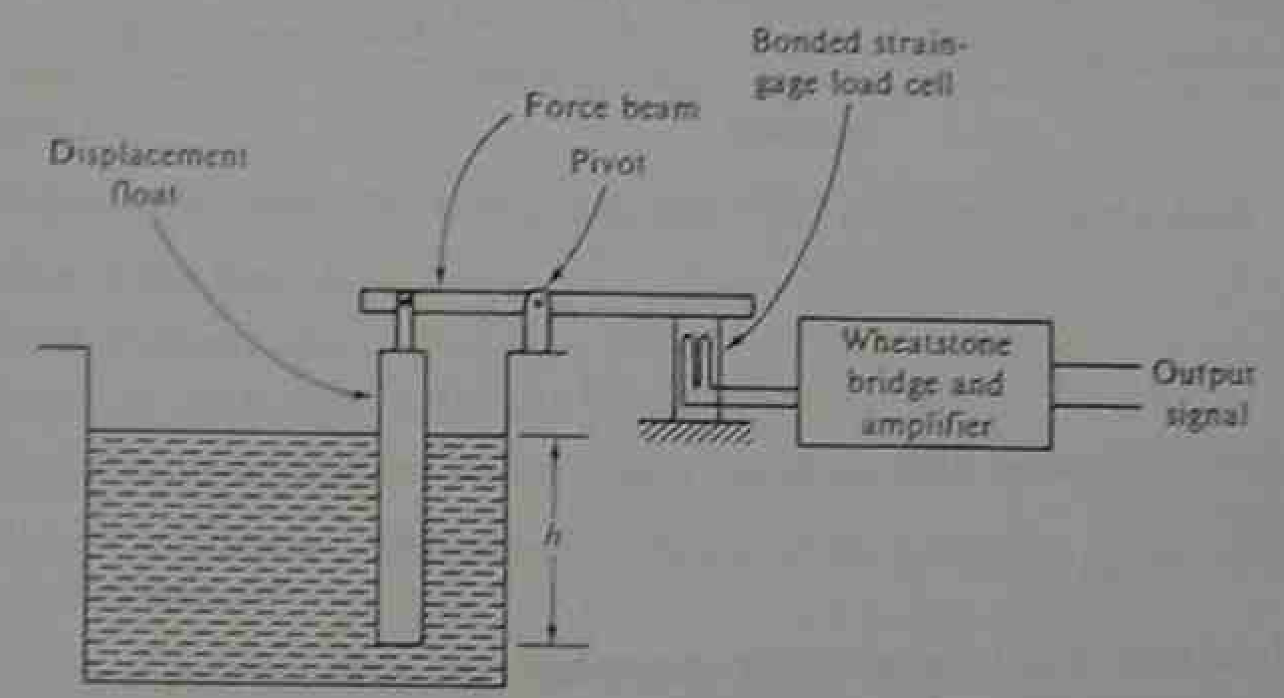


Figure 10.13 A float and a force or displacement sensor use the buoyant force on the float as a measure of the liquid level.

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

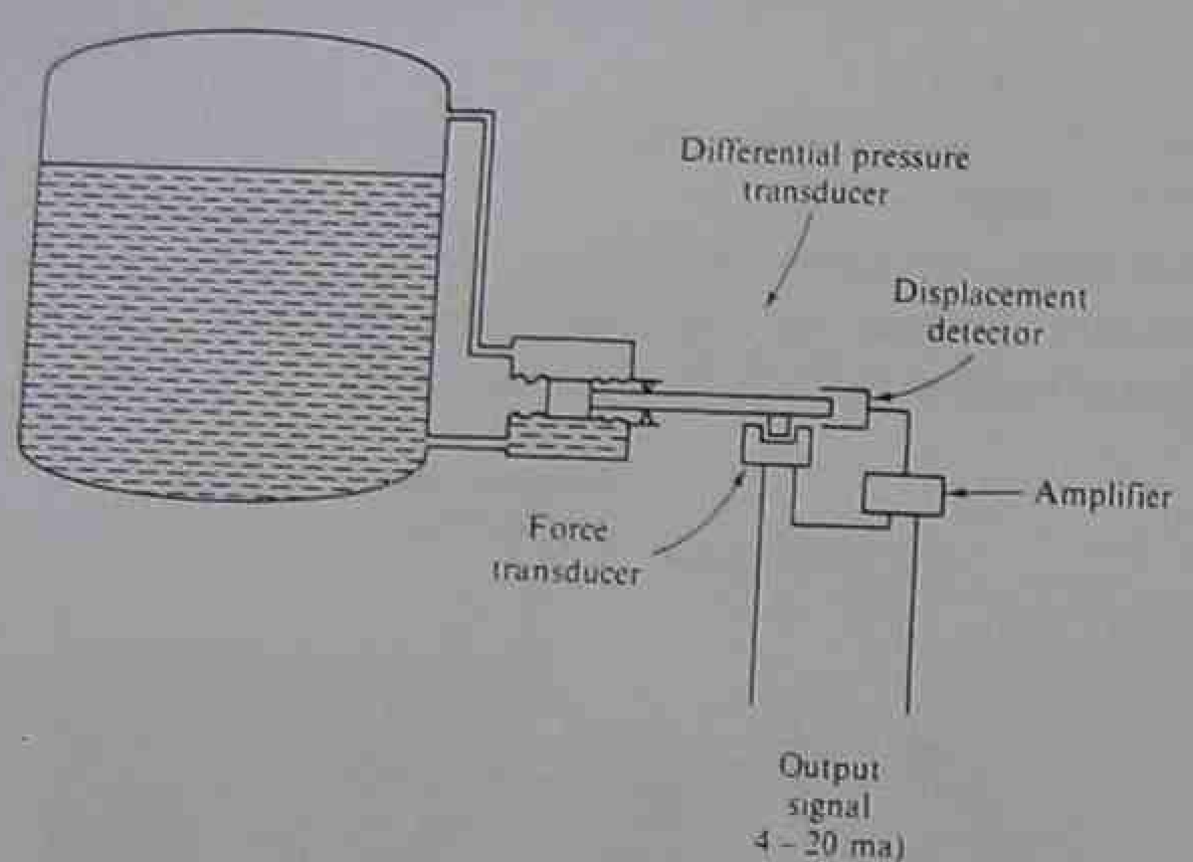


Figure 10.15 A differential pressure sensor can be used to measure liquid level in a closed tank that is under high pressure.

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.

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

$$p = \rho g h$$

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

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

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 cm^3 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 cm^2
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 cm^2
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/cm^2?

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 cm^2
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 $\frac{\text{kilogram}}{\text{meter}^3}$	Absolute Viscosity pascal second	Thermal Conductivity $\frac{\text{watt}}{\text{meter kelvin}}$	Heat Capacity $\frac{\text{joule}}{\text{kg kelvin}}$
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	2,070
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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Theory Notes

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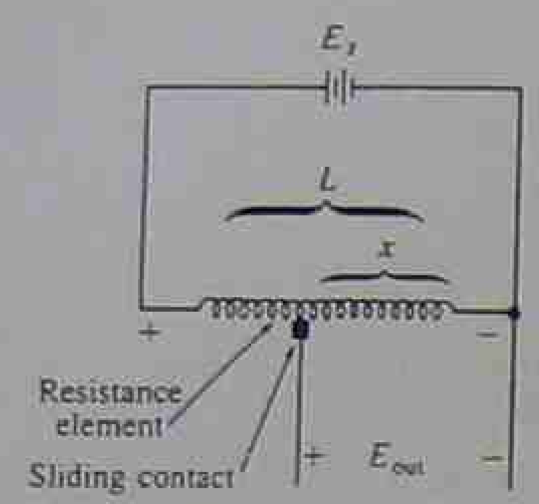
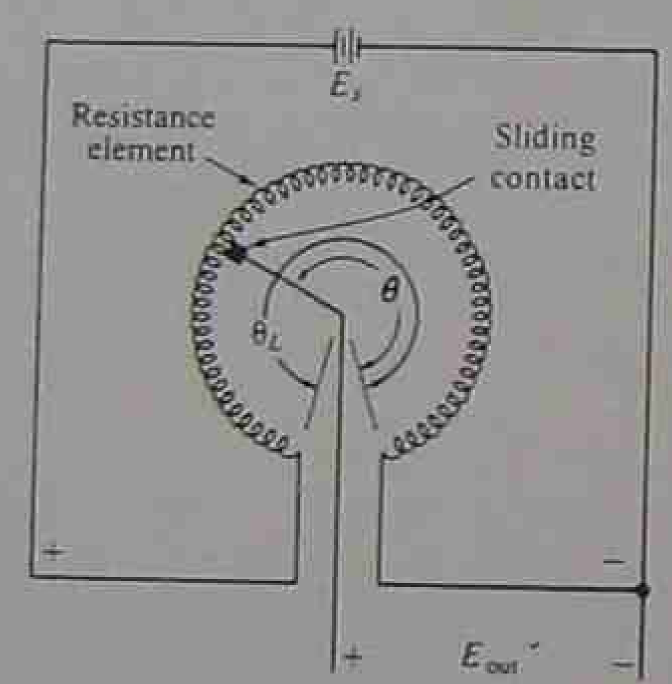
Potentiometers

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:

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

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

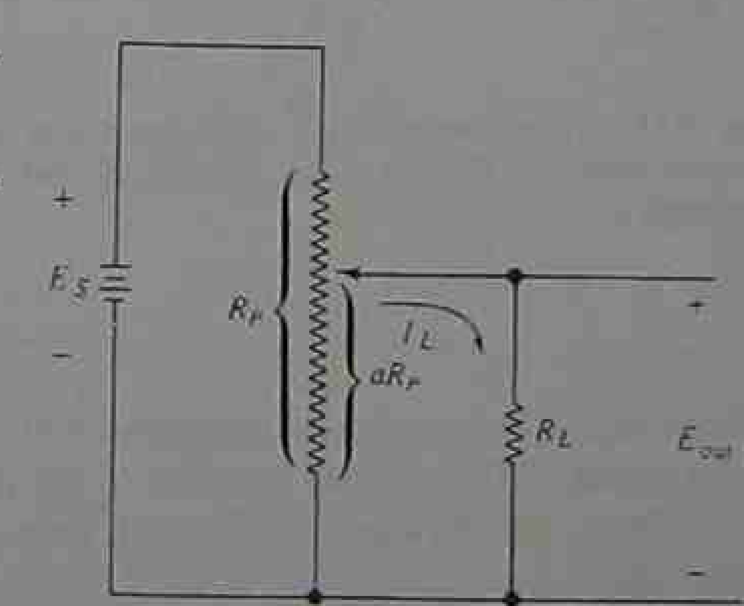
where

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

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.

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

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.

$$E_1 = KE_3 \cos\theta \tag{9.7}$$
$$E_2 = KE_3 \sin\theta \tag{9.8}$$

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:

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

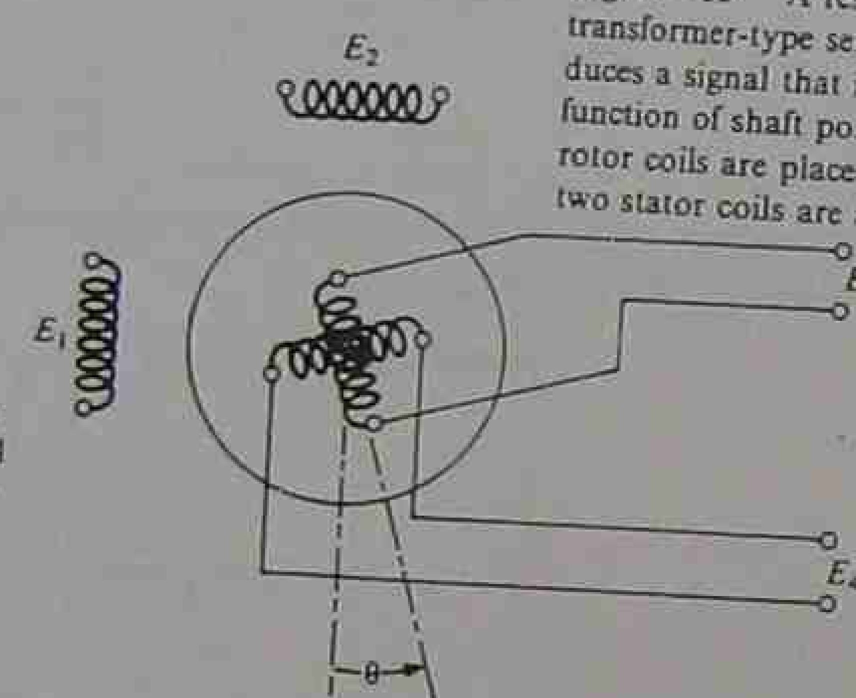


Figure 9.10 A resolver is a transformer-type sensor that produces a signal that is a trigonometric function of shaft position, θ. The two rotor coils are placed 90° apart. The two stator coils are also 90° apart.

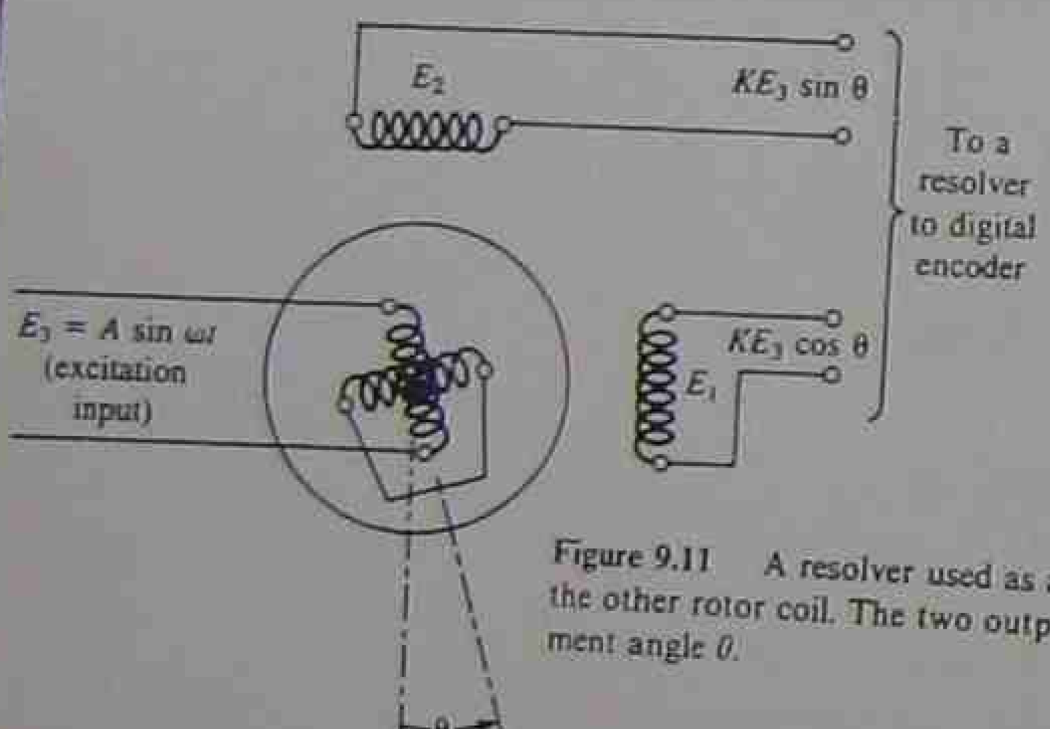


Figure 9.11 A resolver used as a sensor has one of its rotor coils shorted. An ac voltage is applied to the other rotor coil. The two output voltages, E_1 and E_2, are trigonometric functions of the displacement angle θ.

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,

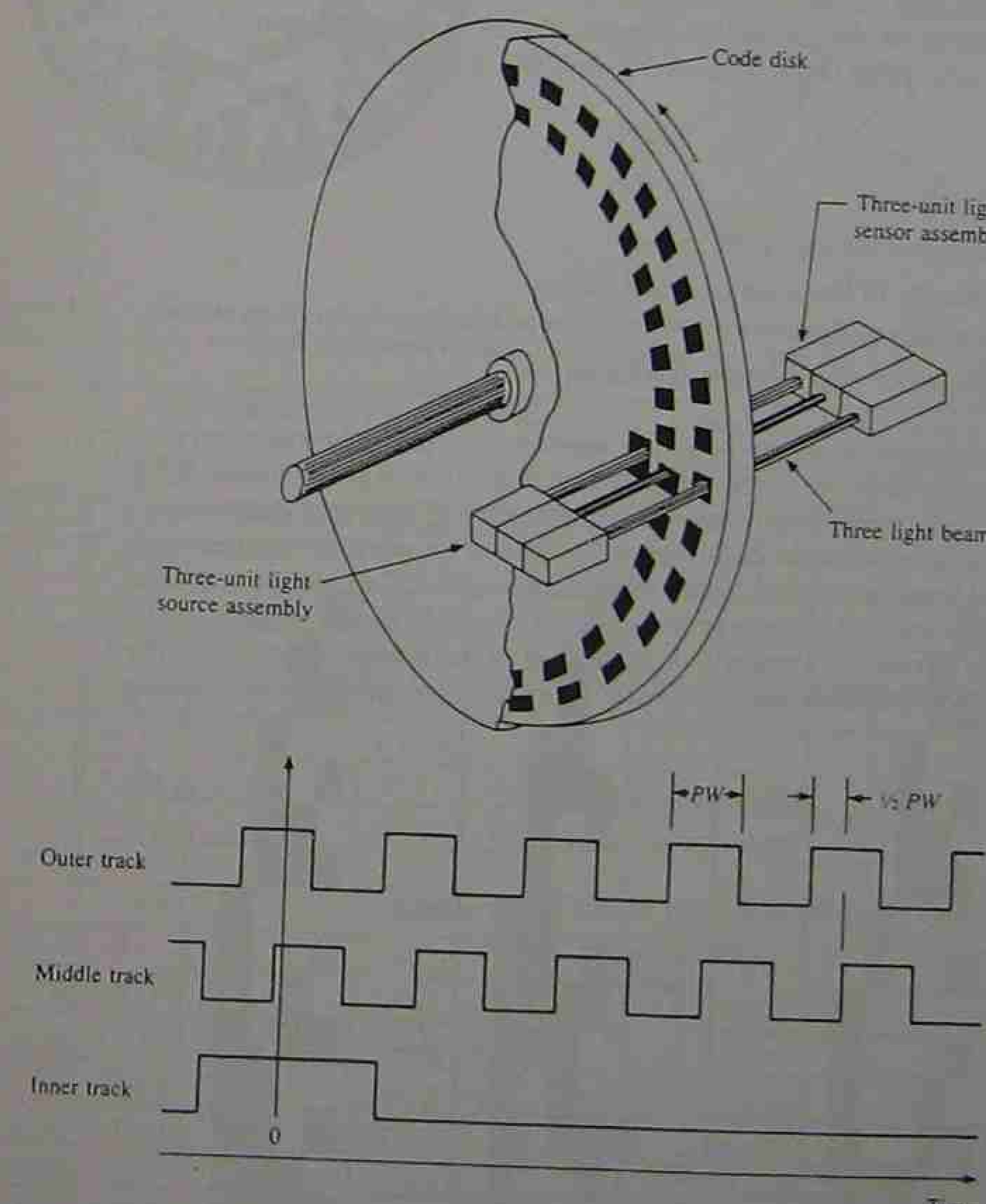


Figure 9.12 The incremental encoder has three tracks. The inner track provides a reference signal to locate the home position. The middle track provides information about the direction of rotation. In one direction, the middle track lags the outside track; in the other direction, the middle track leads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

$$x = \frac{\pi d N_T}{N_R} \tag{9.9}$$

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:

$$\text{Number of positions} = 2^N \tag{9.10}$$

$$\text{Resolution} = 1 \text{ part in } 2^N \tag{9.11}$$

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

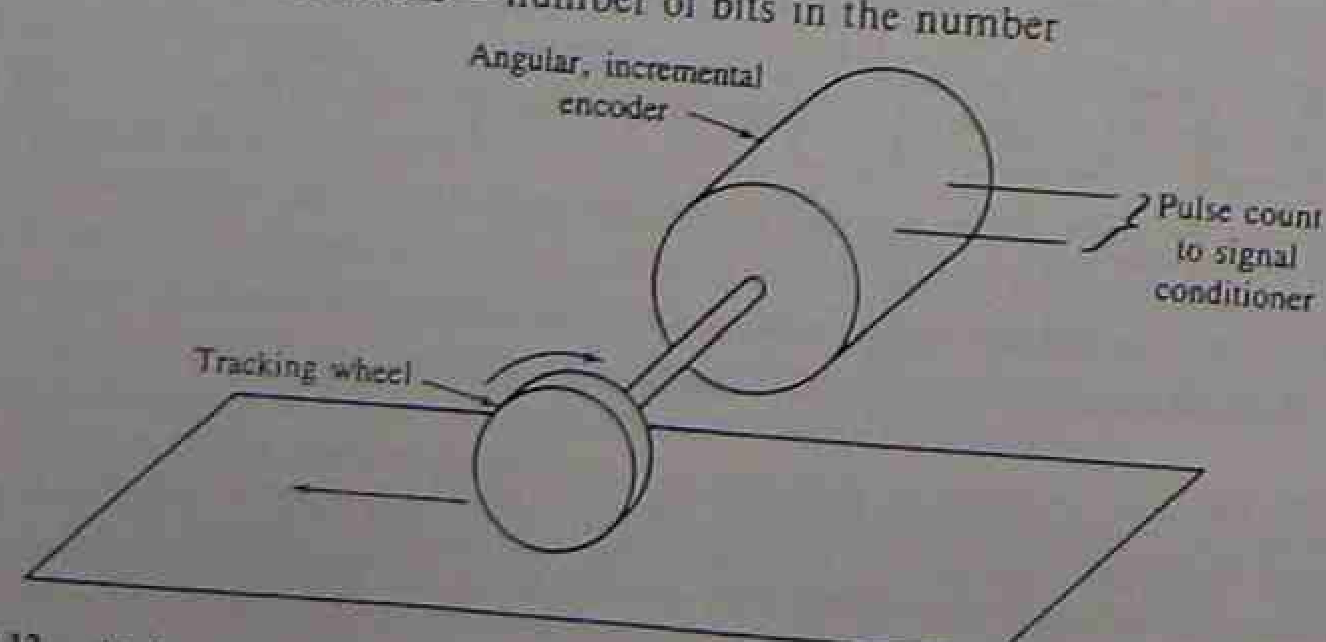


Figure 9.13 An incremental encoder coupled to a tracking wheel is used to measure linear displacement.

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.

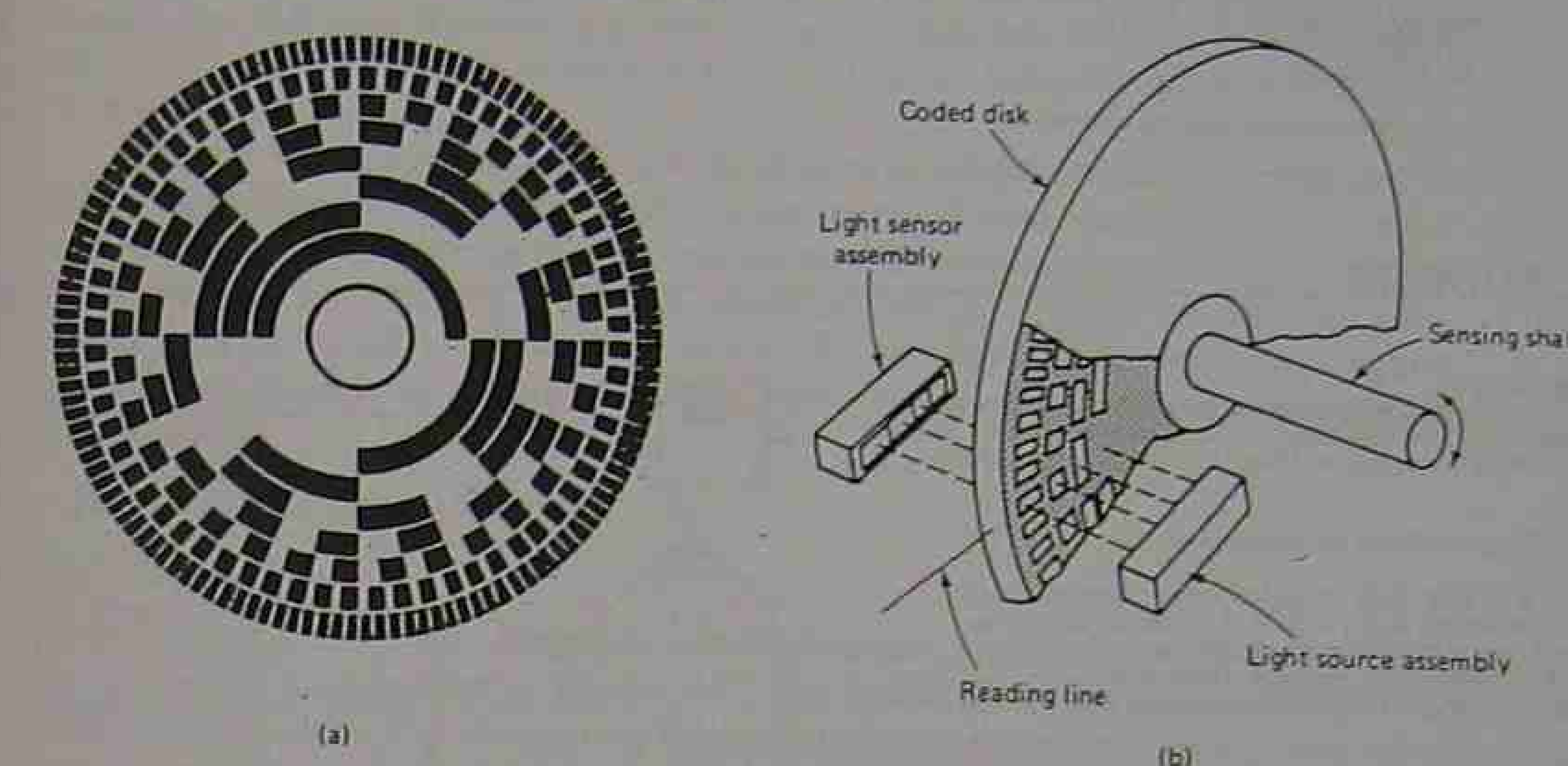
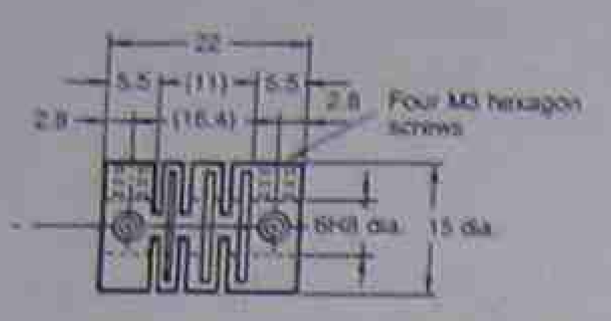


Figure 9.14 Absolute optical encoder: (a) typical code disk; (b) encoder elements. [From H. Norton, *Sensor and Analyzer Handbook* (Englewood Cliffs, N.J.: Prentice-Hall, Inc., 1982), Fig. 1-44, p. 107.]

Arabic number	(Natural) Binary		Gray (Binary)		Binary Coded Decimal (BCD)			
	Digital number	Code pattern	Digital number	Code pattern	Digital number		Code pattern	
					Tens	Units	Tens	Units
	8 4 2 1	2^3 2^2 2^1 2^0		G_3 G_2 G_1 G_0	8 4 2 1	8 4 2 1	2^0	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.

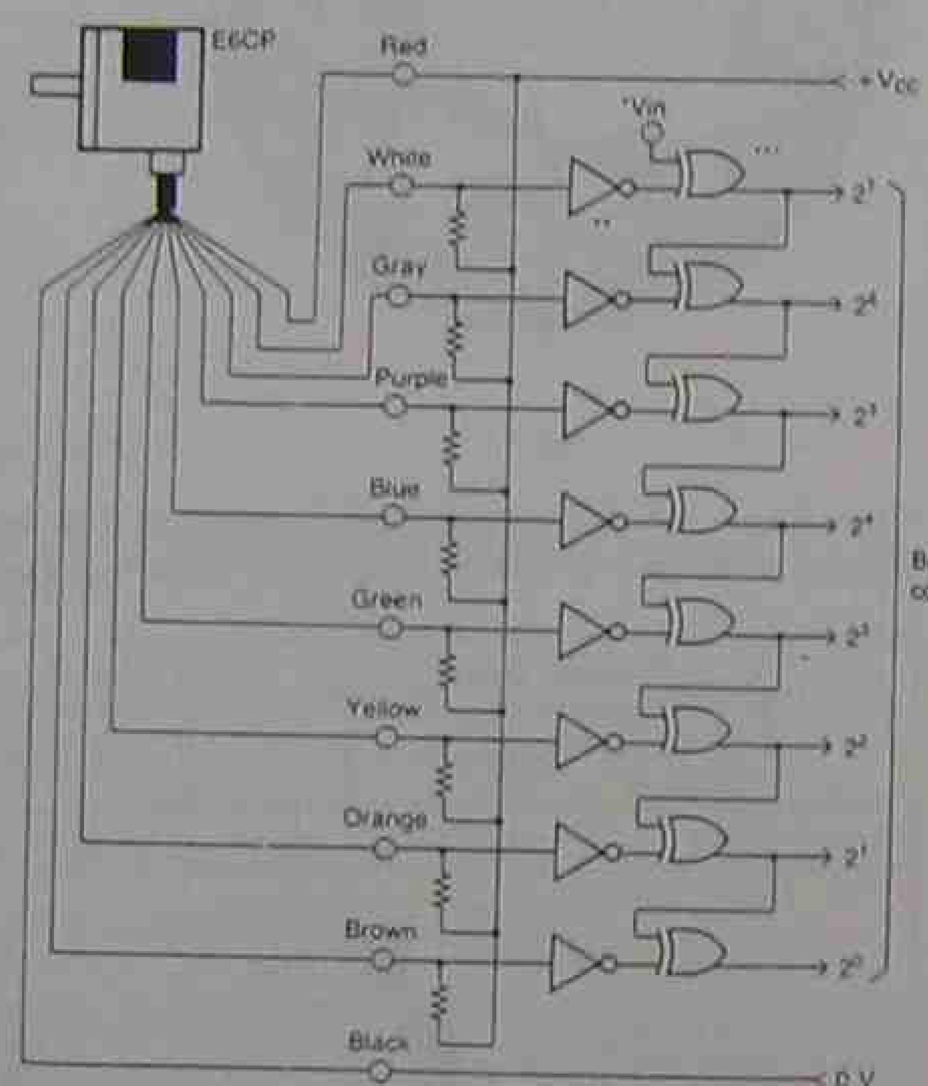
Reference (Gray-to-binary converter circuit)
Binary code
Binary code is a basic code for digital signal processing and consists of numerals 0 and 1 only. It is, however, difficult to change two or more digits simultaneously when a number represented by binary code changes. Consequently, the reading timing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

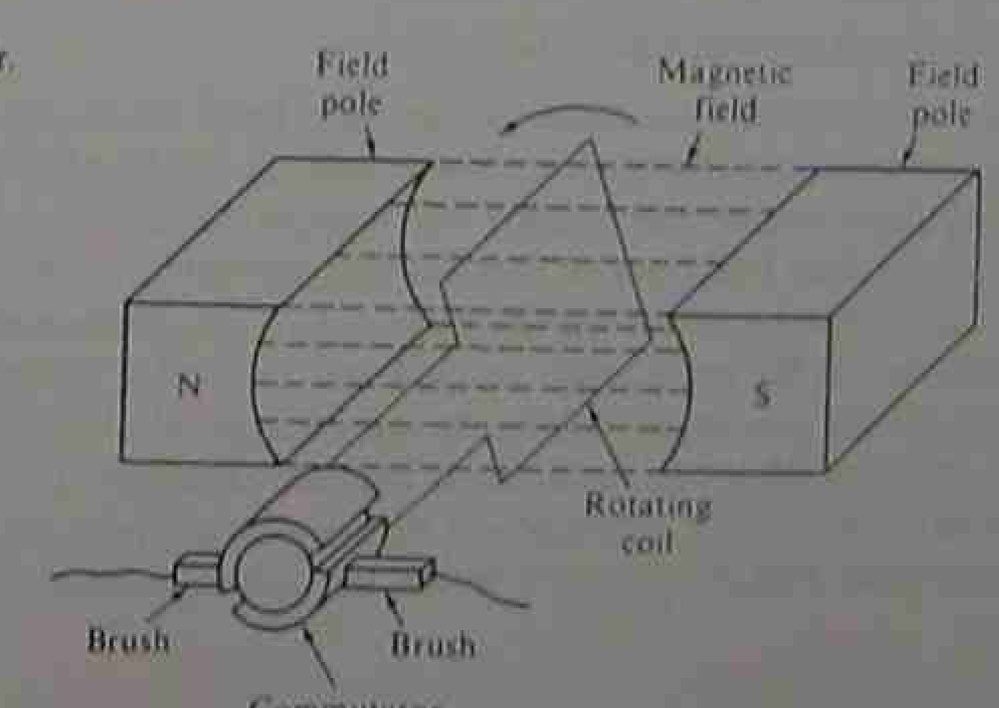
DC Tachometers

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.

$$E = K_E S = \frac{30 K_E \omega}{\pi} \tag{9.12}$$

$$K_E = \frac{2\pi RBNL}{60} \tag{9.13}$$

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.

$$S = \frac{60C}{NT_c} \tag{9.14}$$

$$C = \frac{SNT_c}{60} \tag{9.15}$$

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.

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

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

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

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:

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

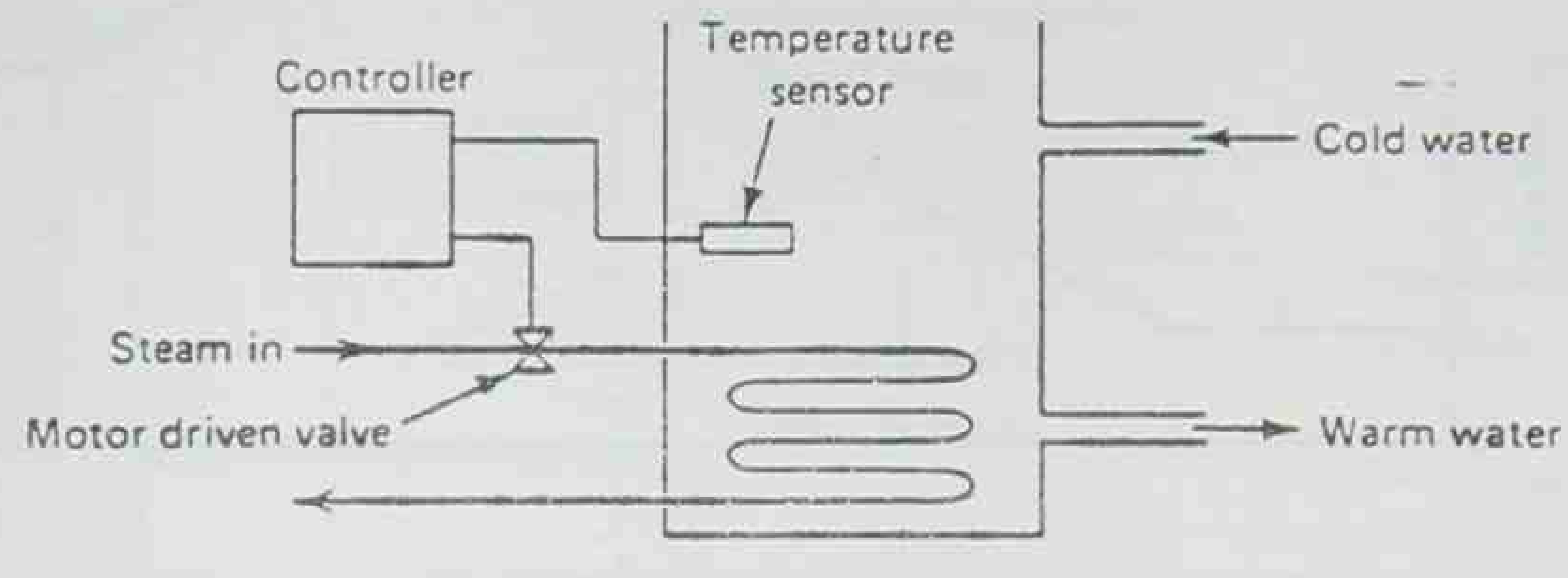
b. Equation (9.14) applies:

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

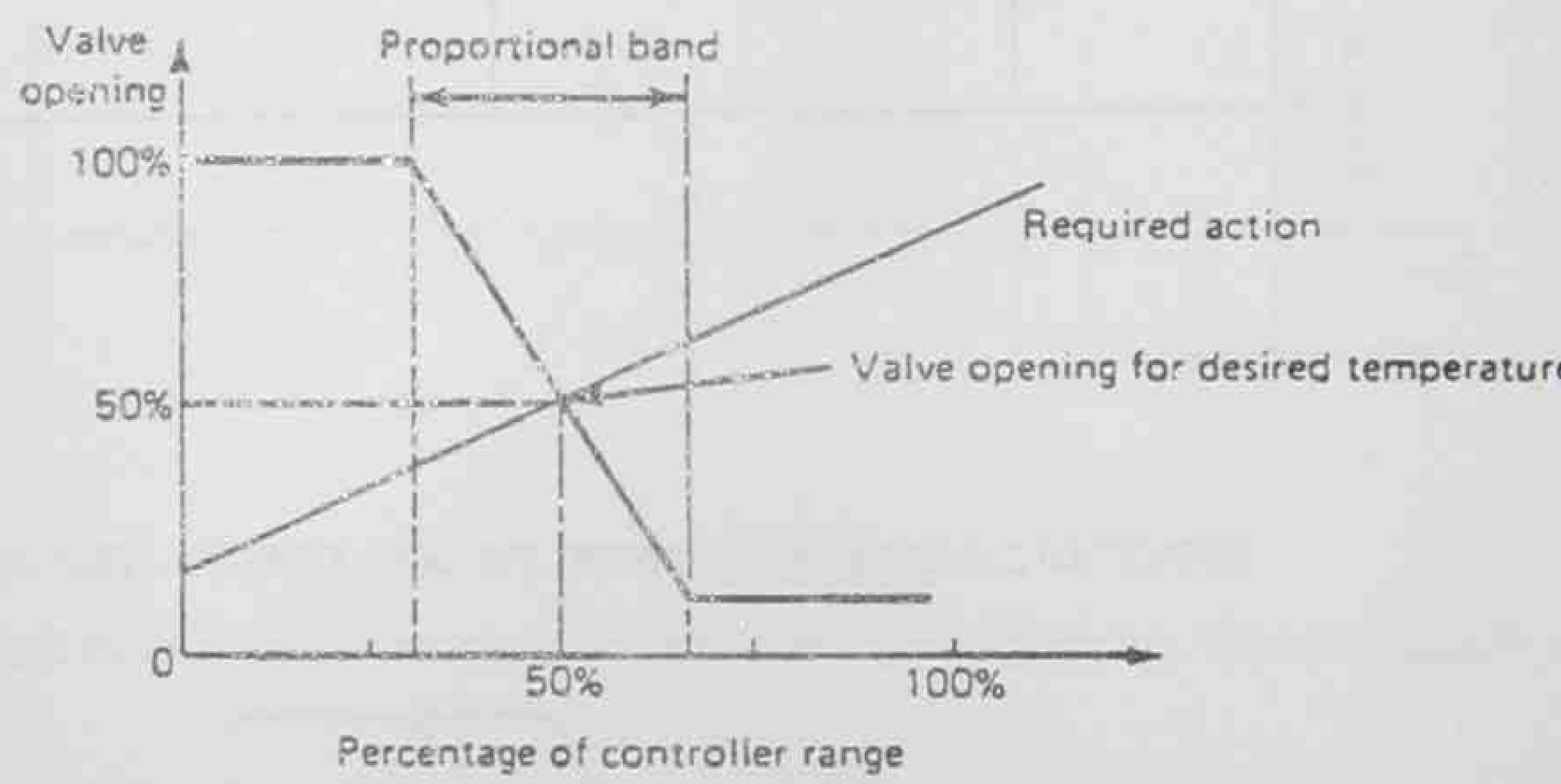
Proportional control

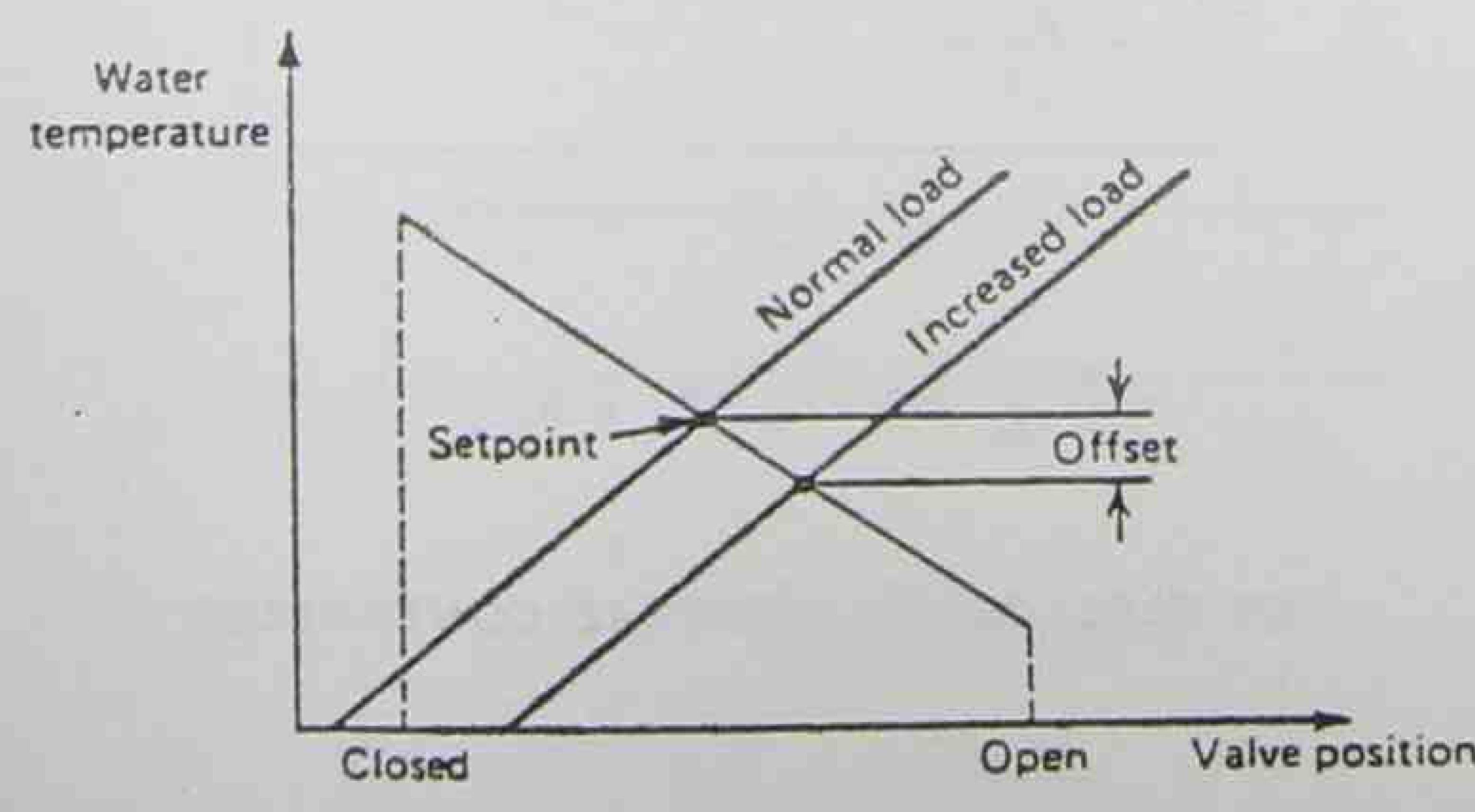
PROPORTIONAL control produces a change in the controller output that is proportional to the error signal. There is a fixed linear relationship between the controlled variable and the position of the final control element.

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.



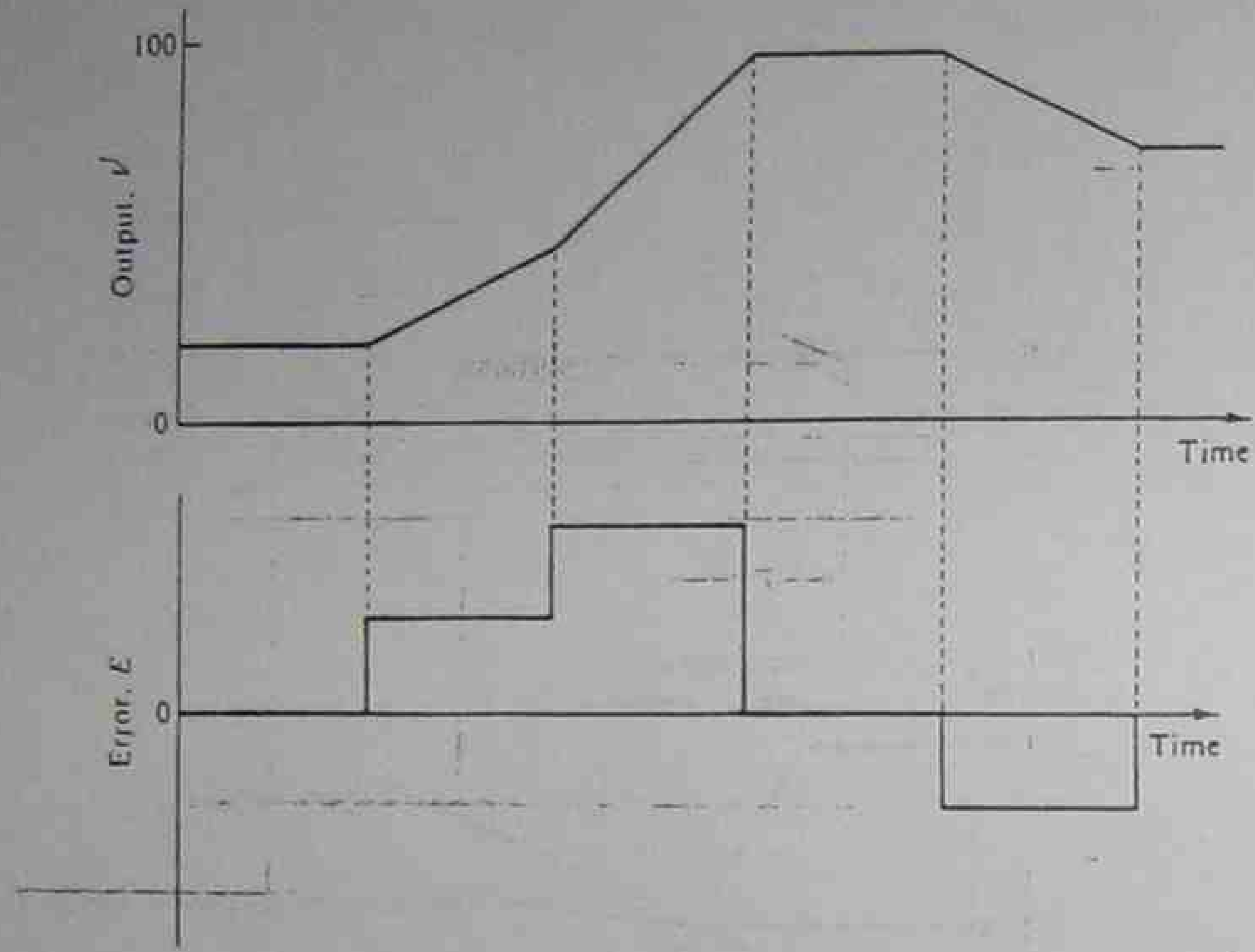
(a)



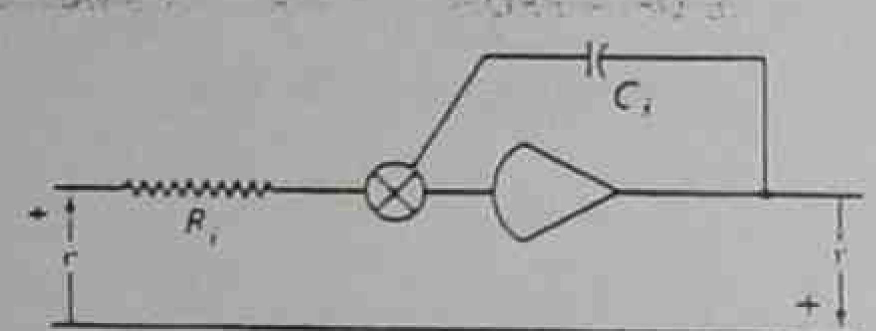


(b)

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

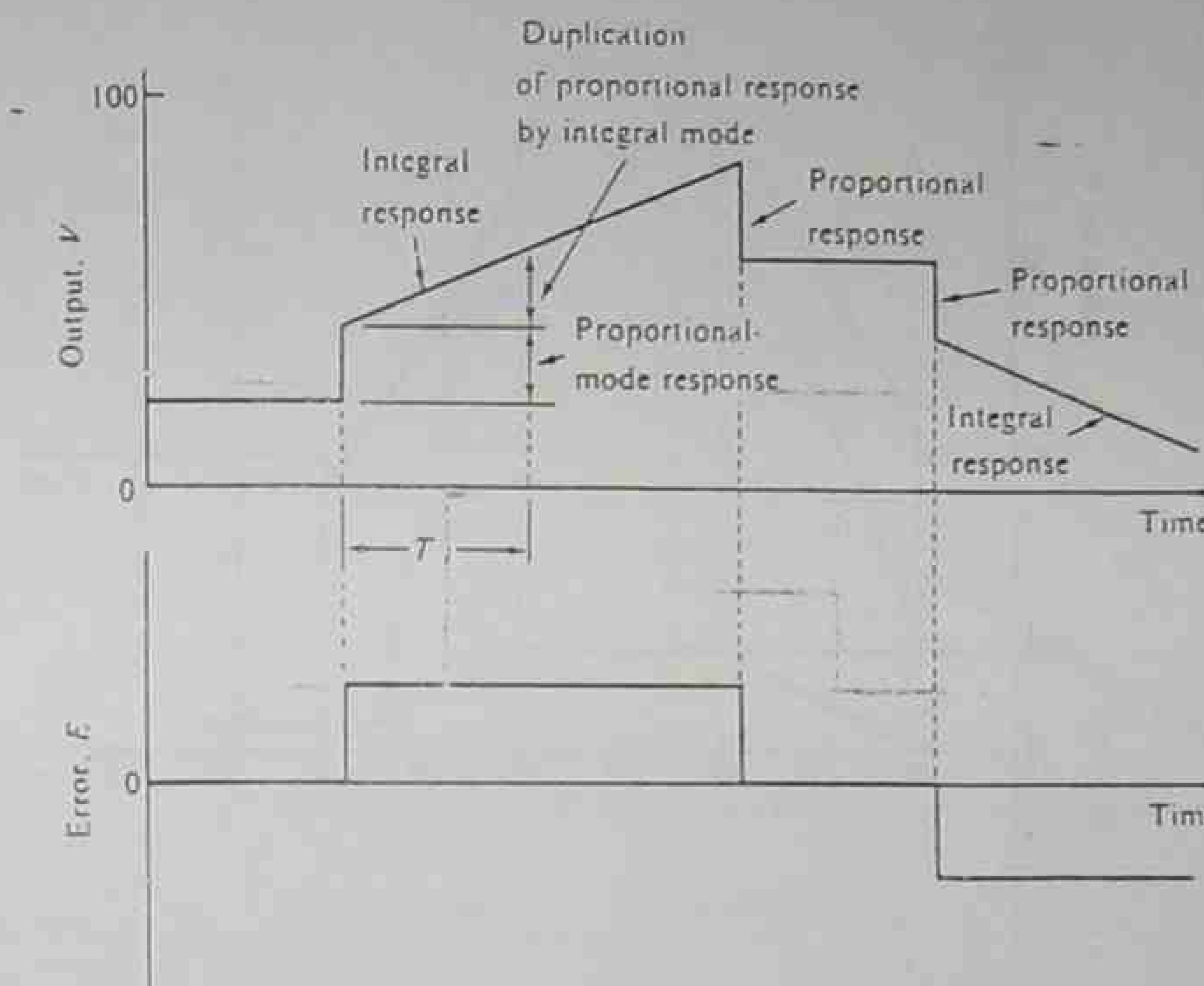


$$r = \frac{1}{T_i}\int_0^t e\,dt + r_0 ;\ T_i = R_iC_i$$

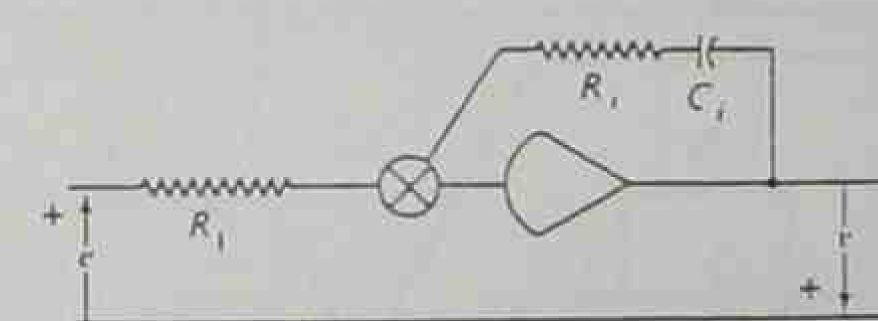
AN ELECTRONIC INTEGRAL CONTROLLER

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

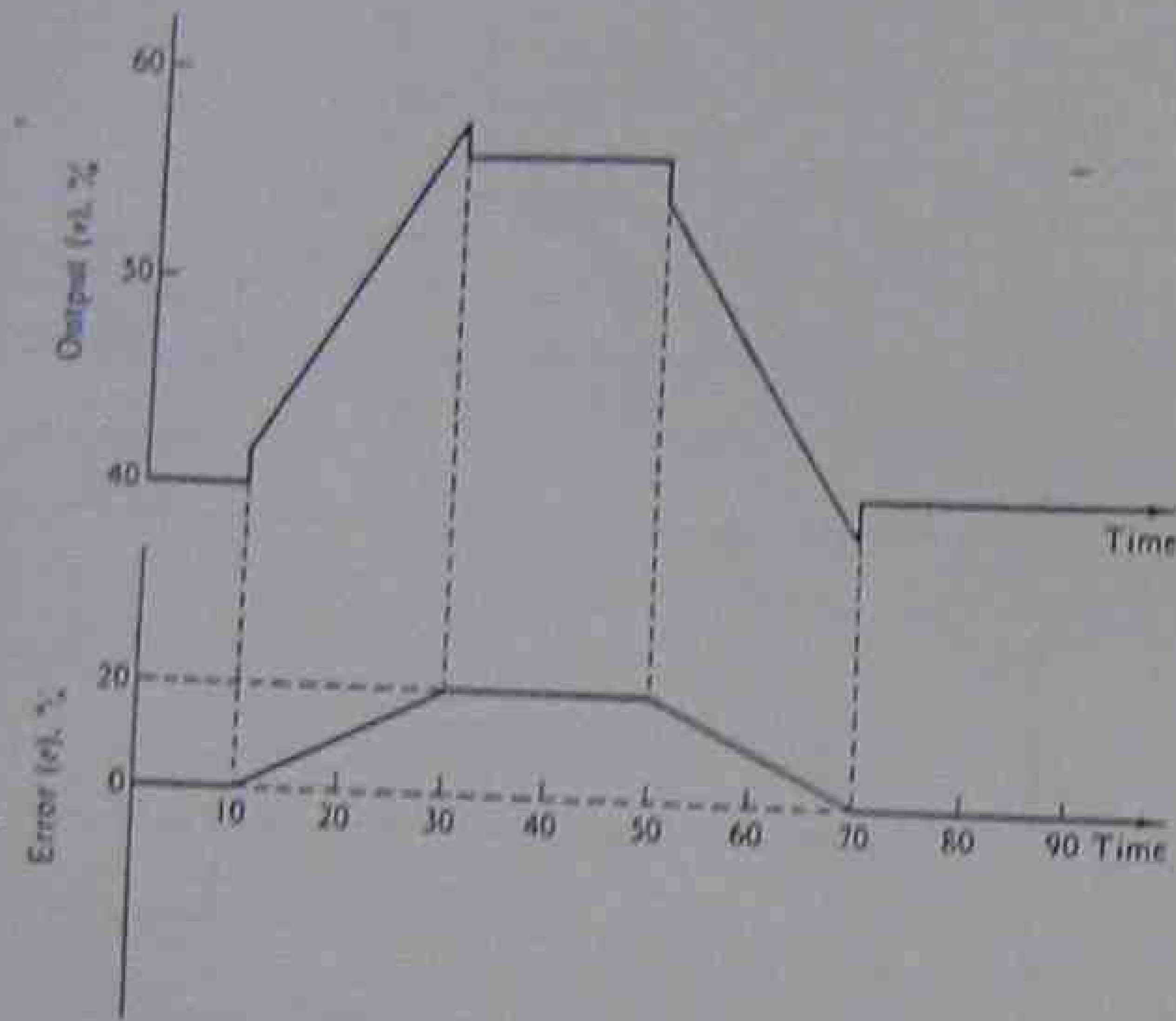


$$r = Ke + \frac{K}{T_i}\int_0^t e\,dt + r_0$$

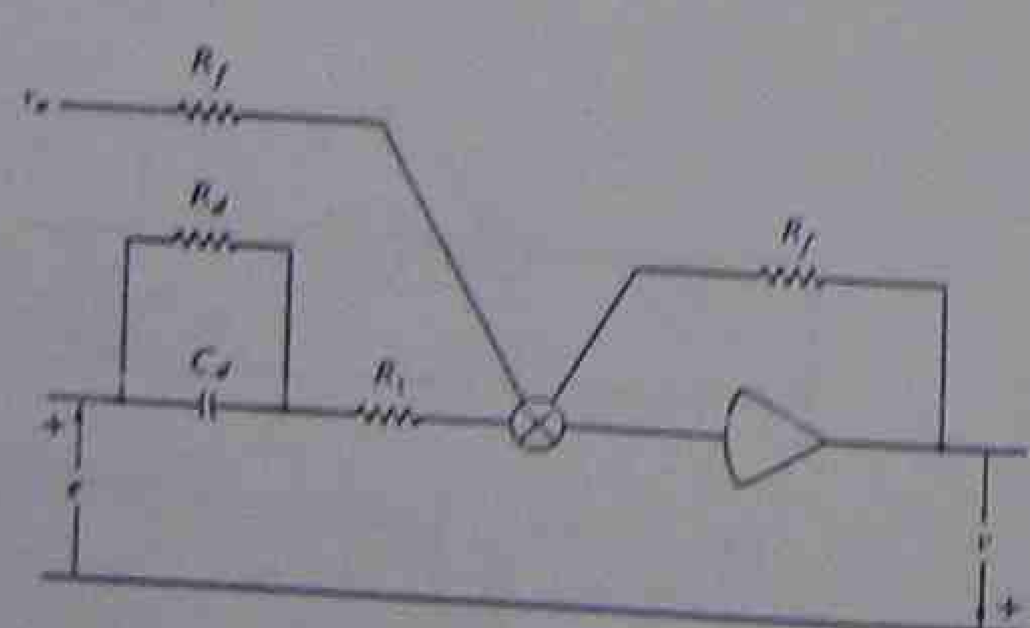
$$T_i = R_iC_i ;\ K = R_i/R_1$$

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.



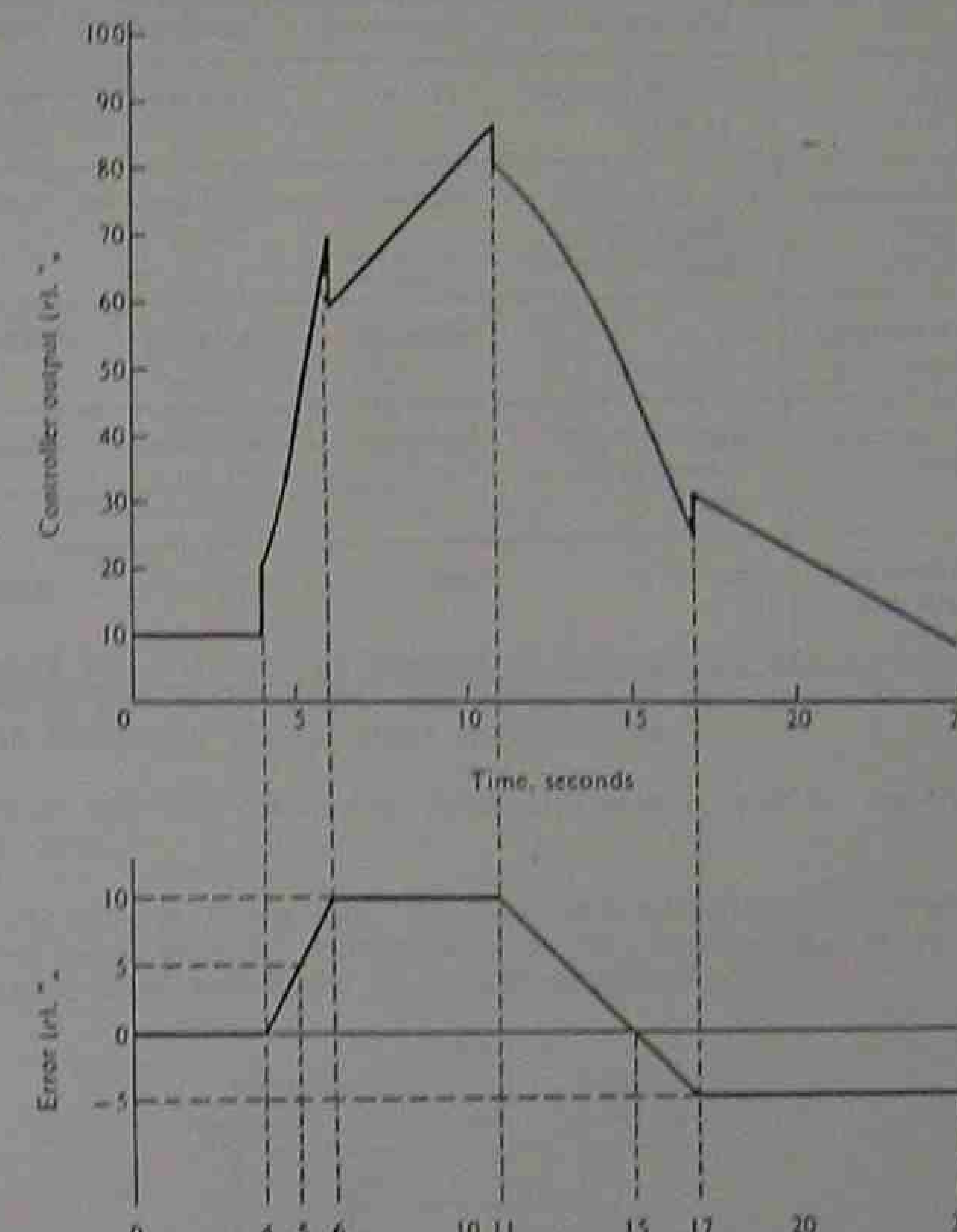
ERROR AND CONTROLLER OUTPUT GRAPHS

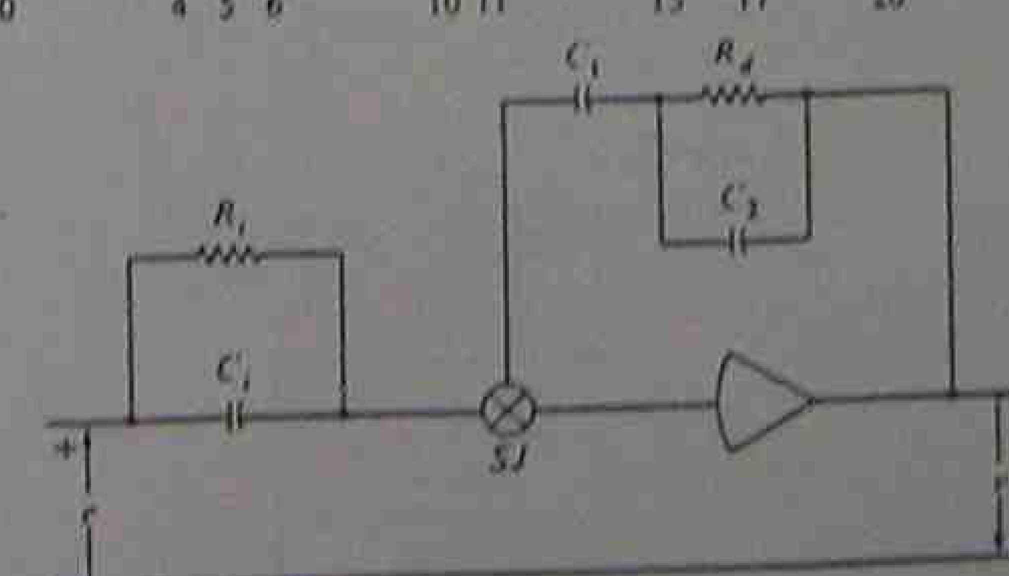


AN ELECTRONIC PROPORTIONAL-PLUS-DERIVATIVE CONTROLLER

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

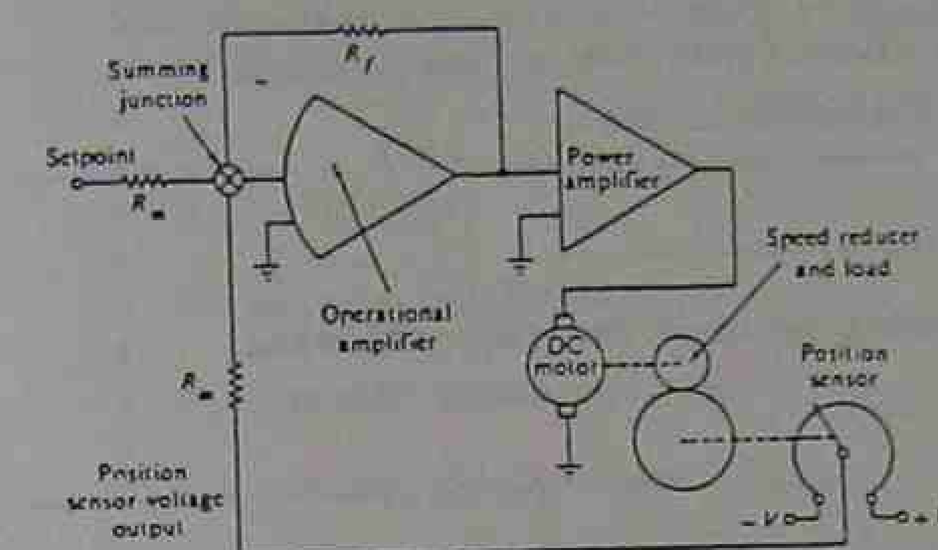
Control mode	Process reaction delay (minimum)	Transfer lag (maximum)	Dead time (maximum)	Size of load disturbance (maximum)	Speed of load disturbance (maximum)
On-Off	Long only (cannot be short)	Very short	Very short	Small	Slow
Proportional only	Long or moderate (cannot be too short)	Moderate	Moderate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

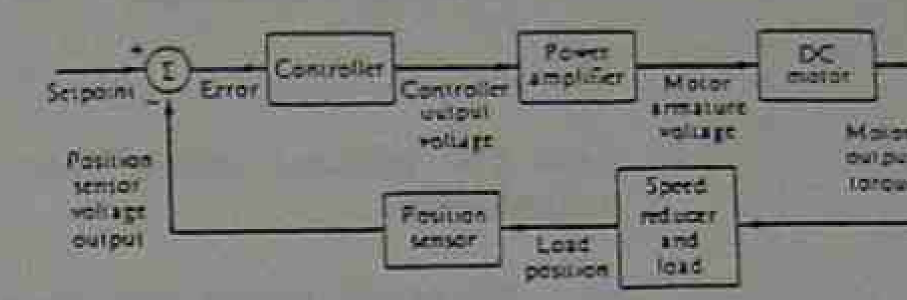
The signals in a control system can be divided into two general categories, Analog or Digital.

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



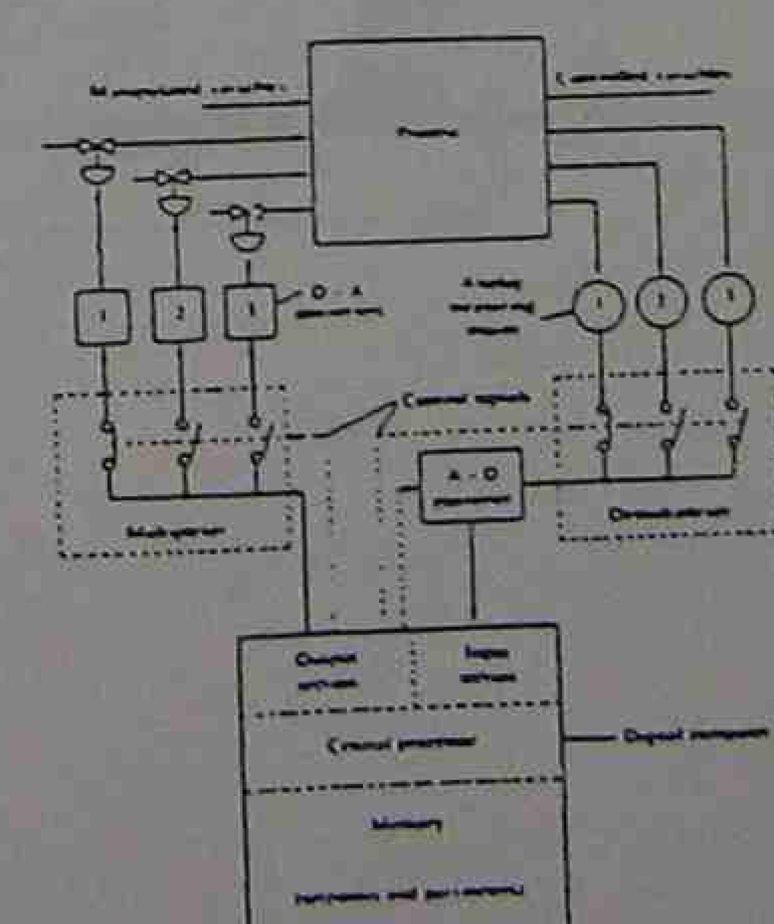
b) Block diagram

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) ________

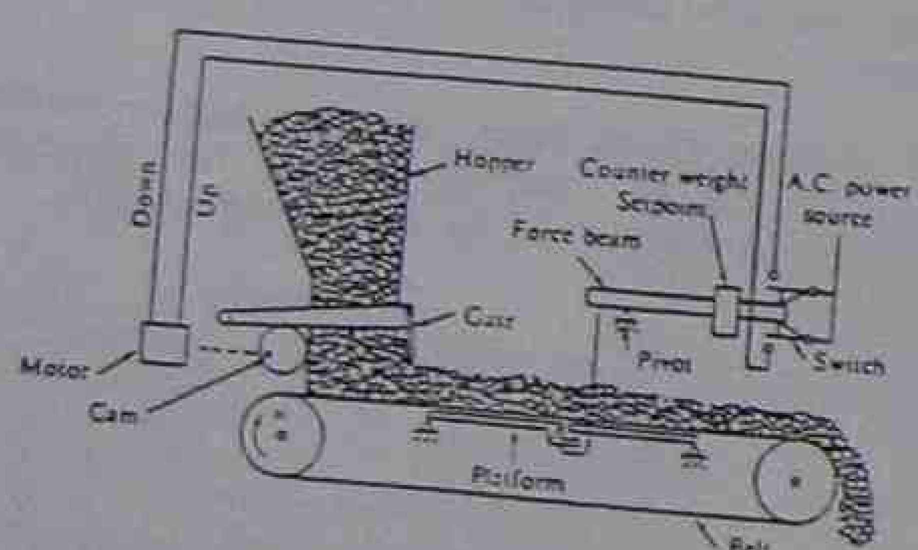
b) List two control modes that could be used for the following situations.

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

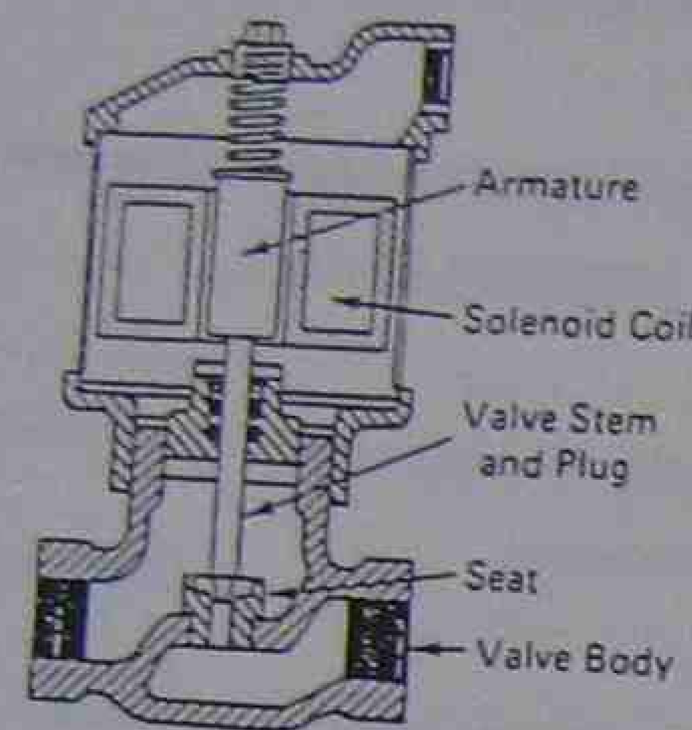
COURSE : ADVANCED CERTIFICATE of INDUSTRIAL ELECTRONICS (6016)
STAGE : 2 (ELECTIVE)
SUBJECT : *INDUSTRIAL CONTROL* (6016K)
WEEK : 3
TOPIC : FINAL CORRECTING I

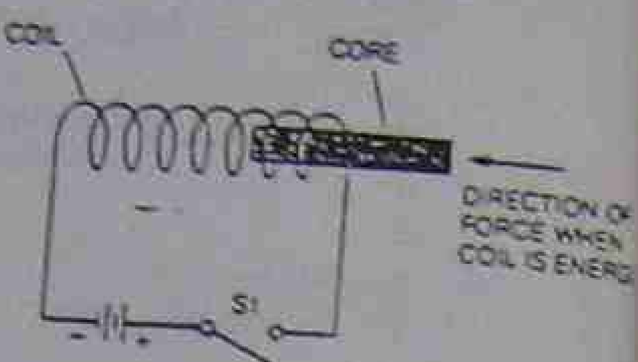
FINAL CORRECTING DEVICES I

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.

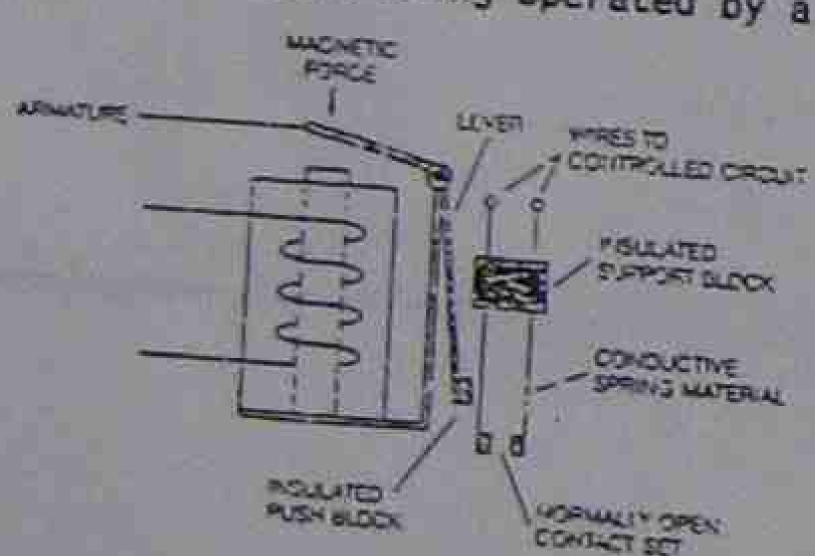




RELAYS AND CONTACTORS

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.

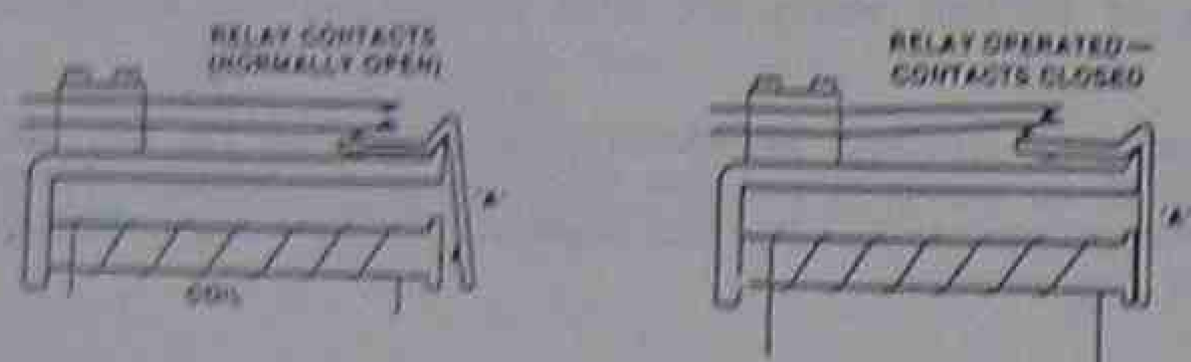


Figure 1. A relay with normally open contacts shown unoperated at left and operated at right. Note the over travel of the contact leaves. The armature is stopped by the core here.

Figure 2. A relay with normally closed contacts. Note that the contact leaves are preloaded and the plunger on the armature operates the upper contact leaf.

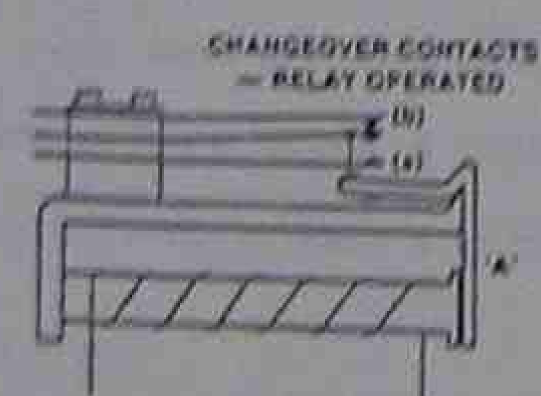


Figure 3. Relay with changeover contacts — one set of normally open contacts and one set of normally closed contacts. The armature is shown here in the operated position.

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 is 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.

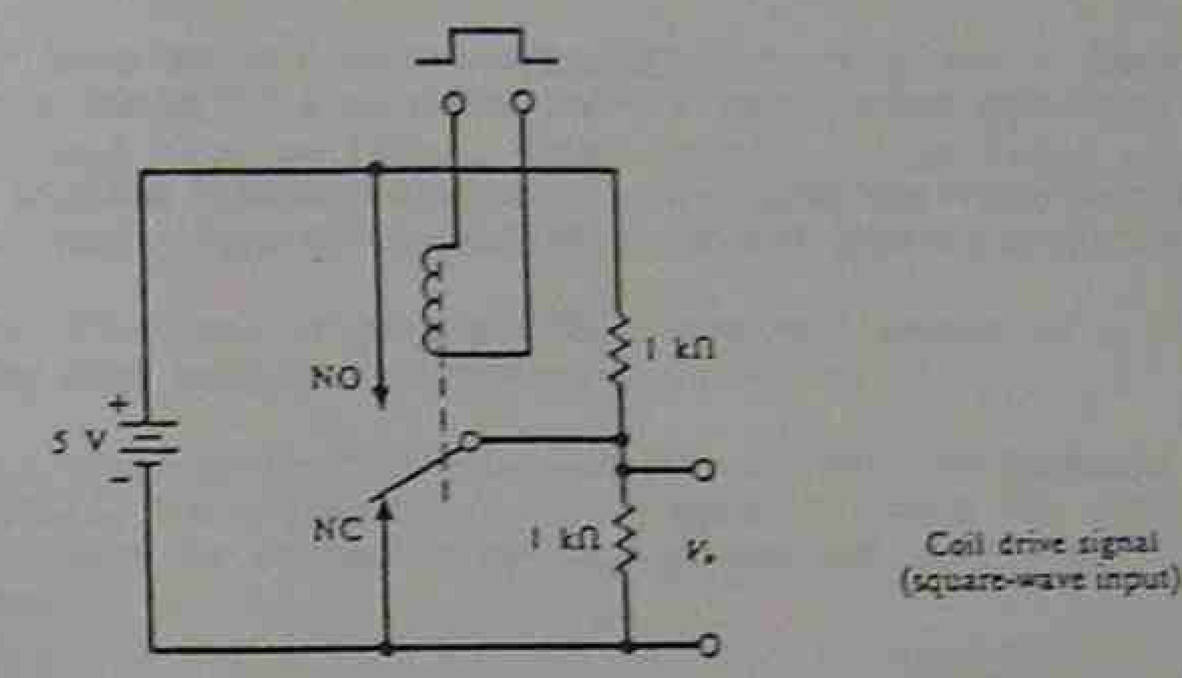
Improved protection can be obtained by using a diode with a resistor/capacitor. The diode should have a PIV of 800 volts, the capacitor should be approximately 1uf per ampere switched with a 400VW rating, the resistor should be approximately 100k per 2 Watts of load.

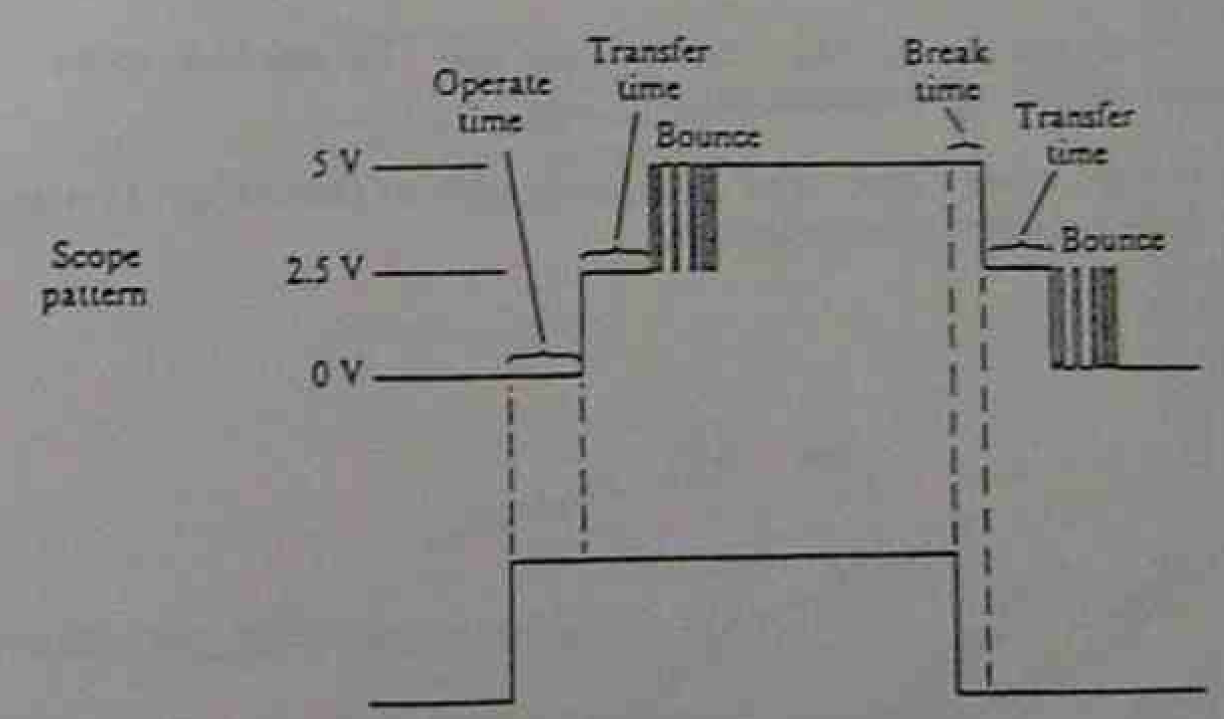
RELAY CONTACTS IN DIGITAL AND COUNTING CIRCUITS

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

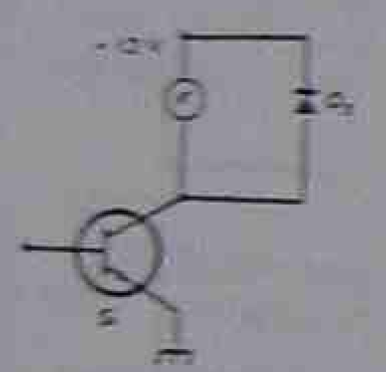
The circuitry that interfaces an inductive load such as a relay or coil to system via electronic components could be damaged by the back emf produced b the collapsing field when the inductive load is switched off.

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 involves connecting a reverse biased diode across the coil.

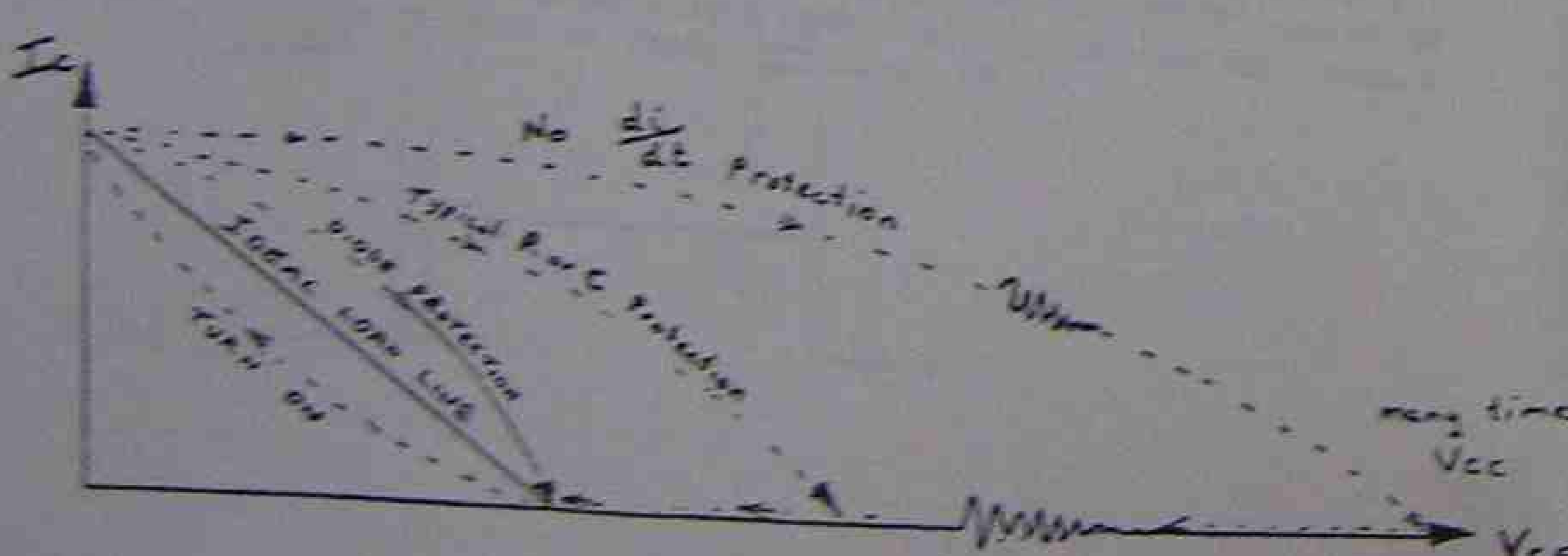


The advantages of diode protection are :-

. 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

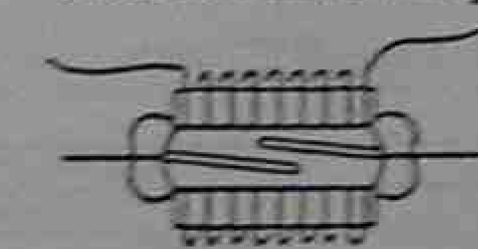
REED RELAYS

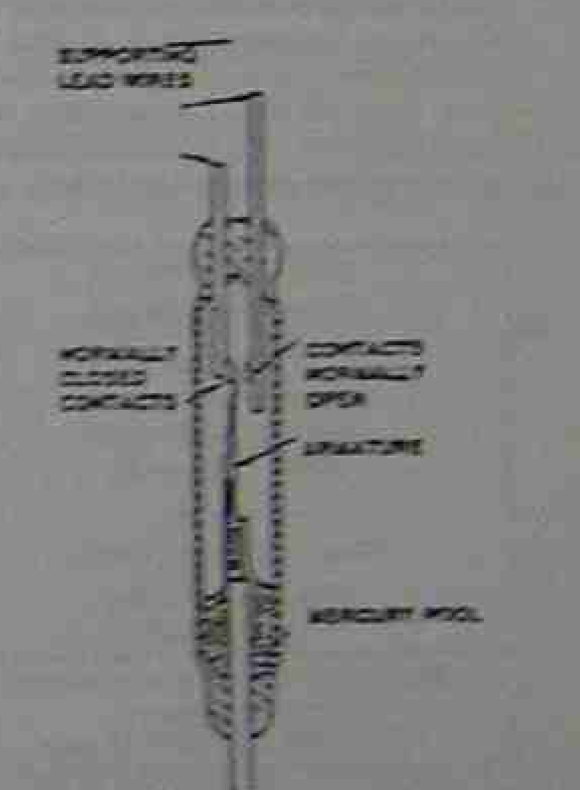
Reed relays use reed switches, these consist of at least two metal reeds sealed in a glass tube filled with inert gas. The reed relay operates by placing the reed switch in the proximity of a magnetic field.

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[illegible] 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

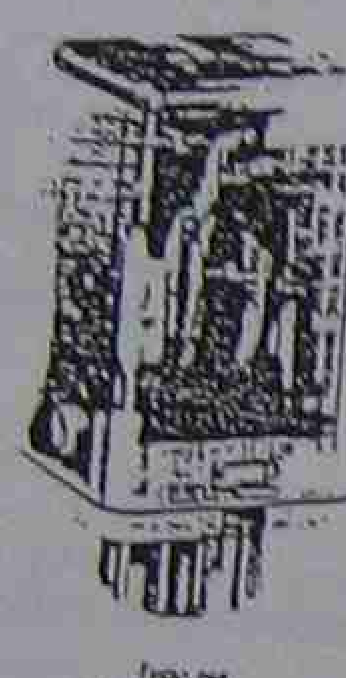
Description — The Bulletin 700 Type H line of general purpose relays includes the Type HA tube base relay, Type HB square base-relay, Type HC miniature-square base relay, Type HT single range timing relay and Type HTM-multi-range-timing relay. A variety of options and accessories are available for each of these relays.

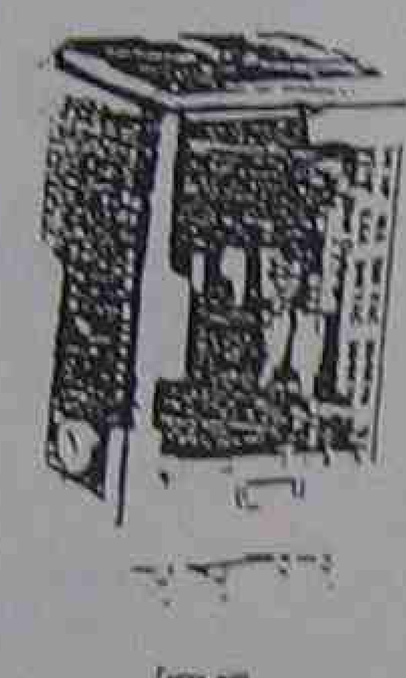
Type HA and HB relays — Type HA and HB relays are rated at 10 amperes and available with AC or DC coils. Type HA relays are available in 2 or 3 pole versions with tube base plug-in terminations. Type HB relays are available in 2 or 3 pole versions with square base, plug-in quick-connect solder terminations.

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	[illegible]	24 VDC	10
240	15	1.5	10□	10□	[illegible]		

□ 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	19		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.

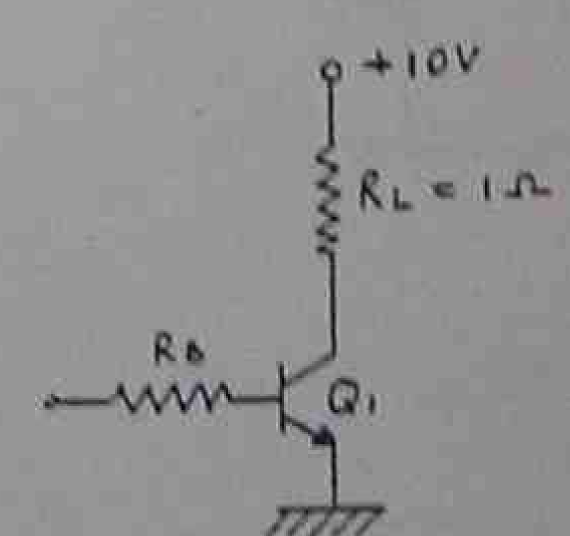
The *ON - OFF* conditions have a power dissipation of :-

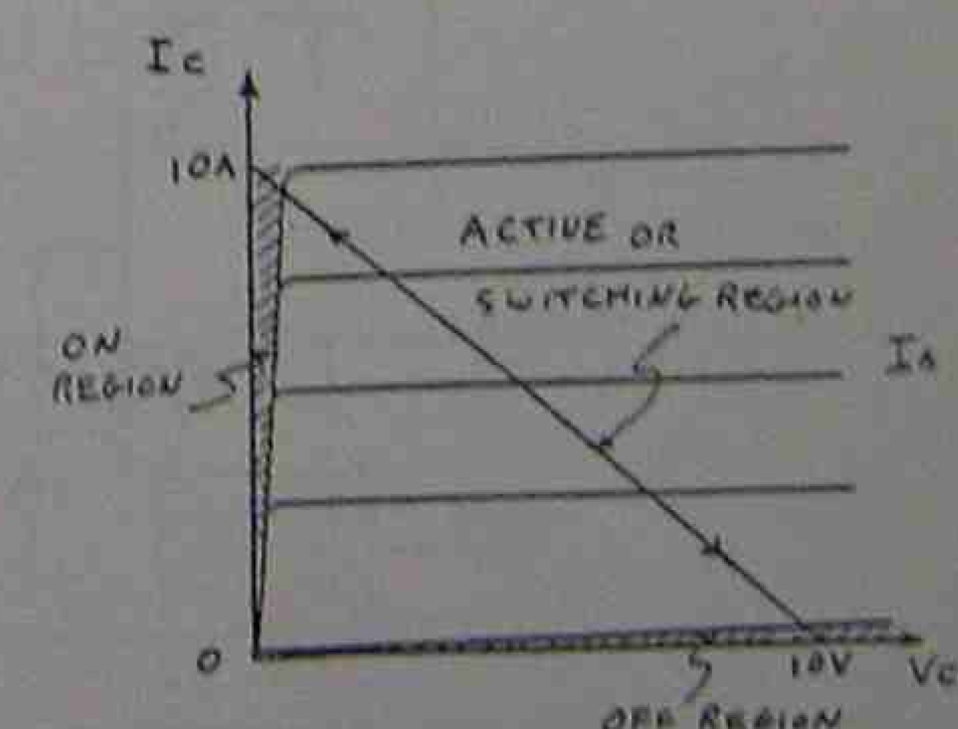
. *POWER OFF = VCC x I_{ce}* . *POWER ON = VCE sat x Ic max*

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.



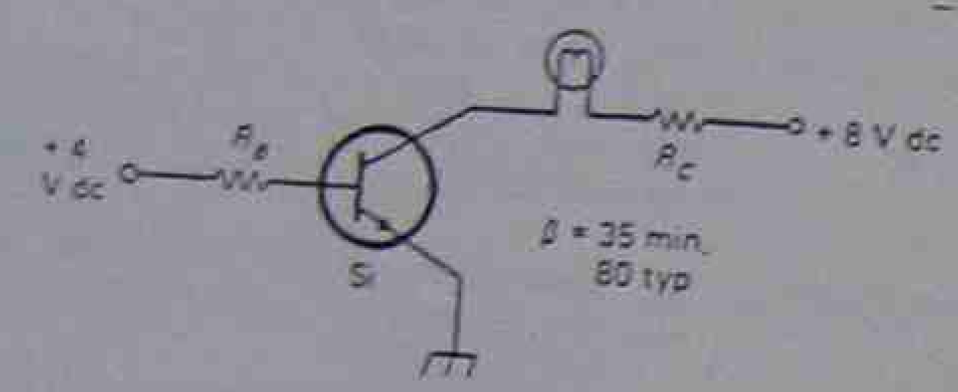


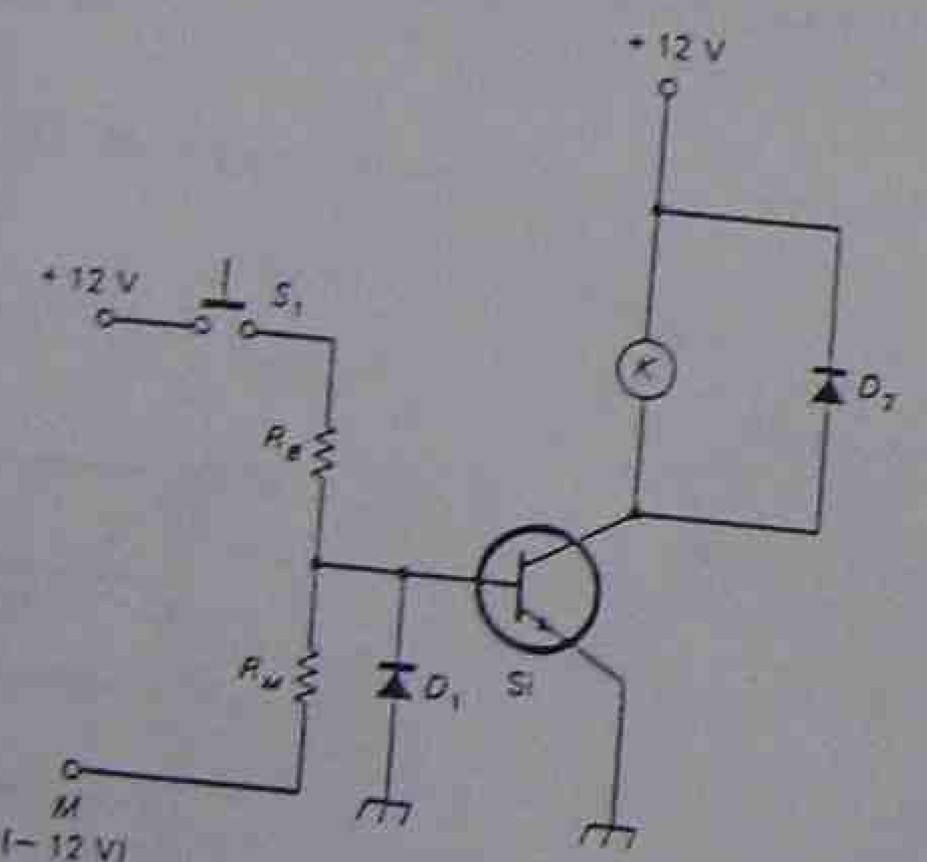
DESIGN OF TRANSISTOR DRIVE CIRCUITS

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.



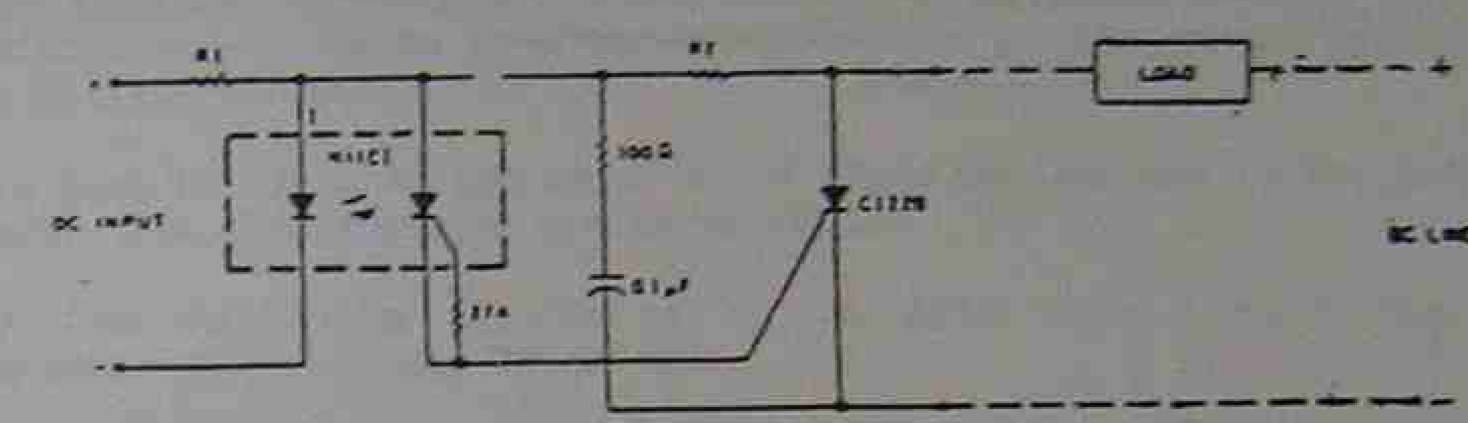


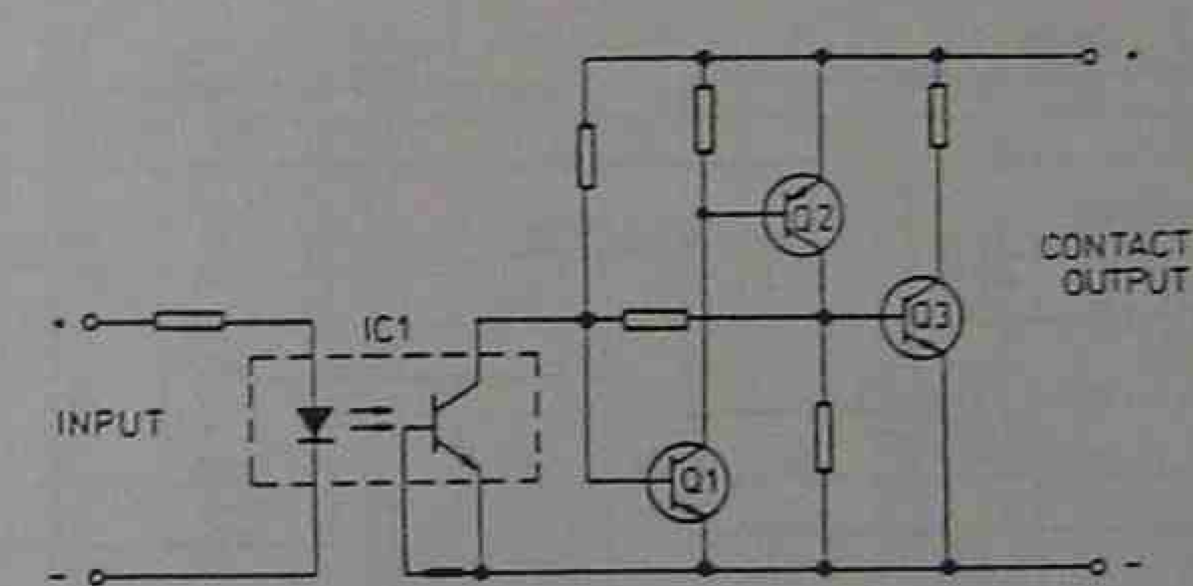
DC OUTPUT MODULES

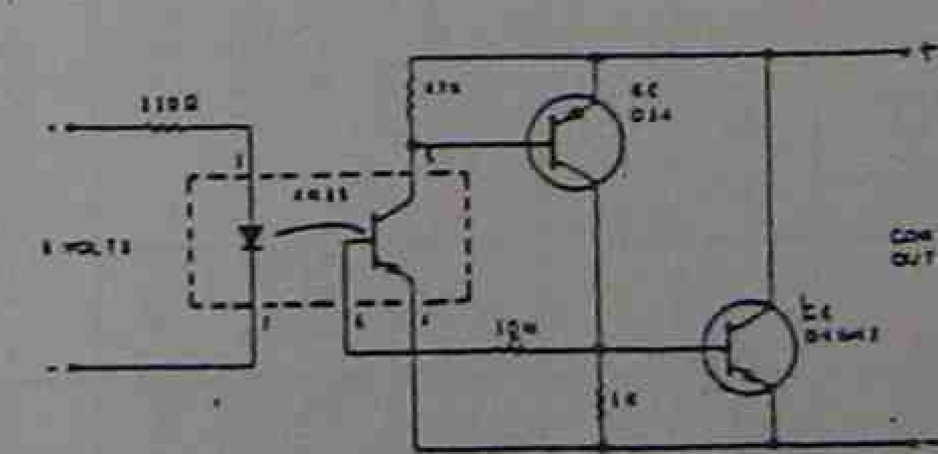
Modules using discrete circuitry can be commercially obtained or constructed for a specific control system.

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.



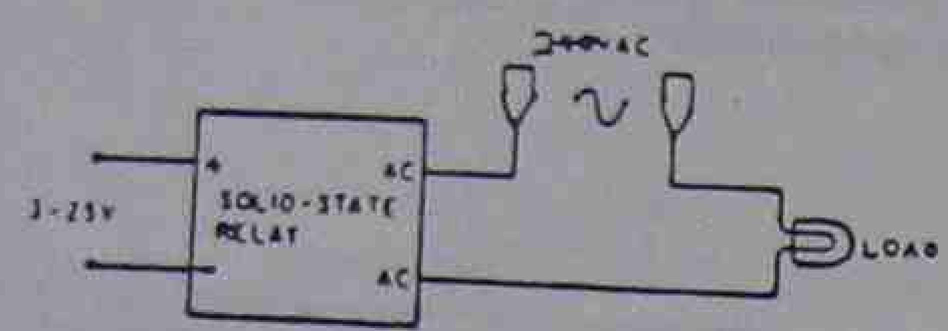


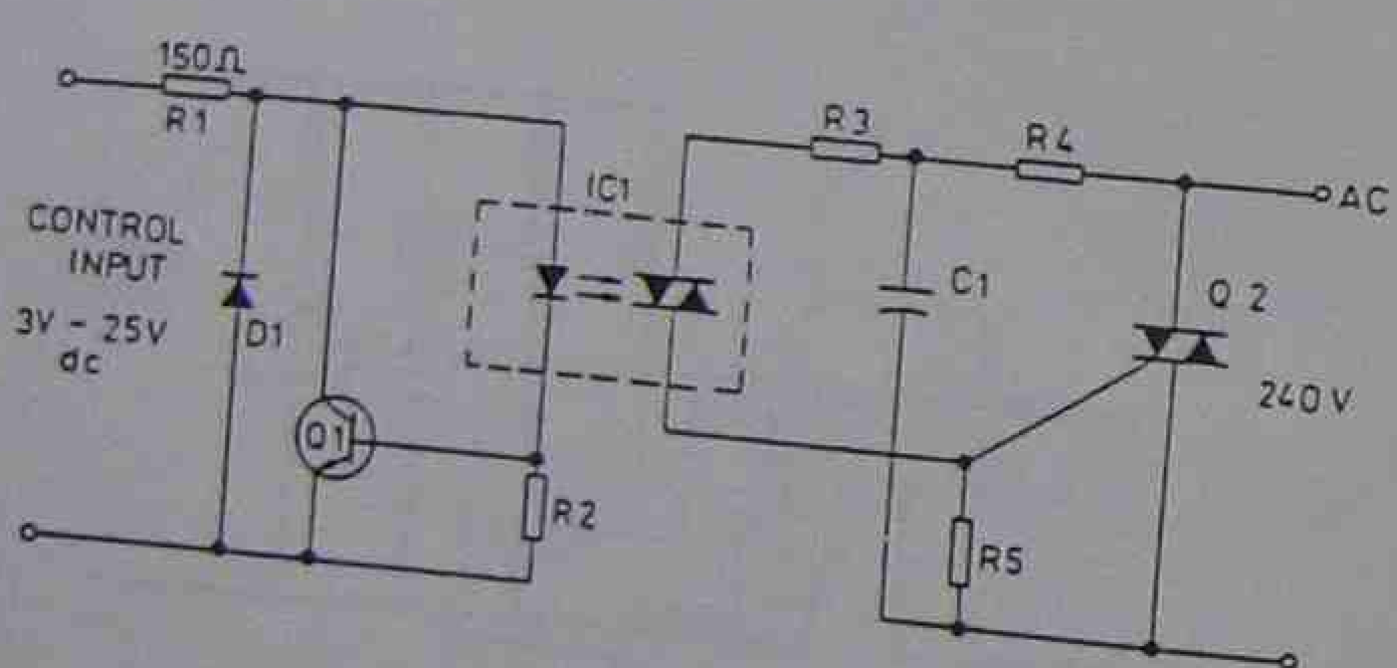


AC OUTPUT MODULES

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.

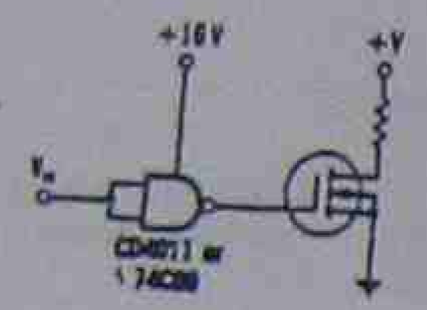
Care must be taken in the handling of these devices due to the possibility of damage from electrostatic discharge (ESD).

For the following sample of a MOS data sheet, note the following characteristics :- Vds, Id and Rds.

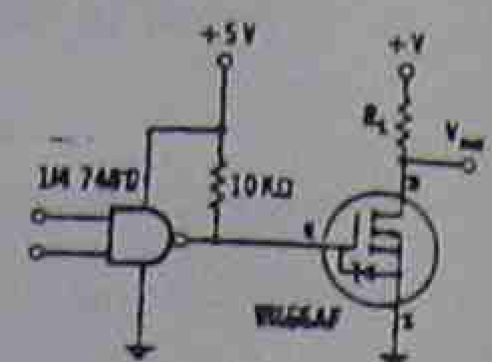
V_{DS} V	$R_{DS(on)}$ Ohm	TO 220 BUZ	I_D/A	TO 218 BUZ	I_D/A	TO 3 BUZ	I_D/A	TO 238 BUZ	I_D/A
50	0,10 0,12 0,10 0.04 0,06 0,03	71 71A 71L 11 11A	13,0 13,0 12,0 30,0 25,0	348 347	39,0 42,0	14 15	39,0 45,0	17 18	32,0 37,0
60	0.04	1152	30,0						
100	0.2 0,25 0.20 0,085 0,6	72 72A 20 21	10,0 9,0 12,0 21,0	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 39	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	450 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

USING POWER FET's

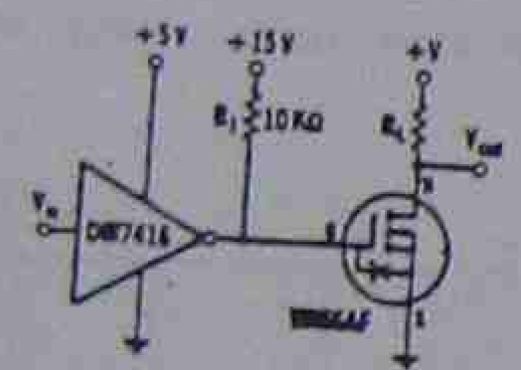
Mosfet's prove extremely versatile in their applications and since the inputs are voltage driven, with virtually zero current, their use with digital circuits are ideal.



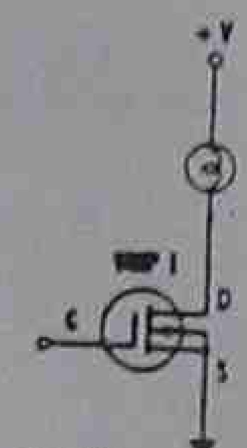
Driving VMOS from CMOS



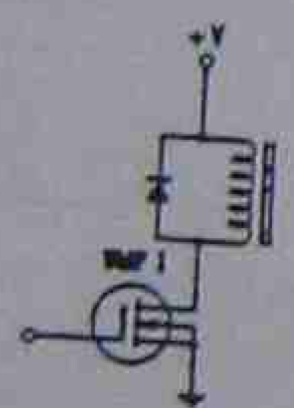
Driving VMOS from TTL



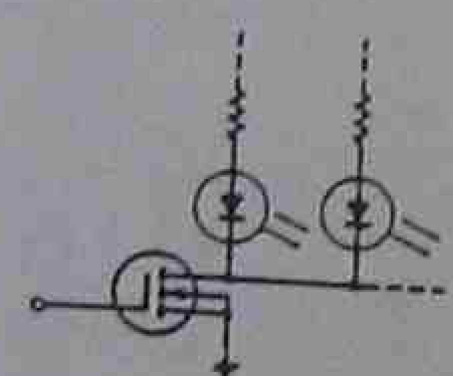
Driving VMOS from o/c TTL



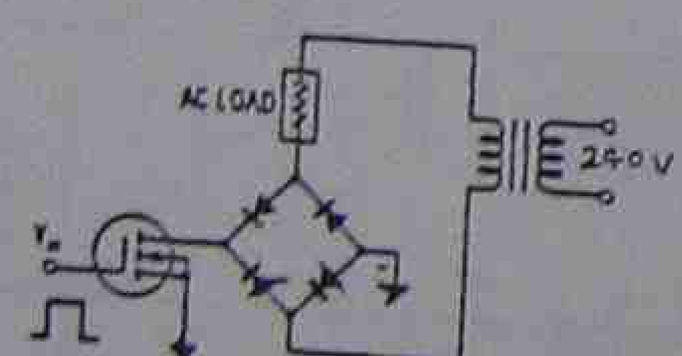
Lamp driver



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 : ____________________

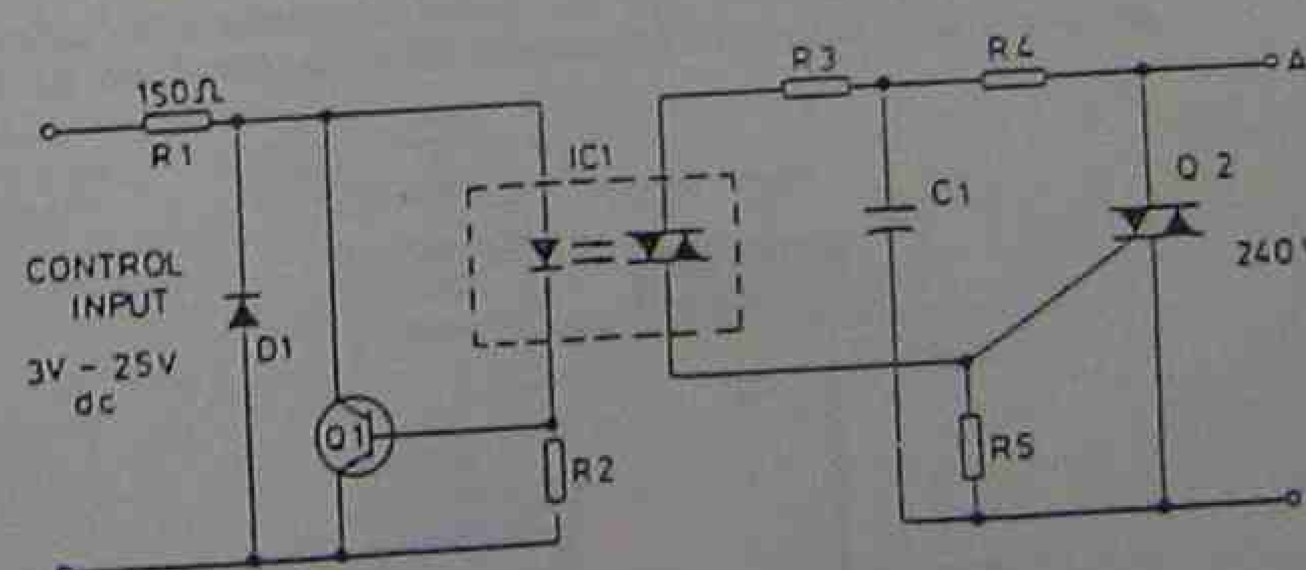
. IC1 : ____________________

. 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

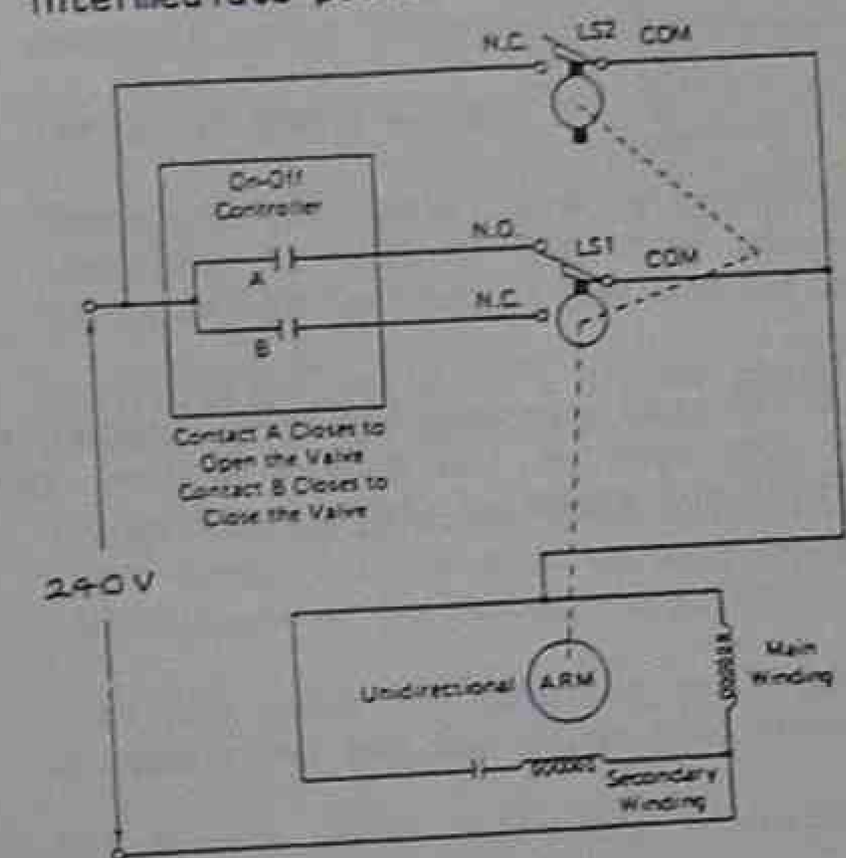
When a large valve is to control high fluid pressure, a motor must be connected to a mechanical linkage to operate the opening and closing.

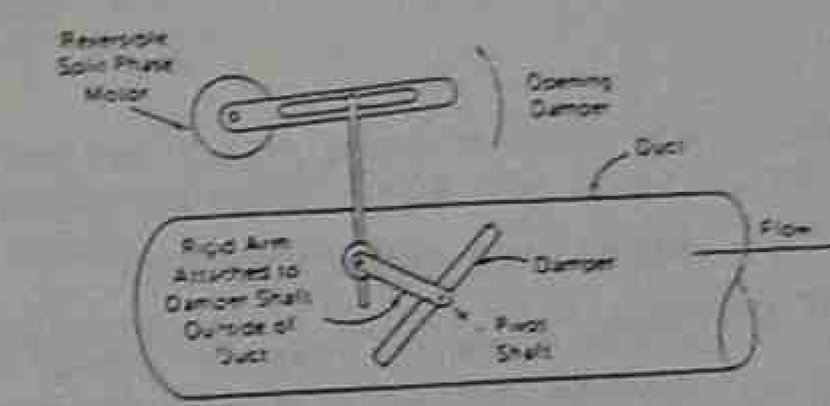
Most two position valves of this type are operated by a unidirectional split phase motor. The motor is geared down to provide a slow moving output shaft speed with high torque.

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.

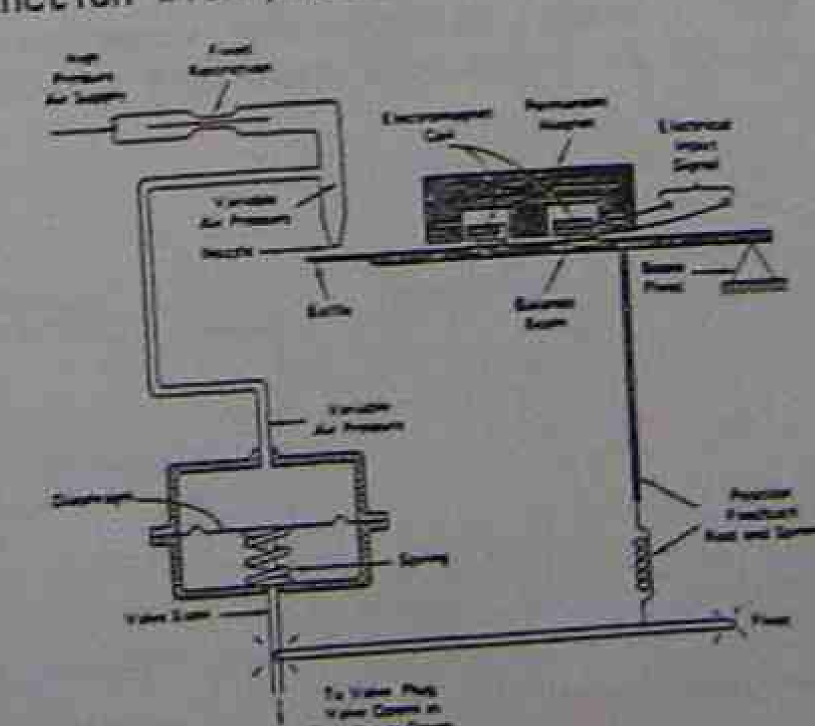
For proportional control the positioning of the control vale will be in an intermediate position.





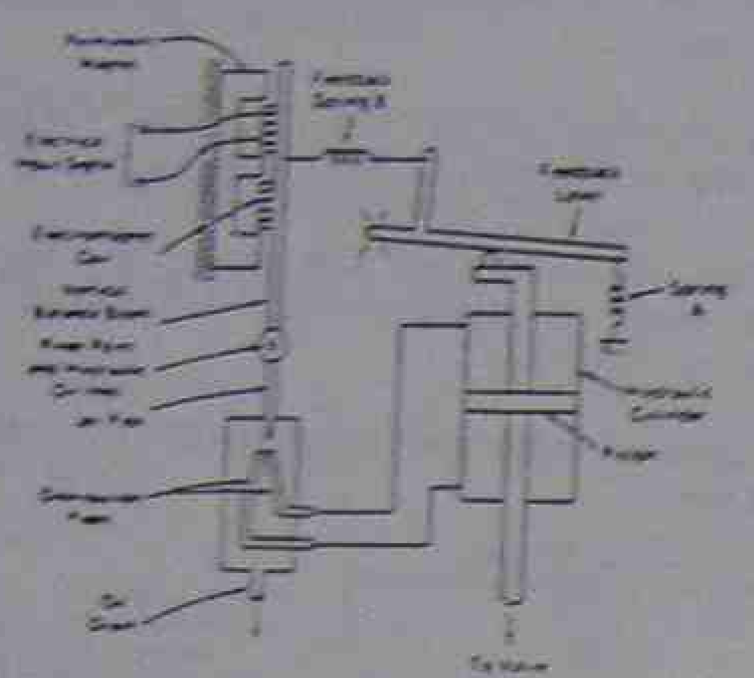
ELECTROPNEUMATIC VALVES

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

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.

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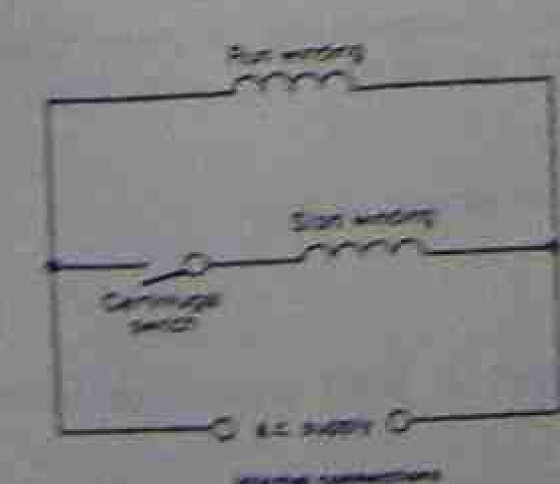
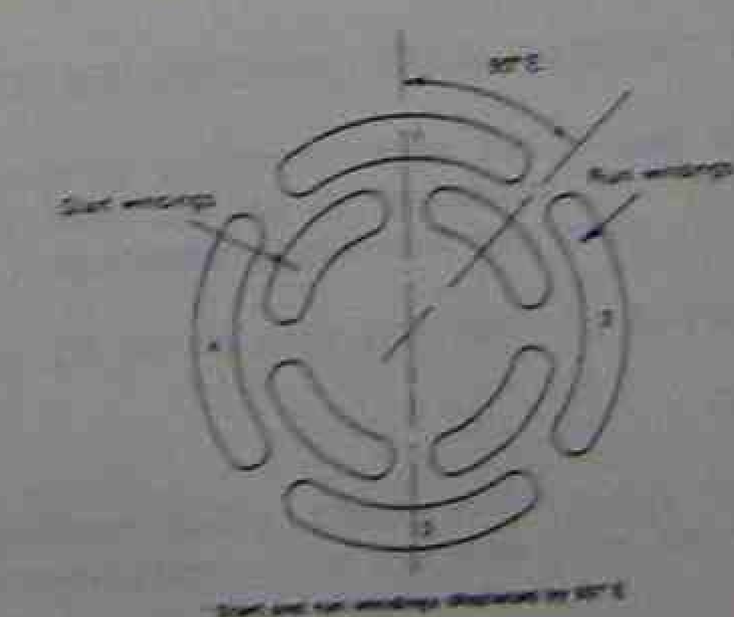
The single phase motor has a vibration of twice the line frequency and this makes it noisy in operation.

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.

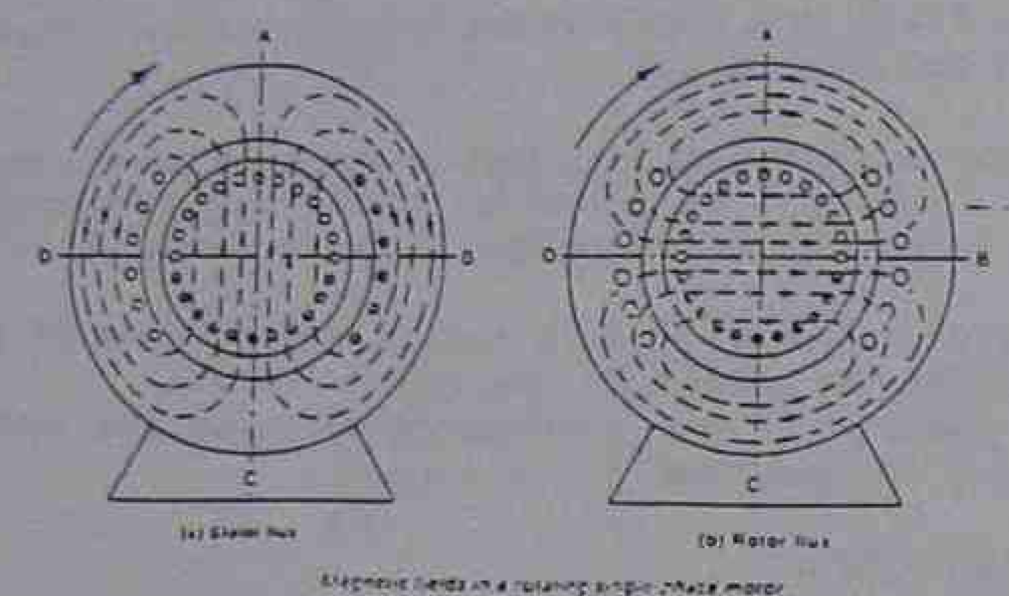




The startor flux rotates at a speed determined by the supply frequency and the number of poles in a winding. The direction of rotation is in the same direction as that of the magnetic field.

$$n = \frac{120 \times f}{p}$$

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.

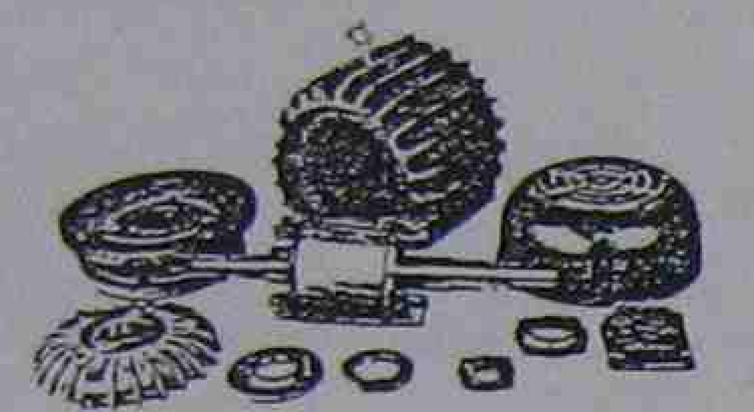
For large motors the load should not be removed when the motor is running at speed but in small motors the internal losses such as friction and windage will limit the speed to a safe value.

USES

Portable appliances such as saws and drills, sewing machines, food mixers and vacuum cleaners.

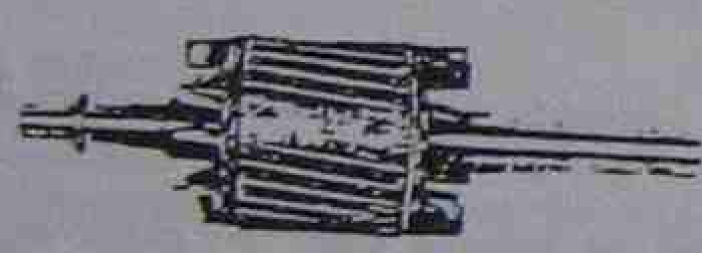
THREE PHASE INDUCTION MOTORS

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.

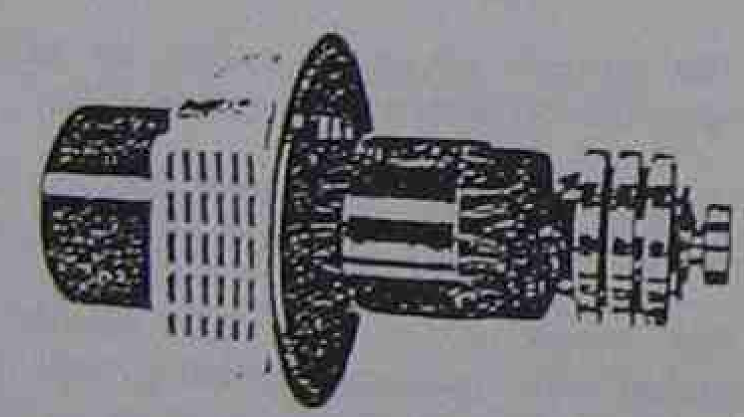


Rotors with bars in slots rather than windings are known as squirrel cage rotors and have a lower starting torque than the wound rotor.

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.

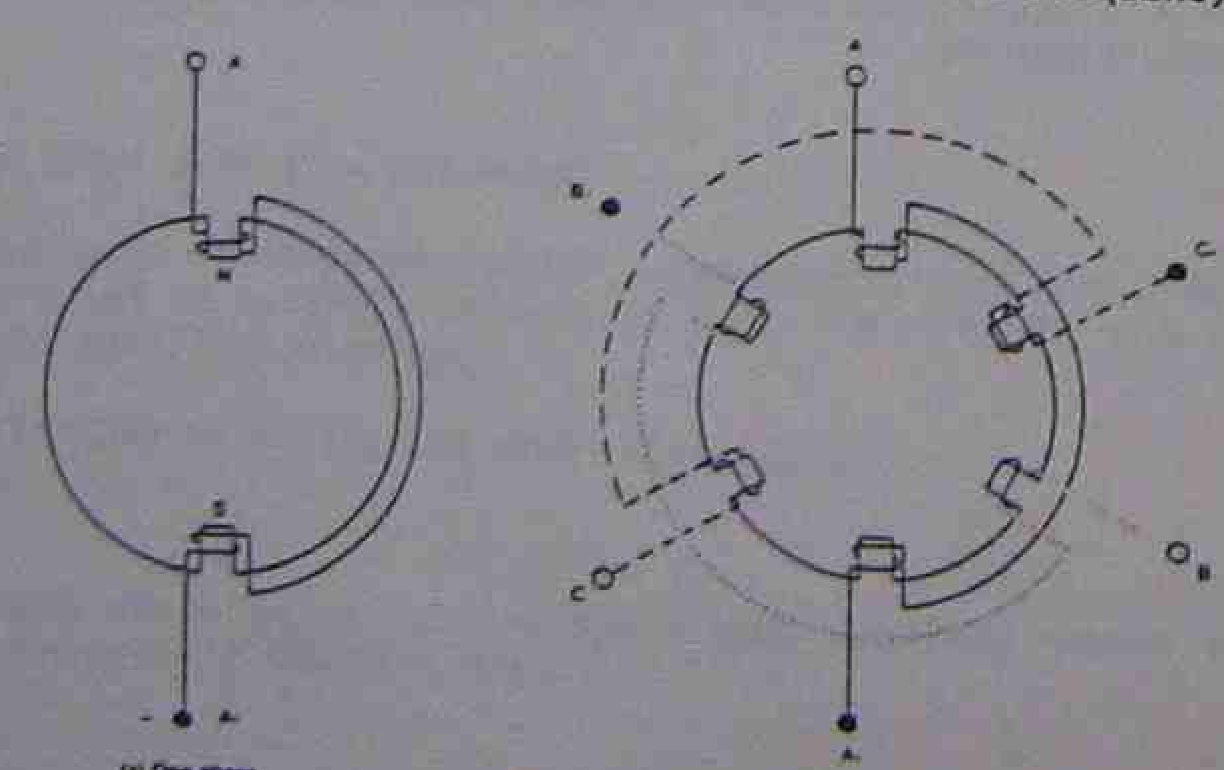


Squirrel-cage rotor for an induction motor



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.



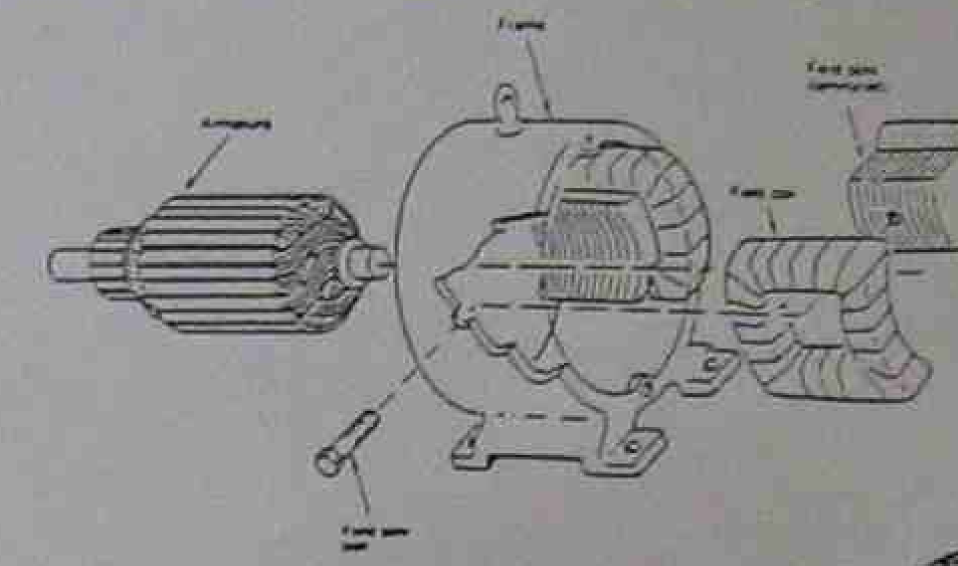
(a) One phase

(b) Three phases

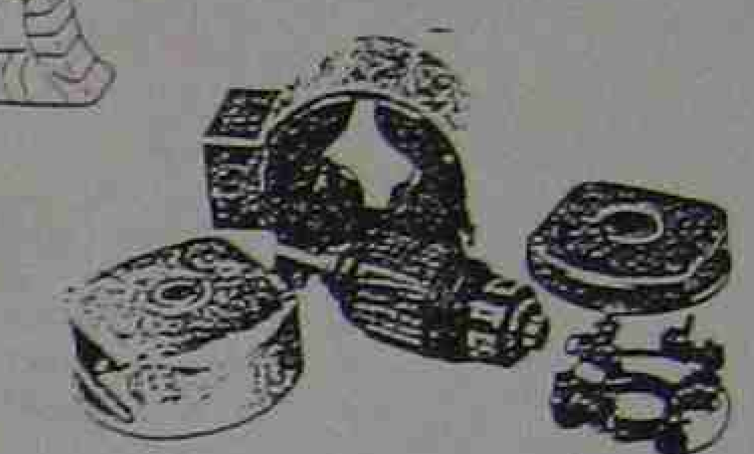
Polarities and connections in a two-pole, three-phase motor

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.

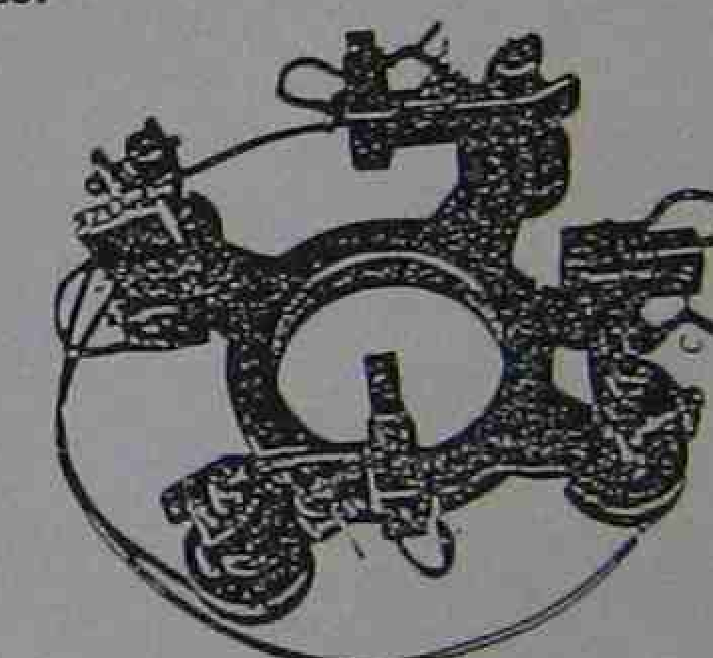


Component parts of a d.c. machine



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.

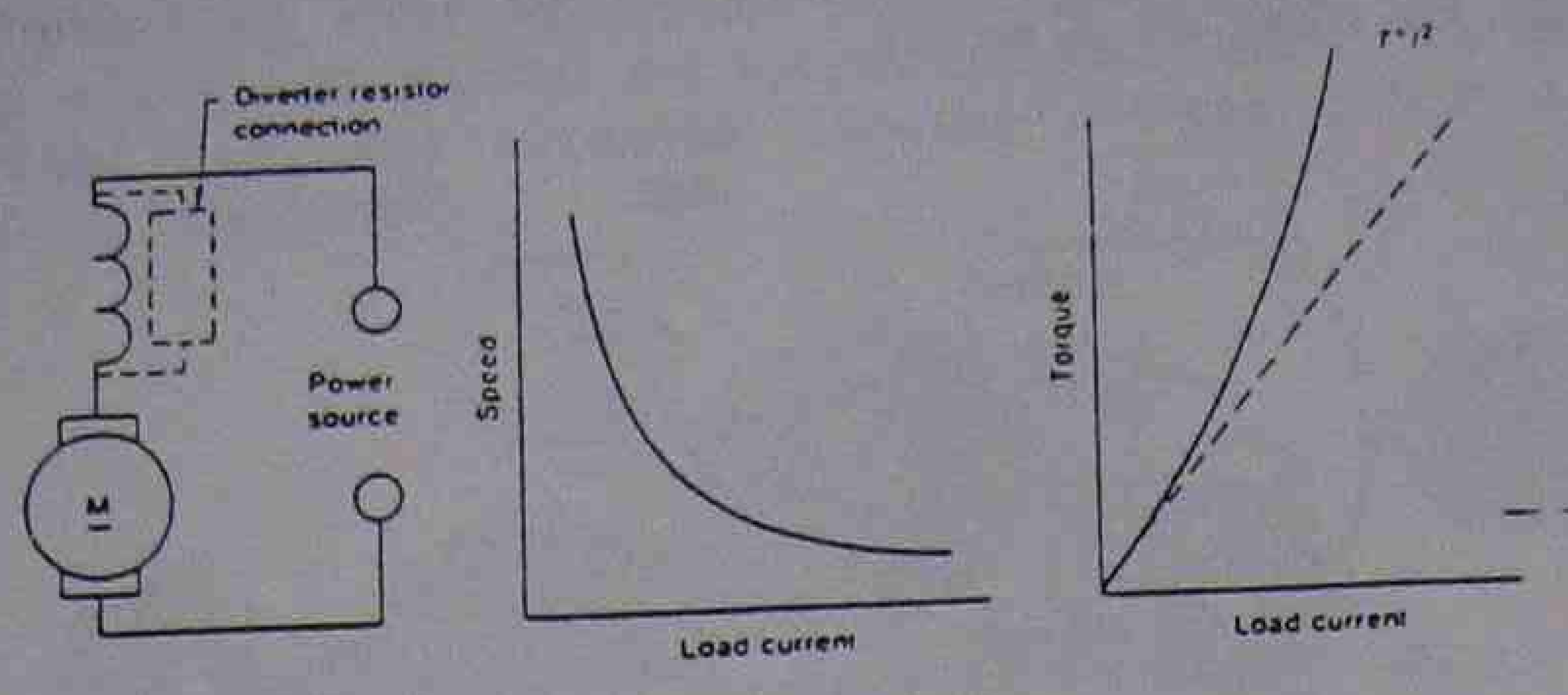
D.C motors can use permanent magnets for their fields, these are usually limited to toys using batteries such trains or radio controlled planes.

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.

SHUNT EXCITED MOTORS

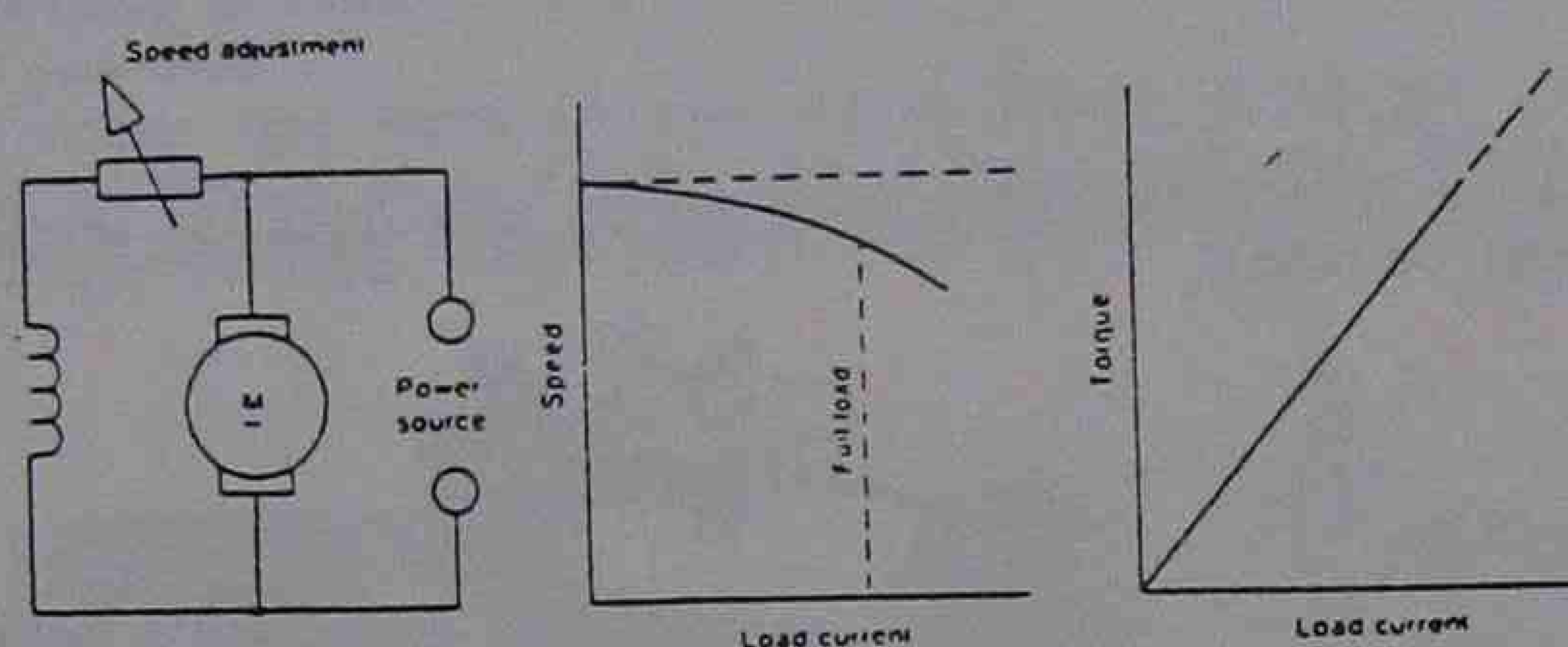
The shunt excited motor is used for its fairly constant speed characteristic.



Series motor characteristics

SERIES EXCITED MOTORS

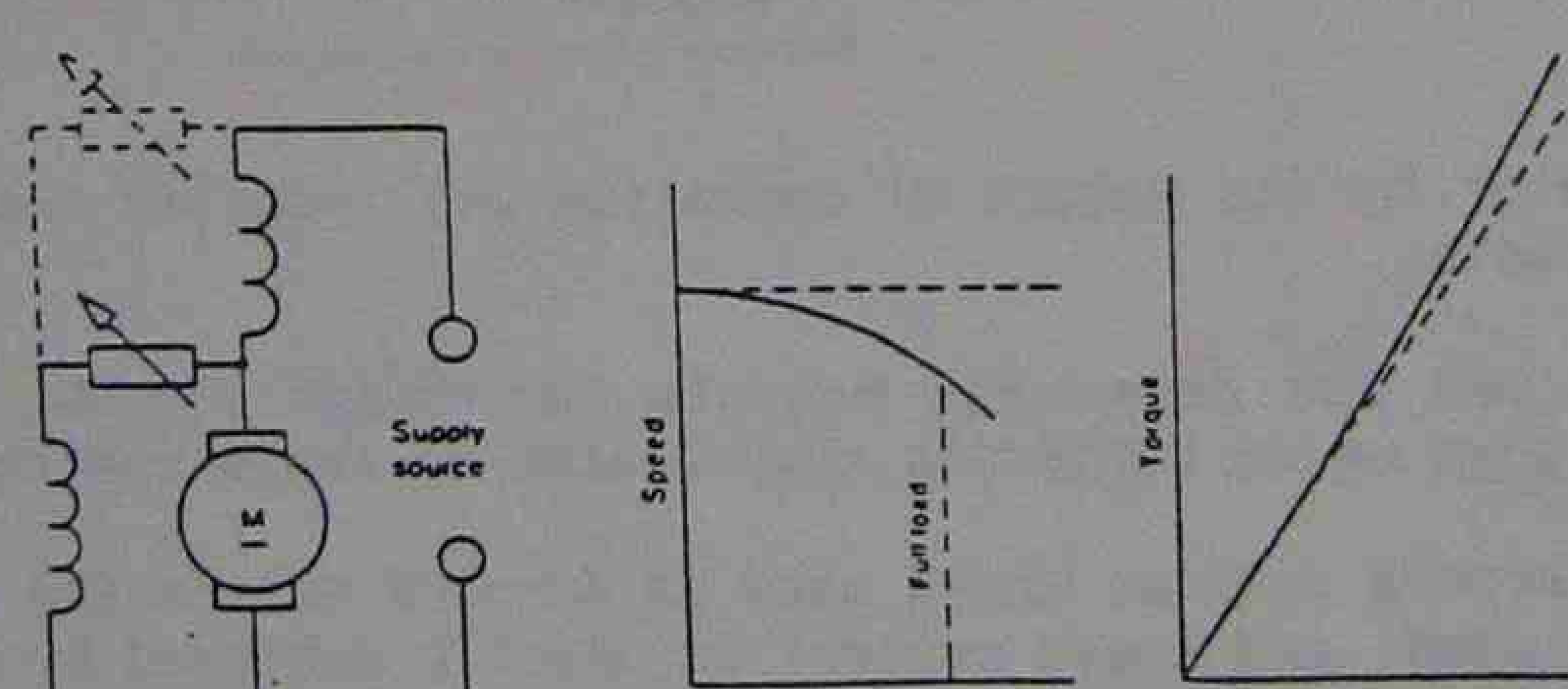
The series excited motor is used for its good starting torque characteristic. eg. trams, elevators, cranes,starter motors.



Shunt motor characteristics

COMPOUND MOTORS

The cumulatively compounded motor combines the characteristics of both the series and shunt motors. Applications are punches, shears, rolling mills.



Compound motor characteristics