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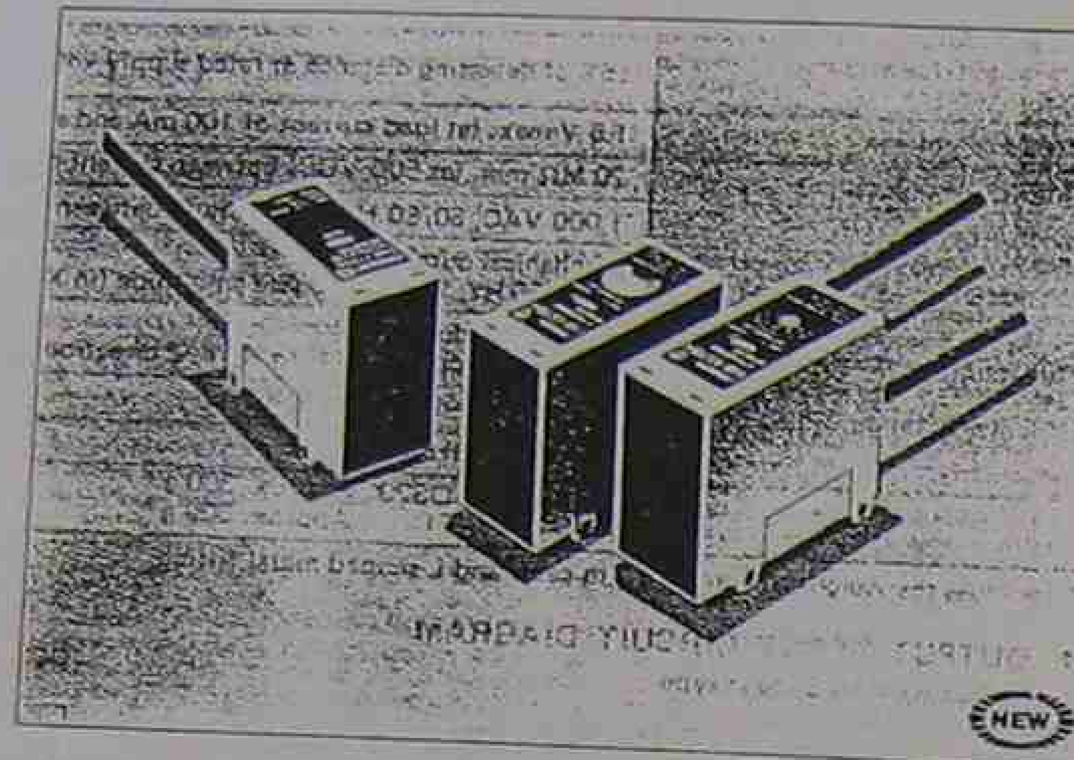
ULTRASONIC PROXIMITY SWITCH

Cat. No. D40-E1-2

Model E4B

Ultrasonic Sensor Capable of Detection Zone Setting of Ultrasonic Beam



■ FEATURES

- Stable detection regardless of target color, transparency, or material composition
- 8° beam width assures detection of target without interference by surrounding obstructions
- Detects targets as small as 2 x 2 cm
- 200 kHz ultrasonic wave frequency provides high noise immunity
- Sealed to resist dust and dirt
- STABILITY indicator aids beam axis and sensitivity adjustments

■ AVAILABLE TYPES

Method of detection *		Separate type (beam break)		Focusable type		Selectable zone type **
Detecting distance		50 cm	1 m	5 to 20 cm	20 to 70 cm	20 to 70 cm***
DC switching type	NPN output	E4B-TS50E4	E4B-T1E4	E4B-LS20E4	E4B-LS70E4	E4B-RS70E4
	PNP output	E4B-TS50F4	E4B-T1F4	E4B-LS20F4	E4B-LS70F4	E4B-RS70F4

* For details on the methods of detection, refer to "HINTS ON CORRECT USE" on the lower portion of page 6.
** Five detecting zones are selectable, at intervals of 10 cm.
*** Type E39-R1 reflector is available as an optional accessory for the selectable zone type switches.

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■ SPECIFICATIONS

- RATINGS

Item / Type		E4B-TS50E4 E4B-TS50F4	E4B-T1E4 E4B-T1F4	E4B-LS20E4 E4B-LS20F4	E4B-LS70E4 E4B-LS70F4	E4B-RS70E4 E4B-RS70F4
Supply voltage (operating voltage range)		12 to 24 VDC (10.8 to 26.4 VDC) Ripple (p-p): 10% max.				
Current consumption	at 12 VDC	Transmitter: 120 mA max. Receiver: 30 mA max.		100 mA max.		
	at 24 VDC	Transmitter: 70 mA max. Receiver: 30 mA max.		50 mA max.		
Standard target		Flat plate (4 x 4 cm)				
Detecting distance		50 cm	1 m	5 to 20 cm	20 to 70 cm	20 to 70 cm
Differential travel		—		20% max. of detecting distance		3 cm max.
Beam width *		8° max.				
Response frequency		50 Hz	10 Hz	50 Hz		20 Hz
Output operation mode		ON/OFF state of output transistor when the beam is incident is selectable (by switching the polarity of the power supply)				
Control output		100 mA (residual voltage: 1.5 V max.), output resistance: 4.7 kΩ				
LED indicators		SENSING indicator: Red LED STABILITY indicator: Green LED				
Beam speed compensation function		Not provided				Provided
Material of housing		ABS				
Cable length		2 m				
Degree of protection**		IP66 (IEC 144), NEMA types 1, 4, 6, 12, 13				

* Denotes the maximum beam width deviating from the optimum center line when the received signal level is −6 dB
** Denotes the degree of protection provided by the housing of the switch, not the operating conditions of the device

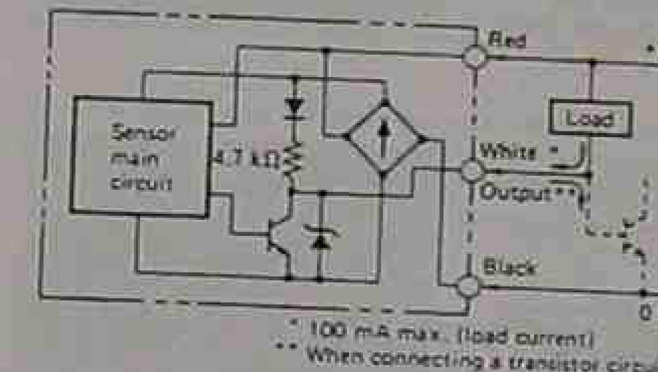
- CHARACTERISTICS

Variation due to temperature fluctuation	±10% of detecting distance at 20°C when operated within a temperature range of −10° to 55°C
Variation due to voltage fluctuation	±5% of detecting distance at rated supply voltage when operated within ±10% of rated supply voltage
Residual voltage	1.5 V max. (at load current of 100 mA and standard cable length of 2 m)
Insulation resistance	20 MΩ min. (at 500 VDC) between current-carrying part and housing
Dielectric strength	1,000 VAC, 50/60 Hz for 1 minute between current-carrying part and housing
Vibration	Mechanical durability: 10 to 55 Hz, 1.5 mm double amplitude (in X, Y, Z directions, respectively for 2 hours)
Shock	Mechanical durability: 500 m/s² (approx. 50 G in X, Y, Z directions, respectively 3 times)
Ambient temperature	Operating: −10° to 55°C
Humidity	35 to 95% RH
Weight	E4B-□S70/-□S20 Approx. 300 g* E4B-TS50/-T1: Approx. 300 g*

* Includes the weight of standard 2 m cable and L-shaped metal fitting.

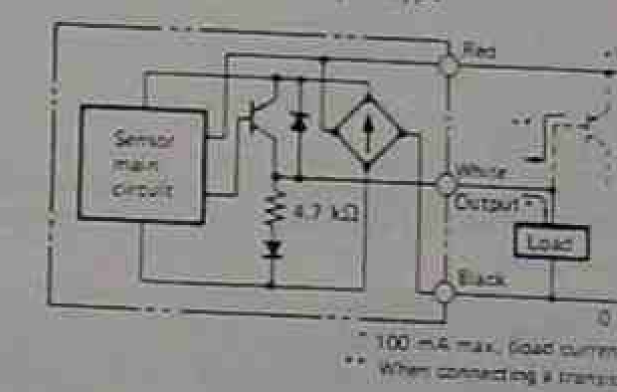
- OUTPUT STAGE CIRCUIT DIAGRAM

DC switching NPN output type



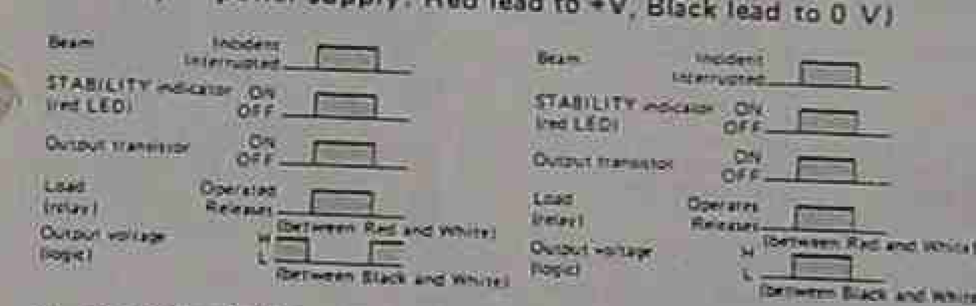
* 100 mA max. (load current)
** When connecting a transistor circuit

DC switching PNP output type

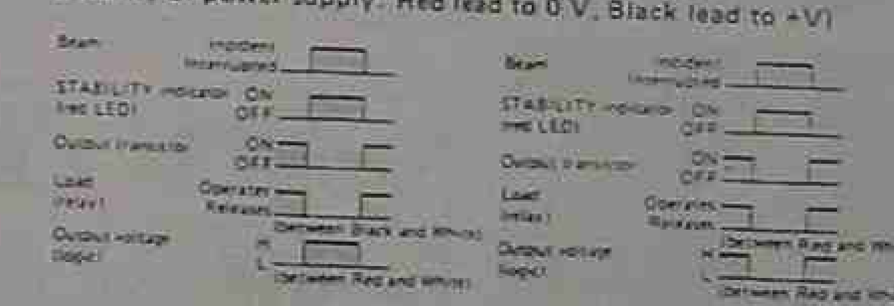


* 100 mA max. (load current)
** When connecting a transistor circuit

When beam is incident, output transistor turns on.
(Polarity of power supply: Red lead to +V, Black lead to 0 V)

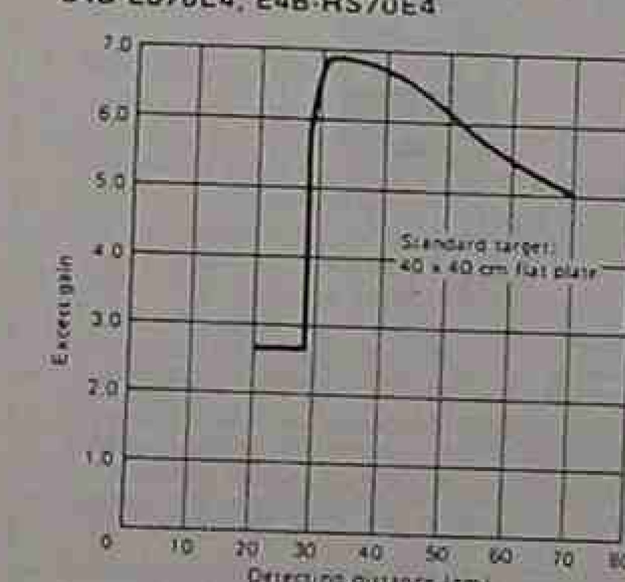


When beam is interrupted, output transistor turns on.
(Polarity of power supply: Red lead to 0 V, Black lead to +V)

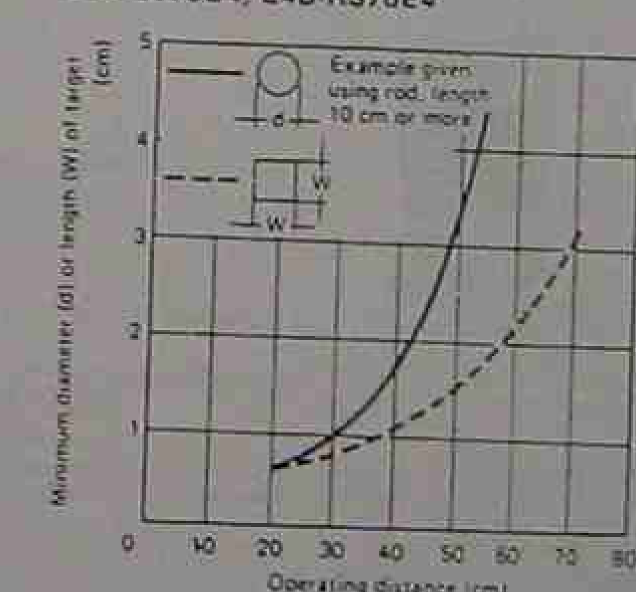


- CHARACTERISTIC DATA (Typical examples)

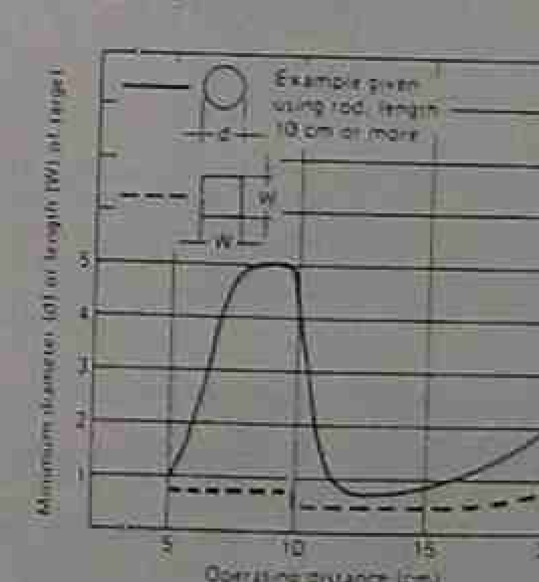
Excess Gain
E4B-LS70E4, E4B-RS70E4



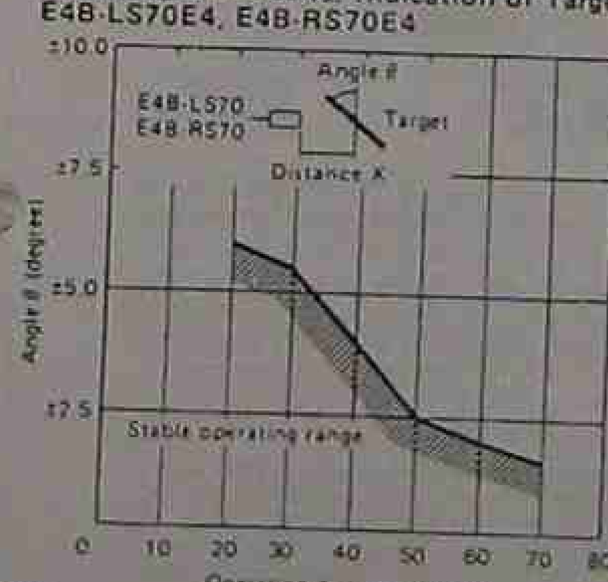
Operating Distance vs. Size of Target
E4B-LS70E4, E4B-RS70E4



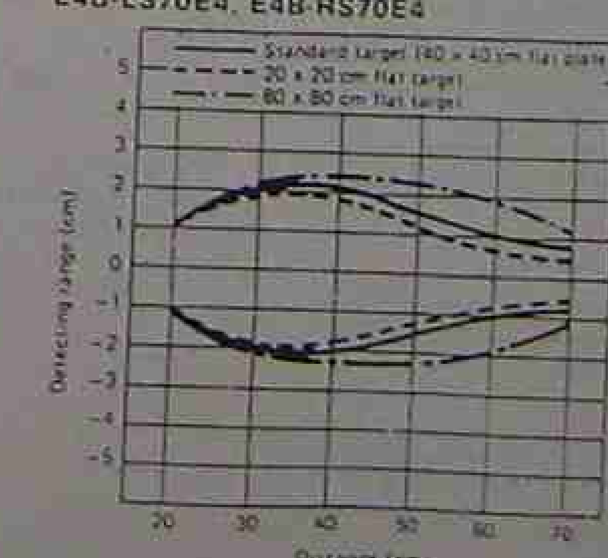
Operating Distance vs. Size of Target
E4B-LS20E4



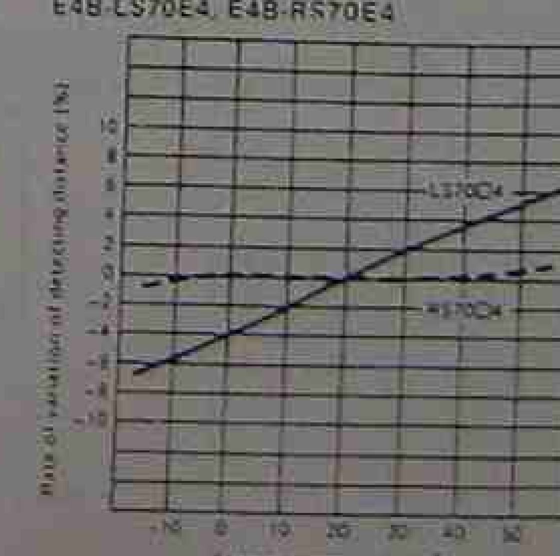
Operating Distance vs. Indication of Target
E4B-LS70E4, E4B-RS70E4



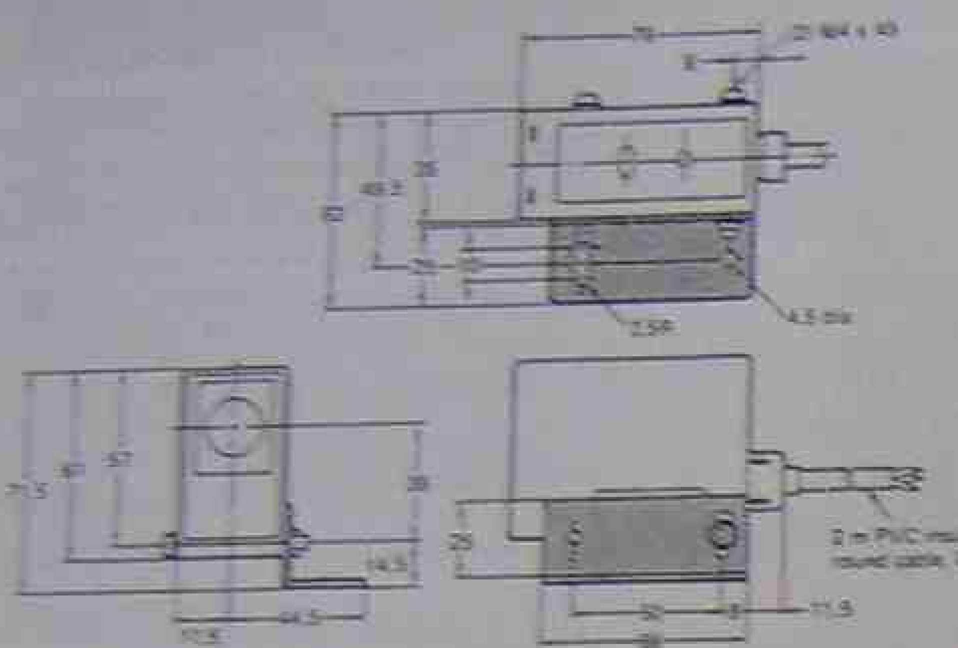
Operating Range Diagram
E4B-LS70E4, E4B-RS70E4



Rate of Variation of Detecting Distance vs. Ambient Temperature
E4B-LS70E4, E4B-RS70E4



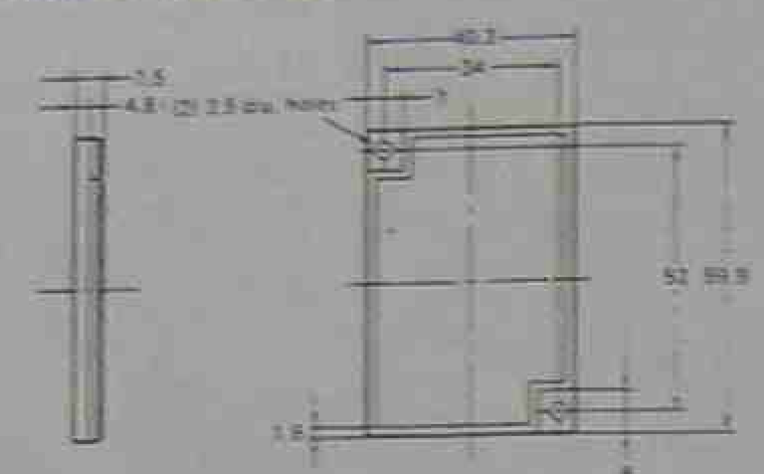
■ DIMENSIONS



■ ACCESSORY (available on request)

- E39-R1 REFLECTOR

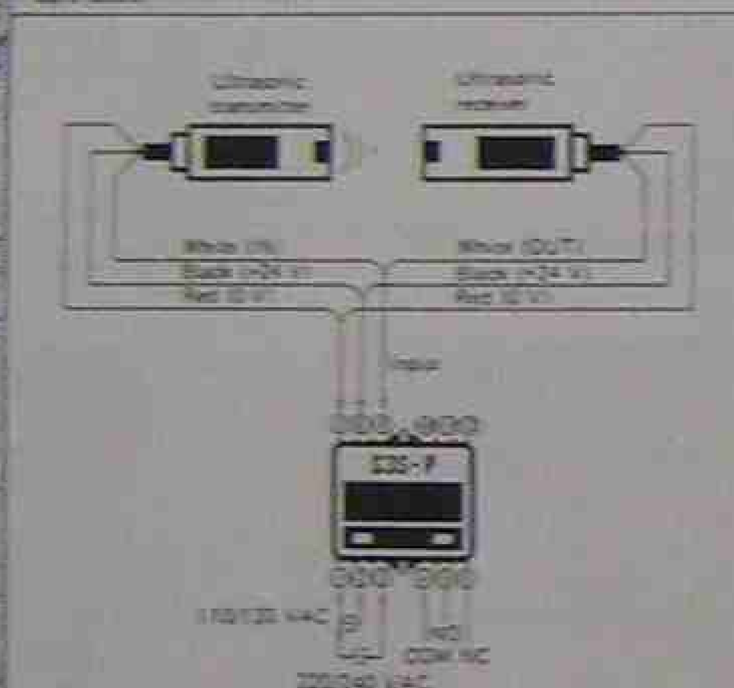
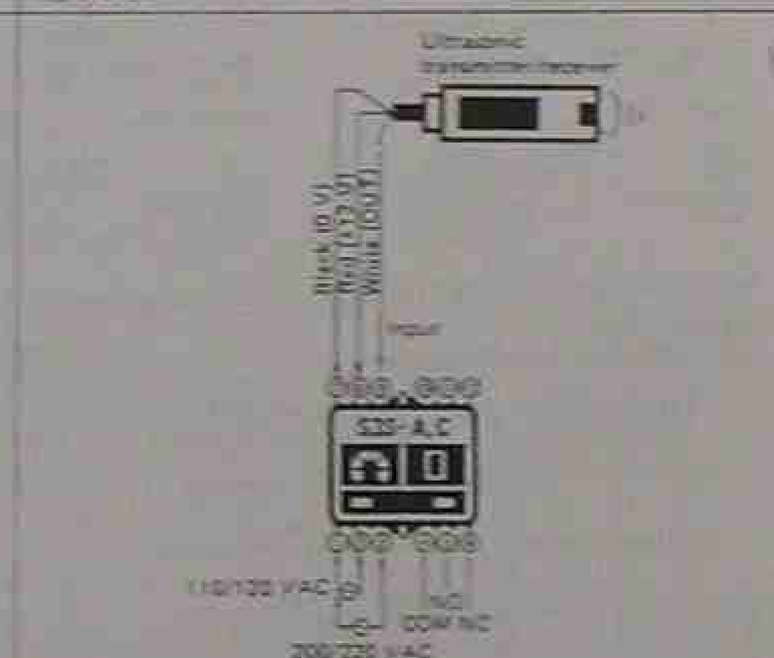
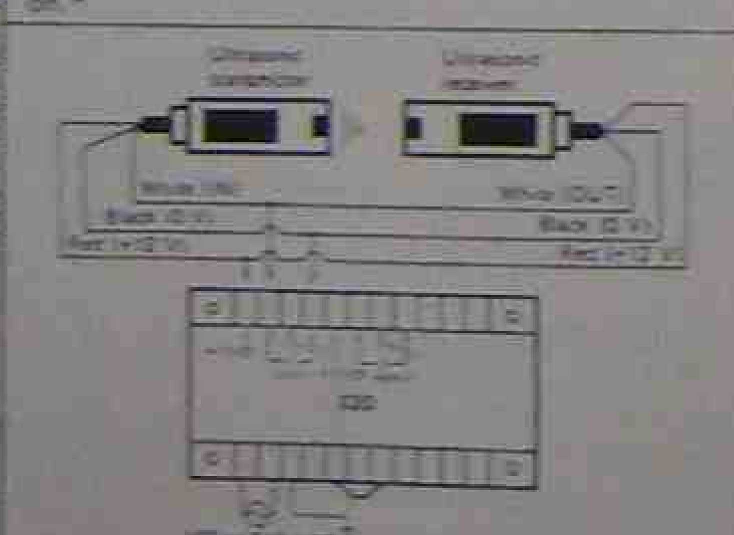
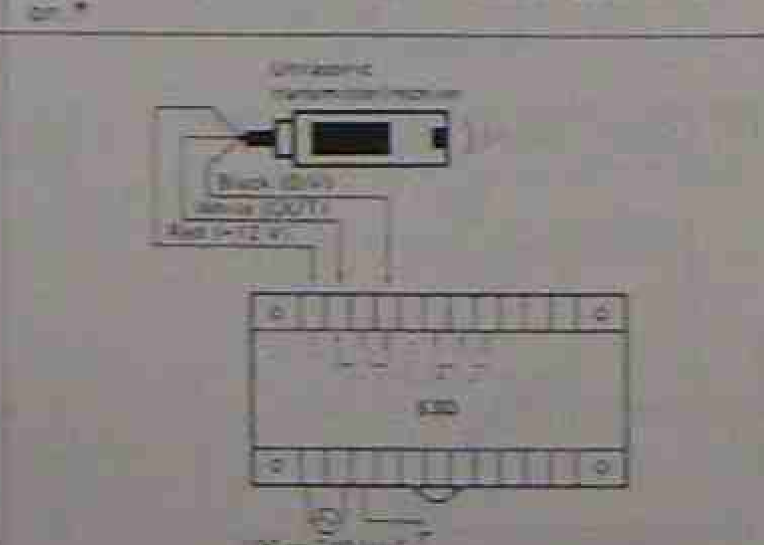
This optional accessory may be used with Types E4B-RS70E4 and E4B-RS70F4 ultrasonic sensors to aid in detecting targets.



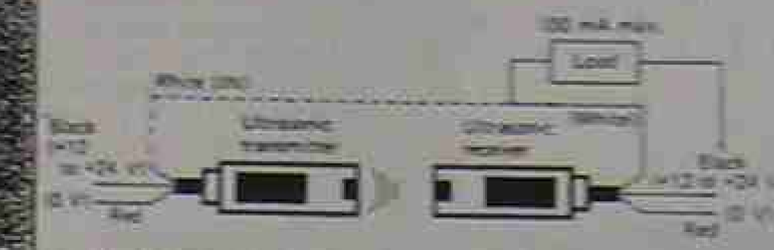
■ CONNECTIONS

Either Model S3S controller unit or S3D sensor controller may be used, depending on the application. For details on Models S3S and S3D, refer to their individual catalogs.

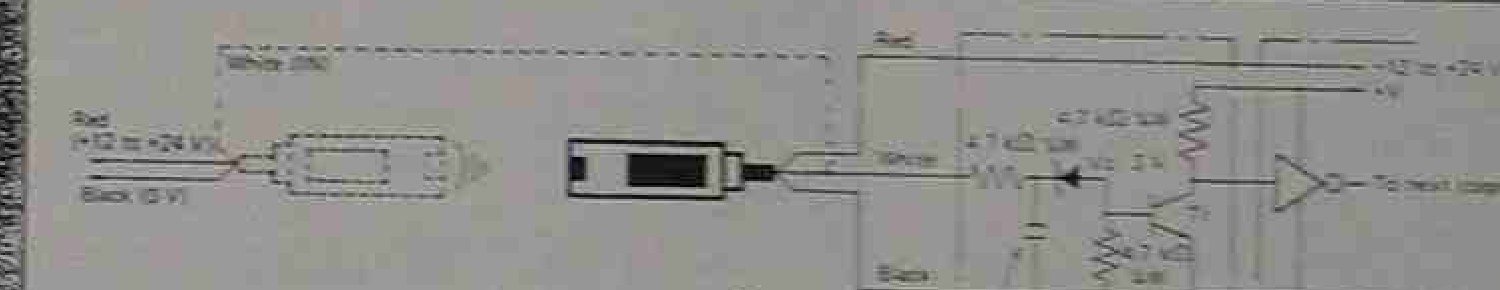
1. When using controller unit

Applicable controller	[illegible]	Focusable and Selectable zone types E4B-LS20E4 E4B-LS70E4 E4B-RS70E4
[illegible]	When the beam is interrupted, the output relay operates.	When the beam is interrupted, the output relay operates.*
S3S-P10 / S3S-A10-5 / S3S-C10		 * This operation mode is specified when the mode selector switch of the controller unit is set to the L-ON position. NOTE: When the mode selector switch is set to the L-ON position, the output relay operates upon incidence of the beam.
[illegible]	When the beam is incident, the input of S3D turns on.*	When the beam is incident, the input of S3D turns on.*
Connection S3D	 * To change the operation mode (i.e., the input turns on when the beam is interrupted), use the [illegible] key on the setting unit of S3D. NOTE: A maximum of two ultrasonic sensors can be connected to the controller unit.	 * To change the operation mode (i.e., the input turns on when the beam is interrupted), use the [illegible] keys on the setting unit of S3D. NOTE: A maximum of two ultrasonic sensors can be connected to the controller unit.

2. When driving a load (e.g., relay) directly

Item	[illegible]	Focusable and Selectable zone types E4B-LS70E4 E4B-RS70E4
Operating state of load	When the beam is interrupted, the load operates.	When the beam is incident, the load operates.
Connection	 NOTE: To operate the load when the beam is incident, connect the black lead to 0 V and the red lead to +12 to +24 V.	 NOTE: To operate the load when the beam is interrupted, connect the red lead to 0 V and the black lead to +12 to +24 V.
Remarks	When the white lead of the ultrasonic transmitter is connected to the white lead of the ultrasonic receiver as shown by the dotted line, the red LED on the receiver functions as a SENSING indicator. When the white and black leads of the receiver are interconnected, the red LED functions as a POWER indicator.	—

3. When applying NPN output to solid-state circuit

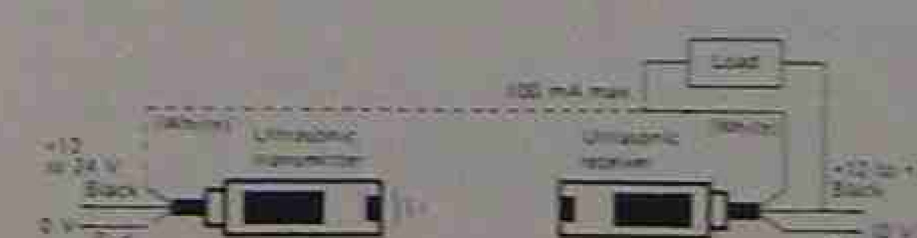
Applicable type	Separate type (beam break), focusable and selectable zone types (with suffix E4 in the type number)
Operating state of output	When the beam is interrupted, the output voltage goes high.
Connection	 NOTE: To turn the output voltage high when the beam is incident, connect the black lead to +12 to +24 V and the red lead to 0 V.
Remarks	When connecting the NPN output of the ultrasonic sensor to a CMOS IC or TTL, provide an interface circuit between the sensor and a solid-state circuit in the following stage as shown above and connect the sensor to the solid-state circuit.

■ HINTS ON CORRECT USE

- INDICATION
- LED indicators

1. STABILITY indicator (green LED)
   When this LED illuminates, it indicates that the amount of ultrasonic beam incident on the receiver is sufficiently large (or the amount of the beam interrupted is small enough) that the sensor is stably operating.
2. SENSING indicator (red LED)
   When this LED illuminates, it indicates that the ultrasonic beam is being received by the receiver.

- Indicator on ultrasonic transmitter (Separate type switches only)
  The LED on the transmitter may be used as a SENSING indicator that illuminates when the beam is incident by connecting the white lead of the transmitter to the white lead of the receiver.



NOTE: Be sure to connect the red leads of both transmitter and receiver to the power terminals of the same polarity (i.e., +V or 0 V) and the black leads of both transmitter and receiver to 0 V or +V to keep the polarity of both leads identical. Otherwise, the LED on the transmitter may illuminate when the beam is interrupted.

When the transmitter and receiver cannot be interconnected to obtain the above SENSING indication (because they are operating on a separate power supply) the LED on the transmitter may be used as a POWER indicator that illuminates upon power application, by connecting the white lead of the ultrasonic transmitter as shown in the table below.

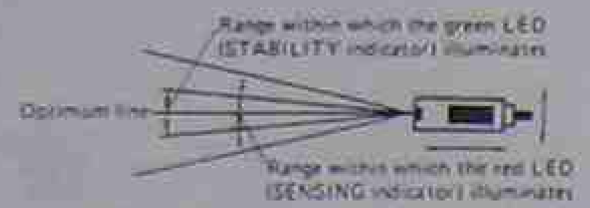
Connection of White lead

Power supply connection \ Output	NPN (-E4)	PNP (-F4)
When Red lead is connected to +V, and Black lead to 0 V	Connect white lead to black lead (0 V)	Connect white lead to red lead (+V)
When Red lead is connected to 0 V, and Black lead to +V	Connect white lead to black lead (+V)	Connect white lead to red lead (0 V)

NOTE: In other than the above combinations, the LED indicator may not illuminate.

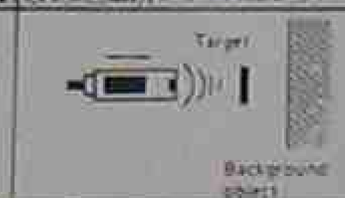
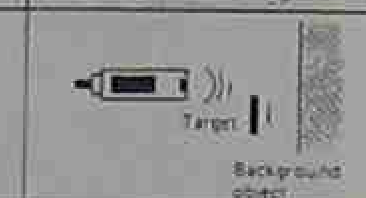
• BEAM AXIS AND SENSITIVITY/DISTANCE ADJUSTMENTS

1. **Separate type (beam break) (E4B-T1□, E4B-TS50□)**
- Turn the SENSITIVITY adjusting VR of the ultrasonic transmitter to the MAX position.
- Move the transmitter and receiver vertically and horizontally until the SENSING indicator of the receiver illuminates and then affix both transmitter and receiver at the midpoint of the range within which the STABILITY indicator illuminates.



- While passing the target through between the transmitter and receiver, adjust the sensitivity so that the SENSING indicator illuminates or goes out positively with the presence or absence the target with the STABILITY indicator in the illuminating st
- If the STABILITY indicator stops illuminating while the sensor is in operation, it is a warning signal for possible detection error. Check or readjust the sensitivity as required.
- When the ultrasonic transmitter and receiver are to be set with a distance shorter than the rated detecting distance, use the ultrasonic sensor by lowering its sensitivity within the range in which the STABILITY indicator illuminates. This action will increase the stability of the sensor against noise.

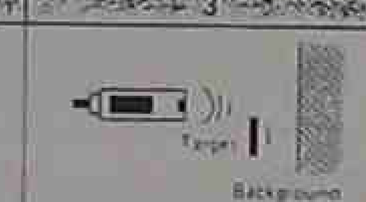
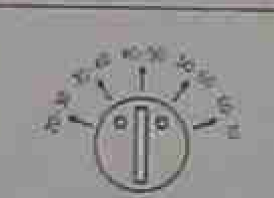
2. **Focusable type (E4B-LS20□/-LS70□)**
- Place the ultrasonic sensor so that both the STABILITY and SENSING indicators illuminate when the target is in position and that the STABILITY indicator illuminates but the SENSING indicator goes out when the target is removed.

Step	1	2	3
Detecting condition			
Distance adjustment		—	—
Adjustment procedure	With the target placed in position, gradually turn the DISTANCE adjusting VR clockwise (toward the MAX position) until both the SENSING and STABILITY indicators illuminate.	Move the ultrasonic sensor vertically and horizontally and affix the sensor at the midpoint of the range within which the STABILITY indicator illuminates.	Remove the target from the front of the sensor and confirm that the SENSING indicator goes out and the STABILITY indicator is still illuminating.

NOTE: If the STABILITY indicator stops illuminating while the sensor is in operation, it is a warning signal for possible detecting error. Check or readjust the sensitivity as required.

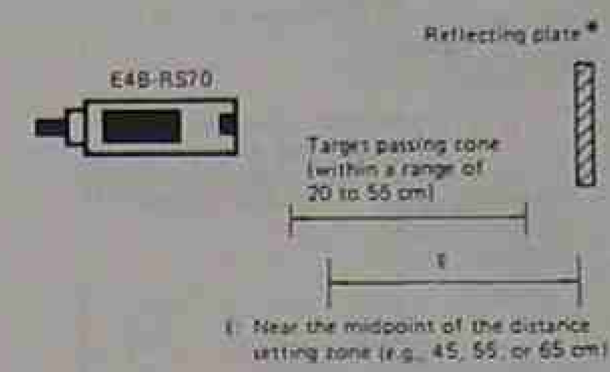
3. **Selectable zone type (E4B-RS70□)**

When the switch is to be used for general purposes
- Place the ultrasonic sensor so that both the STABILITY and SENSING indicators illuminate when the target is in position and that the STABILITY indicator illuminates but the SENSING indicator goes out when the target is removed.

Step	1	2	3
Detecting condition			
DISTANCE selector switch		—	—
Adjustment procedure	With the target placed in position, gradually step up the DISTANCE selector switch from the minimum value until both the STABILITY and SENSING indicators illuminate.	Move the ultrasonic sensor vertically and horizontally and affix the sensor at the midpoint of the range within which the STABILITY indicator illuminates.	Remove the target from the front of the sensor and confirm that the SENSING indicator goes out and the STABILITY indicator is still illuminating.

NOTE: 1. If the STABILITY indicator stops illuminating while the sensor is in operation, it is a warning signal for possible detection error.
2. If the background object is within a distance of 1.5 m from the detecting head of the switch, the SENSING indicator may go out and the STABILITY indicator may illuminate despite of the absence of any target.

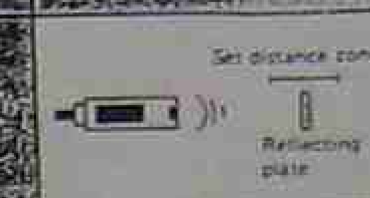
When the ultrasonic sensor is to be used as a retroreflective type
- The ultrasonic sensor used as a retroreflective type assures stable detection of any targets under irregular and unstable reflecting conditions that take place, for example, when the targets are nonuniform in the shape of this surface, when they are of different size, when they are being passed diagonally, and when they are warped or move wavily.
- Set the ultrasonic sensor so that both the STABILITY and SENSING indicators illuminate in the absence of any target.



ℓ: Near the midpoint of the distance setting zone (e.g., 45, 55, or 65 cm)

* The reflecting plate should preferably be of sheet metal, measuring 4 x 4 cm or more in size. The E39-R1 reflector may also be used for this purpose.

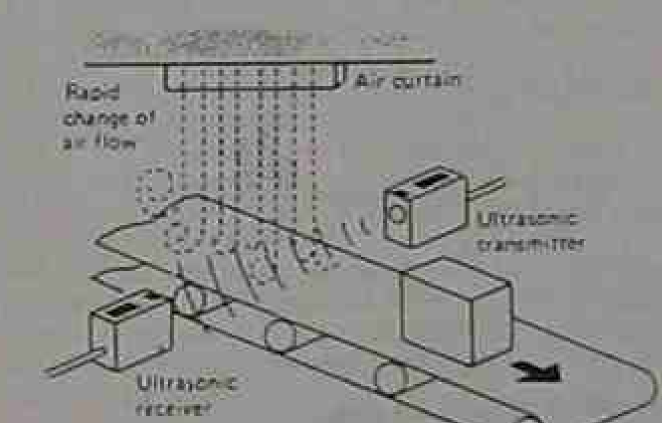
NOTE: Set the target so that its reflecting surface is positioned within the distance setting zone.

Step	1	2	3
Detection condition			
DISTANCE selector switch		—	—
Adjustment procedure	Set the DISTANCE selector switch to the desired zone position, affix the reflecting plate at or near the midpoint of the set distance zone, and confirm that both the STABILITY and SENSING indicators are illuminating.	Move the ultrasonic sensor vertically and horizontally and affix the sensor at the midpoint of the range within which the STABILITY indicator illuminates.	With the target placed in position, confirm that the SENSING indicator is not illuminating and the STABILITY indicator is illuminating.

NOTE: 1. If the STABILITY indicator stops illuminating while the sensor is in operation, it is a warning signal for possible detecting error.
2. If the target is in position A parallel to the reflecting plate as shown above, the SENSING indicator may go out and the STABILITY indicator may illuminate depending on the position of the target. In such a case, give priority to the adjustment of the STABILITY indicator when the beam is incident by means of the reflecting plate.

• ENVIRONMENTAL CONDITIONS

1. Because the ultrasonic sensor uses air as a media of sound wave transmission, avoid use of the sensor in a location where the difference in temperature exists locally or in a location subject to significant air convection. Note that malfunctioning of the sensor may be caused by a rapid change in air flow (for example, a blast by an air curtain) which takes place within the operating range of the sensor.

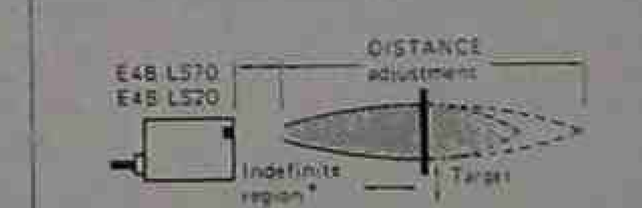
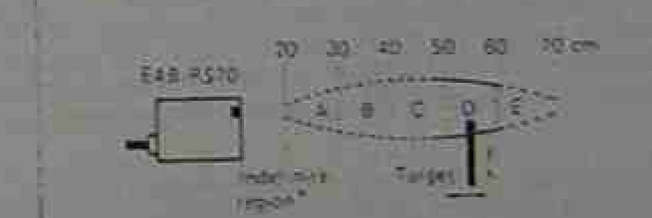


2. Jet sound generating from an air nozzle contains various frequency components which have an adverse effect on the operation of the ultrasonic sensors. Before using the sensor in the vicinity of such sound or similar noises, be sure that the sensor is free of malfunction.
3. With the focusable and selectable zone types, note that these ultrasonic sensors may not be capable of detecting sound-absorbing materials such as fine pulverulent surface cotton.

• CONNECTION

1. When the cable length of the ultrasonic sensor is to be extended, be sure to use a cable having a conductor cross-sectional area of 0.3 mm^2 and keep the extended cable length of 100 m or less.
2. When using a commercially available switching regulator, ground the FG (Frame Ground) and G (Ground) terminals to assure more stabilized sensor operation.

• METHOD OF DETECTION

Separate type (beam break)	Focusable type	Selectable zone type
This type of ultrasonic sensor detects the attenuation or interrupted condition of the ultrasonic beam caused by the target passing between the ultrasonic transmitter and receiver. 	This type of ultrasonic sensor detects only the wave reflected from the target existing within the detecting distance range set by the DISTANCE adjusting VR. 	This type of ultrasonic sensor detects only the wave reflected from the target existing in the detection zone set by the DISTANCE selector switch. 

* The indefinite region is outside the DISTANCE adjustment range. However, the focusable or selectable zone type ultrasonic sensor may be caused to perform a detecting operation by multiple reflection depending on the type of target. The ultrasonic sensor operation becomes unstable in the indefinite region. Therefore, do not use the sensor in that region.

ALL DIMENSIONS SHOWN ARE IN MILLIMETERS.
To convert millimeters into inches multiply by 0.03937. To convert grams into ounces multiply by 0.03527.

LEVEL SENSORS

GLOSSARY

Capacitance (Electrostatic capacity)
Assume that the potential difference between two conductors is V volts and equal positive and negative electric charges are given to each of the two conductors. The potential difference is proportional to the electric charge and is calculated by the following formula:

$$V = \frac{1}{C} \cdot Q$$

where C is a proportional constant that depends on the size, dimensions, arrangement and ambient conditions of the two conductors. This constant is generally called the "capacitance between two conductors" and is represented in the units of coulombs/volt or Farads.

Dielectric constant
Assuming that the electrostatic capacity of a conductor when it exists in the vacuum (virtually equal to air) is C_0 Farads and that of the conductor when it exists in a certain insulator is C_1 Farads, the following relationship exists between C_0 and C_1:

$$C_1 = \varepsilon_s \cdot C_0$$

where ε_s is a proportional constant called the "dielectric constant", which is dependent upon the insulating material.

Electroconductive detectable substances
Those which conduct electricity. For example, water (containing acid, alkali, or salts), etc.

Dielectric detectable substances
Those which conduct almost no electricity. For example, plastic, glass, rubber, etc.

Sensitive electrode
The conductor which forms a capacitance to ground and detects a change in the level of the substance by a change in capacitance. (See Fig. 1).

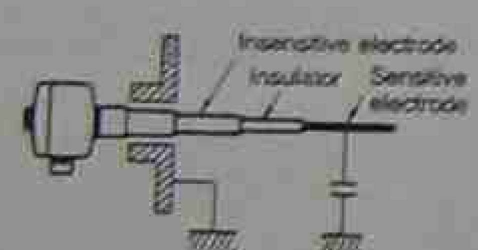


Fig. 1

Resistance inoperative type
The type of level sensor which detects a change in capacitance only, without being affected by the decrease in the resistance between the sensitive and insensitive electrodes caused by the adherence of the substance subject to control, to the electrode section. This type of level sensor also detects electroconductive substances in a stable manner (See Fig. 2.).

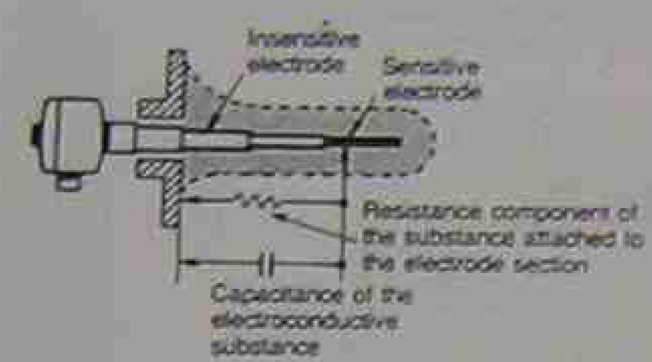


Fig. 2

Insensitive electrode
The grounded electrode section which is connected to a tank through a pipe thread or flange when mounting the switch head section to the tank. The length of the insensitive electrode is selected so that it protrudes from the inner wall of the tank. This prevents unstable switch operation caused by adherence of residual substances to the electrode section. (See Fig. 1.)

Mutual interference
When two or more level sensors are installed in the same tank, the switches may not operate in a stable manner due to insufficient distance between the two electrodes. This condition is referred to as "mutual interference". (See Fig. 3.)

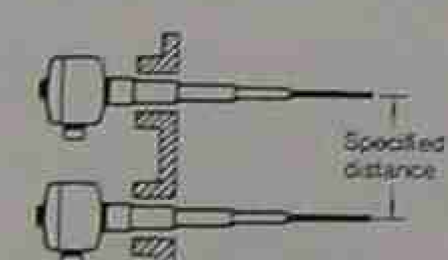


Fig. 3

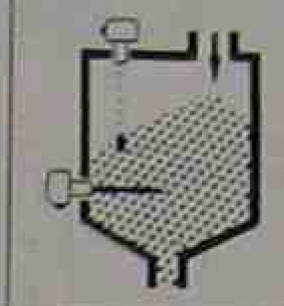
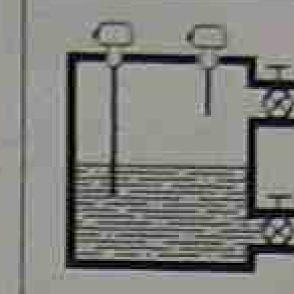
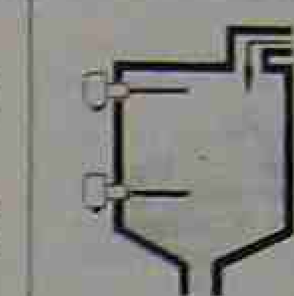
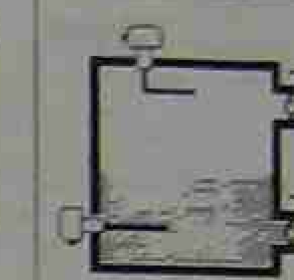
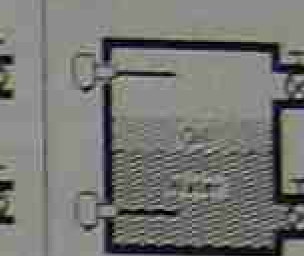
Operating sensitivity
Assuming that the shape and dimensions of the electrode are not changed, the change in capacitance required for the level sensor to operate differs according to the substance subject to level control (i.e., dielectric constant ε_s). Therefore, it is necessary to set in advance the operating point of the level sensor according to the substance to be controlled. This set value is called the "operating sensitivity", which is represented by possible setting range of capacitance.

Stable operating sensitivity
The required minimum value of the change in capacitance for the level sensor to operate, determined by taking into account such factors as variations due to temperature and voltage fluctuations and changes in the dielectric constant of the substances to be detected. This represents the maximum sensitivity which may be set.

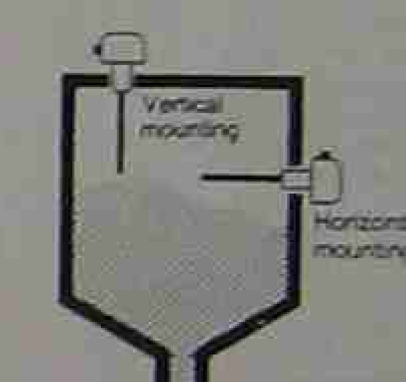
Deoiling (separation of oil and water)
The detection of the boundary between the oil and water is possible since a difference in dielectric constant ε_s exists between these two substances. Therefore, if oil and water are mixed in the same tank, a capacitance type level sensor allows detection of the separation boundary.

Plastic covered electrode
If the material to be detected does not provide adequate resistance between sensitive and insensitive electrodes, covering the sensitive electrode with a plastic material can be used to increase the resistance to an acceptable level.

Electrode Selection According to Substance and Tank Shape

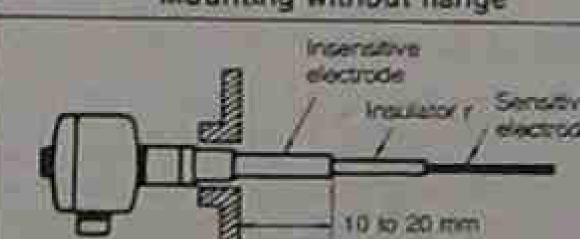
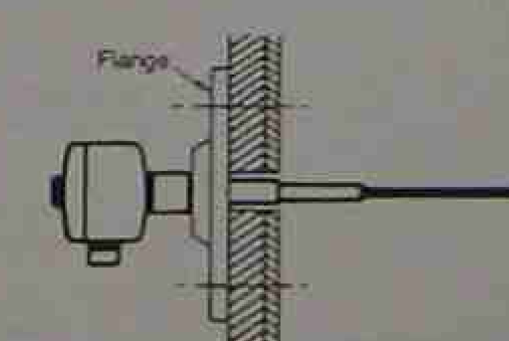
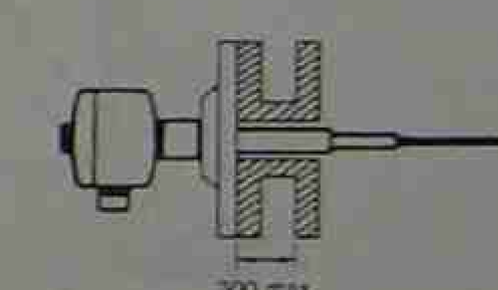
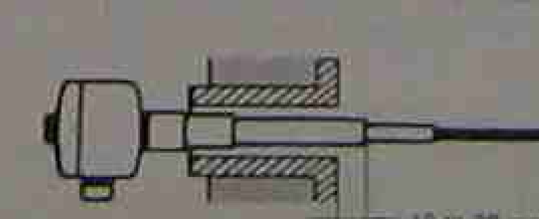
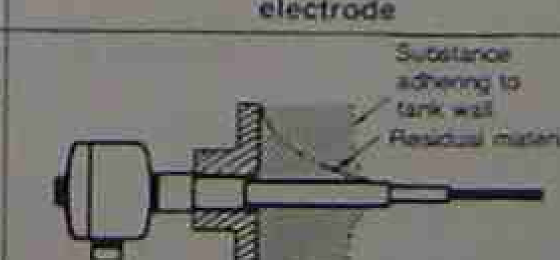
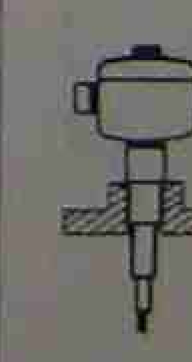
	Fodder	Substances related to the chemical industry	Substances related to the steel industry	Substances related to the food industry	Deoiling (separation of oil and water)
Substance subject to level control	Maize (ε = 2.3 to 2.6) Soybean cake (ε = 2.7 to 2.8) Wheat flour ε = 2.2 to 3.0)	Trichlene (ε = 3.4) Barium nitrate (ε = 5.0 to 6.0) Sulfur (liquid) (ε = 3.4)	Silica sand (ε = 2.5 to 3.5) Aluminum oxide (ε = 2.0 to 2.2) Gravel (ε = 5.4 to 5.6) Ferric oxide (ε = 1.4 to 1.8)	Edible oil (ε = 1.8 to 2.2) Margarine (liquid) (ε = 5.4 to 5.6) Granulated sugar (powdered) (ε = 1.5 to 2.2)	Heavy oil, crude oil, vegetable oil (ε = 2 to 10) Industrial water, sea water, pure water (ε = 70 to 80)
Tank Shape & Mounting location					
Electrode	For upper-limit control: Wire electrode (3,000 mm max.) is most suitable. For lower-limit control: 20-mm dia. load resistant electrode applicable to loads of up to 73 kg (safety factor 4).	Coated electrode (with a coating of ethylene trifluoride resin or hexafluoride ethylene resin) or chemical-resistant electrode (SUS316, hastelloy B)	Load-resistant electrode	Standard electrode or L-shaped electrode (ideal for applications where higher precision is vital)	Standard electrode or coated electrode (where oil contains solvent or corrosive ingredients)

Electrode Installation
The electrode can be mounted in one of two ways: horizontal (side) mounting and vertical (upper surface) mounting.
For the horizontal mounting, the electrode is always parallel with the fluctuating surface of the substance to be detected. This permits accurate level control by providing a greater change in the capacitance of the electrode versus a slight variation of the surface. However, electroconductive substances are likely to adhere to the electrode and can result in unstable switch operation. It is important to minimize the adhesion of such substance to the electrode. A longer insensitive electrode will solve this adhesion problem.
For the vertical mounting, electroconductive substances usually will not adhere to the electrode, thus permitting stable switch operation. However, this method results in reduced detecting accuracy and requires a longer electrodes for the lower-limit control in large tanks.



Vertical mounting
① Reduced electroconductive substance adhesion effects
② Reduced detecting accuracy
③ Longer electrode required for lower-limit control in large tanks

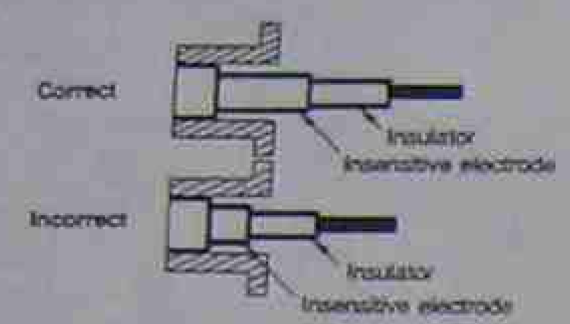
Horizontal mounting
① Increased detecting accuracy
② Reduced stability due to the adhesion of electroconductive substances to the electrode

Mounting without flange	Mounting with flange	Mounting with flange mounting base
 For the electrode mounting section, taper pipe thread (PT3/4) is provided. When mounting the electrode section directly to the tank, use a conduit tube and screw it directly into the pipe thread to attach the electrode section to the tank. In this case, make sure that the insensitive electrode protrudes 10 to 20 mm from the inner wall of the tank.		 Determine the length of the flange pipe section of the flange mounting base in connection with the length of the insensitive electrode.
Mounting without a flange on a thick wall	**Substance is likely to adhere to the electrode**	**Vertical mounting**
 When a longer mounting thread is necessary, be sure to use a longer insensitive electrode accordingly.	 When an excessive amount of residual substance adheres to the electrode section, be sure to extend the length of the insensitive electrode as much as possible	 This method is effective when the tank has a small diameter, the shape of the tank is unfit for horizontal mounting, or when viscous substances are subject to level control.

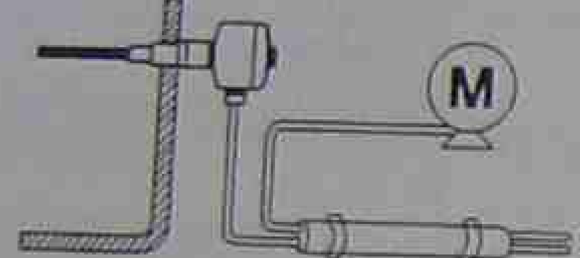
Precautions

When using the electrostatic capacitance type level sensor, pay attention to the following items:

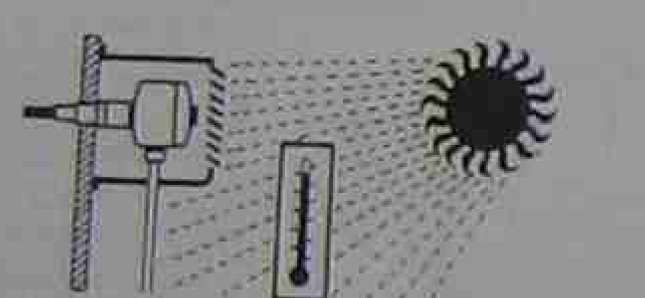
Install the level sensor so that the insensitive electrode protrudes from the inner wall of the tank.



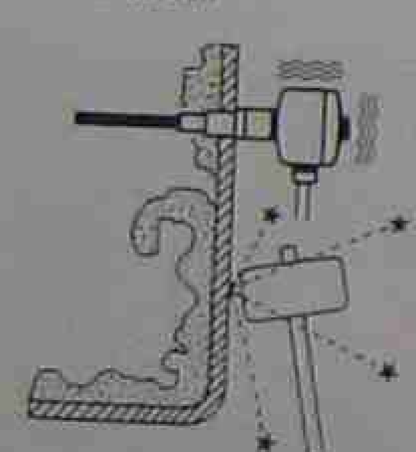
The level sensor is provided with adequate protection against surge current. However, to improve swtich reliability, isolate the switch cable by not routing it through the same duct or conduit tube as the power line.



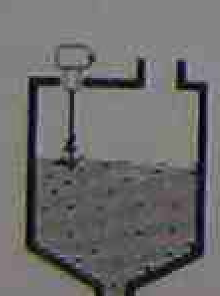
For outdoor applications, be sure to use a cover to prevent the temperature of the switch circuitry from rising as a result of direct sunlight.



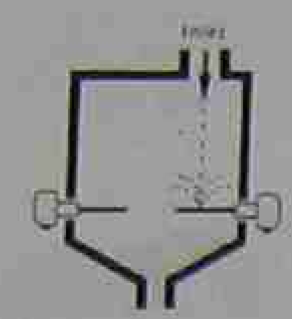
Before striking the tank wall with a wooden mallet to remove the substance adhering to it, remove the main circuit section (i.e., head section) of the level sensor.



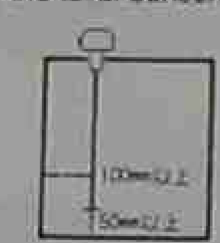
When using the level sensor for the level control of substances that tend to leave deposits such as caustic soda, lime water, paint, etc., perform periodic cleaning and inspection of the electrode.



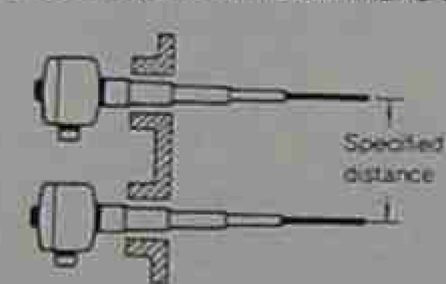
Avoid installing the level sensor in locations where the electrode is directly exposed to the inflow of the substance subject to level control.



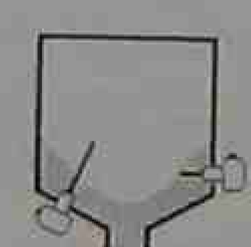
Allow for the following distances between the side and bottom walls of the tank and the level sensor.



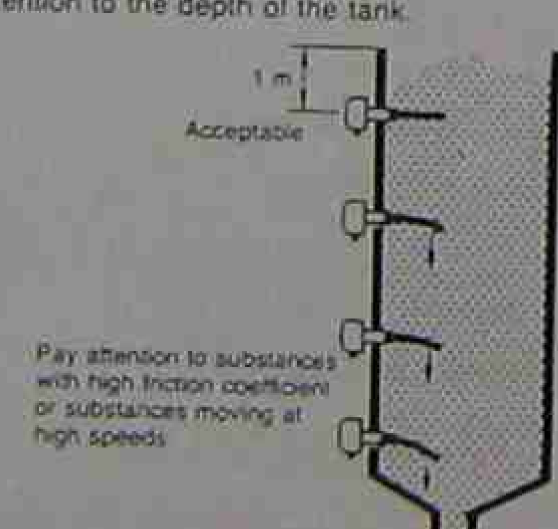
When two or more level sensors are to be used in the same tank, be sure to provide the specified distance between the two sensitive electrodes to prevent mutual interference.



If the substance to be detected is likely to adhere to the bottom or inner wall of the tank, install the level sensor upside down as shown below, instead of horizontally.

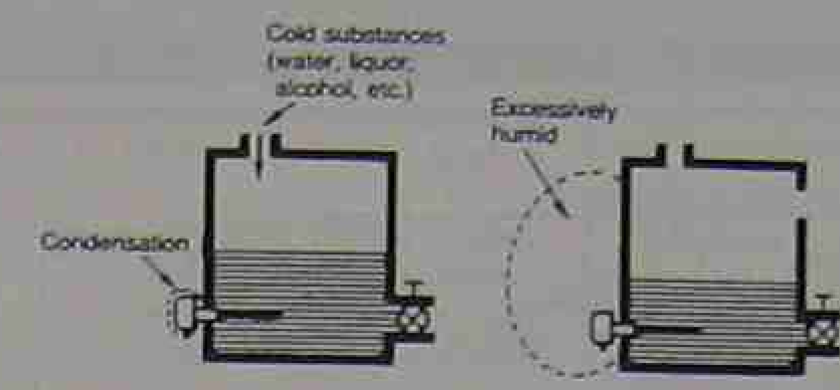


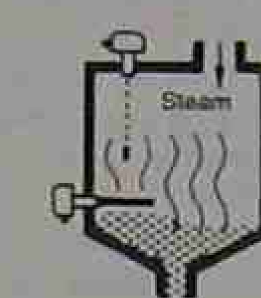
Pay attention to the depth of the tank.



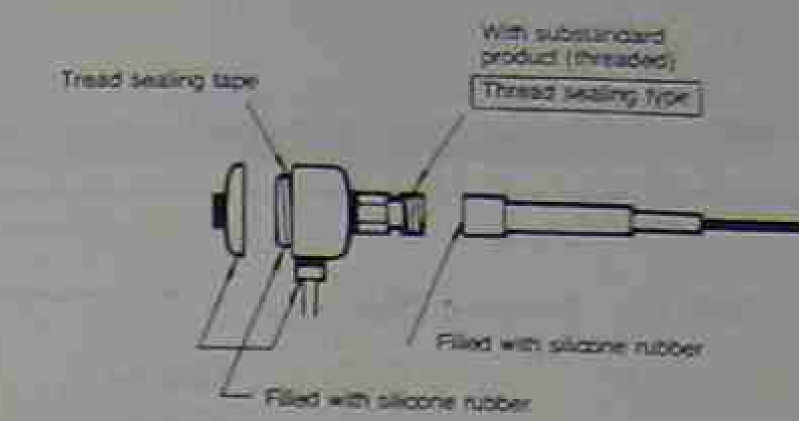
In locations where vapor is present, a type of level sensor with adequate protection against respiration is necessary.

The following prcautions are necessary for quick cooling and low-temperature substances (with a difference of 10°C from ambient temperature) and excessively humid condition.



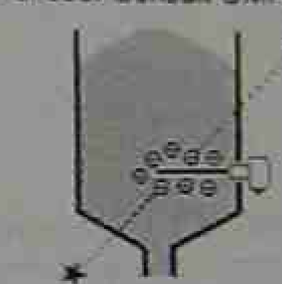


Countermeasures



If the substance to be detected is a highly corrosive liquid, use a level switch with a polyacetar insulator, a silicone rubber packing, and SUS304 electrode.
Models with a teflon packing and teflon-coated electrode are also available. Consult OMRON for detail.

If the substance to be detected generates significant static electricity, such as plastic, use of an improved model of level sensor is recommended. Consult OMRON for details.



Specific Inductive Capacities of Selected Materials

Material	ε_s	Material	ε_s	Material	ε_s
Air	1.000264	Gasoline	2.0 to 2.2	Polyester resin	2.8 to 8.1
Carbon dioxide	1.000985	Paraffin	1.9 to 2.5	Polyacetal	3.6 to 3.7
Water	80	Vaseline	2.2 to 2.9	Styrene resin	2.3 to 3.4
Aqueous solution	50 to 80	Glass	3.7 to 10	Polyvinyl chloride resin	2.8 to 3.1
Glycerine	47	Porcelain	5 to 7	Acrylic resin	2.7 to 4.5
Methanol	33	Salt	6	Tetrafluoroethylen resin	2.0
Ethylene glycol	38.7	Sand	3 to 5	Silicon varnish	2.8 to 3.3
Nitrobenzene	36	Sulfur	3.4	Shellac varnish	2.5 to 4.7
Ethanol	24	Shell lime	1.2	Sugar	3.0
Ammonia	12 to 25	Soybean oil	2.9 to 3.5	Cereal	3 to 5
Acetone	19.3	Liquid air	1.5	Wood	2 to 6
Carbon tetrachloride	2.2	Liquid carbon dioxide	1.6		(10 to 30 when wet)
Aniline	6.9	Liquid Chlorine	2.0	Press boad	2 to 5
Toluene	2.3	Phenol resin	4 to 12	Flied ash	1.5 to 1.7
Benzene	2.3	Urea resin	5 to 8	Polypropylene	2.0 to 2.2
Propane liquid	1.6 to 1.9	Metamin resin	4.7 to 10.2	Nylon	4 to 5
Petroleum	2.0 to 2.2	Epoxy resin	2.5 to 6	Flour	2.5 to 3.0

Ordering Information for Level Sensor
The applicable model depends upon the methods of installation and detection.
(When you place your first order, be careful to specify the model number correctly because if the wrong model is used to detect an inappropriate substance, the level sensor may mulfunction. This also applies to the selection of the substance to be controlled by the switch since inappropriate substances may cause some level sensors to mulfunction (e. g. electrode adhesion).

Refer to the above table to determine your choice of a level sensor and the substances it can be used with. For further information, consult OMRON.

Industrial Sensors.

Topic -

Table 9.2 Dielectric Constants

Material	Dielectric Constant	Material	Dielectric Constant
Cereal	3–5	Salt	6
Ethanol	24	Sand	3–5
Flour	2.5–3	Sugar	3
Gasoline	2.2	Water	80
Nylon	4–5	Wood (dry)	2–6
Paper	1.6–2.6	Wood (wet)	10–30

Table 9.1 Representative Specifications for Inductive Proximity Sensors

Sensing Range (mm)	Diameter (mm)	Length (mm)	Switching Speed (Hz)	Typical Part Number
1	8	40	5000	IP–1
2	12	40	1000	IP–2
5	18	40	400	IP–3
10	30	50	200	IP–4
20	47	60	40	IP–5

Ultrasonic Detection

TABLE 3-2 ACOUSTIC DATA FOR SEVERAL COMMON MATERIALS

Material	Sound velocity (m/s)	Mass density (kg/m)	Acoustic impedance, Z (kg/m-s)
Air	332	1.281	425.3
Aluminum	5102	2643	13.48×10^6
Brass	3499	8553	29.93×10^6
Copper	3557	8906	31.68×10^6
Hydrogen	1269	0.090	114.2
Iron	5000	7100	35.50×10^6
Clay, ceramic	3000–5000	1500–2500	4.5×10^6–12.5×10^6
Water	1461	998	1.46×10^6
Wood	3048–4572	480–800	1.5×10^6–3.7×10^6

8.6 Material characteristics

TURCK

Practical excess gain calculations

The maximum range of a sensor in a particular situation can be estimated from the excess gain curves and applying a simple formula. Two examples are given below.

For a **diffuse reflective sensor** (object sensors), the maximum range is dependent on firstly the reflectivity (table 1), and secondly the environmental conditions (table 2).

The Multibeam sensor MB14-LU1-NPEX has been chosen for this example. Its excess gain curve is given in fig. 16.

Printed cardboard was selected as the object to detect, and the environment is slightly dirty.

Figures obtained from both tables are:
Reflectivity 70% - factor 1.3
Dirt factor 5

Both factors multiplied together give the standard value for determining maximum operating range from the excess gain curve:

1.3 x 5 = factor 6.5

The maximum range of the chosen sensor under these conditions is therefore
85 cm at factor 6.5

The **opposed mode** of photoelectric detection offers very high integrity even when operating in unfavourable environments. This is achieved because the systems feature high excess gain and is totally unaffected by object reflectivity.

The excess gain curve for an opposed sensor (MB 52/62) is given in fig. 17. Its maximum sensing range under optimised conditions is 33 m at factor 2.

If we consider the environment to be very dirty, table 2 gives us a factor of 50. From the excess gain curve this factor gives us a maximum range of 6 m.

Table 1

Material	Reflectivity	Factor for excess gain
Kodak test card	90 %	1
white paper	80 %	1.1
newspaper (with print)	55 %	1.6
tissue paper	47 %	1.9
cardboard	70 %	1.3
sone-pine wood, clean	75 %	1.2
rough wood pallet, clean	20 %	4.5
beer foam	70 %	1.3
clear plastic bottle	40 %	2.3
translucent (brown) plastic bottle	60 %	1.5
opaque white plastic	87 %	1.0
opaque black plastic	14 %	6.4
black neoprene	4 %	22.5
black foam carpet backing	2 %	45
black rubber tire wall	1.5 %	60
natural aluminium unfinished	140 %	0.6
natural aluminium straightlined	105 %	0.9
black anodized aluminium, unfinished	115 %	0.8
black anodized aluminium, straightlined	50 %	1.8
stainless steel, microfinish	400 %	0.2
cork	35 %	2.9

Table 2

correction factor	environmental conditions
1.5 x	clean air no dirt built-up on lenses or reflectors
5 x	slightly dirty by mist, dust, oil on lenses or reflectors (lenses are cleaned on a regular schedule)
10 x	moderately dirty by mist, dust, oil on lenses or reflectors (lenses are cleaned occasionally or when necessary)
50 x	very dirty by heavy fog, dust, smoke or oil film (minimal or no cleaning of lenses)

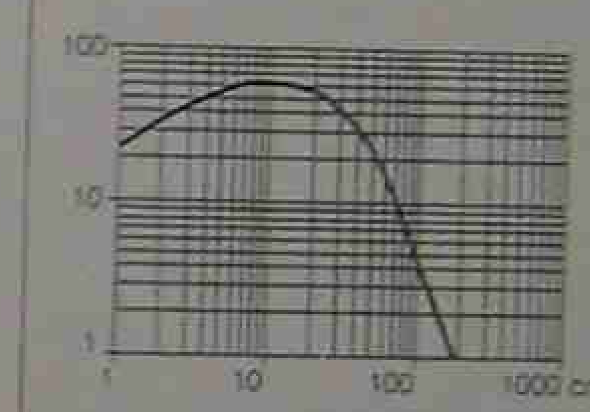


Fig. 16 Excess gain curve MB14-LU1-NPEX

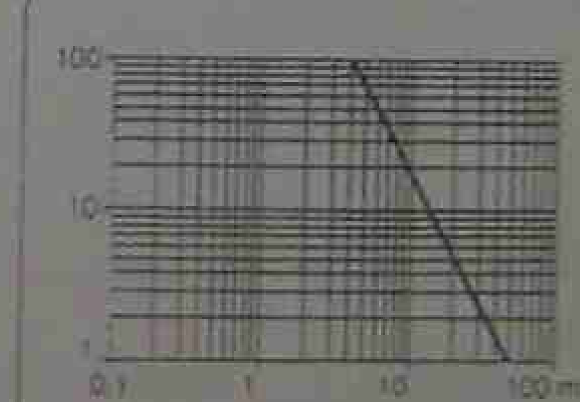


Fig. 17 Excess gain curve MB52/62

Photoelectric

Sydney Institute of Technology

School of Electrotechnology

Industrial Sensors

Theory Notes

Area ____________________

Topic ____________________

Session No.

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

7-6 INSTRUMENTATION AMPLIFIER

In many industrial and consumer applications the measurement and control of physical conditions are very important. For example, measurements of temperature and humidity inside a dairy or meat plant permit the operator to make necessary adjustments to maintain product quality. Similarly, precise temperature control of a plastic furnace is needed to produce a particular type of plastic.

Generally, a transducer is used at the measuring site to obtain the required information easily and safely. The *transducer* is a device that converts one form of energy into another. For example, a strain gage when subjected to *pressure* or force (physical energy) undergoes a change in its *resistance* (electrical energy). An instrumentation system is used to measure the output signal produced by a transducer and often to control the physical signal producing it. Figure 7-11 shows a simplified form of such a system. The input stage is composed of a preamplifier and some sort of transducer, depending on the physical quantity to be measured. The output stage may use devices such as meters, oscilloscopes, charts, or magnetic recoders.

In Figure 7-11 the connecting lines between the blocks represent *transmission lines*, used especially when the transducer is at a remote test site monitoring hazardous conditions such as high temperatures or the liquid levels of flammable chemicals. These transmission lines permit signal transfer from unit to unit. The length of the transmission lines depends primarily on the physical quantities to be monitored and on system requirements.

The signal source of the instrumentation amplifier is the output of the transducer. Although some transducers produce outputs with sufficient strength to permit their use directly, many do not. To amplify the low-level output signal of the transducer so that it can drive the indicator or display is the major function of the *instrumentation amplifier*. In short, the instrumentation amplifier is intended for precise, low-level signal amplification where low noise, low thermal and time drifts, high input resistance, and accurate closed-loop gain are required. Besides, low power consumption, high common-mode rejection ratio, and high slew rate are desirable for superior performance.

There are many instrumentation operational amplifiers, such as the μA725, ICL7605, and LH0036, that make a circuit extremely stable and accurate. These ICs are, however, relatively expensive; they are very precise special-purpose circuits in which most of the electrical parameters, such as offsets, drifts, and power consumption, are minimized, whereas input resistance, CMRR, and supply range are optimized. Some instrumentation amplifiers are even available in modular form to suit special installation requirements.

Obviously, the requirements for instrumentation op-amps are more rigid than those for general-purpose applications. However, where the requirements are not too strict, the general-purpose op-amp can be employed in the differential mode. We will call such amplifiers *differential instrumentation amplifiers*. Since most of the instrumentation systems use a transducer in a bridge circuit, we will consider a simplified differential instrumentation system arrangement using a transducer bridge circuit.

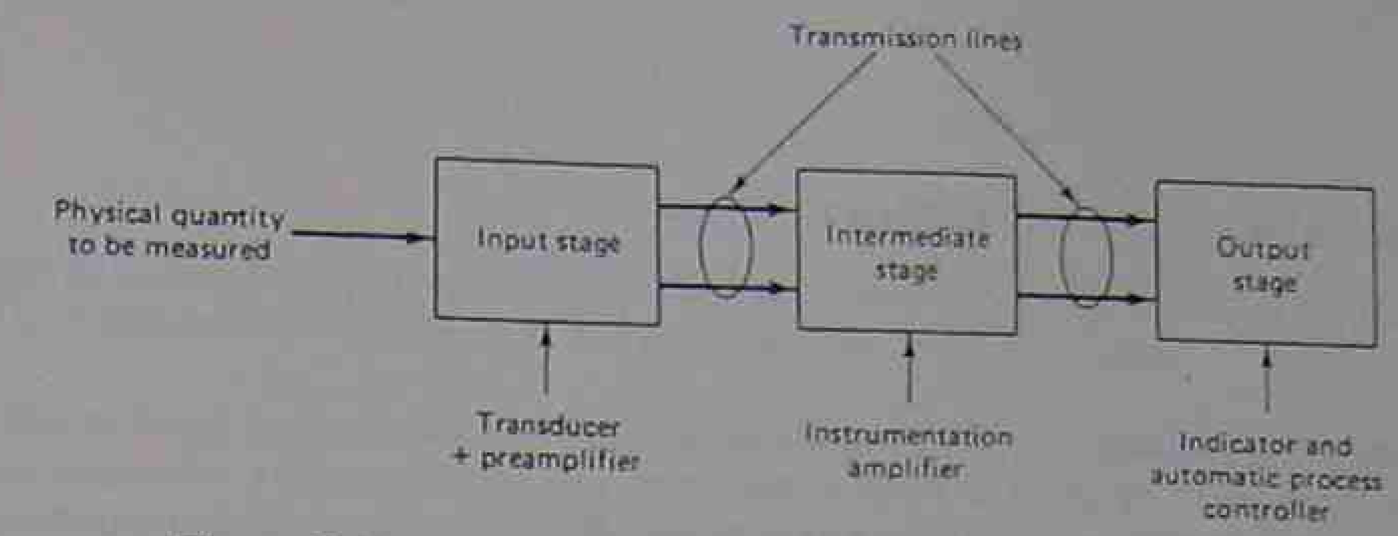


Figure 7–11 Block diagram of an instrumentation system.

7-6.1 Instrumentation Amplifier Using Transducer Bridge

Figure 7-12 shows a simplified differential instrumentation amplifier using a transducer bridge. A *resistive transducer* whose resistance changes as a function of some physical energy is connected in one arm of the bridge with a small circle around it and is denoted by $(R_T \pm \Delta R)$, where R_T is the resistance of the transducer and ΔR the change in resistance R_T.

The bridge in the circuit of Figure 7-12 is dc excited but could be ac excited as well. For the balanced bridge at some reference condition,

$$V_b = V_a$$

or, IN TERMS OF THE RESISTORS & V_{DC}

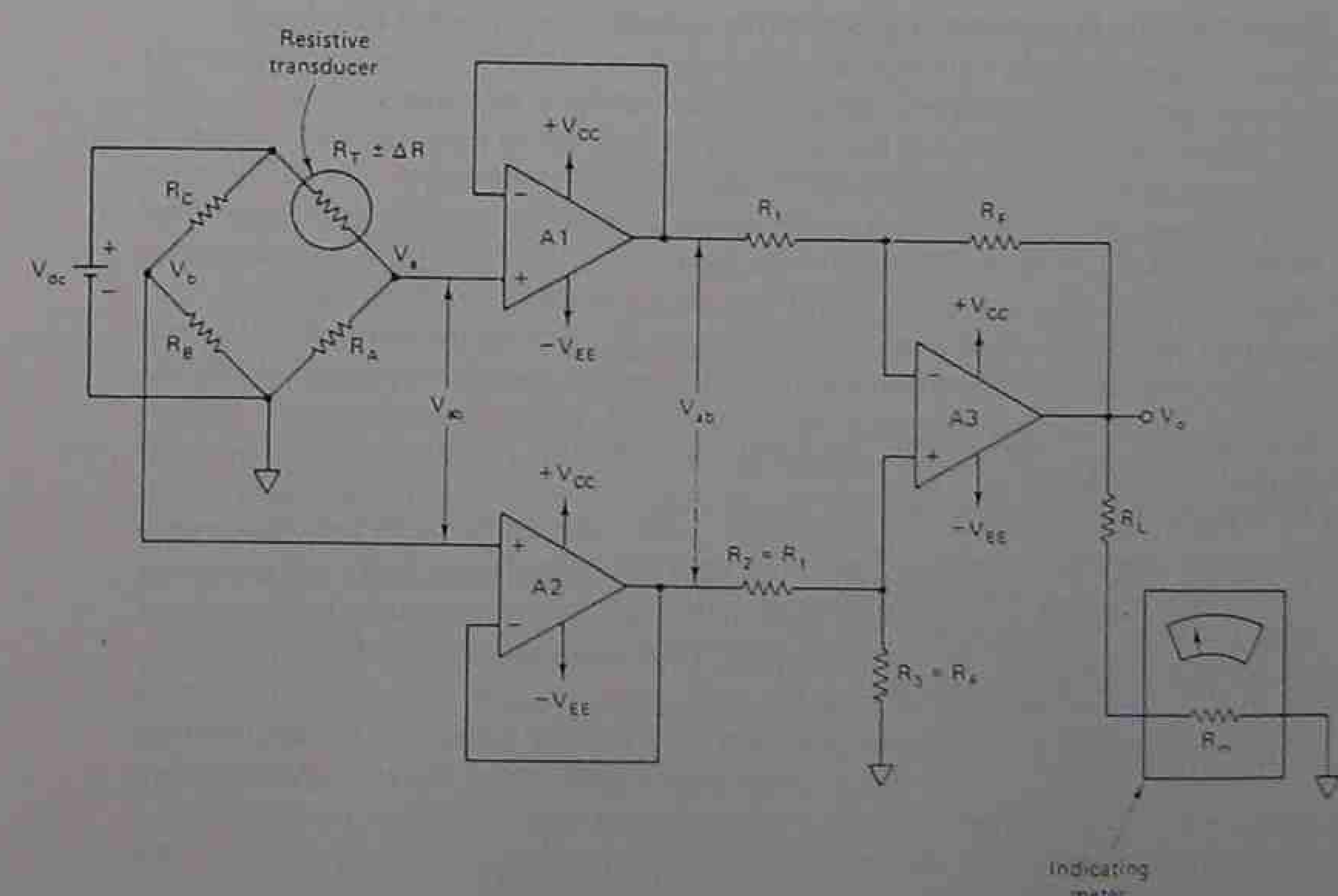


Figure 7–12 Differential instrumentation amplifier using a transducer bridge.

That is, A RATIO OF RESISTORS:—

(7-13)

Generally, resistors R_A, R_B, and R_C are selected so that they are equal in value to the transducer resistance R_T at some *reference condition*. The reference condition is the specific value of the physical quantity under measurement at which the bridge is balanced. This value is normally established by the designer and depends on the transducer's characteristics, the type of physical quantity to be measured, and the desired application.

The bridge is balanced initially at a desired reference condition. However, as the physical quantity to be measured changes, the resistance of the transducer also changes, which causes the bridge to unbalance ($V_a \neq V_b$). The output voltage of the bridge can be expressed as a function of the change in resistance of the transducer, as described next.

Let the change in resistance of the transducer be ΔR. Since R_B and R_C are fixed resistors, the voltage V_b is constant. However, voltage V_a varies as a function of the change in transducer resistance. Therefore, according to the voltage-divider rule,

$$V_a = \underline{\qquad\qquad}$$

$$V_b = \underline{\qquad}$$

Consequently, the voltage V_{ab} across the output terminals of the bridge is

$$V_{ab} = V_a - V_b$$

$$= \underline{\qquad\qquad} - \underline{\qquad}$$

However, if $R_A = R_B = R_C = R_T = R$, then

$$V_{ab} = - \underline{\qquad\qquad} \qquad (7\text{-}14)$$

The negative (−) sign in this equation indicates that $V_a < V_b$ because of the increase in the value of R_T.

The output voltage V_{ab} of the bridge is then applied to the differential instrumentation amplifier composed of three op-amps (see Figure 7-12). The voltage followers preceding the basic differential amplifier help to eliminate loading of the bridge circuit. The gain of the basic differential amplifier is (− /); therefore, the output V_o of the circuit is

$$V_o = V_{ab}\left(-\,\frac{\quad}{\quad}\right) = \underline{\qquad\qquad}\ \text{—} \qquad (7\text{-}15a)$$

Generally, the change in resistance of the transducer ΔR is very small. Therefore, we can approximate $(2R + \Delta R) \cong 2R$. Thus, the output voltage,

$$V_o = \frac{\quad}{\quad} V_{dc} \qquad (7\text{-}15b)$$

This equation indicates that V_o is directly proportional to the change in resistance ΔR of the transducer. Since the change in resistance is caused by a change in physical energy, a meter connected at the output can be calibrated in terms of the units of that physical energy.

Before proceeding with specific bridge applications, let us briefly consider the important characteristics of some resistive types of transducers. In these resistive types of transducers the resistance of the transducer changes as a function of some physical quantity. The thermistors, photoconductive cells, and strain gages are some of the most commonly used resistive transducers; hence they will be further discussed here.

Thermistors are essentially semiconductors that behave as resistors, usually with a *negative temperature coefficient of resistance*. That is, as the temperature of a thermistor increases, its resistance decreases. The temperature coefficient of resistance is expressed in ohms per unit change in degrees Celsius (°C). Thermistors with a high temperature coefficient of resistance are more sensitive to temperature change and are therefore well suited to temperature measurement and control. Thermistors are available in a wide variety of shapes and sizes; However, thermistor beads sealed in the tips of glass rods are most commonly used because they are relatively easy to mount.

The photoconductive cell belongs to the family of photodetectors (photosensitive devices) whose resistance varies with an incident radiant energy or with light. As the intensity of incident light increases, the resistance of the cell decreases. The resistance of the photoconductive cell in darkness is typically on the order of 100 kΩ. Generally, the resistance of the cell in darkness and at particular light intensities is listed on the data sheet. The intensity of light is expressed in meter candles (lux).

Materials such as cadmium sulfide and silicon, whose conductivity is a function of incident radiant energy, are used for photoconductive cells. Some cells are extremely sensitive to light and hence can be used into the ultraviolet and infrared regions. The photoconductive cell is typically composed of a ceramic base, a layer of photoconductive material, a moisture-proof enclosure, and metallic leads. Photoconductive cells are also known as photocells or light-dependent resistors (LDRs).

Another important resistive transducer is the strain gage, whose resistance changes due to elongation or compression when an external stress is applied. The stress is defined as force per unit area [newtons/(meter)^2] and can be related to pressure, torque, and displacement. Therefore, a strain gage may be used to monitor change in applied pressure, torque, and displacement by measuring the corresponding change in the gage's resistance.

Two basic types of strain gages are the *wire* and *semiconductor*. Semiconductor strain gages are much more sensitive than the wire type and therefore provide better accuracy and resolution. The *sensitivity* of a strain gage is defined as unit change in resistance per unit change in length and is a dimensionless quantity.

The thermistor, photocell, and strain gage are all passive transducers, meaning that they require external voltage (ac or dc) for their operation.

7-6.1(a) Temperature indicator. The circuit of Figure 7-12 can be used as a temperature indicator if the transducer in the bridge circuit is a thermistor and the output meter is calibrated in degrees Celsius or Fahrenheit. The bridge can be balanced at a desired reference condition, for instance 25°C. As the temperature varies from its reference value, the resistance of the thermistor changes and the bridge becomes unbalanced. This unbalanced bridge in turn produces the meter movement. The meter can be calibrated to read a desired temperature range by selecting an appropriate gain for the differential instrumentation amplifier. In Figure 7-12 the meter movement is dependent on the amount of imbalance in the bridge, that is, the change ΔR in the value of the thermistor resistance. The ΔR for the thermistor, however, can be determined as follows: ΔR = (temperature coefficient of resistance) (final temperature—reference temperature.) See Example 7-7.

EXAMPLE 7-7

In the circuit of Figure 7-12, $R_1 = 1\ \text{k}\Omega$, $R_F = 4.7\ \text{k}\Omega$, $R_A = R_B = R_C = 100\ \text{k}\Omega$, $V_{dc} = +5$ V, and op-amp supply voltages $= \pm 15$ V. The transducer is a thermistor with the following specifications: $R_T = 100\ \text{k}\Omega$ at a reference temperature of 25°C; temperature coefficient of resistance $= -1\ \text{k}\Omega/°\text{C}$ or 1%/°C. Determine the output voltage at 0°C and at 100°C.

SOLUTION At 25°C, $R_A = R_B = R_C = R_T =$ ...

Obviously, the requirements for instrumentation op-amps are more rigid than those for general-purpose applications. However, where the requirements are not too strict, the general-purpose op-amp can be employed in the differential mode. We will call such amplifiers *differential instrumentation amplifiers*. Since

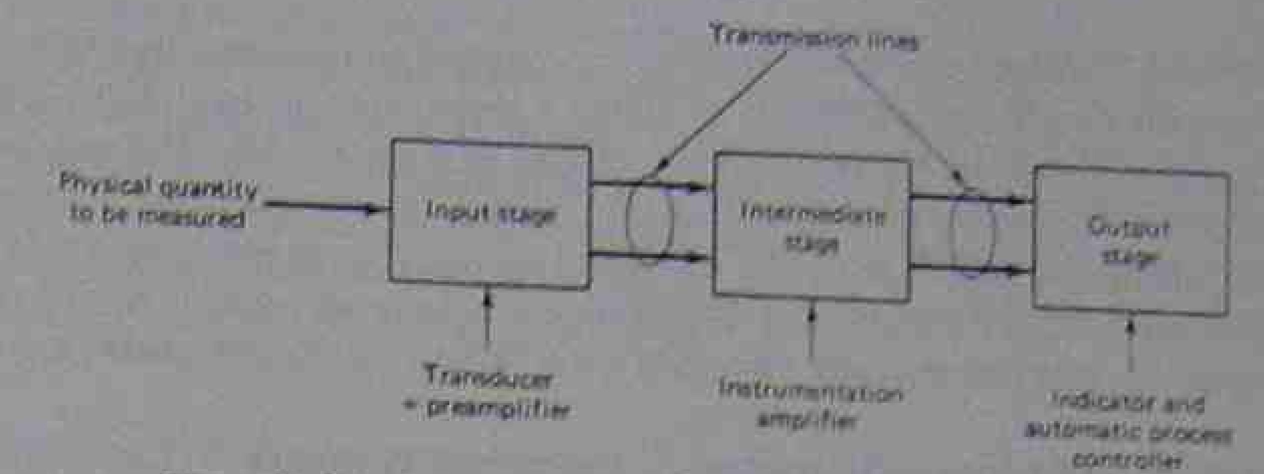


Figure 7-11 Block diagram of an instrumentation system.

most of the instrumentation systems use a transducer in a bridge circuit, we will consider a simplified differential instrumentation system arrangement using a transducer bridge circuit.

7-6.1 Instrumentation Amplifier Using Transducer Bridge

Figure 7-12 shows a simplified differential instrumentation amplifier using a transducer bridge. A *resistive transducer* whose resistance changes as a function of some physical energy is connected in one arm of the bridge with a small circle around it and is denoted by $(R_T \pm \Delta R)$, where R_T is the resistance of the transducer and ΔR the change in resistance R_T.

The bridge in the circuit of Figure 7-12 is dc excited but could be ac excited as well. For the balanced bridge at some reference condition,

$$V_b = V_a$$

or

$$\frac{R_B(V_{dc})}{R_B + R_C} = \frac{R_A(V_{dc})}{R_A + R_T}$$

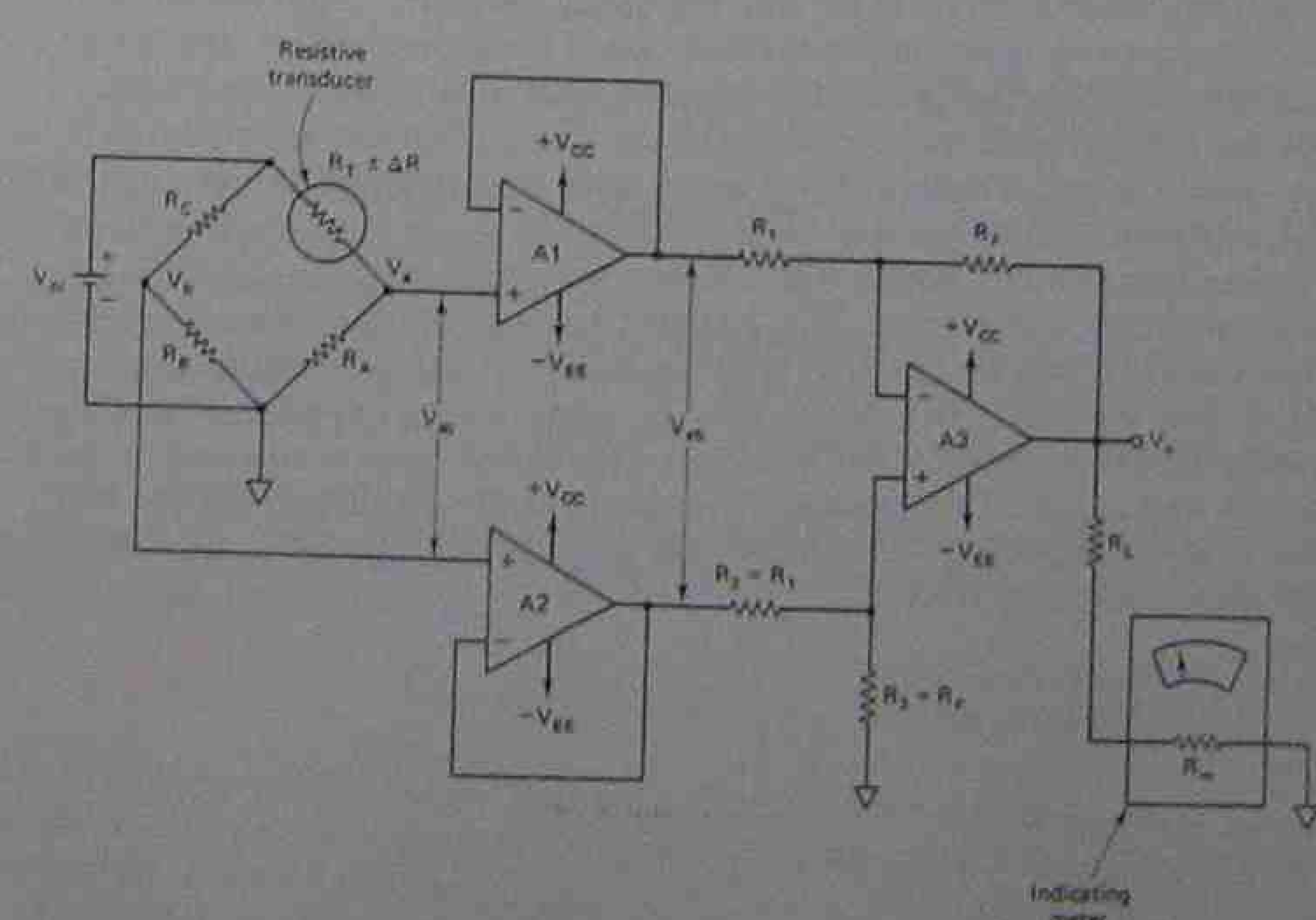


Figure 7-12 Differential instrumentation amplifier using a transducer bridge.

7-6.1(b) Temperature controller. A simple and inexpensive temperature control circuit may be constructed by using a thermistor in the bridge circuit and by replacing a meter with a relay in the circuit of Figure 7-12. The output of the differential instrumentation amplifier drives a relay that controls the current in the heat-generating circuit. A properly designed circuit should energize a relay when the temperature of the thermistor drops below a desired value, causing the heat unit to turn on, BUT CONTINUES TO HEAT TO A PRESET HIGHER TEMP, I.E. HYSTERESIS.

7-6.1(c) Light-intensity meter. The circuit in Figure 7-12 can be used as a light-intensity meter if a transducer is a photocell. The bridge can be balanced for darkness conditions. Therefore, when exposed to light, the bridge will be unbalanced and cause the meter to deflect. The meter can be calibrated in terms of *lux* to measure the change in light intensity.

The light-intensity meter using an instrumentation bridge amplifier is more accurate and stable than single-input inverting or noninverting configurations because the common-mode (noise) voltages are effectively rejected by the differential configuration.

7-6.1(d) Measurement of flow and thermal conductivity. A *flow meter* or a *thermal conductivity meter* may be constructed using the circuit of Figure 7-12, provided that two thermistors are used adjacent to each other in the bridge. For instance, assume that R_C and R_T represent two identical thermistors. For the flow measurement, one thermistor is sealed in a small-cavity copper cylinder and the other installed in a small copper pipe. When no air flows through the pipe, the bridge can be balanced so that the output voltage is zero. When air flows over the thermistor, its temperature decreases and in turn the resistance increases, causing the bridge to be unbalanced. This unbalanced voltage is then amplified by the differential instrumentation amplifier and applied to the meter. Thus the amount of meter deflection is proportional to the flow rate of the air in the pipe. The meter can be calibrated in (meter)^3/second to accommodate a desired flow rate range.

For the thermal conductivity measurement, two thermistors are mounted in separate small copper cylinders. With air in both cylinders, the bridge can be balanced, and hence the output will be zero. When air in one cylinder is replaced by a medium being tested, which has a different thermal conductivity than air, the bridge becomes unbalanced. This happens because the temperature of the thermistor changes, thus changing its resistance. Suppose that the medium being tested is carbon dioxide (CO_2). Because of its lower thermal conductivity, CO_2 will increase the temperature of the thermistor, which will cause a decrease in the thermistor's resistance. This results in unbalancing the bridge, which in turn produces a meter deflection. The meter can be calibrated in terms of relative thermal conductivity (cal/s-cm-°C).

7-6.1(e) Analog weight scale. By connecting a strain gage in the bridge, the circuit of Figure 7-12 can be converted into a simple and inexpensive *analog weight scale*.

In the analog weight scale, strain gage elements are connected in all four arms of the bridge. The elements are mounted on the base of the weight platform so that, when an external force or weight is applied to the platform, one pair of elements in the opposite arms elongates, whereas the other pair of elements in the opposite arms compresses. In other words, when the weight is placed on the platform, R_{T1} and R_{T3} both decrease in resistance, and R_{T2} and R_{T4} both increase in resistance, or vice versa (see Figure 7-13).

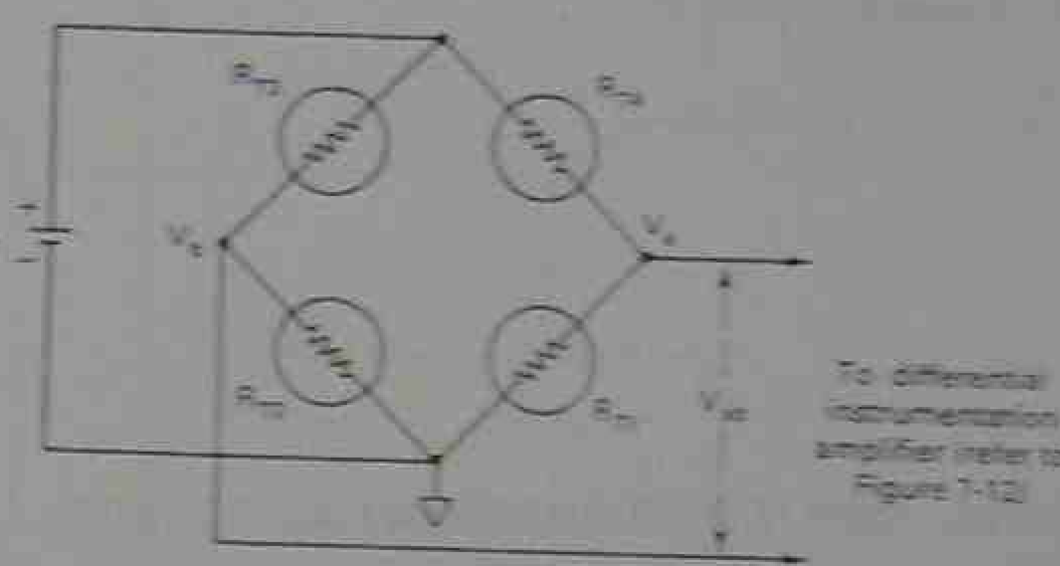


Figure 7–13 Strain-gage bridge circuit for analog weight scale. $R_{T1} = R_{T2} = R_{T3} = R_{T4}$.

When no weight is placed on the platform, the bridge is balanced, $R_{T1} = R_{T2} = R_{T3} = R_{T4} = R$, and the output voltage of the weight scale can be zero. When a weight is placed on the scale platform, the bridge becomes unbalanced. Assuming that R_{T1} and R_{T3} decrease in resistance and R_{T2} and R_{T4} increase in resistance by the same number of ohms ΔR, the unbalanced voltage V_{ab} is given by

$$V_{ab} = -V_{dc}\left(\text{—}\right) \qquad (7\text{-}16)$$

IN TERMS OF ΔR & R, AND VDC.

where V_{dc} = dc excitation voltage of the bridge
$R = R_{T1} = R_{T2} = R_{T3} = R_{T4}$ = unstrained gage resistance
ΔR = change in gage resistance

Remember that, if the decrease in gage resistance R_{T1} and R_{T3} is ΔR, the increase in resistance of R_{T2} and R_{T4} is also ΔR. Therefore, with this assumption, the voltage $V_a < V_b$ and the output voltage V_{ab} is negative, as indicated by Equation (7-16).

The voltage V_{ab} is then amplified by the differential instrumentation amplifier, which drives the meter. Since the gain of the amplifier is (– /), the output voltage V_o is

$$V_o = V_{dc}\left(\text{—}\right)\text{—} \qquad (7\text{-}17)$$

IN TERMS OF ΔR, R, VDC, & GAIN.

The gain of the amplifier can be selected according to the sensitivity of the strain gage and the full-scale deflection requirements of the meter. The meter can be calibrated in terms of kilograms.

For better accuracy and resolution, a microprocessor-based digital weight scale may be constructed. However, such a scale is much more complex and expensive than the analog scale.

EXAMPLE 7-8

The circuit of Figure 7-12 is used as an analog weight scale with the following specifications. The gain of the differential instrumentation amplifier = -100. Assume that $V_{dc} = +10$ V and that the op-amp supply voltages = ± 10 V. The unstrained resistance of each of the four elements of the strain gage is 100 Ω. When a certain weight is placed on the scale platform, the output voltage V_o = 1 V. Assuming that the output is initially zero, determine the change in the resistance of each strain-gage element.

SOLUTION Using Equation (7-17),

7-7 DIFFERENTIAL INPUT AND DIFFERENTIAL OUTPUT AMPLIFIER

In all the applications discussed so far, the op-amp is used with a single-ended or unbalanced output. However, in certain applications a differential output is required. The differential input and differential output amplifier is most commonly used as a preamplifier and in driving push-pull arrangements.

Figure 7-14(a) shows one possible arrangement of a differential input and differential output amplifier using two identical op-amps, that is, a dual op-amp. The connection diagrams of the 8-pin mini DIP and 14-pin DIP are shown in Figure 7-14(b). The analysis of the circuit in Figure 7-14(a) can be accomplished by determining the output of each op-amp due to the differential input. Using the superposition theorem, the output V_{ox} due to inputs V_x and V_y is

(7-18a)

(7-18b)

$$\boxed{V_o = \left(1 + \frac{2R_F}{R_1}\right) V_{in}} \qquad (7\text{-}18c)$$

This means that the differential input and output are in phase or of the same polarity provided that $V_{in} = V_x - V_y$ and $V_o = V_{ox} - V_{oy}$.

The differential input and output amplifier of Figure 7-14(a) is very useful in noisy environments, especially if the input signal is relatively smaller, because it rejects the common-mode noise voltages.

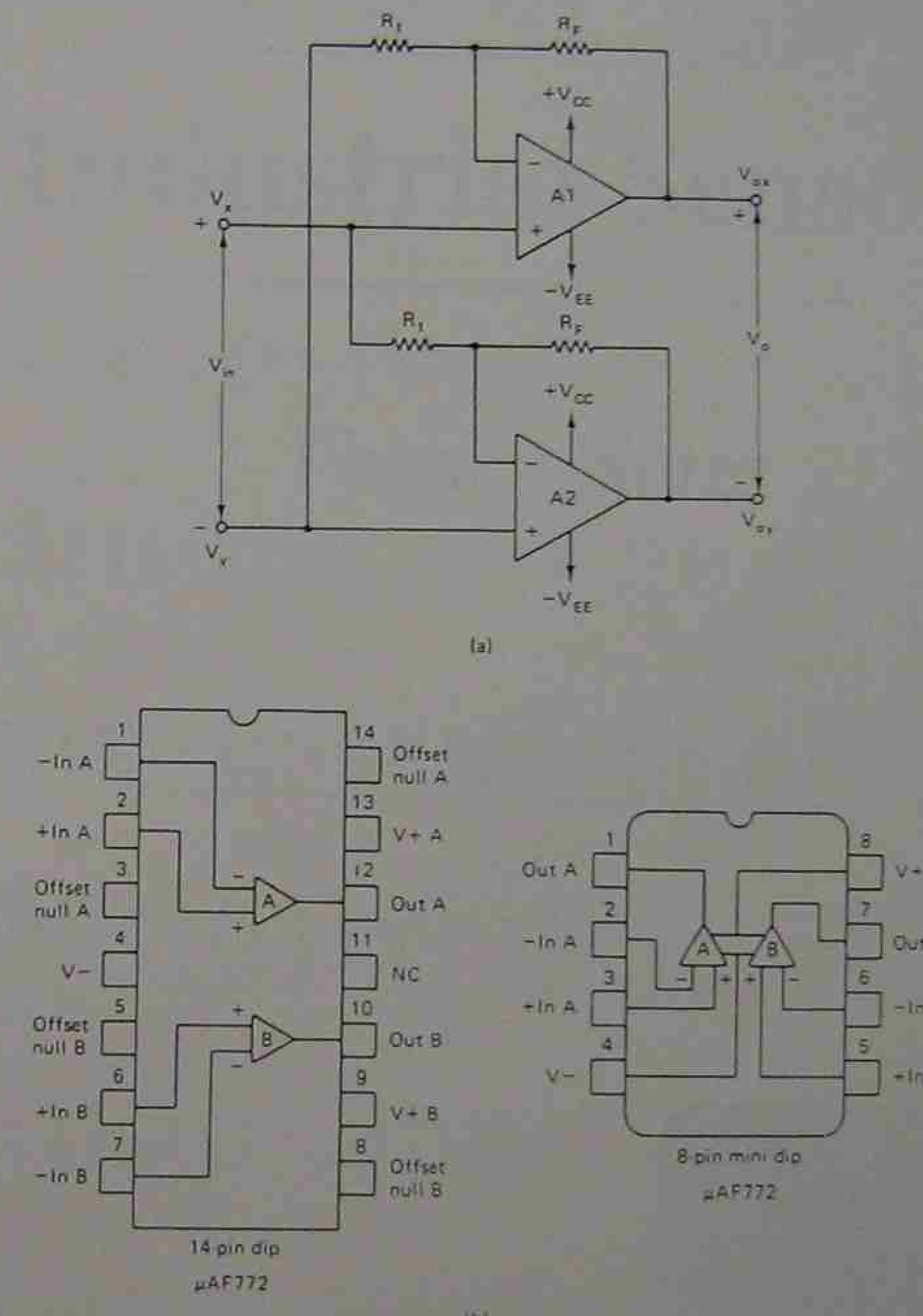


Figure 7–14 (a) Differential input and differential ouput amplifier using a dual op-amp. (b) The connection diagram of a typical dual op-amp. (Note especially the difference between the supply voltage and offset-null connections.) (Courtesy of Fairchild Camera and Instrument Corporation.)

EXAMPLE 7-9

The differential input and output amplifier of Figure 7-14(a) is used as a preamplifier and requires a differential output of at least 3.7 V. Determine the gain of the circuit if the differential input V_{in} = 100 mV.

SOLUTION Using Equation (7-18c), we get

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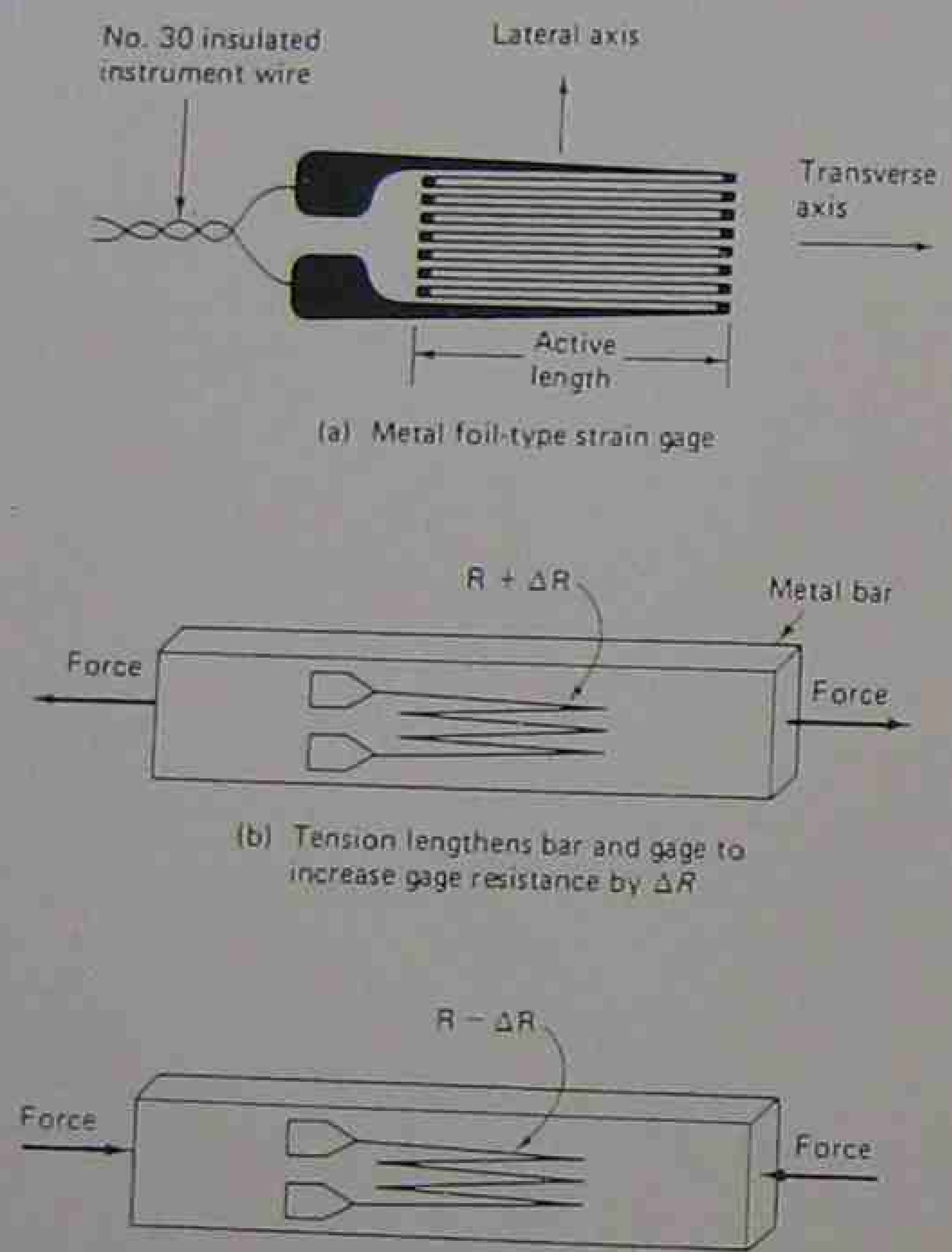
8-8.1 Introduction to the Strain Gage

A strain gage is a conducting wire whose resistance changes by a small amount when it is lengthened or shortened. The change in length is small, a few millionths of an inch. The strain gage is bonded to a structure so that the percent change in length of the strain gage and structure are identical.

A foil-type gage is shown in Fig. 8-13(a). The active length of the gage lies along the transverse axis. The strain gage must be mounted so that its transverse axis lies in the same direction as the structure motion that is to be measured [see Fig. 8-13(b) and (c)]. Lengthening the bar by tension lengthens the strain gage conductor and increases its resistance. Compression reduces the gage's resistance because the normal length of the strain gage is reduced.



(a) Metal foil-type strain gage

(b) Tension lengthens bar and gage to increase gage resistance by ΔR

(c) Compression shortens bar and gage to reduce gage resistance by ΔR

FIGURE 8-13 Using a strain gage to measure the change in length of a structure.

8-8.2 Strain-Gage Material

Strain gages are made from metal alloy such as constantan, Nichrome V, Dynaloy, Stabiloy, or platinum alloy. For high-temperature work they are made of wire. For moderate temperature, strain gages are made by forming the metal alloy into very thin sheets by a photoetching process. The resultant product is called a foil-type strain gage and a typical example is shown in Fig. 8-13(a).

8-8.3 Using Strain-Gage Data

In the next section we show that our instrumentation measures only the gage's *change* in resistance ΔR. The manufacturer specifies the unstrained gage's resistance R. Once ΔR has been measured, the ratio $\Delta R/R$ can be calculated. The manufacturer also furnishes a specified *gage factor* (GF) for each gage. The gage factor is the ratio of the percent change in resistance of a gage to its percent change in length. These percent changes may also be expressed as decimals. If the ratio $\Delta R/R$ is divided by gage factor G, the result is the ratio of the *change* in length of the gage ΔL to its original length L. Of course the structure where the gage is mounted has the same $\Delta L/L$. An example will show how gage factor is used.

Example 8-10

A 120-Ω strain gage with a gage factor of 2 is affixed to a metal bar. The bar is stretched and causes a ΔR of 0.001 Ω. Find $\Delta L/L$.

Solution

The ratio $\Delta L/L$ has a name. It is called *unit strain*. It is the unit strain data (we have developed from a measurement of ΔR) that mechanical engineers need. They can use this unit strain data together with known characteristics of the structural material (modulus of elasticity) to find the *stress* on the beam. *Stress is the amount of force acting on a unit area*. The unit for stress is pounds per square inch (psi). If the bar in Example 8-10 were made of mild steel, its stress would be about 125 psi. *Strain is the deformation of a material resulting from stress*, or $\Delta L/L$.

8-8.4 Strain-Gage Mounting

Before mounting a strain gage the surface of the mounting beam must be cleaned, sanded, and rinsed with alcohol, Freon, or methyl ethyl ketone (MEK). The gage is then fastened permanently to the cleaned surface by Eastman 910, epoxy, polyimide adhesive, or ceramic cement. The manufacturer's procedures should be followed carefully.

8-8.5 Strain-Gage Resistance Changes

It is the *change* of resistance in a strain gage ΔR that must be measured and this change is *small*. ΔR has values of a few milliohms. The technique employed to measure small resistance change is discussed next.

8-9 MEASUREMENT OF SMALL RESISTANCE CHANGES

8-9.1 Need for a Resistance Bridge

To measure resistance we must first find a technique to convert the *resistance change* to a current or voltage for display on an ammeter or voltmeter. If we must measure a small *change* of resistance, we will obtain a very small voltage *change*. For example, if we passed 5 mA of current through a 120-Ω strain gage, the voltage across the gage would be 0.600 V. If the resistance *changed* by 1 mΩ, the voltage *change* would be 5 μV. To display the 5-μV change, we would need to amplify it by a factor of, for example, 1000 to 5 mV. However, we would also amplify the 0.6 V by 1000 to obtain 600 V plus 5 mV. It is difficult to detect a 5-mV difference in a 600-V signal. Therefore, we need a circuit that allows us to amplify only the *difference* in voltage across the strain gage caused by a *change* in resistance. The solution is found in the bridge circuit.

8-9.2 Basic Resistance Bridge

The strain gage is placed in one arm of a resistance bridge, as shown in Fig. 8-14. Assume that the gage is unstrained, so that its resistance = R. Also assume that R_1, R_2, and R_3 are all precisely equal to R. (This unlikely assumption is dealt with in Section 8-10.) Under these conditions $E_1 = E_2 = E/2$ and $E_1 - E_2 = 0$. The bridge is said to be *balanced*. If the strain gage is *compressed*, R would decrease by ΔR and the differential voltage $E_1 - E_2$ would be given by

$$E_1 - E_2 = E\frac{\Delta R}{4R} \tag{8-7}$$

This approximation is valid because $2\,\Delta R \ll 4R$ for strain gages.

Equation (8-7) shows that E should be made large to maximize the bridge differential output voltage, $E_1 - E_2$.

Example 8-11

If $\Delta R = 0.001\ \Omega$, $R = 120\ \Omega$, and $E = 1.0$ V in Fig. 8-14, find the output of the bridge, $E_1 - E_2$.

Solution From Eq. (8-7),

We conclude that a voltage E and bridge circuit plus an instrumentation amplifier can convert a change in resistance of 1 mΩ to an output voltage change of 22 mV.

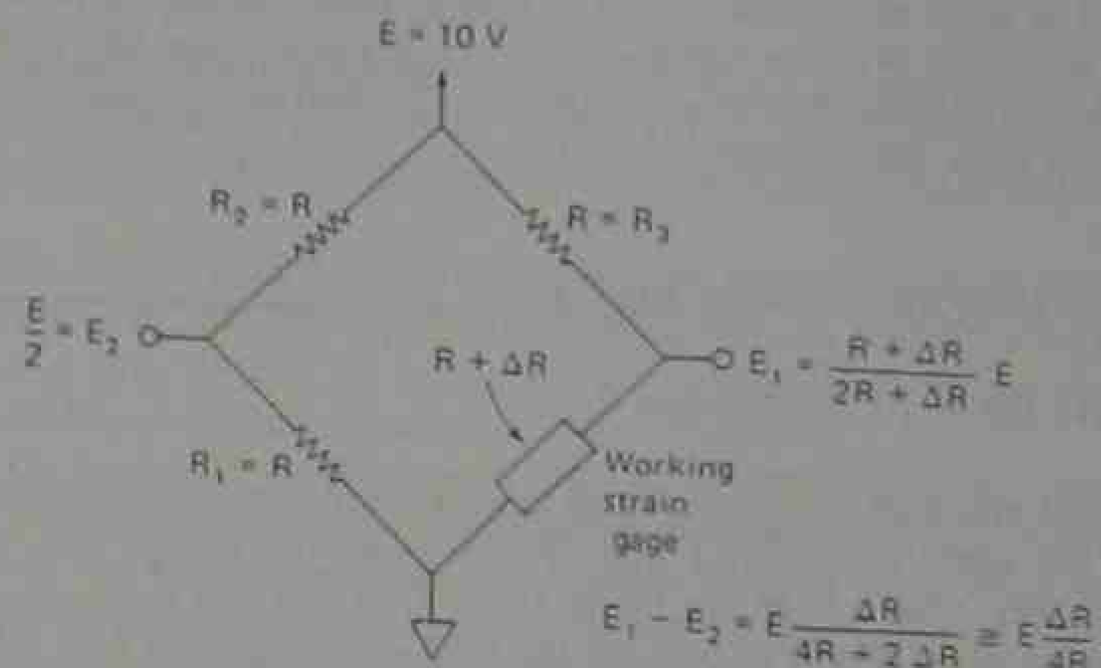


FIGURE 8-14 The resistor bridge arrangement and supply voltage E convert a resistance change in the strain gage ΔR to a differential output voltage $E_1 - E_2$. If $R = 120\ \Omega$, $E = 10$ V, and $R = 1$ mΩ, $E_1 - E_2 = 22\ \mu$V.

8-9.3 Thermal Effects on Bridge Balance

Even if you succeed in balancing the bridge circuit of Fig. 8-14, it will not stay in balance because slight temperature changes in the strain gage cause resistance change equal to or greater than those caused by strain. This problem is solved by mounting another identical strain gage immediately adjacent to the working strain gage so that both share the same thermal environment. Therefore, as temperature changes, the added gage's resistance changes exactly as the resistance of the working gage. The added gage provides automatic temperature compensation, and is appropriately called the *temperature-compensation* or *dummy* gage.

The *temperature-compensation gage* is mounted with its transverse axis perpendicular to the transverse axis of the working gage, as shown in Fig. 8-15. This

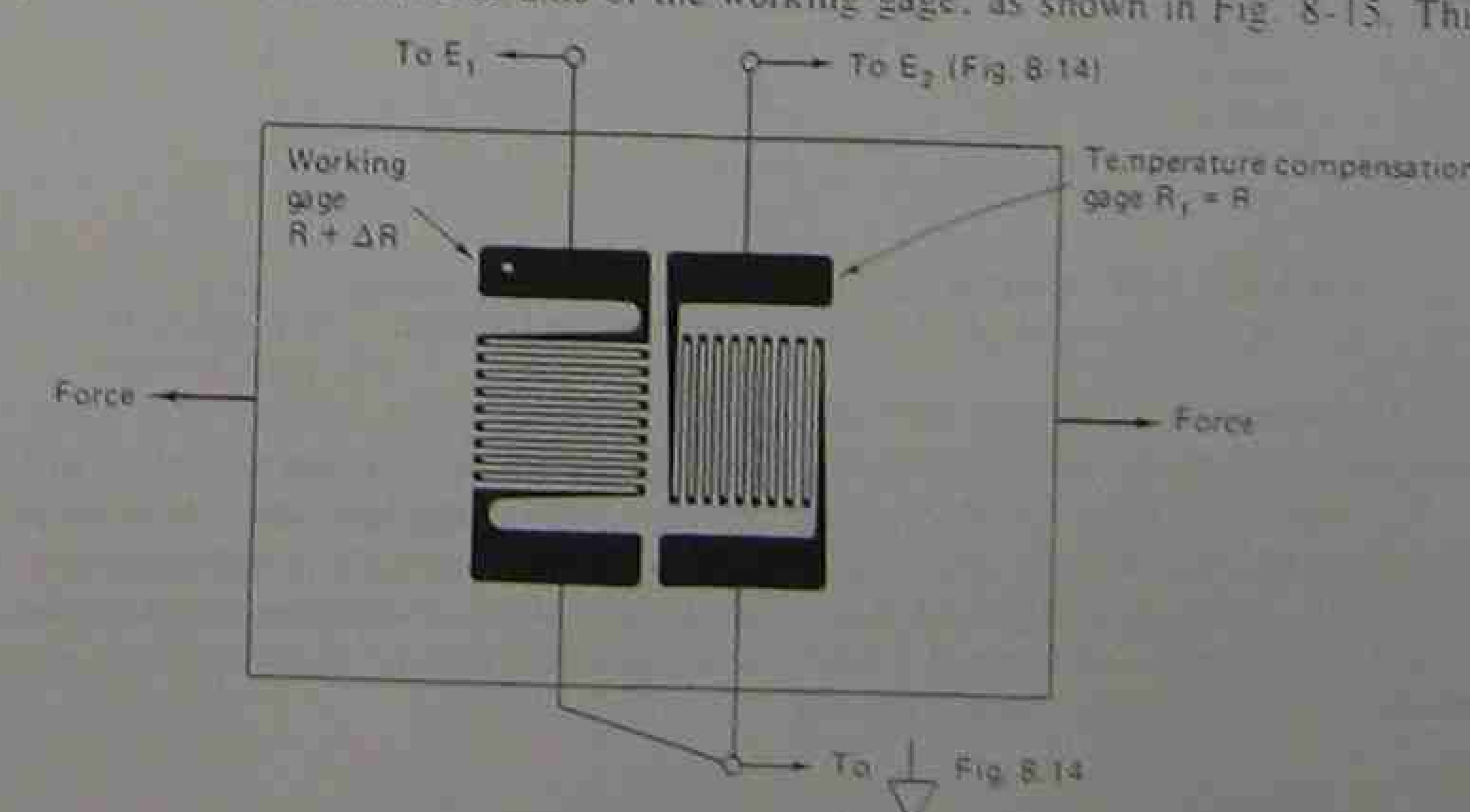


FIGURE 8-15 The temperature-compensation gage has the same resistance changes as the working gage with changes in temperature. Only the working gage changes resistance with strain. By connecting in the bridge circuit of Fig. 8-14 as shown, resistance changes due to temperature changes are automatically balanced out.

type of standard gage arrangement is available from manufacturers. The new gage is connected in place of resistor R_1 in the bridge circuit of Fig. 8-14. Once the bridge has been balanced, R of the temperature-compensation gage and working gage track one another to hold the bridge in balance. Any unbalance is caused strictly by ΔR of the working gage due to strain.

8-10 BALANCING A STRAIN-GAGE BRIDGE

8-10.1 The Obvious Technique

Suppose that you had a working gage and temperature-compensation gage in Fig. 8-16 that are equal to within 1 mΩ. To complete the bridge, you install two 1%, 120-Ω resistors. One is high by 1% at 121.200 Ω and one is low by 1% at 118.800 Ω. They must be equalized to balance the bridge. To do so, a 5-Ω 20-turn balancing pot is installed, as shown in Fig. 8-16. Theoretically, the pot should be set as shown to equalize resistances in the top branches of the bridge at 122.500 Ω.

Further assume that an instrumentation amplifier with a gain of 1000 is connected to the bridge of Fig. 8-16. From Example 8-11, the *output* of the instrumentation amplifier (IA) will be about 22 mV per milliohm of unbalance. This means that the 5-Ω pot must be adjusted to within 1 mΩ of the values shown, so that $E_1 - E_2$ and consequently, V_o of the IA will equal 0 V ±22 mV.

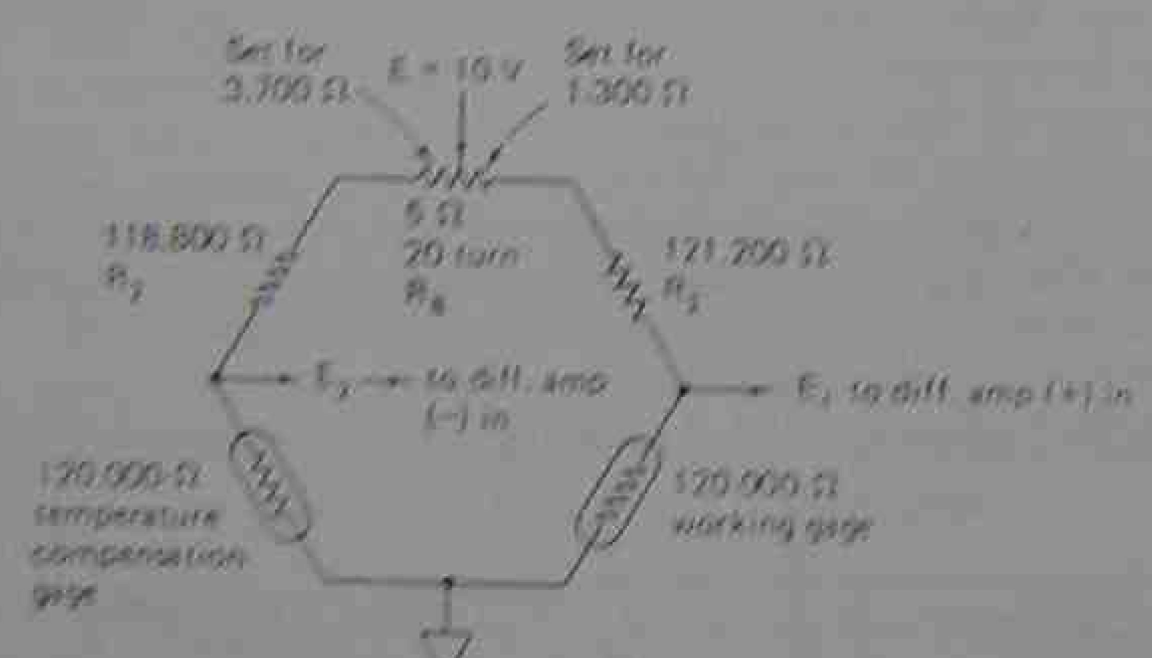


FIGURE 8-16 Balance pot R_B is adjusted in an attempt to make $E_1 - E_2 = 0$ V.

Unfortunately, it is very difficult in practice to adjust for balance. This is because each turn of the pot is worth 5 Ω/20 turns = 250 mΩ. When you adjust the pot it is normal to expect a backlash of $\pm\frac{1}{50}$ of a turn. Therefore, your best efforts result in an unbalance at the pot of about ±5 mΩ. You observe this unbalance at the IA's output, where V_o changes by ±0.1 V on either side of zero as you fine-tune the 20-turn pot. It turns out there is a better technique that uses an ordinary linear potentiometer ($\frac{3}{4}$ turn) and a single resistor.

8-10.2 The Better Technique

To analyze operation of the balance network in Fig. 8-17, assume that the R_2 and R_3 bridge resistors are reasonably equal, to within ±1%. The strain gage's resistance should have equal resistances within several milliohms if the working gage is not under strain.

Resistor R_{B1} is an ordinary $\frac{3}{4}$-turn linear pot. Its resistance should be about $\frac{1}{10}$ or less than resistor R_{B2} so that the voltage fE depends only on E and the decimal fraction f. Values of f vary from 0 to 1.0 as the pot is adjusted from one limit to the other. R_{B1} should be 10 or more times the gage resistance.

Resistor R_{B2} is chosen to be greater than 10 or more times R_{B1}. Under these conditions R_{B2} does not load down the voltage-divider action of R_{B1}. Also, the size of R_{B2} determines the maximum balancing current that can be injected into, or extracted from, the E_2 node. The pot setting f determines how much of that maximum current is injected or extracted.

Balancing action is summarized by observing that if $f > 0.5$, a small current is injected into the E_2 node and flows through the temperature gage to ground; this makes E_2 more positive. If $f < 0.5$ current is extracted from the E_2 node; this increases current through R_2 to make E_2 less positive.

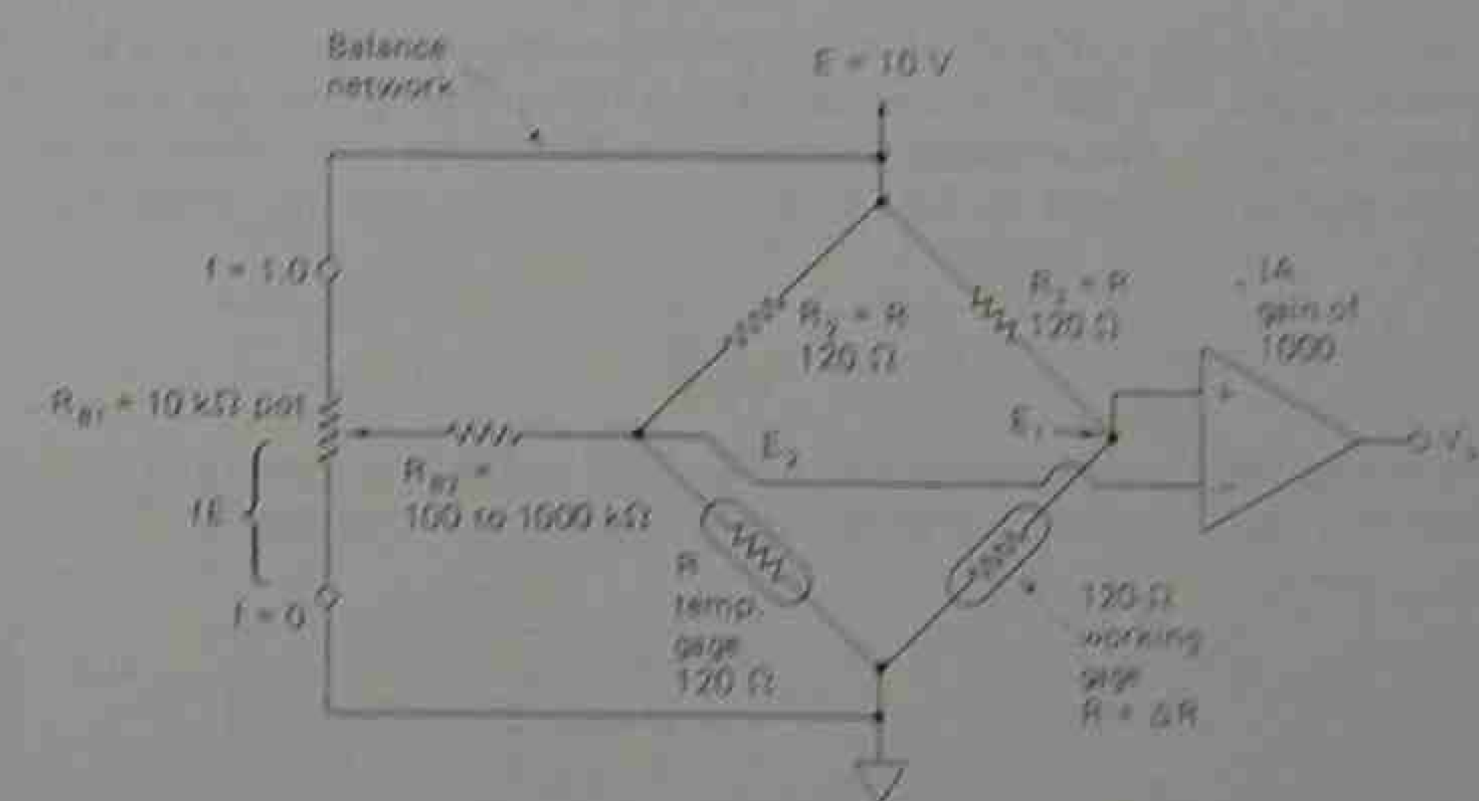


FIGURE 8-17 Improved balance network R_{B1} and R_{B2} allow easy adjustment of V_o to 0 V.

In a real bridge setup, begin with R_{B2} = 100 kΩ and R_{B1} = 10 kΩ. Monitor V_o of the IA and check the balancing action. If the variation in V_o is larger than you want, increase R_{B2} to 1000 kΩ and recheck the balance action. The final value of R_{B2} is selected by experiment and depends on the magnitude of unbalance between R_2 and R_3.

8-11 INCREASING STRAIN-GAGE BRIDGE OUTPUT

A single working gage and temperature-compensation gage were shown to give a differential bridge output in Fig. 8-14 of

$$E_1 - E_2 = E\frac{\Delta R}{4R} \tag{8-7}$$

This bridge circuit and placement of the gages is shown again in Fig. 8-18(a).

The bridge output voltage $E_1 - E_2$ can be doubled by doubling the number of working gages, as in Fig. 8-18(b). Gages 1–2 and 5–6 are the working gages and will increase resistance (tension) if force is applied as shown. By arranging the working gages in opposite arms of the bridge and the temperature gages in the other arms, the bridge output is

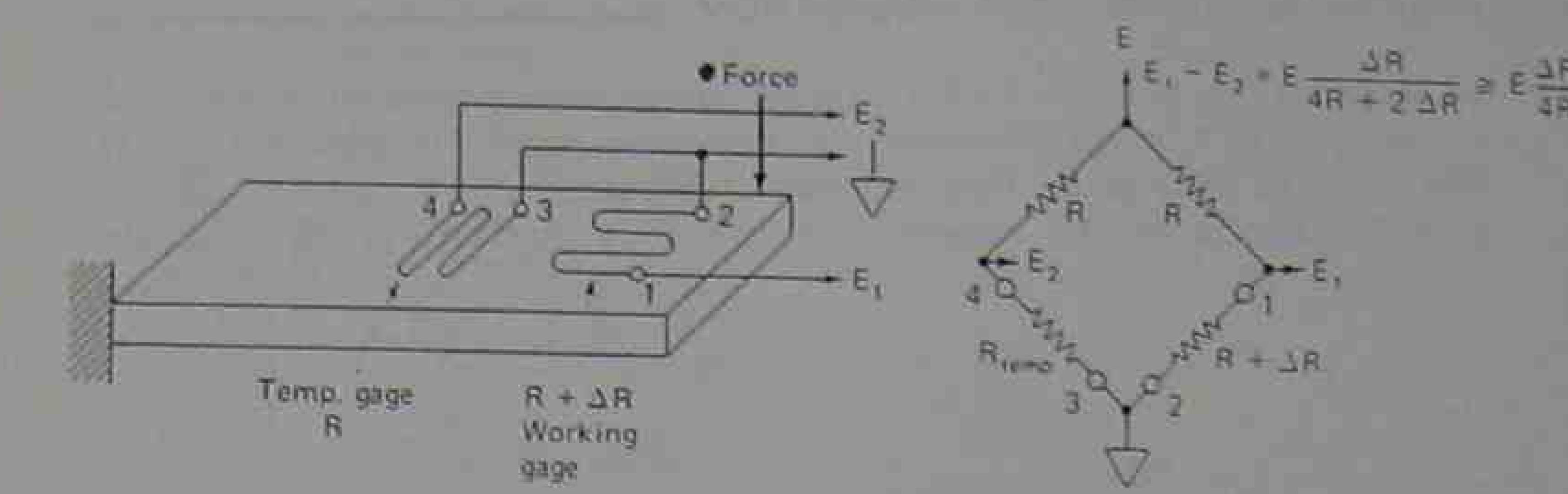
$$E_1 - E_2 = E\frac{\Delta R}{2R + \Delta R} = E\frac{\Delta R}{2R}$$

If the structural member experiences bending as shown in Fig. 8-18(c), even greater bridge sensitivity can be obtained. The upper side of the bar will lengthen (tension) to increase resistance of the working strain gages by $(+)\Delta R$. The lower side of the bar will shorten (compression) to decrease the working strain gages by $(-)\Delta R$.

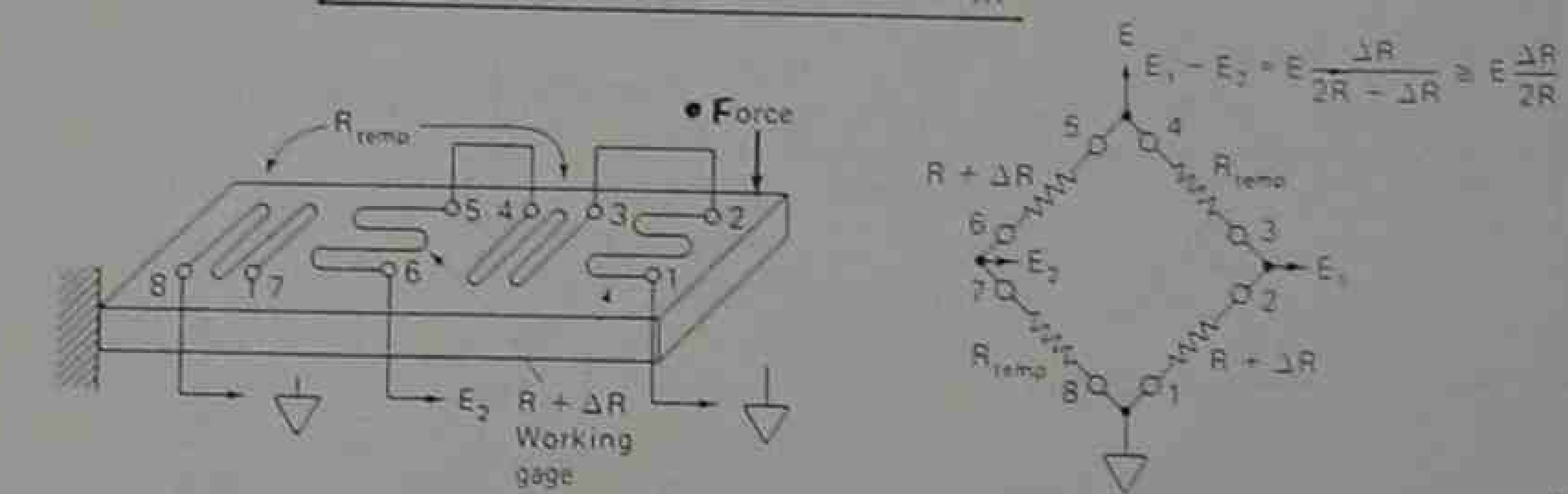
The tension gages 1–2 and 5–6 are connected in opposite arms of the bridge. Compression gages 3–4 and 7–8 are connected in the remaining opposite arms of the bridge. The gages also temperature-compensate one another. The output of the four-strain-gage arrangement in Fig. 8-18(c) is quadrupled over the single-gage bridge to

$$E_1 - E_2 = E\frac{\Delta R}{R}$$

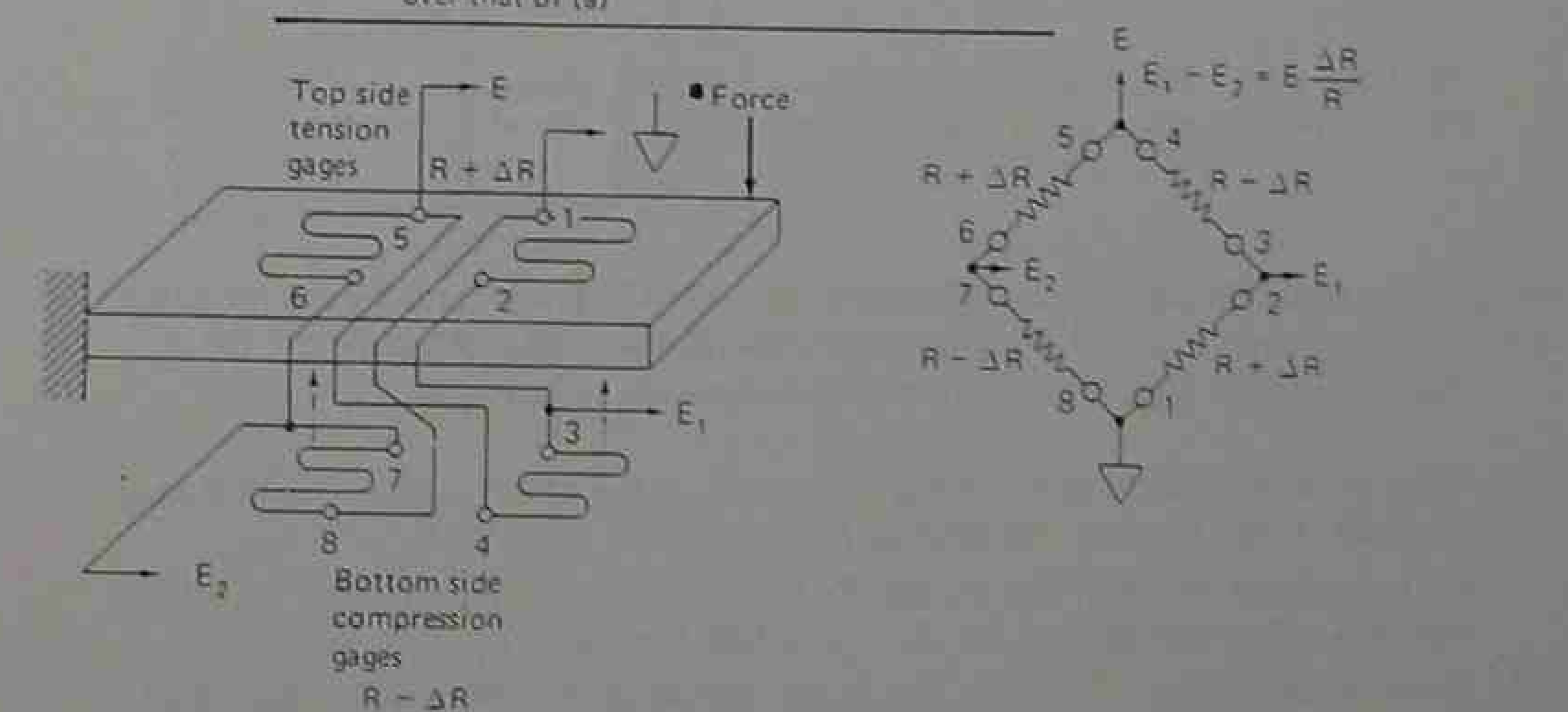
Of course, each bridge arrangement in Fig. 8-18 should be connected to a balance network [which, for clarity, was not shown (see Fig. 8-17 and Section 8-10)].



(a) Single working gage gives $E_1 - E_2 = E\frac{\Delta R}{4R}$



(b) Two working gages double the $E_1 - E_2$ output over that of (a)



(c) Four working gages quadruple the $E_1 - E_2$ over that of (a)

FIGURE 8-18 Comparison of sensitivity for three stain-gage bridge arrangements. (ΔR is small with respect to R for foil strain gages.)

8-12 A PRACTICAL STRAIN-GAGE APPLICATION

As shown in Fig. 8-19, an AD521 (Analog Devices) instrumentation amplifier (IA) is connected to a bridge arrangement of four strain gages. The gages are 120-Ω, SR4, foil-type strain gages. They are mounted on a steel bar in accordance with Fig. 8-18(c). Also the balance network of Fig. 8-17 is connected to the strain-gage

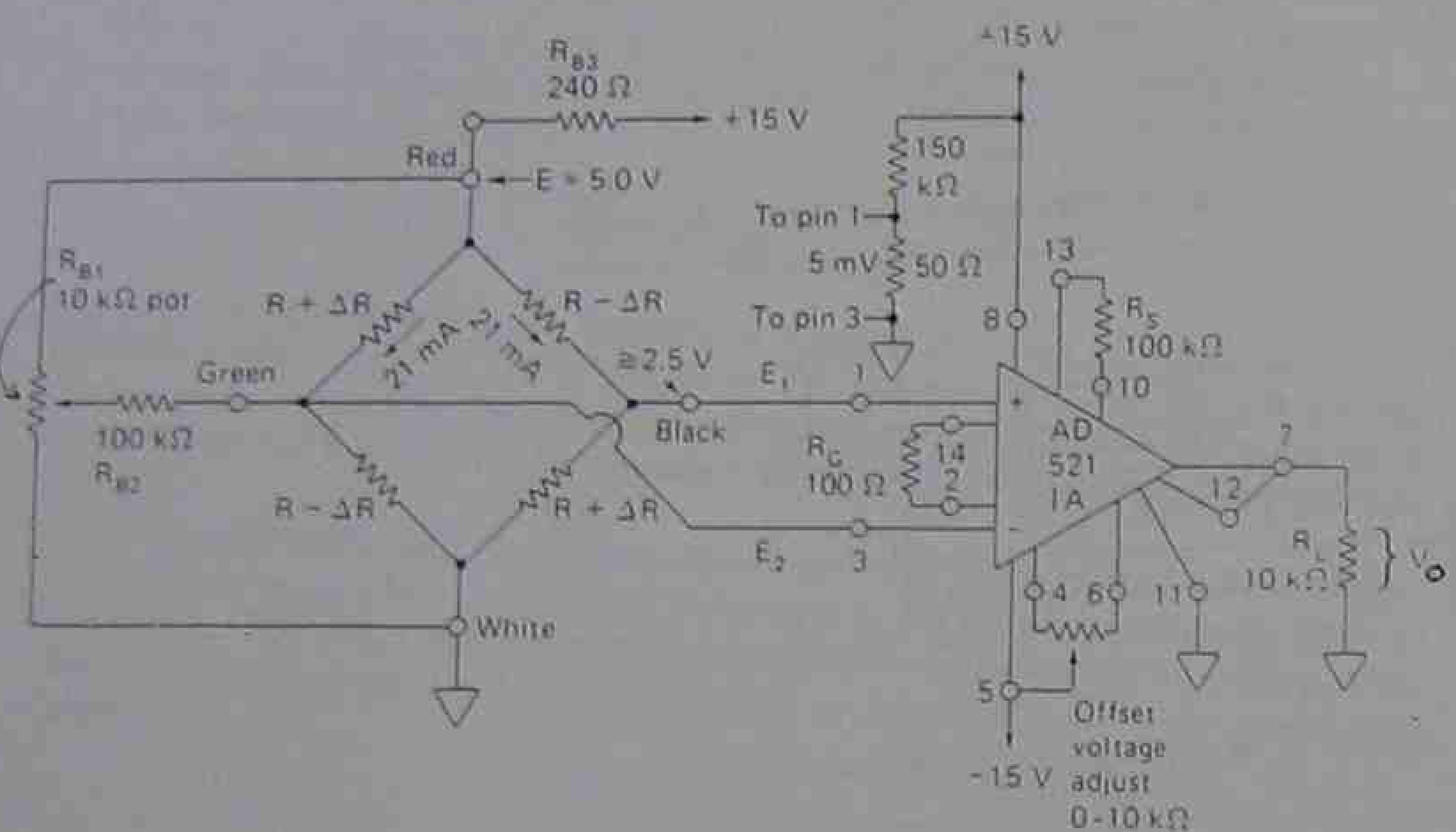
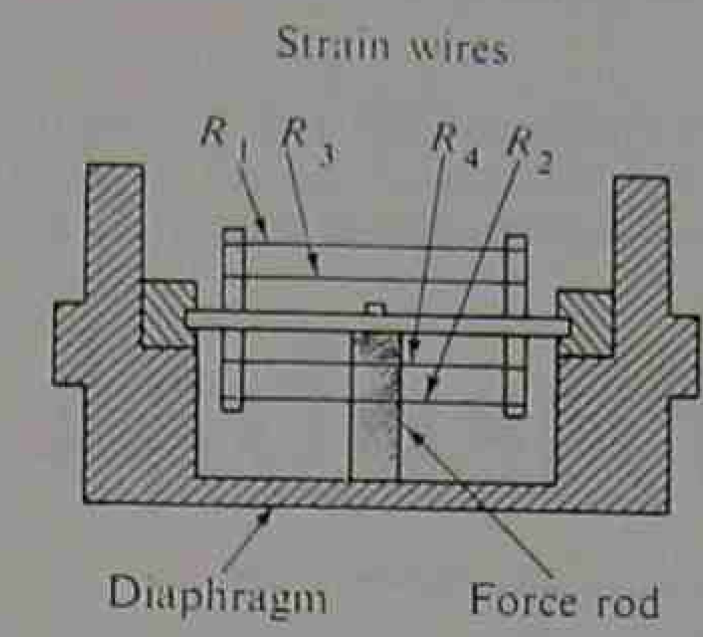


FIGURE 8-19 The AD521 instrumentation amplifier is used to amplify output of the four working strain gages [see Fig. 8-18(c)].

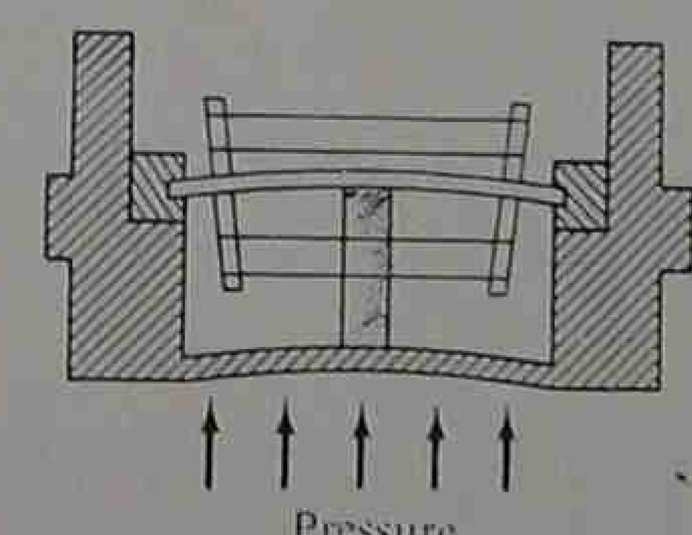
bridge. R_{B2} was selected, after experiment, as 100 kΩ. Strain gages were mounted in strict accordance with the manufacturer's instructions (BLH Electronics, Inc.)

The calibration procedure for the instrumentation is as follows:

1. *Gain resistors*. Select gain setting resistors $R_{scale}(R_S)$ and $R_{gain}(R_G)$ to give a gain of 1000. Gain is set by the ratio of R_S/R_G, or 100 kΩ/100 Ω.
2. *Offset voltage adjust*. With R_S and R_G installed, ground inputs 1 and 3. Adjust the IA's offset voltage pot for $V_o = 0$ V (see Chapter 9).
3. *Gain measurement*. Connect the 5-mV signal from the 150-kΩ and 50-Ω voltage divider as $E_1 - E_2$ to pins 1 and 3. Measure V_o; calculate gain = $V_o/(E_1 - E_2)$.
4. *Zeroing the bridge network*. Connect E_1 and E_2 to pins 1 and 3 of the IA. Adjust R_{B1} for $V_o = 0$ V when the gages are *not* under strain.



a) Pressure not applied



b) Pressure applied

Example 8-12

The test setup of Fig. 8-19 is used to measure the strain resulting from deflection of a steel bar. V_e is measured to be 100 mV. Calculate (a) ΔR; (b) $\Delta R/R$; (c) $\Delta L/L$. Assume that the calibration procedure has been followed and that the gain from step 3 is 1000. The gage factor is 2.0.

Solution (a) Find $E_1 - E_2$ from

8-13 MEASUREMENT OF PRESSURE, FORCE, AND WEIGHT

Example 8-12 illustrated how pressure could be measured by a strain-gage system. The mechanical engineers can be given $\Delta L/L$ by electrical personnel, who can measure $\Delta R/R$ and look up the gage factor. From the value of $\Delta L/L$, the mechanical engineers and technicians can calculate pressure on a structure. Since pressure is force per unit area, they can calculate force by measuring the structure's area.

Furthermore, the weight of an object exerts a force on any supporting structure. By installing a strain gage on the supporting structure, you can weigh very heavy objects such as a gravel-filled truck or a 747 aircraft.

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5-3 APPLICATIONS OF THE PIEZOELECTRIC TRANSDUCER

One company that specializes in the design and construction of the multielement piezoelectric structure is the Vernitron Corporation in Bedford, Ohio. Their trade name for this style of element is Bimorph. Figure 5-8 illustrates applications of this construction type in some products in common use. In the paragraphs that follow we discuss other applications of piezoelectric transducers and how they operate.

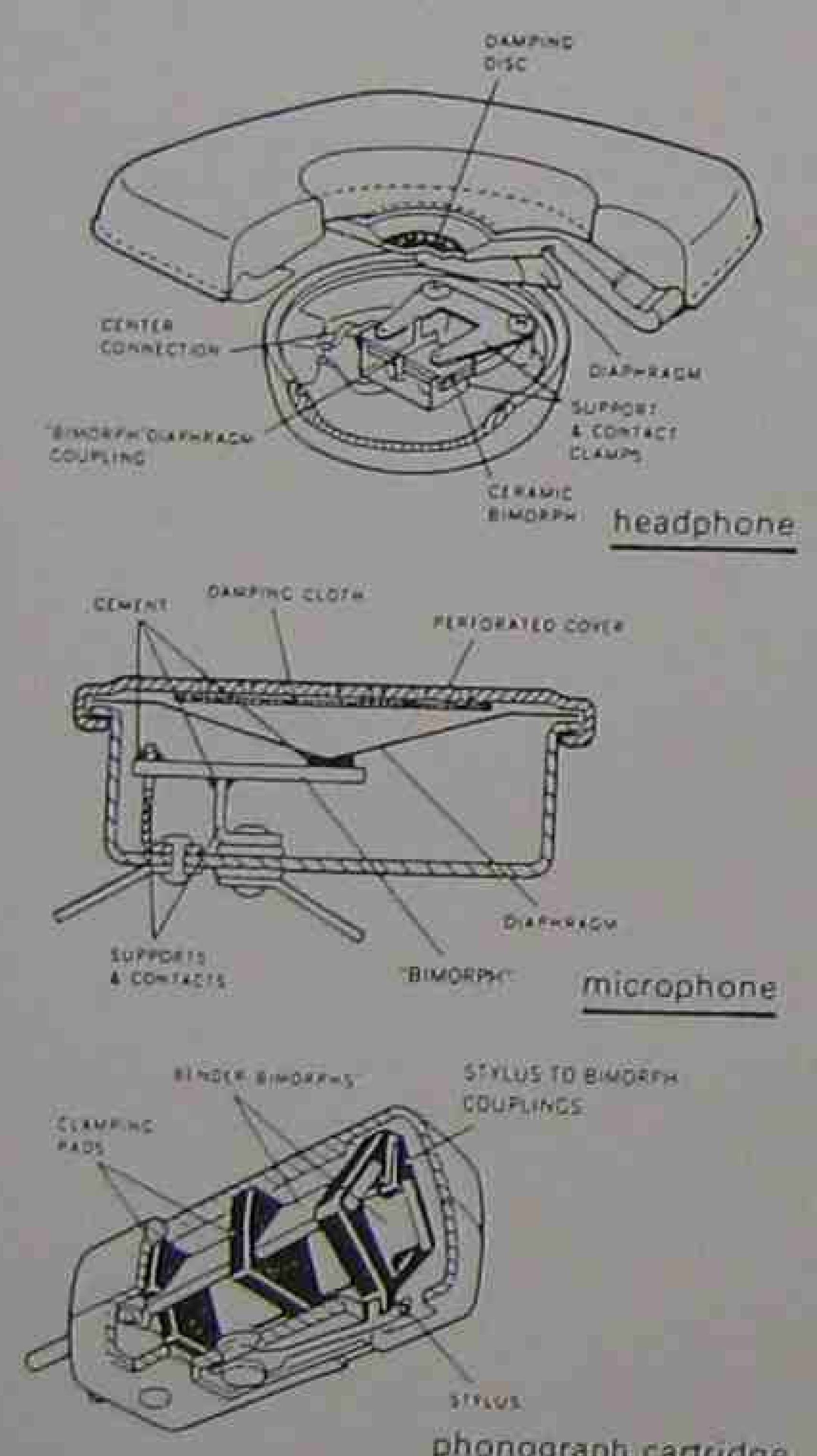

Figure 5-8 Bimorph crystals. (Courtesy of Morgan Matroc, Inc., Vernitron Division, Bedford, OH.)

5-2 THEORY AND DESIGN OF PIEZOELECTRIC SENSORS

5-2.1 Single-Element Crystal

Certain naturally occurring crystalline materials display the piezoelectric phenomenon. Recall from the discussion that a piezoelectric substance generates an electric charge when subjected to a sudden change in stress, such as a sudden blow to the crystal. Conversely, when this substance is exposed to a changing electrical charge, the physical shape of the crystalline substance is altered. Piezoelectric materials can also be produced artificially. Table 3-1 lists both human-made and naturally occurring materials.

Figure 5-1 shows the natural crystal shapes of the more popular piezoelectric crystals used. Notice that an axis coordinate system has been assigned to each crystal shape. The z-axis is the longitudinal axis in each case (although in the case of the lithium sulfate crystal, it is somewhat difficult to ascertain which is the longitudinal axis; notice that the z-axis in this case is parallel to the lines of intersection of the surfaces comprising the crystal's sides). The x and y axes can only be identified from the shape of the crystal and are therefore not as easily determined as the z-axis.

Figure 5-2 shows how a piezoelectric crystal can be fitted with an electrode plate on opposite faces so that a crystal can accommodate and respond to a particular stress direction. In addition, the output sensitivity of the crystal can be enhanced for certain directions merely by altering the crystal housing's design to allow it to move in that direction, as shown in Figure 5-3. Therefore, in reality, there are only two

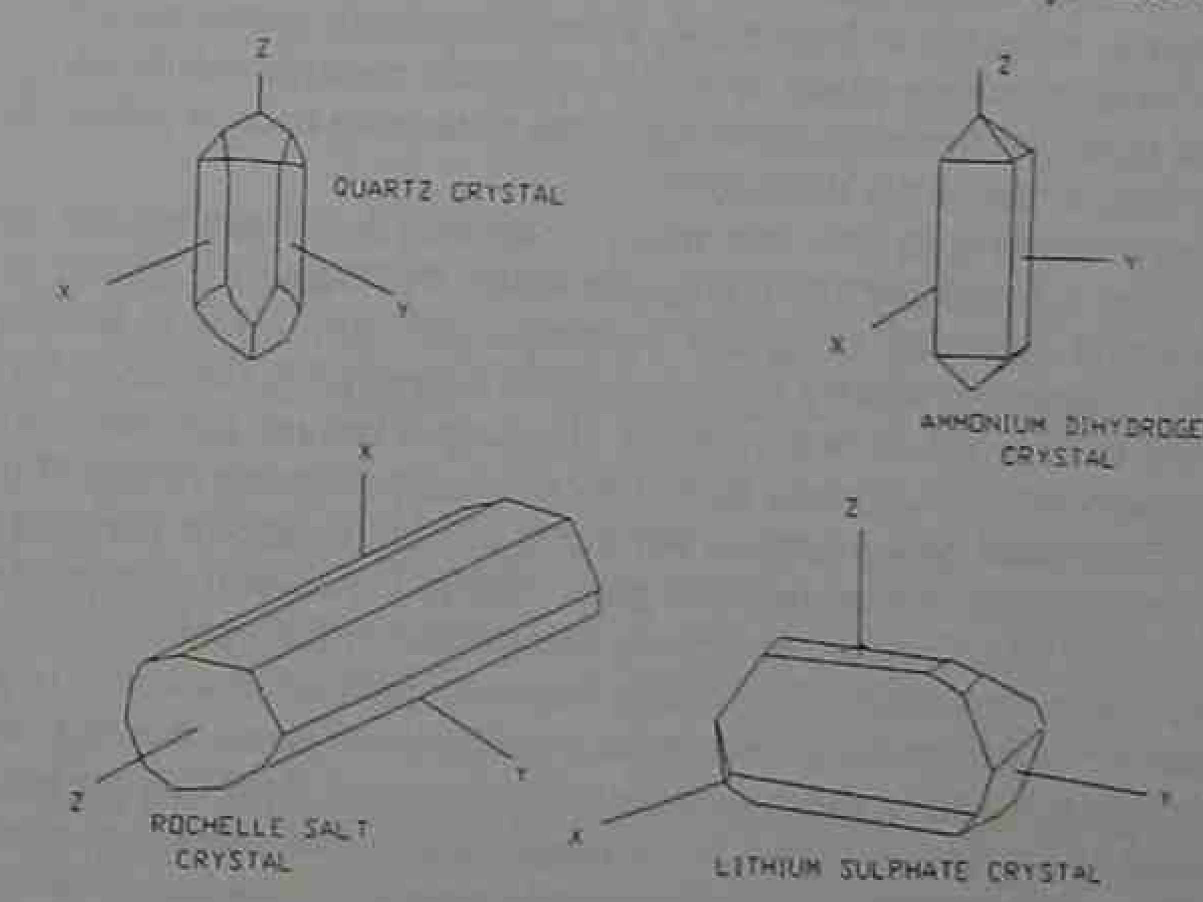

Figure 5-1 Natural crystal shapes of piezoelectric substances.

modes of deformation associated with a piezoelectric crystal: shear and compression (or tension).

The crystal can be identified by the type of cut that has been made with a crystal slab cutter. These cutters are designed for cutting very hard crystalline materials and are usually impregnated with diamonds or other extremely tough wear-resistant cutting material. When a cut has been made through a crystal, resulting in a slab whose two major surfaces are perpendicular to, say, the y-axis, it is called a *y-cut* crystal. The other axes cuts are identified in a similar manner. It is also possible to cut slabs at angles other than perpendicular to an axis. The angle of cut has a significant effect on the crystal's behavior as a piezoelectric sensor, and crystal manufacturers have spent much time researching this effect.

The type of cut made in a certain crystal also has a dramatic effect on the crystal's output. For instance, a y-cut made from lithium sulfate has a very strong thickness expansion characteristic. An x-cut made from Rochelle salts has one of the highest outputs known.

In the case of *PZT* (lead zirconate–barium titanate) or "ceramic" crystals, the axes used for positioning and location information are determined by a means somewhat different from that described above. These axes are found by the direction of a dc polarizing field used to polarize the ceramic element after manufacture. This process is called *poling*. The poling direction is determined by the application of a high dc voltage during the crystal's manufacture. This direction goes from plus to minus on the polarizer's electrodes. Once polarized, the axes are identified, usually by a numbering system rather than by letters, so that we have axis 1, axis 2, axis 3, and so on. The directions of the applied stress and the polarization axis are referred to as the *mode* of the ceramic material. The mode is described by the numbered axes. For instance, referring to Figure 5-4, if the polarization direction were along axis 3 and the stress axis were along axis 1, the ceramic material would be operating as a mode 31 type (the first digit referring to the polarizing field or poling direction, the second digit to the stress direction).

It is interesting to note that when an applied dc voltage has the same polarity as that used for creating the poling direction, the element expands in the poling direction and contracts along the perpendicular axes. This is illustrated in Figure 5-5(a). However, when a force in tension is applied in the same direction as the polarizing field in this ceramic crystal, the output voltage generated by the crystal has a polarity which is *opposite* that of the original polarizing voltage [Figure 5-5(c)]. On the other hand, when a signal voltage is applied to an axis perpendicular to the poling voltage axis, shearing takes place within the crystal element, as shown in Figure 5-5(b). The direction of shear is determined by the applied voltage's polarity. Figure 5-5 also shows the relationships between other combinations of poling directions and applied force directions. Also, equations are given showing the mathematical relationships among these variables. Single-crystal slabs that are placed between electrodes as implied in Figure 5-5 are referred to as *sandwiched* crystals.

We have referred to the piezoelectric element's electrodes several times but have not described how these electrodes are constructed. In the case of the older quartz crystals these electrodes were nothing more than plates of copper that were spring-loaded so as to press against opposite sides of the crystal's body. In the case of ceramic crystals the electrodes are either plated or sprayed onto the crystal element's body. Graphite and copper are two commonly used materials.

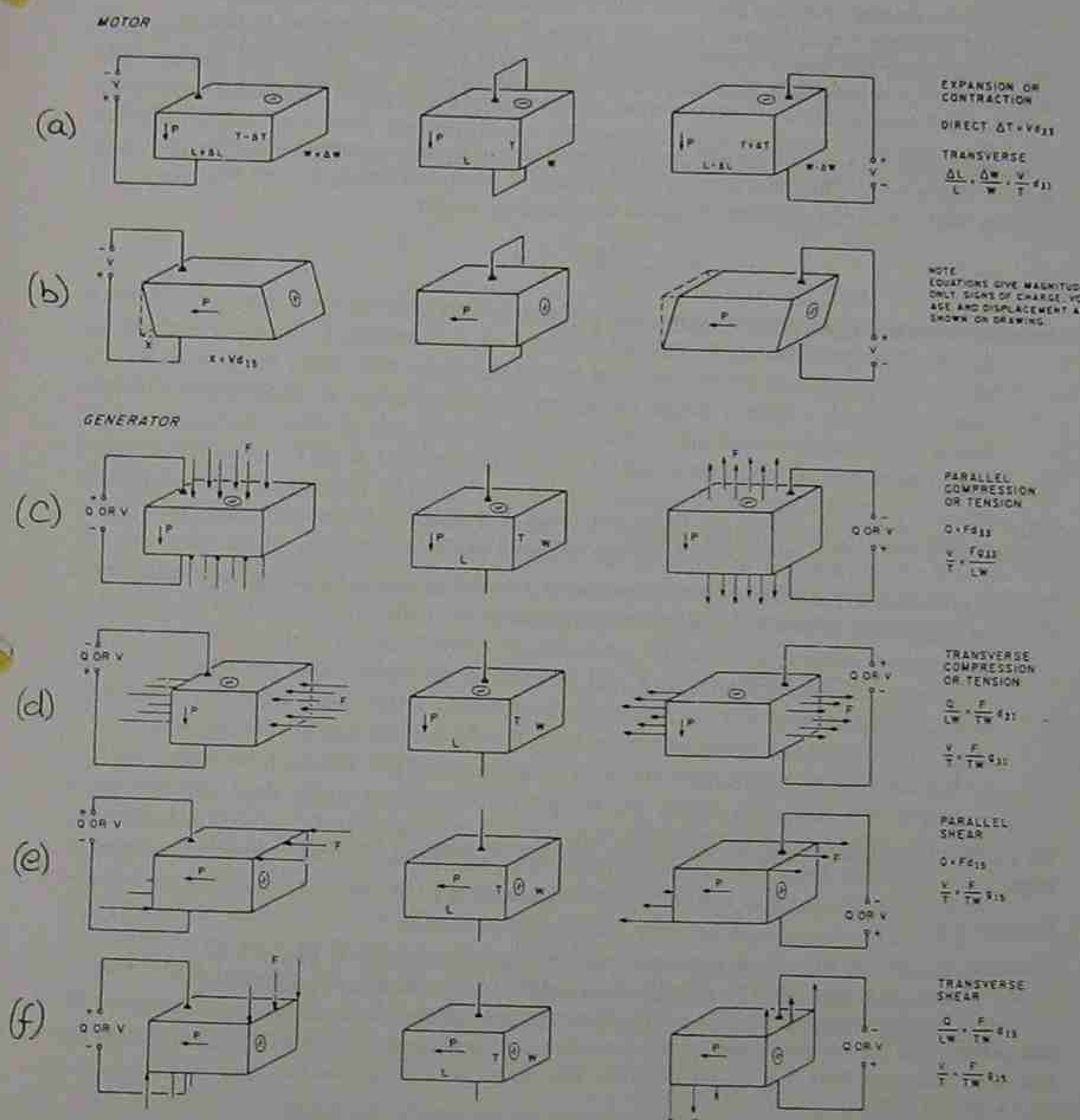

Figure 5-5 Generator and motor modes of a "sandwiched" crystal showing applied forces and voltages and poling directions. (Courtesy of *Sensors*, North American Technology, Inc., Peterborough, NH, from the article, "Piezoelectric design notes," by Piezo Electric Products, Inc., March 1984, pp. 20–27.)

Example 5-1

Calculate the amount of voltage generated with a cantilevered series-connected ceramic crystal of thickness 3 mm, length 5 mm, and width 20 mm. The force applied to the sensor containing this crystal is 5 N. The force is a direct force, with no associated frequency; it is applied one time only. The transverse voltage coefficient, g_{31}, is -11.1×10^{-3} V-m/N.

Solution: Since we are working with a cantilevered crystal in its generator mode, we refer to Figure 5-6 and find that part (a) seems to depict the crystal description we are working with. Using the equation alongside the diagram, we get

$$V = \frac{3}{2}\frac{FLg_{31}}{WT} \tag{5-1}$$

Inserting the values given in the problem, we get

Example 5-2

What would the output voltage be if the crystal in Example 5-1 were used in a parallel-connected circuit rather than a series-connected circuit?

Solution: We will now use Figure 5-6(b) instead of part (a) for our system analysis, and will consequently use part (b)'s equation to calculate V.

$$V = \frac{3}{4}\frac{FLg_{31}}{WT} \tag{5-2}$$

We see that the output is only approximately one-half as much as the output in Example 5-1.

Example 5-3

Calculate the amount of voltage generated with the same crystal cut as above (i.e., having the same dimensions) and having the same applied force. However, assume that the crystal is now "sandwiched" (i.e., not cantilevered, but mounted between compression plates), that the force is applied in the same direction as the polarization field, and that both are along axis 3 (i.e., the Y-axis). The direct voltage coefficient, g_{33}, is 26.1×10^{-3} V-m/N.

Solution: Since we are working again with a crystal in its generator mode, but now sandwiched, we refer to Figure 5-5 and find that part (c) seems to describe the crystal design we are dealing with here. Therefore, we rearrange this equation to solve for the voltage, V:

$$V = \frac{Fg_{33}T}{LW} \tag{5-3}$$

Inserting our known given values, we get

5-2.4 General Characteristics of Piezoelectric Materials

One serious problem associated with a piezoelectric sensor is its characteristically high output impedance. A hint of this characteristic can be seen while making a qualitative inspection of sorts of a typical crystal output condition. For a given crystal voltage output, very little current output will be observed. Because impedance, Z, and voltage, E, are related by means of the current, I, through a variation of Ohm's law, that is,

$$Z = \frac{E}{I} \tag{5-4}$$

it can be seen that for a given E and an extremely small I, a rather large impedance, Z, might well result. Furthermore, to transfer this high-impedance signal from the crystal to a signal cable, the cable itself must have an impedance that matches the crystal's impedance in order to obtain maximum signal transfer. Unfortunately, high-impedance cable, in the form of coaxial cable, suffers from a condition called *triboelectric noise*. This is a condition unique to this type of cable. Because of the dielectric material used in its construction and because of the possibility of relative motion existing between the cable's dielectric and the outer shield, electrostatic charges may easily develop. The net result is the creation of extraneous electrical noise such as popping or hissing. This is a condition brought on by cable movement as would be experienced in the case of using a piezoelectric crystal for vibration or impact measurements. A particularly troublesome vibrational frequency range for these cables seems to be below about 20 or 30 Hz. As a result, particular care must be used in selecting a low-noise high-impedance cable for this type of application.

Piezoelectric crystals are readily affected by temperature. As a matter of fact, any piezoelectric material will lose its charge generation characteristics if heated to a sufficiently high temperature. The temperature point at which the crystal's piezoelectric quality is lost is called the *Curie point*. This discussion, given in Chapter 3, bears repeating here. Some of the piezoelectric substances listed in Table 3-1, which concerned the discussion of piezoelectric microphones, are repeated in Table 5-2. However, for each of the substance types listed, the Curie points are also given.

As can be seen in Table 5-2, the range of Curie point temperatures is from 45 to 550°C. It is important to be able to identify the Curie point temperature value for the crystal in a given transducer design so that you have some idea what its maximum operating temperature limitations are.

TABLE 5-2 CURIE POINTS FOR CERTAIN PIEZOELECTRIC MATERIALS

Material	Curie point temp. (°C)
ADP	120
PZT	300
Quartz	550
Rochelle salts	45

* Crystals are affected by temperature in another way, too. Their frequency of oscillation is directly related to temperature.

In the case of a vibrating object, we are describing what usually consists of rapid and cyclic changes in the object's velocity *direction*. That is, the object first heads in one direction, then travels in the opposite direction. These reversals in velocity direction imply that for an instant, the velocity of the object drops to zero. The object then reassumes a velocity value, but in the opposite direction. These velocity reversals represent changes in velocity that produce acceleration.

To distinguish between acceleration of the type just described and the acceleration produced by a change in velocity while traveling in one direction only (probably the more familiar type of acceleration), we will refer to the latter as *linear* acceleration.

Figure 5-19 represents a simplified view of how a piezoelectric accelerometer operates. The force shown in this illustration can be a "created" force. That is, it can be created by a mass, called a *seismic mass*, that reacts to acceleration. If that acceleration changes [i.e., $d(ma)$ assumes a value over a change in time, $d(t)$], the crystal produces an output response. Just such an arrangement is depicted in Figure 5-20, where we see a seismic mass attached to a piezoelectric crystal inside a cylinder containment. In one example we see the seismic mass riding on the crystal slab, causing compression. In the other example the seismic mass is attached to the crystal so as to produce shear. In either case, any change in the acceleration of the mass will produce an output at the crystal's terminals. Keep in mind, though, that what is being said here concerning acceleration has to do with the acceleration produced by vibration or shock, not with linear acceleration.

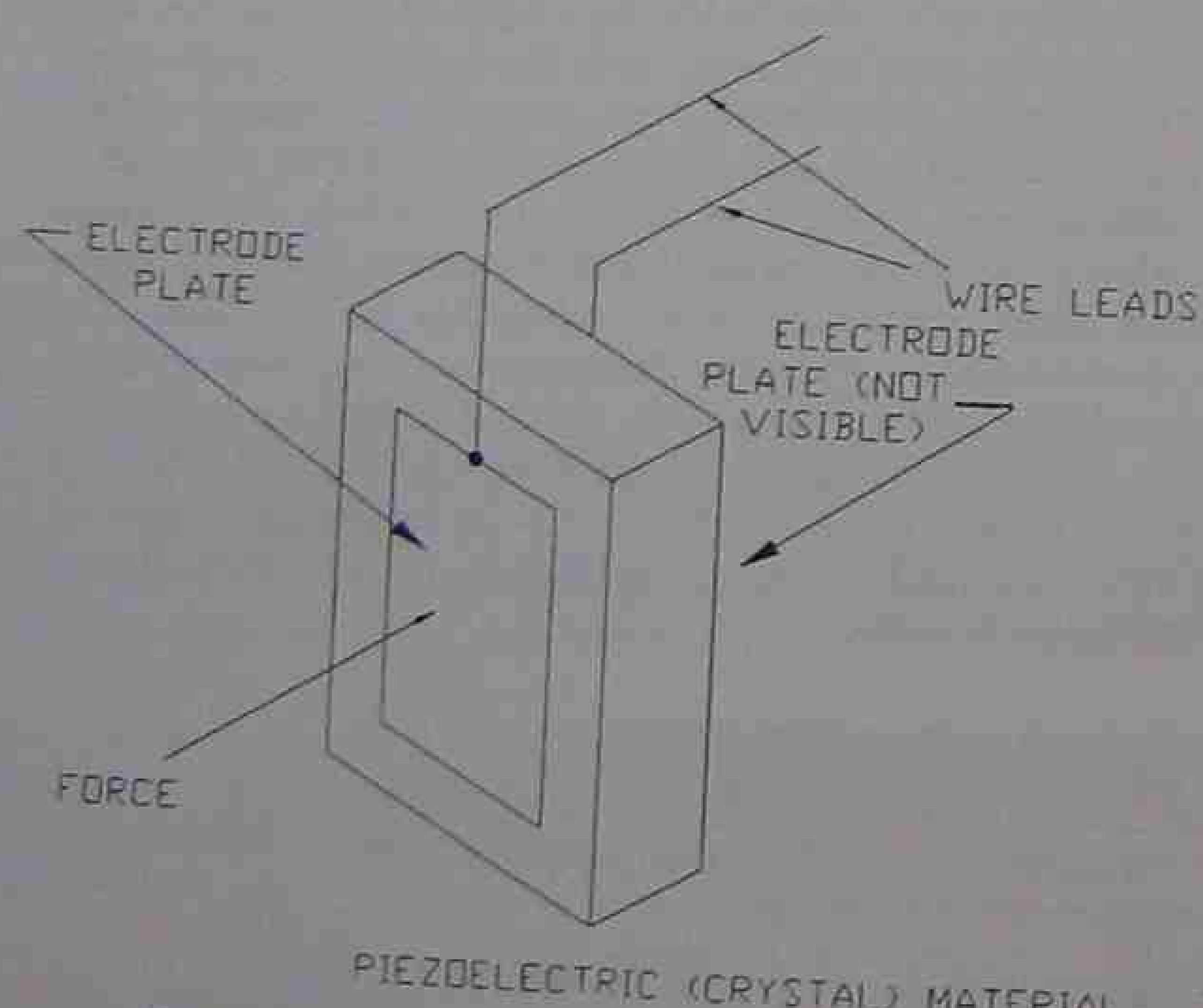

Figure 5-19 How a simple crystal accelerometer functions.

Temperature range: "−40 to +100°C." The temperature range is the range of temperatures within which the transducer has been designed to operate without producing unanticipated results.

Sensitivity versus temperature effect: "less than 0.1%/°C." Since temperature does affect the behavior of most transducer devices, this figure indicates to what degree the transducer's output/input is affected.

Power requirements: "12 to 25 V dc at 4 mA, max." This transducer requires a dc power source that can supply dc voltage anywhere between 12 and 25 V. This power source, should it be able to supply 25 V, must also be able to have a current capacity of at least 4 mA.

Physical configurations

Figure 5-21 shows a typical configuration for an accelerometer. The transducer housing is usually circular in shape and is often constructed of stainless steel. The container itself is usually hermetically sealed to prevent contaminants from entering. The stainless construction makes the transducer virtually corrosion-proof. Physical outline data are also given. The hermetically sealed casing allows the transducer to be installed in certain hazardous locations. These locations are usually given in the specifications in the form of hazardous class types, groups, divisions, and so on. The various categories are usually defined by a national standardization organization.

The accelerometer casing is usually supplied with a connector containing pipe threads for attachment to metal conduit. This is an electrical safety requirement

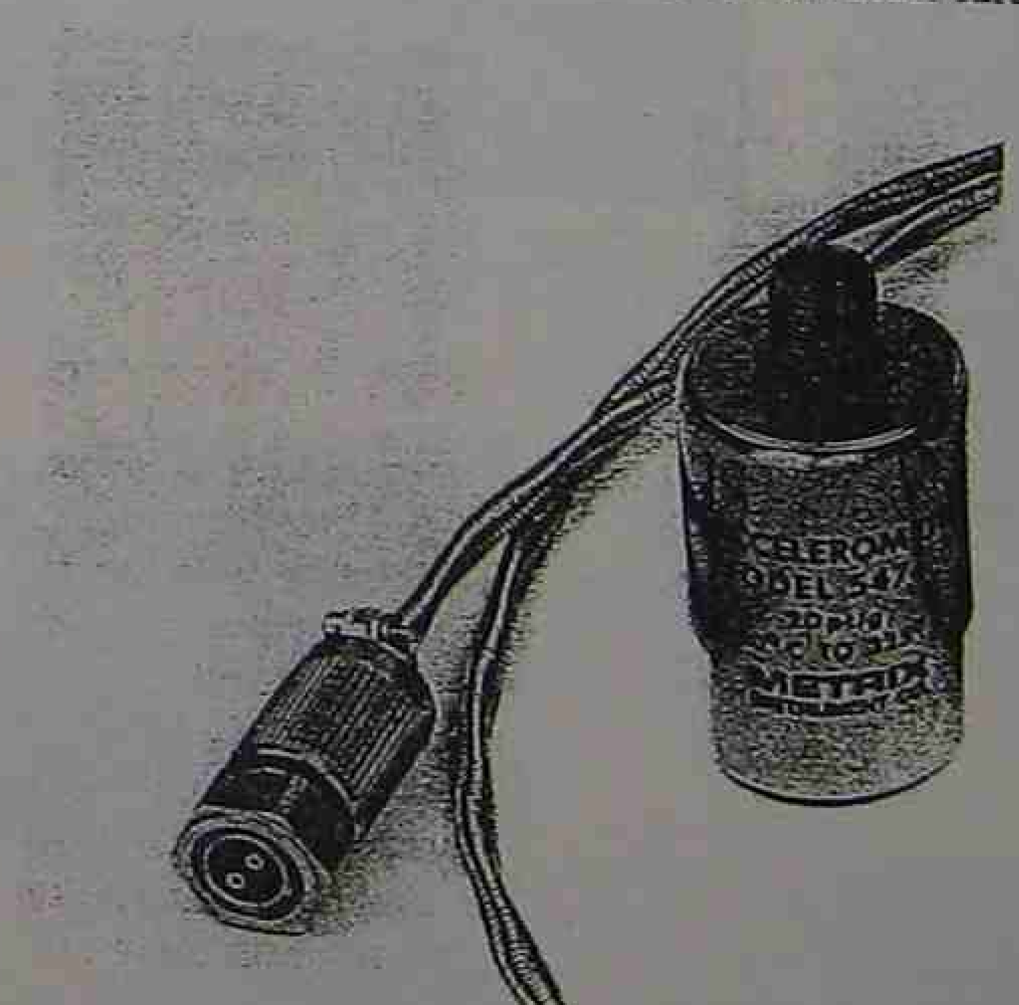

Figure 5-21 Typical accelerometers. (Courtesy of Metrix Instrument Co., Houston, TX.)

to prevent exposed wires from occurring in industrial applications. The wires are for supplying power to the transducer and for conducting the output signal from the transducer to an instrument or processing center. It is often possible to specify some other connector style if the transducer is needed for other than industrial applications.

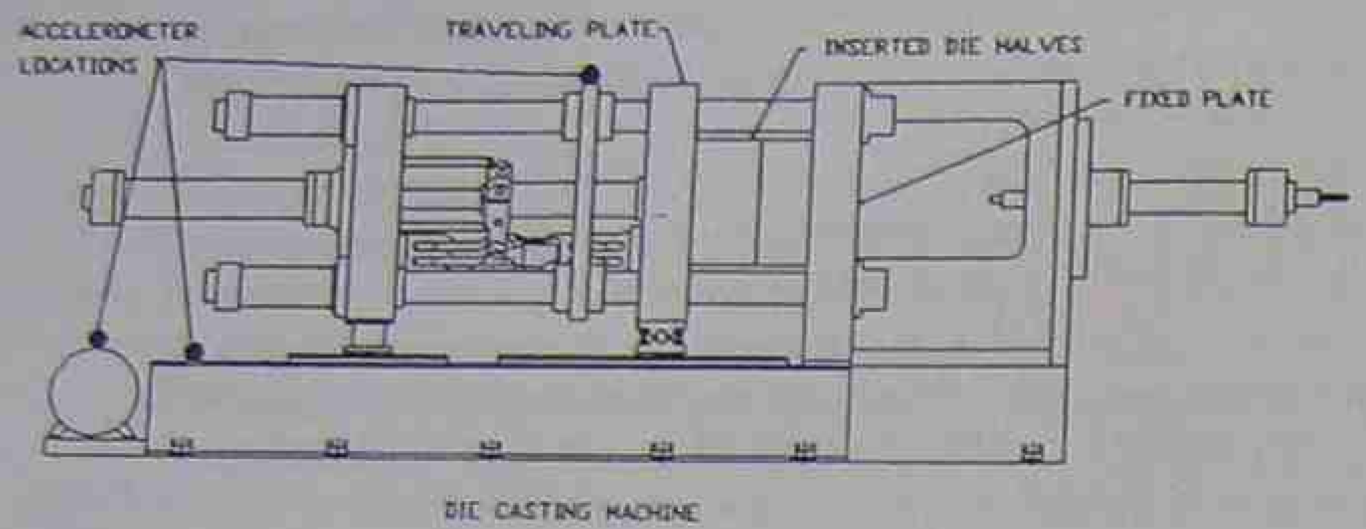

Figure 5-22 Production machine outfitted with an accelerometer for troubleshooting. (Adapted from Wayne Alofs and James R. Carstens, *Mechanical Maintenance and Evaluation of Die Casting Machines*, copyright 1987. Reprinted by permission of the North American Die Casting Association, River Grove, IL.)

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

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

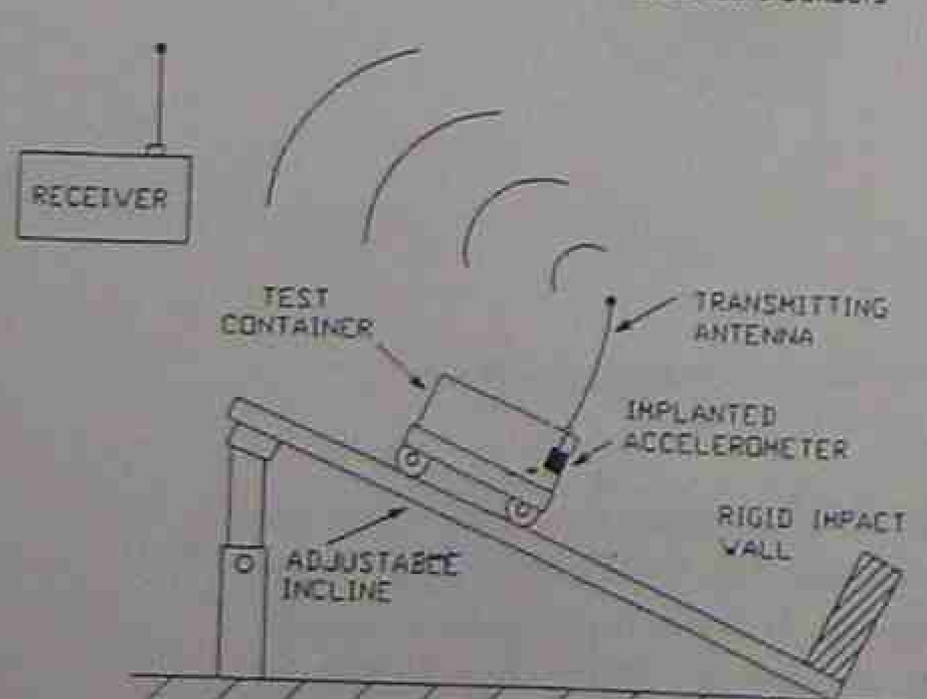

Figure 5-25 Shipping simulator used for the testing of product shipping methods.

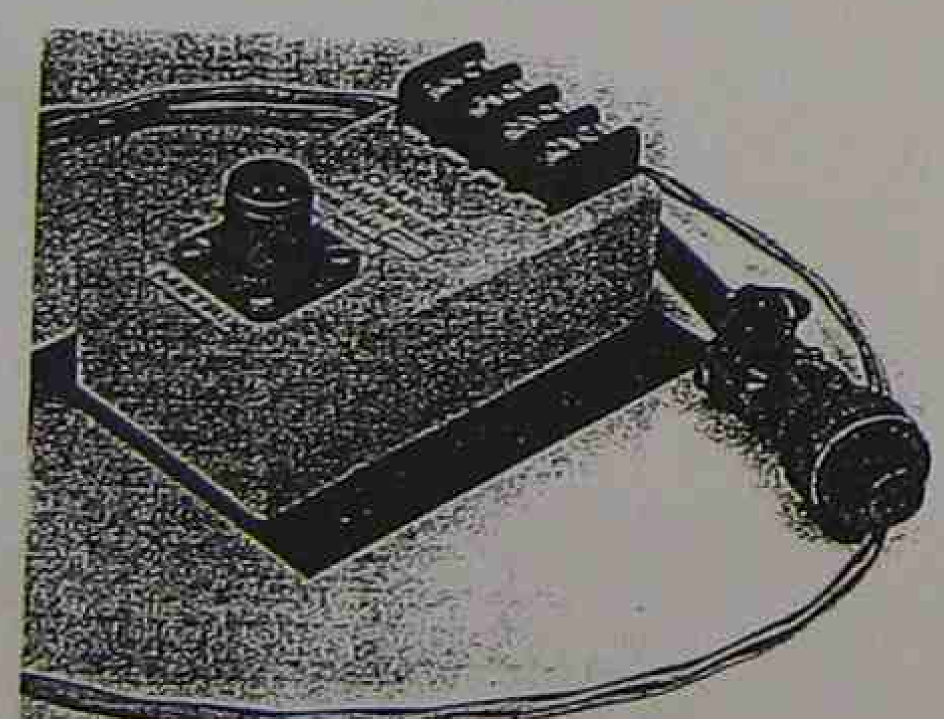

Figure 5-26 Charge amplifier used in conjunction with piezoelectric accelerometers. (Courtesy of Metrix Instrument Co., Houston, TX.)

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

* Not ANSI symbol

FIGURE 7–10
Characteristics of Standard Thermocouple Types

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International Colour Coding for Thermocouple Wires

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

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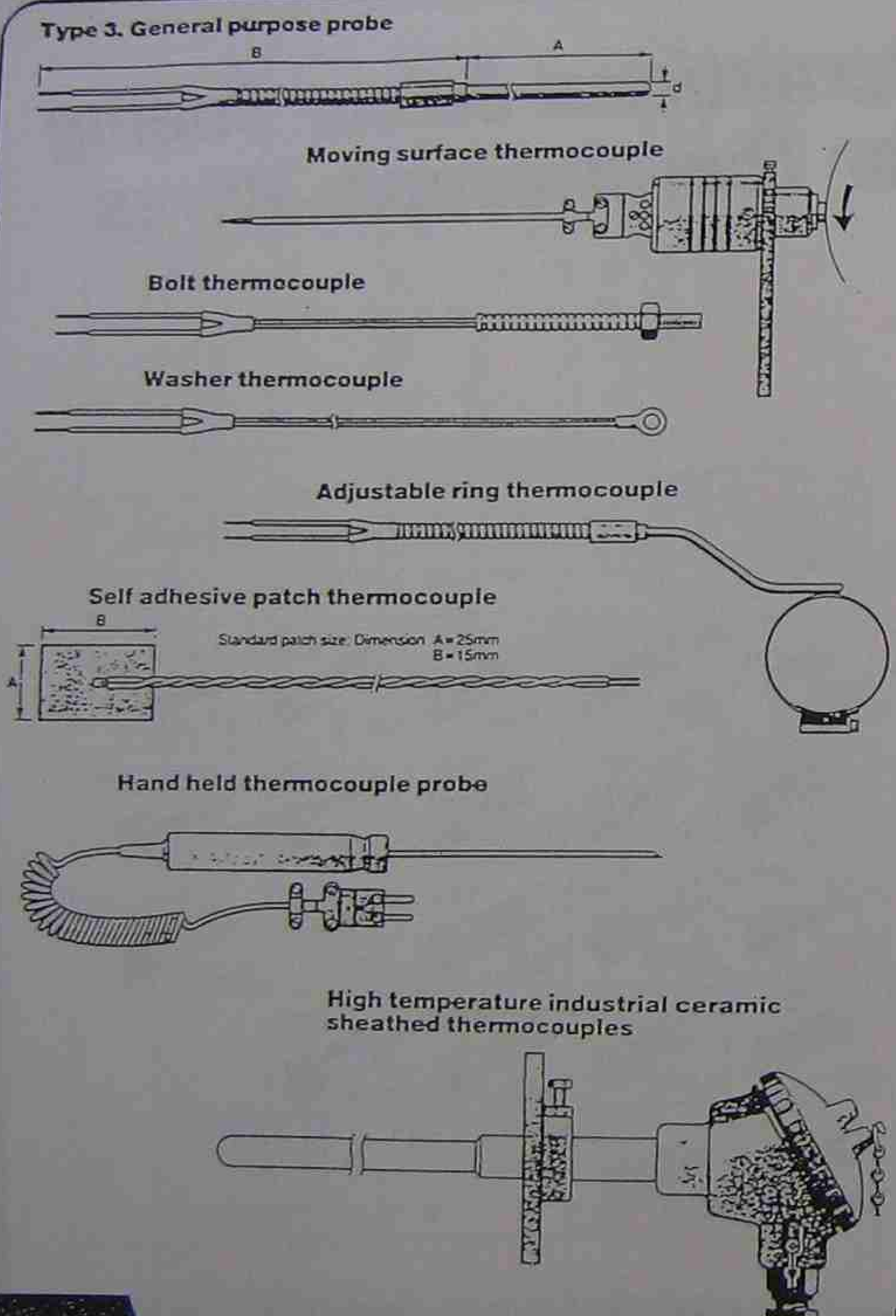


A selection of thermocouple types (courtesy TC Ltd).

THERMOCOUPLE REVERSAL AND DOUBLE REVERSAL.

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

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

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

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

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

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

ASSIGNMENT.

Thermocouple tables

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

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

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

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

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

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

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

Example 4.10

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

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

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

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

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

Example 4.11

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

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

(4.15)

Change of table reference

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

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

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

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

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

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

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

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

400°C results in $V =$

150°C results in $V =$

-90°C results in $V =$

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

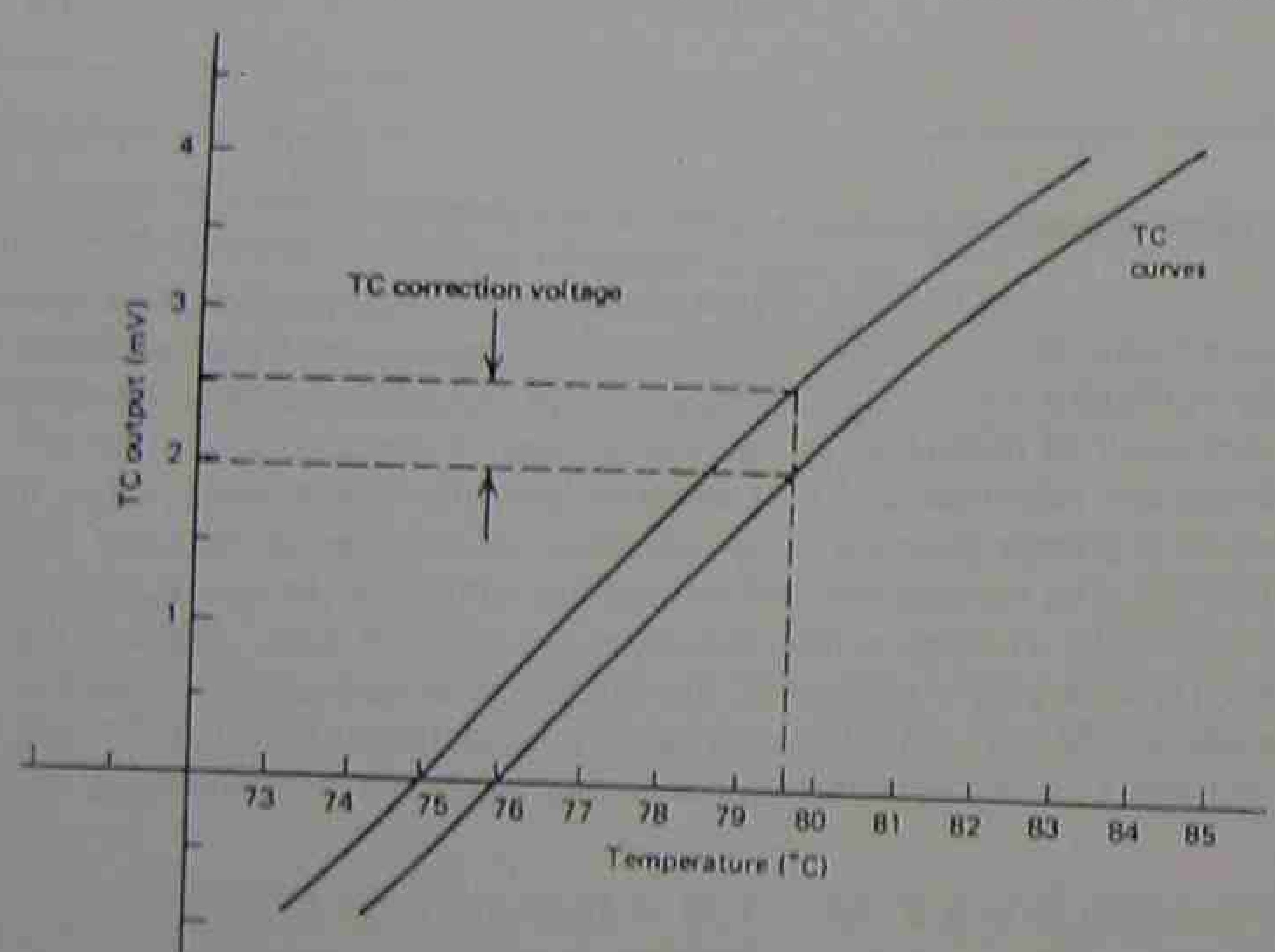


Figure 4.10 When a TC reference is changed, a correction voltage must be applied. If the tables are for 75°C and the actual reference is 76°C, a correction of 0.5 mV is needed.

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

TYPE K THERMOCOUPLES

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

THERMOELECTRIC VOLTAGE IN ABSOLUTE MILLIVOLTS

TYPE K THERMOCOUPLES

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

THERMOELECTRIC VOLTAGE IN ABSOLUTE MILLIVOLTS

THERMOCOUPLE TEMPERATURE CONTROL CIRCUIT

from Maloney "Devices & Systems"
p464-p4655

Brief circuit explanation

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

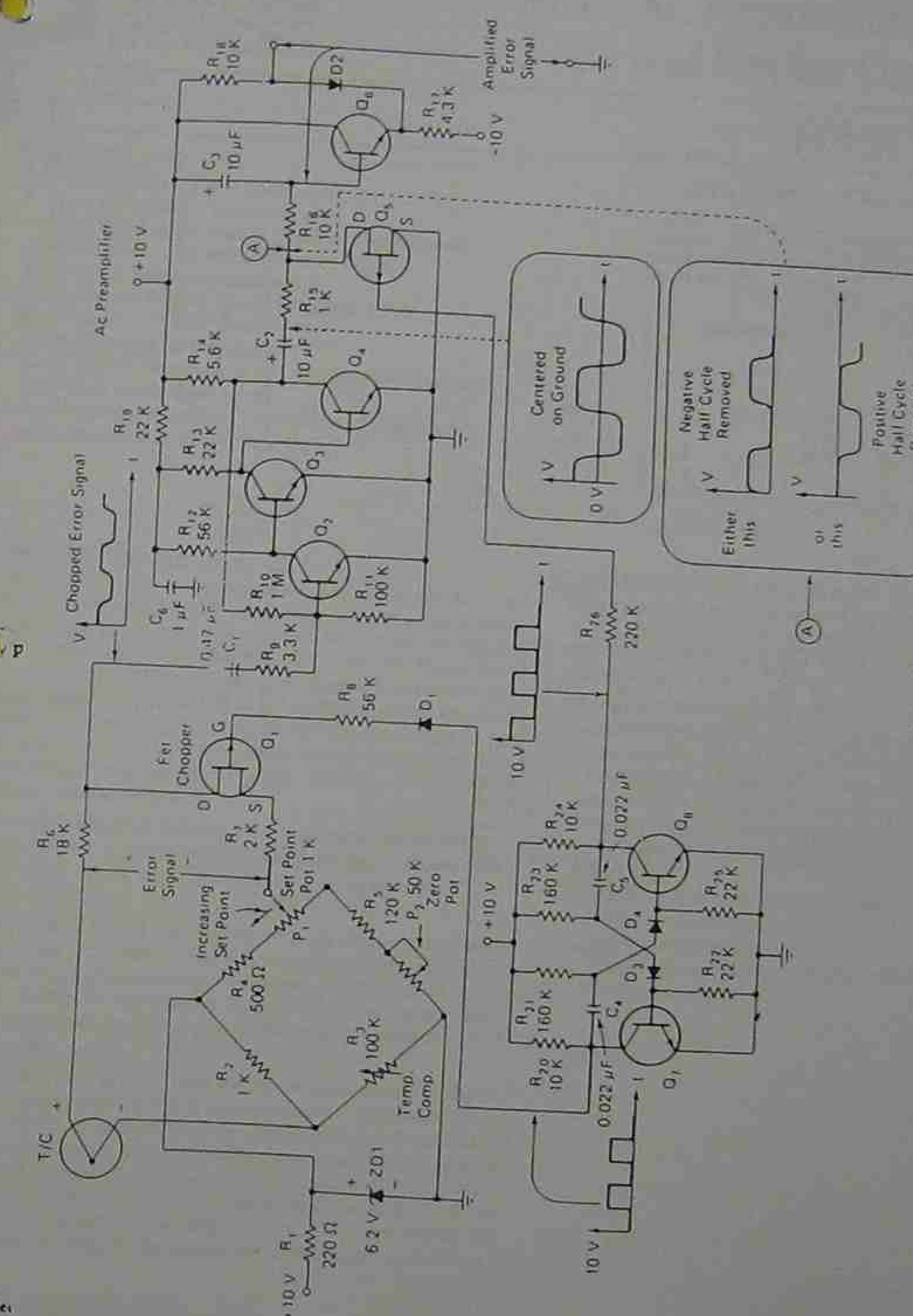
Figure 12.4b is the valve position control circuitry which includes the proportional and integral control of the process.

Figure 12.4a

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

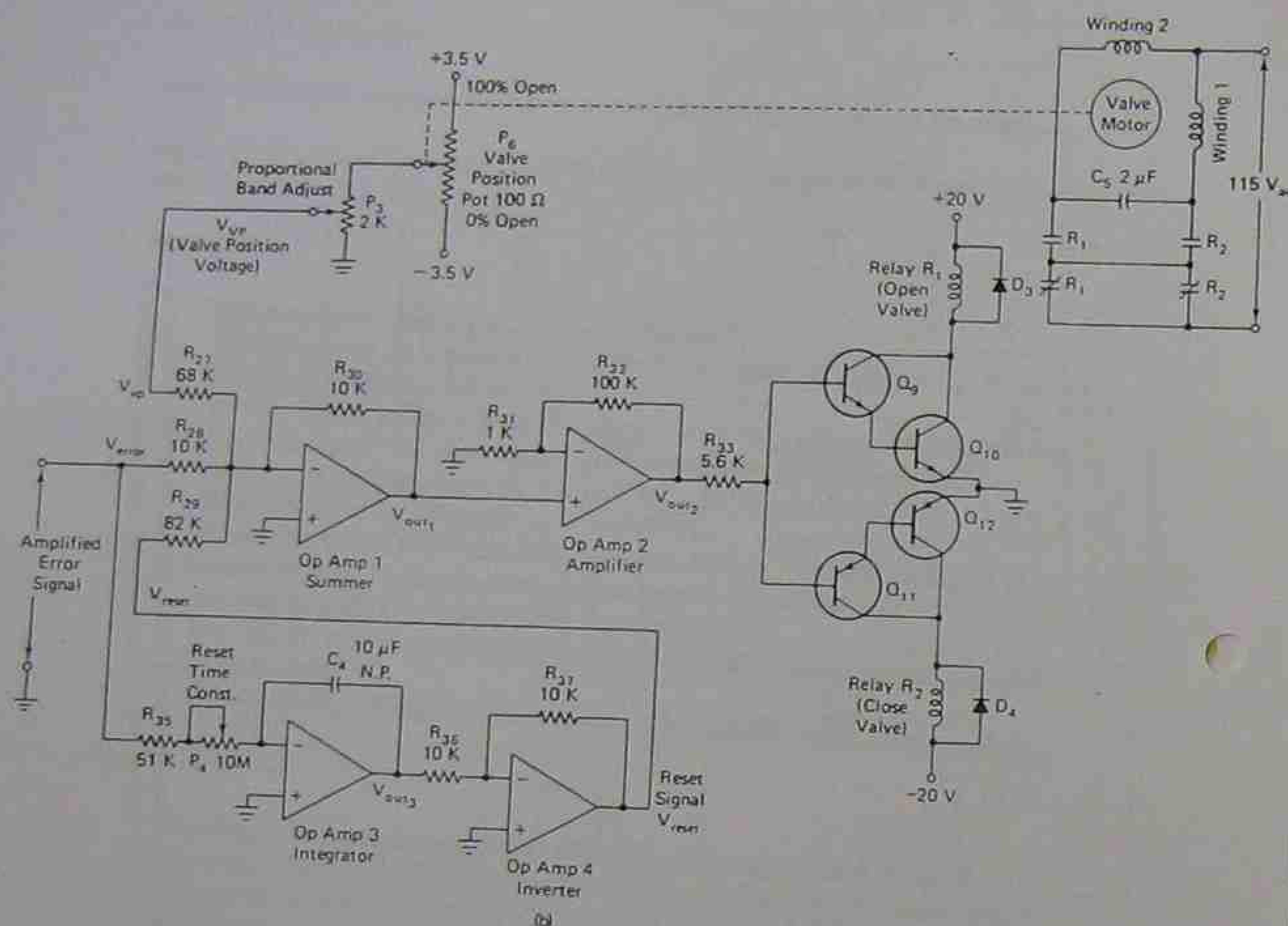
Figure 12.4b

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



(a)

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



(b)

Figure 12-4. (Cont.)

ANALOG DEVICES

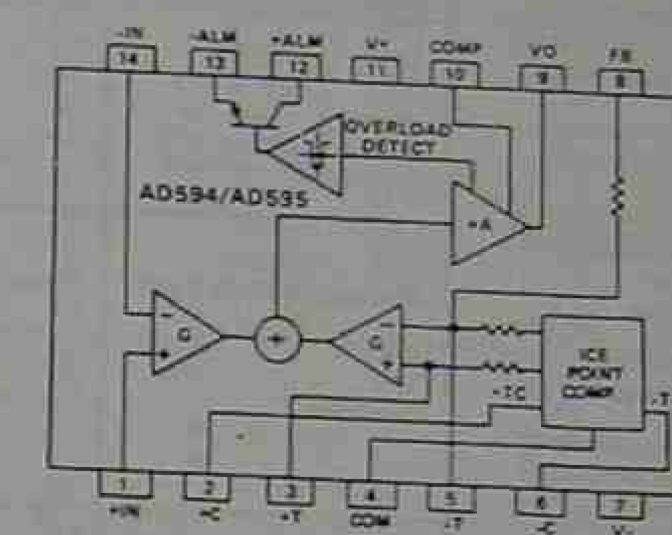
Monolithic Thermocouple Amplifier with Cold Junction Compensation

AD594*/AD595*

FEATURES

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

AD594/AD595 BLOCK DIAGRAM



PRODUCT DESCRIPTION

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

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

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

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

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

PRODUCT HIGHLIGHTS

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

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

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

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

NOTES

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

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

[3] Pin 8 shorted to pin 9.

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

[5] See Section 19 for package outline information.

Specifications subject to change without notice.

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

ORDERING GUIDE

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

Thermocouple Temperature °C	Type J Voltage mV	AD594 Output mV	Type K Voltage mV	AD595 Output mV
−200	−7.890	−1523	−5.891	−1454
−180	−7.402	−1428	−5.550	−1370
−160	−6.821	−1316	−5.141	−1269
−140	−6.159	−1188	−4.669	−1152
−120	−5.426	−1046	−4.138	−1021
−100	−4.632	−893	−3.553	−876
−80	−3.785	−729	−2.920	−719
−60	−2.892	−556	−2.243	−552
−40	−1.960	−376	−1.527	−375
−20	−.995	−189	−.777	−189
−10	−.501	−94	−.392	−94
0	0	3.1	0	2.7
10	.507	101	.397	101
20	1.019	200	.798	200
25	1.277	250	1.000	250
30	1.536	300	1.203	300
40	2.058	401	1.611	401
50	2.585	503	2.022	503
60	3.115	606	2.436	605
80	4.186	813	3.266	810
100	5.268	1022	4.095	1015
120	6.359	1233	4.919	1219
140	7.457	1445	5.733	1420
160	8.560	1659	6.539	1620
180	9.667	1873	7.338	1817
200	10.777	2087	8.137	2015
220	11.887	2302	8.938	2213
240	12.998	2517	9.745	2413
260	14.108	2732	10.560	2614
280	15.217	2946	11.381	2817
300	16.325	3160	12.207	3022
320	17.432	3374	13.039	3327
340	18.537	3588	13.874	3434
360	19.640	3801	14.712	3641
380	20.743	4015	15.552	3849
400	21.846	4228	16.395	4057
420	22.949	4441	17.241	4266
440	24.054	4655	18.088	4476
460	25.161	4869	18.938	4686
480	26.272	5084	19.788	4896
500	27.388	5300	20.640	5107
520	28.511	5517	21.493	5318
540	29.642	5736	22.346	5529
560	30.782	5956	23.198	5740
580	31.933	6179	24.050	5950
600	33.096	6404	24.902	6161
620	34.273	6632	25.751	6371
640	35.464	6862	26.599	6581
660	36.671	7095	27.445	6790
680	37.893	7332	28.288	6998
700	39.130	7571	28.128	7206
720	40.382	7813	29.965	7413
740	41.647	8058	30.799	7619
750	42.283	8181	31.214	7722
760	–	–	31.629	7825
780	–	–	32.455	8029
800	–	–	33.277	8232
820	–	–	34.095	8434
840	–	–	34.909	8636
860	–	–	35.718	8836
880	–	–	36.524	9035
900	–	–	37.325	9233
920	–	–	38.122	9430
940	–	–	38.915	9626
960	–	–	39.703	9821
980	–	–	40.488	10015
1000	–	–	41.269	10209
1020	–	–	42.045	10400
1040	–	–	42.817	10591
1060	–	–	43.585	10781
1080	–	–	44.349	10970
1100	–	–	45.108	11158
1120	–	–	45.863	11345
1140	–	–	46.612	11530
1160	–	–	47.356	11714
1180	–	–	48.095	11897
1200	–	–	48.828	12078
1220	–	–	49.555	12258
1240	–	–	50.276	12436
1250	–	–	50.633	12524

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

INTERPRETING AD594/AD595 OUTPUT VOLTAGES

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

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

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

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

or conversely:

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

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

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

SINGLE AND DUAL SUPPLY CONNECTIONS

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

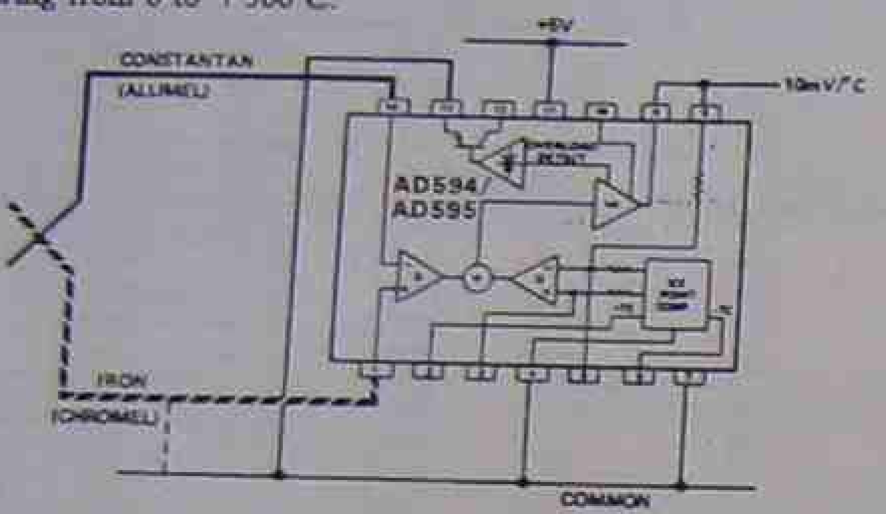


Figure 1. Basic Connection, Single Supply Operation

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

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

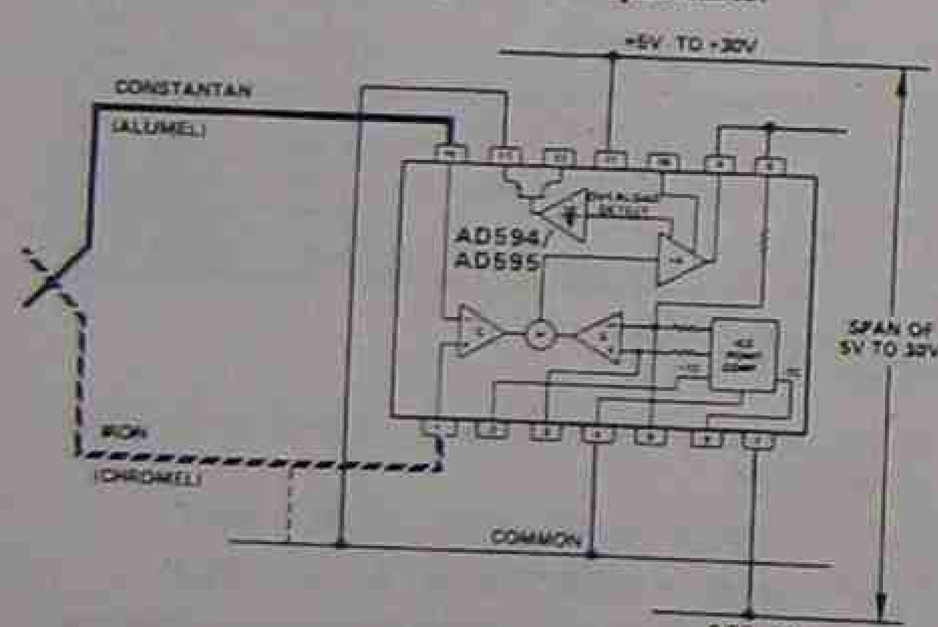


Figure 2. Dual Supply Operation

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

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

THERMOCOUPLE CONNECTIONS

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

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

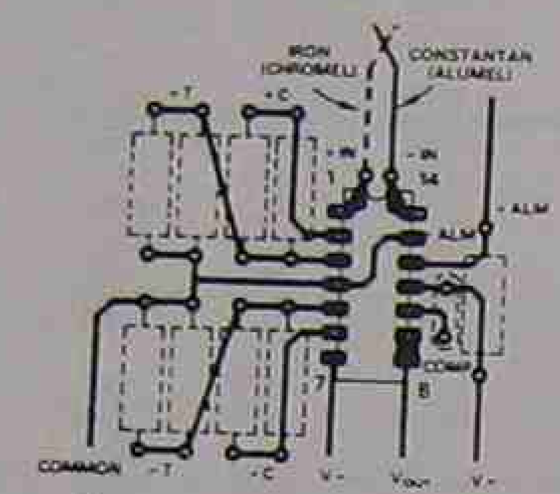


Figure 3. PCB Connections

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

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

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

FUNCTIONAL DESCRIPTION

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

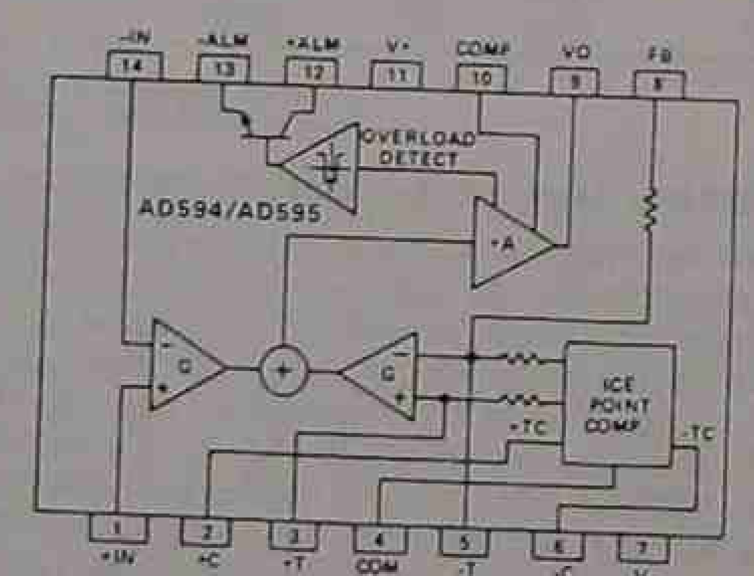


Figure 4. AD594/AD595 Block Diagram

In normal operation the main amplifier output, at pin 9, is connected to the feedback network, at pin 8. Thermocouple signals applied to the floating input stage, at pins 1 and 14, are amplified by gain G of the differential amplifier and are then further amplified by gain A in the main amplifier. The output of the main amplifier is fed back to a second differential stage in an inverting connection. The feedback signal is amplified by this stage and is also applied to the main amplifier input through a summing circuit. Because of the inversion, the amplifier causes the feedback to be driven to reduce this difference signal to a small value. The two differential amplifiers are made to match and have identical gains, G. As a result, the feedback signal that must be applied to the right-hand differential amplifier will precisely match the thermocouple input signal when the difference signal has been reduced to zero. The feedback network is trimmed so that the effective gain to the output, at pins 8 and 9, results in a voltage of 10mV/°C of thermocouple excitation.

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

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

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

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

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

CAUTIONS:

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

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

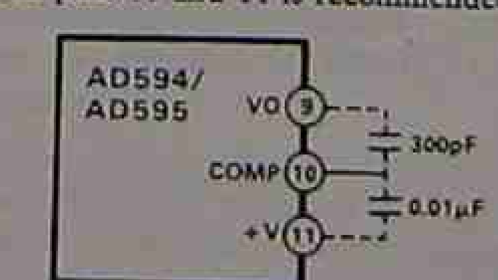


Figure 5. Low Gain Frequency Compensation

RECALIBRATION PRINCIPLES AND LIMITATIONS

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

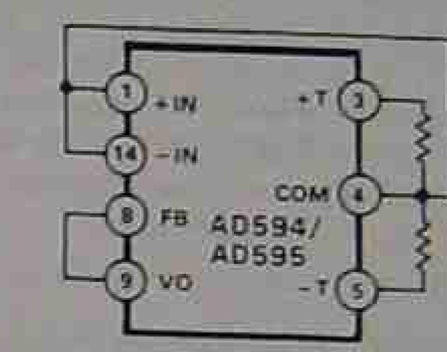


Figure 6. Decreased Sensitivity Adjustment

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

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

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

EXAMPLE: TYPE E RECALIBRATION – AD594/AD595

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

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

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

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

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

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

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

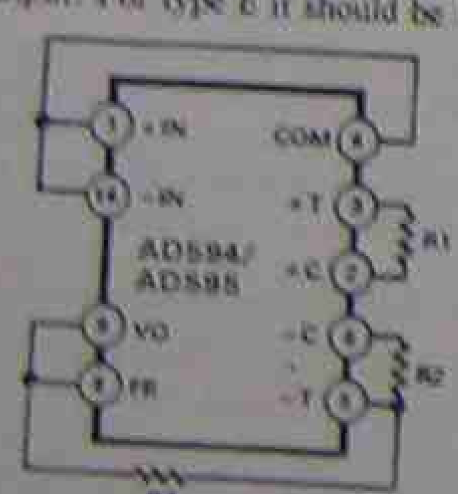


Figure 7. Type E Recalibration

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

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

USING TYPE T THERMOCOUPLES WITH THE AD595

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

STABILITY OVER TEMPERATURE

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

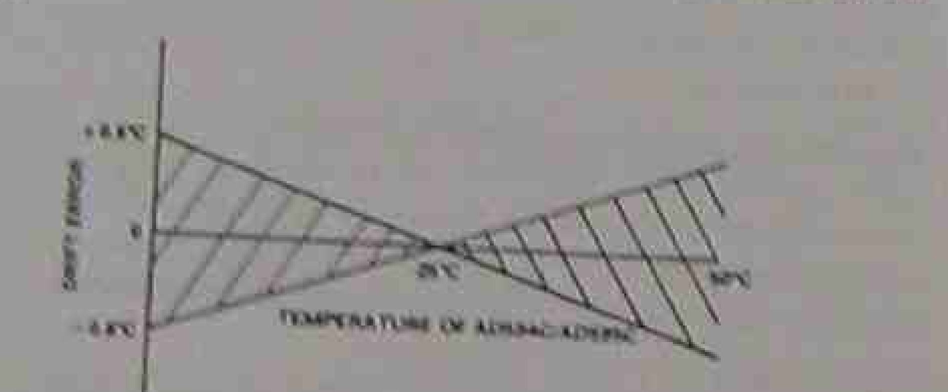


Figure 8. Drift Error vs. Temperature

THERMAL ENVIRONMENT EFFECTS

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

SET-POINT CONTROLLER

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

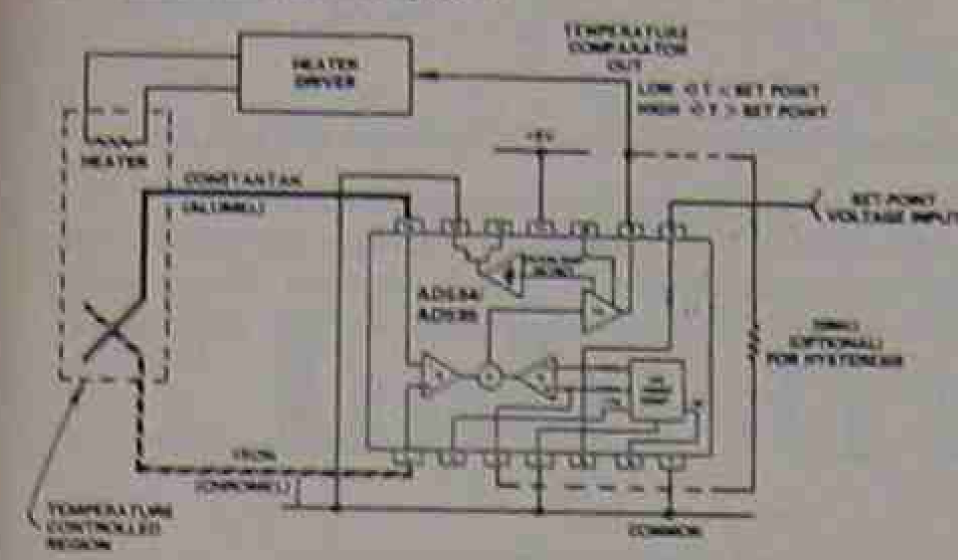


Figure 9. Set-Point Controller

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

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

ALARM CIRCUIT

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

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

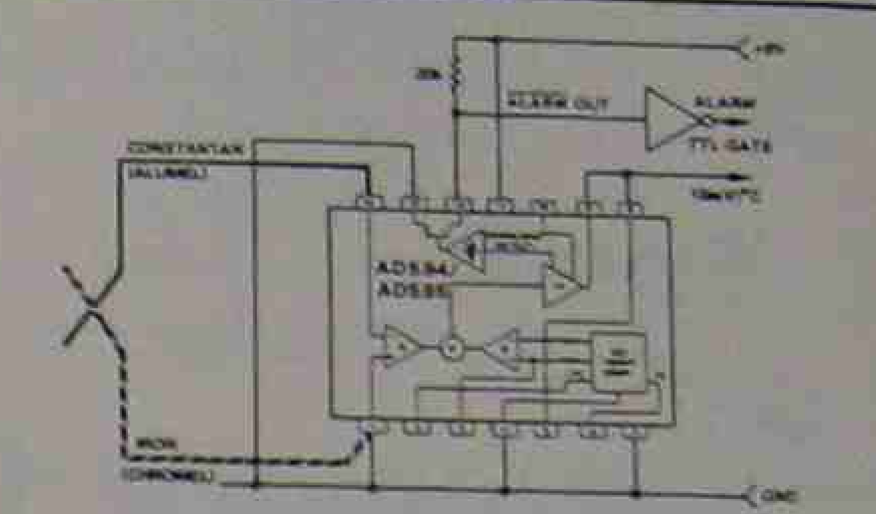


Figure 10. Using the Alarm to Drive a TTL Gate ("Grounded" Emitter Configuration)

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

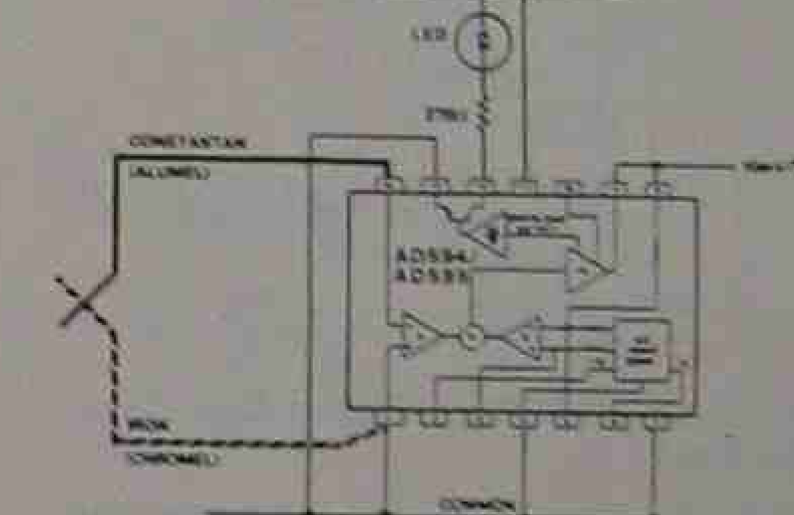


Figure 11. Alarm Directly Drives LED

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

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

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

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

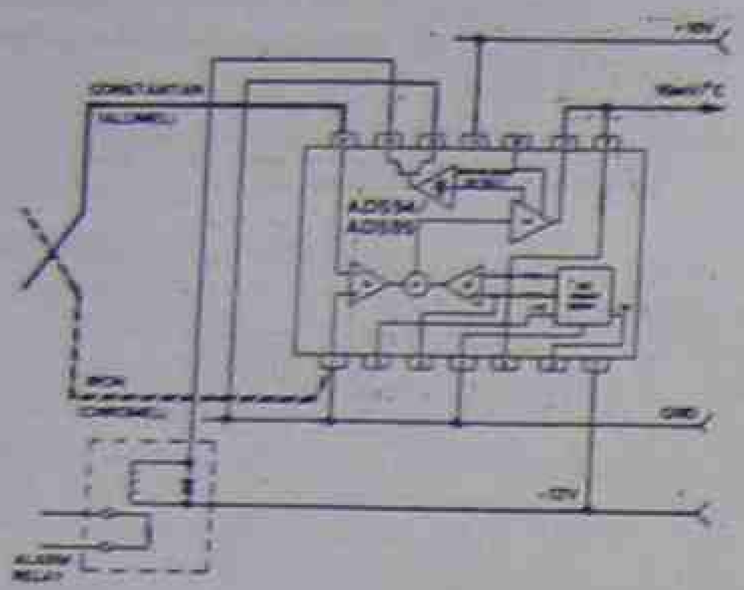


Figure 12. —ALM Driving A Negative Referenced Load

CELSIUS THERMOMETER

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

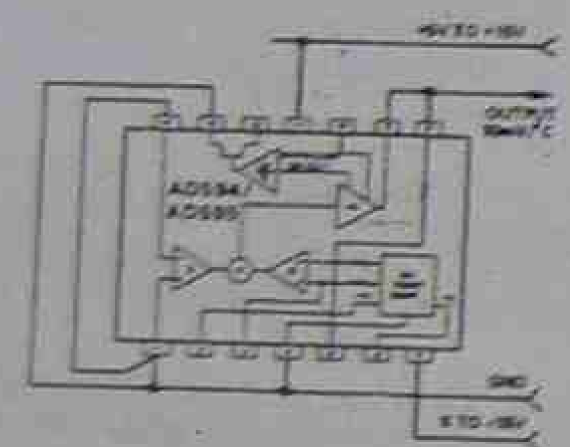


Figure 13. AD594/AD595 as a Stand-Alone Celsius Thermometer

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

THERMOCOUPLE BASICS

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

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

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

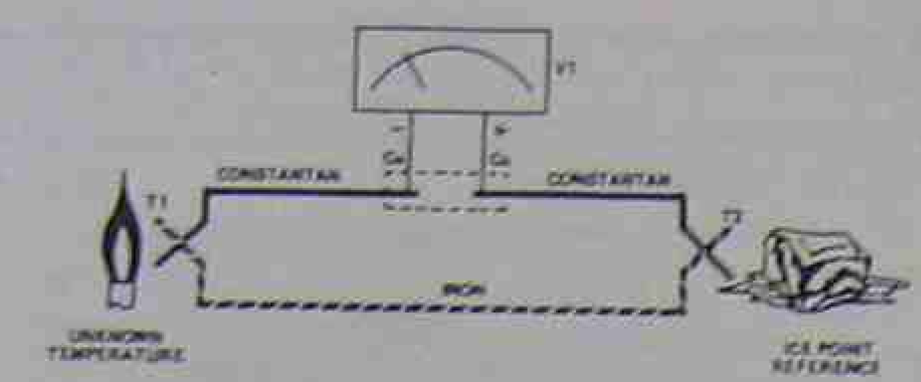


Figure 14. Thermocouple Voltage with 0°C Reference

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

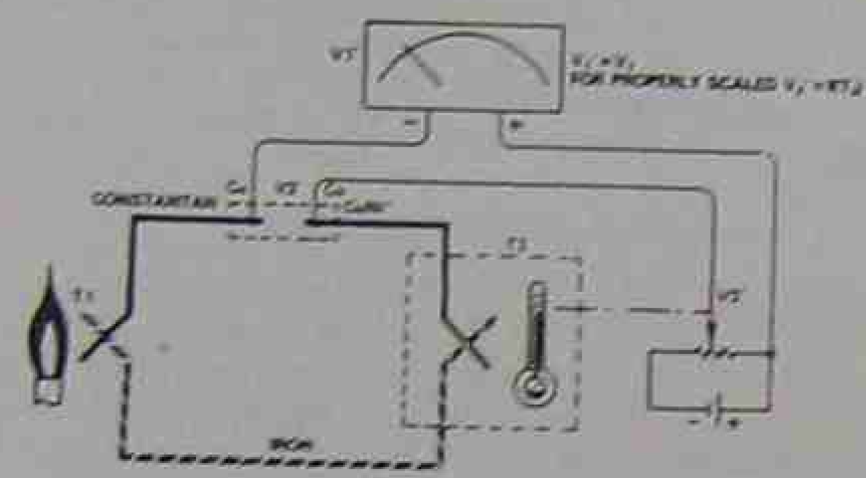


Figure 15. Substitution of Measured Reference Temperature for Ice Point Reference

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

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

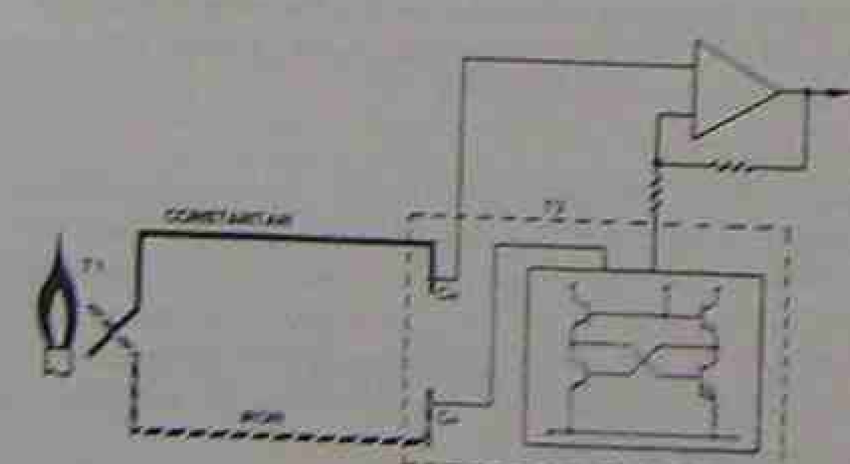


Figure 16. Connecting Isothermal Junctions

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

Sydney Institute of Technology

School of Electrotechnology

Industrial Sensors

Theory Notes

Area ________________

Topic ________________

Session No.

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

13-1 INTRODUCTION

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

13-2 COMMONLY SENSED MEASURANDS

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

13-3 RESISTANCE TEMPERATURE DETECTOR

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

13-3.1 Temperature Coefficients of Metals

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

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

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

where E = thermal energy (J)

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

T = absolute temperature (K)

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

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

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

$$R = \frac{\rho l}{A_{CM}} \tag{13-2}$$

where R = resistance of wire (Ω)

ρ = resistivity constant (Ω-mil/ft)

l = length (ft)

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

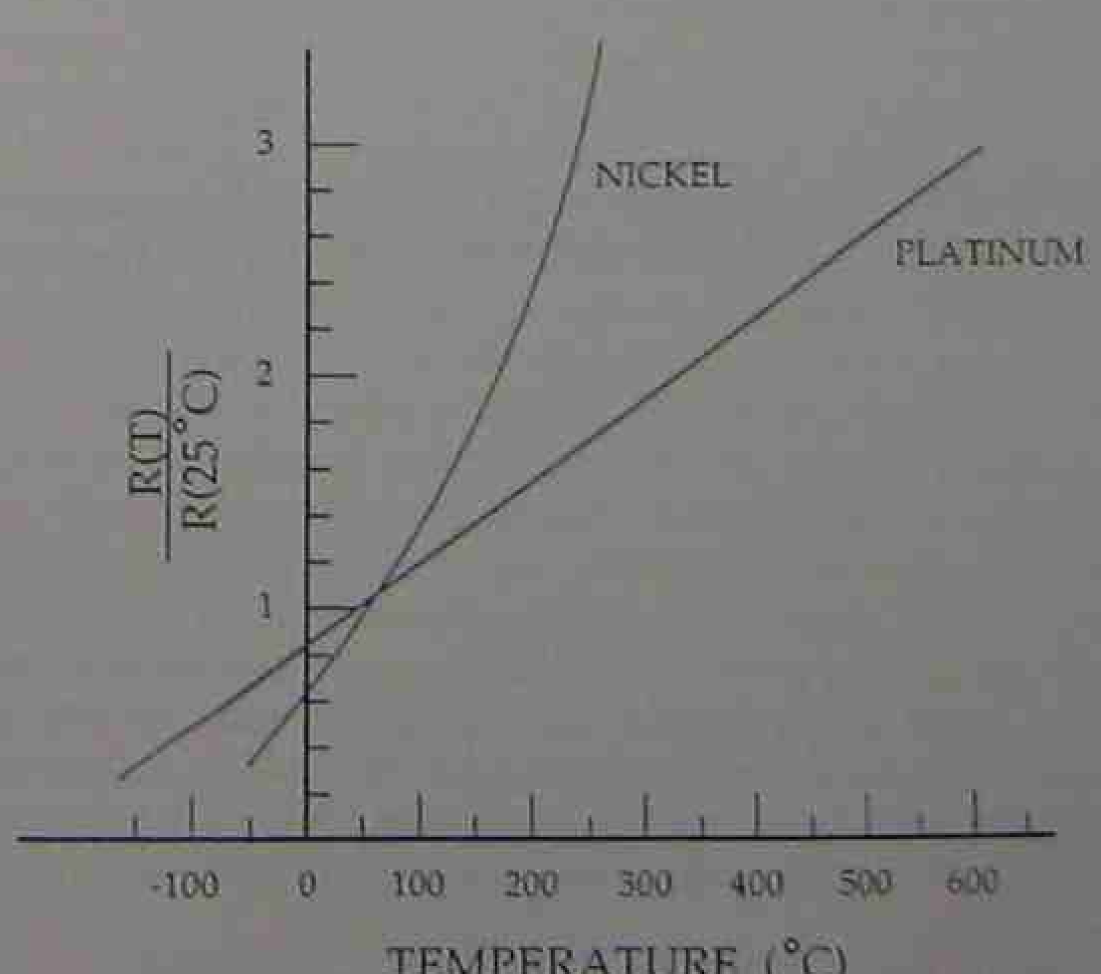


Figure 13-1 Graph of dc resistance for various metals versus temperature

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

Example 13-1

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

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

$$R =$$

$$=$$

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

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

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

T = difference between the operating temperature and 20°C

Example 13-2

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

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

$$R =$$

$$=$$

$$=$$

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

13-3.2 Metals Used for Temperature Sensing

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

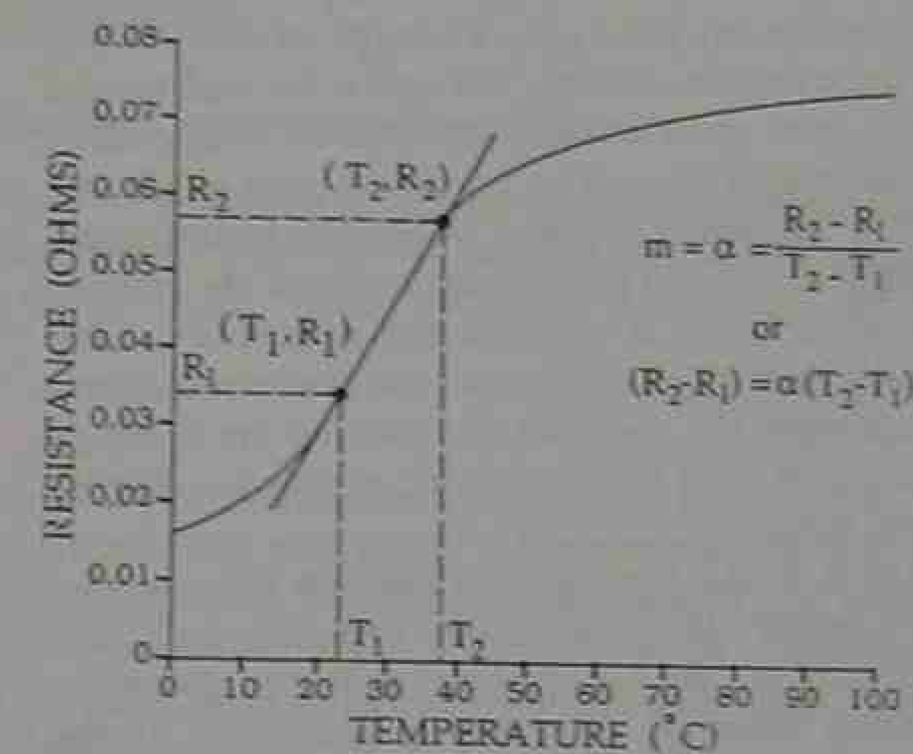


Figure 13-2 Temperature versus dc resistance for platinum.

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

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

13-3.3 Typical Construction

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

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

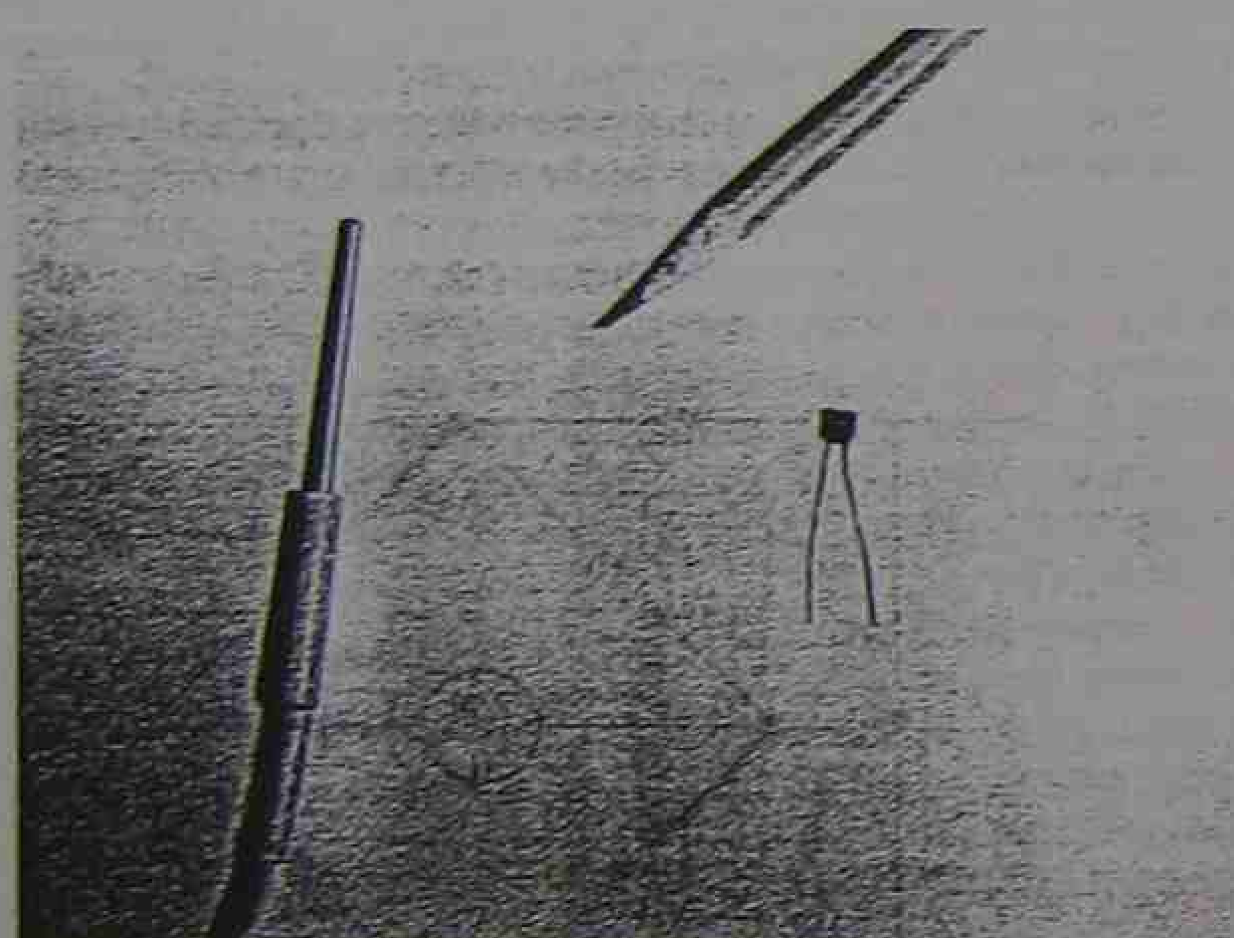


Figure 13-3 Shown at left, wire-wound platinum RTD. At right, deposited-film RTD.

13-3.4 RTD Signal-Conditioning Circuits

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

Two-wire uncompensated RTD circuit

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

$$\frac{R_1}{R_2} = \frac{R_{L1} + R_{L2} + R_3}{R_4} \qquad (13\text{-}4)$$

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

Two-wire compensated RTD circuit

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

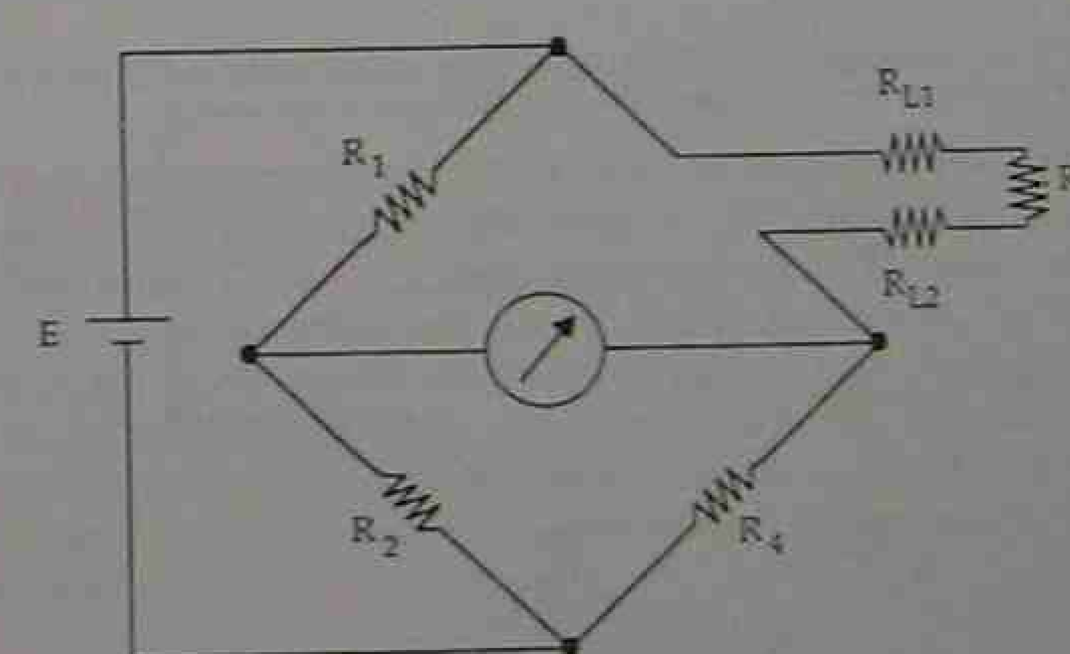


Figure 13-4 Two-wire uncompensated RTD circuit.

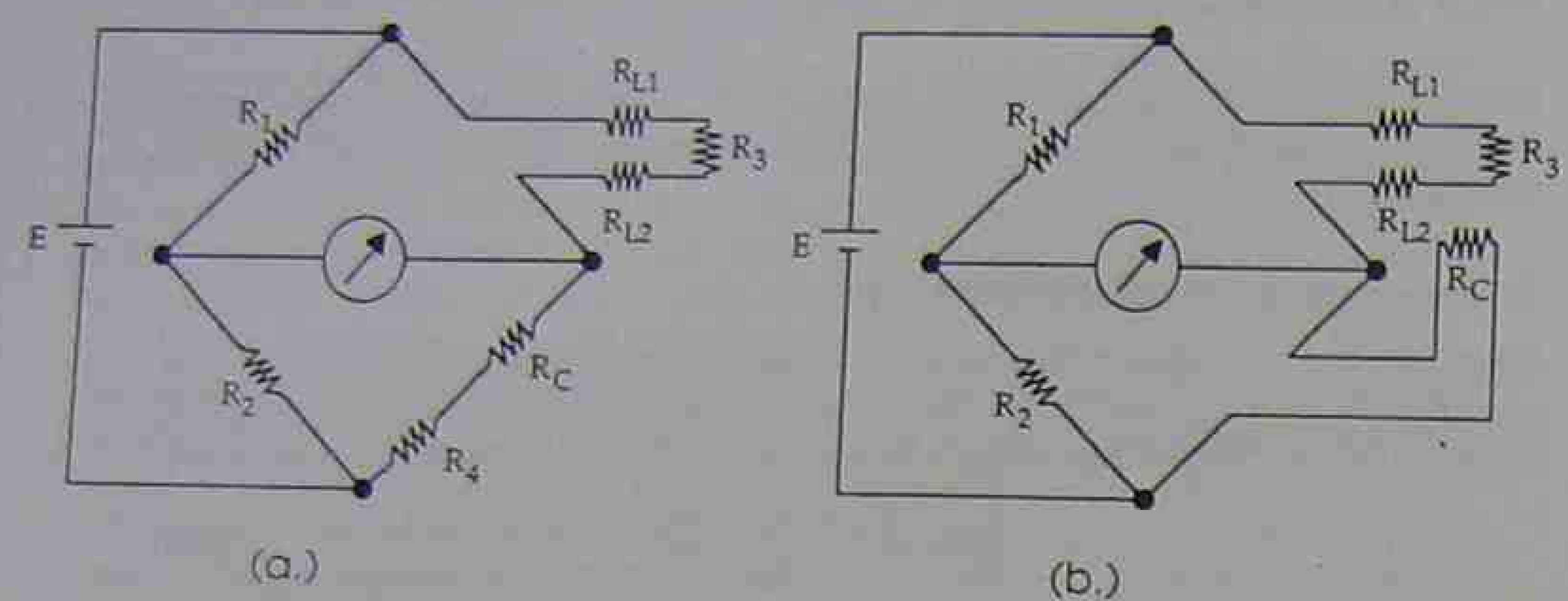


Figure 13-5 Two-wire compensated RTD circuits.

Three-wire RTD circuit

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

Four-wire RTD circuit

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

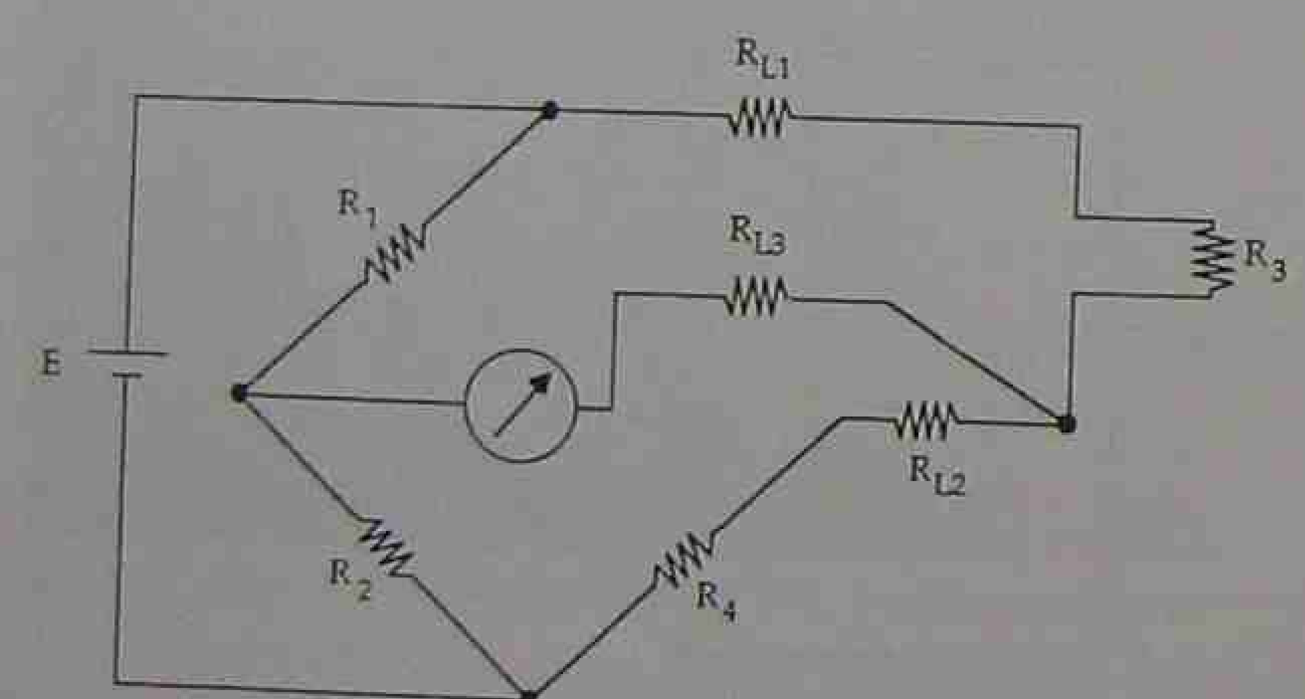


Figure 13-6 Three-wire lead circuit.

13-3.5 Self-Heat

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

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

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

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

P_{SHE} = self-heat error (°C/mW)

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

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

SEE EXAMPLE 13-3

13-3.6 Typical Characteristics and Design Specifications

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

Example 13-3

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

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

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

Now, using eq. (13-5),

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

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

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

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

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

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

Therefore,

$$\text{actual temperature} = 175.84°\text{C} - 3.24°\text{C}$$
$$= 172.60°\text{C}$$

13-3.7 Practical Applications

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

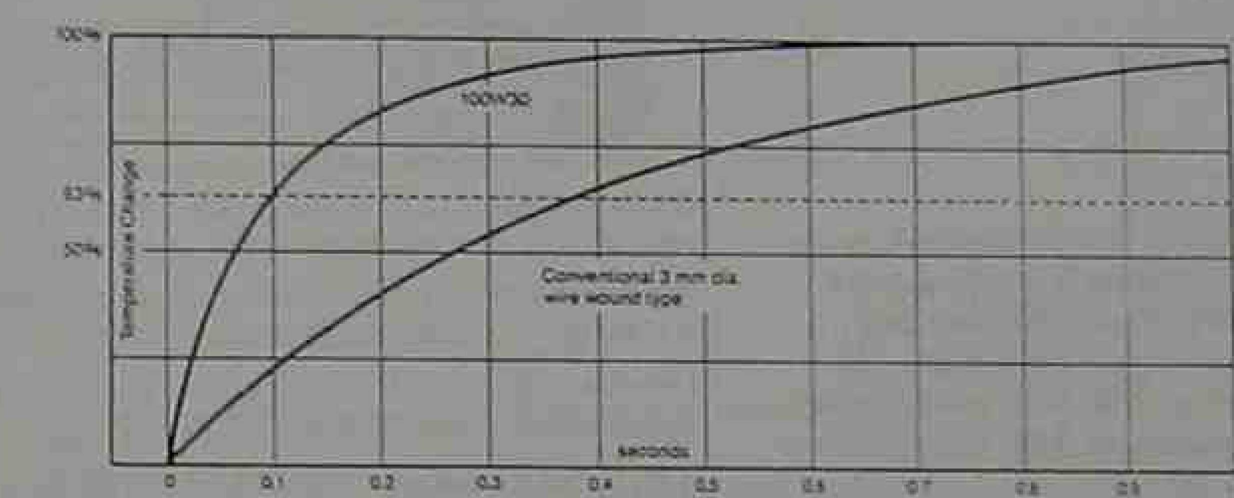


Figure 13-7 Time-response curve for a platinum wire RTD. (REPRODUCED WITH THE PERMISSION OF OMEGA ENGINEERING, INC.)

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

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

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

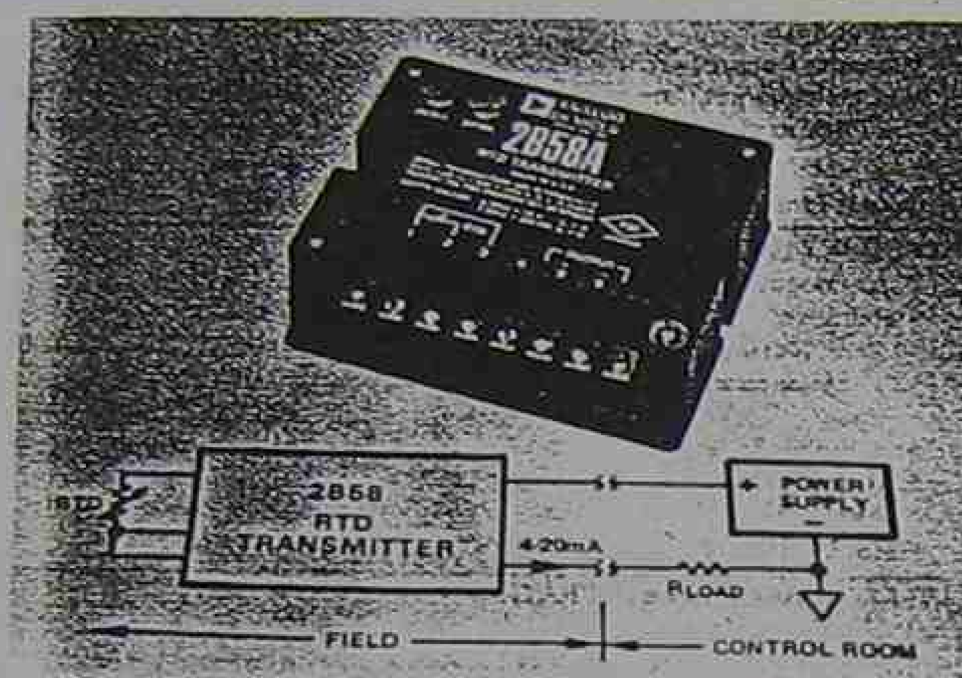
ANALOG DEVICES

Two-Wire, Linearized RTD Temperature Transmitter

MODEL 2B58

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

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



9

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

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

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

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

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

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

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

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

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

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

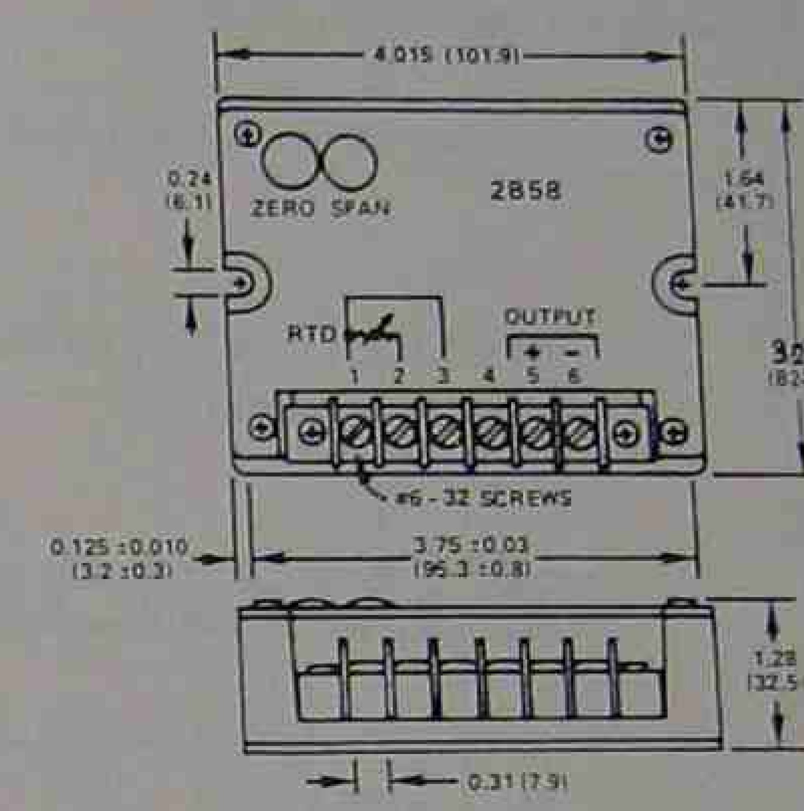
SPECIFICATIONS (typical @ +25°C and V_S = +24V dc unless otherwise noted)

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

NOTES
[1] Protected for reverse polarity and for any input/output connection combination.
[2] Accuracy is specified as a percent of output span (16mA). Accuracy spec includes combined effects of transmitter repeatability, hysteresis and sensor linearization conformity. Does not include sensor error.
[3] See ordering information for measurement temperature ranges 01 through 04.
[4] Per MIL-STD-202E Method 103B.

Specifications subject to change without notice.

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



ORDERING INFORMATION

Example: Model 2B58A - 1 - 1 - 01

Enter Model
2B58A

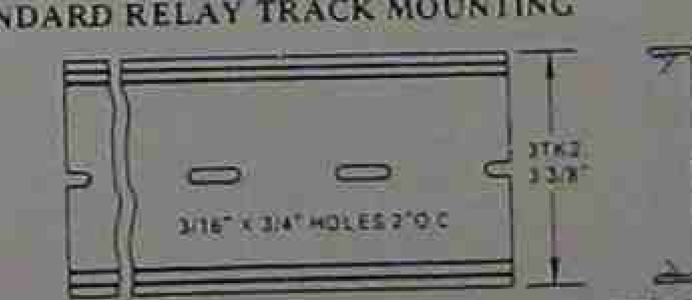
Enter Housing
1-Standard Enclosure

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

Enter Temperature Range
01 Through 04

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

STANDARD RELAY TRACK MOUNTING



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

TEMPERATURE CONTROLS PTY LTD

A.C.N. 003 512 294

EY
(02) 560 0644 Fax (02) 560 0339

MELBOURNE
Phone (03) 364 0677 Fax (03) 364 1461

THERMOCOUPLES • RTD SENSORS • THERMOWELLS • EXTENSION CABLES

Reference table for platinum resistance elements

R.T.D.

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

°C	0	1	2	3	4	5	6	7	8	9
0	100.00	100.39	100.78	101.17	101.56	101.95	102.34	102.73	103.12	103.51
10	103.90	104.29	104.68	105.07	105.46	105.85	106.24	106.63	107.02	107.40
20	107.79	108.18	108.57	108.96	109.35	109.73	110.12	110.51	110.90	111.28
30	111.67	112.06	112.45	112.83	113.22	113.61	113.99	114.38	114.77	115.15
40	115.54	115.93	116.31	116.70	117.08	117.47	117.85	118.24	118.62	119.01
50	119.40	119.78	120.16	120.55	120.93	121.32	121.70	122.09	122.47	122.86
60	123.24	123.62	124.01	124.39	124.77	125.16	125.54	125.92	126.31	126.69
70	127.07	127.45	127.84	128.22	128.60	128.98	129.37	129.75	130.13	130.51
80	130.89	131.27	131.66	132.04	132.42	132.80	133.18	133.56	133.94	134.32
90	134.70	135.08	135.46	135.84	136.22	136.60	136.98	137.36	137.74	138.12
100	138.50	138.88	139.26	139.64	140.02	140.39	140.77	141.15	141.53	141.91
110	142.29	142.66	143.04	143.42	143.80	144.17	144.55	144.93	145.31	145.68
120	146.08	146.44	146.81	147.19	147.57	147.94	148.32	148.70	149.07	149.45
130	149.82	150.20	150.57	150.95	151.33	151.70	152.08	152.45	152.83	153.20
140	153.58	153.95	154.32	154.70	155.07	155.45	155.82	156.19	156.57	156.94
150	157.31	157.69	158.06	158.43	158.81	159.18	159.55	159.93	160.30	160.67
160	161.04	161.42	161.79	162.16	162.53	162.90	163.27	163.65	164.02	164.39
170	164.76	165.13	165.50	165.87	166.24	166.61	166.98	167.35	167.72	168.09
180	168.45	168.83	169.20	169.57	169.94	170.31	170.68	171.05	171.42	171.79
190	172.16	172.53	172.90	173.26	173.63	174.00	174.37	174.74	175.10	175.47
200	175.84	176.21	176.57	176.94	177.31	177.68	178.04	178.41	178.78	179.14
210	179.51	179.88	180.24	180.61	180.97	181.34	181.71	182.07	182.44	182.80
220	183.17	183.53	183.90	184.26	184.63	184.99	185.36	185.72	186.09	186.45
230	186.82	187.18	187.54	187.91	188.27	188.63	189.00	189.36	189.72	190.09
240	190.45	190.81	191.18	191.54	191.90	192.26	192.63	192.99	193.35	193.71
250	194.07	194.44	194.80	195.16	195.52	195.88	196.24	196.60	196.96	197.33
260	197.69	198.05	198.41	198.77	199.13	199.49	199.85	200.21	200.57	200.93
270	201.29	201.65	202.01	202.36	202.72	203.08	203.44	203.80	204.16	204.52
280	204.88	205.23	205.59	205.95	206.31	206.67	207.02	207.38	207.74	208.10
290	208.45	208.81	209.17	209.52	209.88	210.24	210.59	210.95	211.31	211.66
300	212.02	212.37	212.73	213.09	213.44	213.80	214.15	214.51	214.86	215.22
310	215.57	215.93	216.28	216.64	216.99	217.35	217.70	218.05	218.41	218.76
320	219.12	219.47	219.82	220.18	220.53	220.88	221.24	221.59	221.94	222.29
330	222.65	223.00	223.35	223.70	224.06	224.41	224.76	225.11	225.46	225.81
340	226.17	226.52	226.87	227.22	227.57	227.92	228.27	228.62	228.97	229.32
350	229.67	230.02	230.37	230.72	231.07	231.42	231.77	232.12	232.47	232.82
360	233.17	233.52	233.87	234.22	234.56	234.91	235.26	235.61	235.96	236.31
370	236.65	237.00	237.35	237.70	238.04	238.39	238.74	239.09	239.43	239.78
380	240.13	240.47	240.82	241.17	241.51	241.86	242.20	242.55	242.90	243.24
390	243.59	243.93	244.28	244.62	244.97	245.31	245.66	246.00	246.35	246.69
400	247.04									

PROUDLY MANUFACTURING IN AUSTRALIA

Sydney Institute of Technology

School of Electrotechnology

Industrial Sensors

Theory Notes

Area Industrial Sensors

Topic Industrial Sensors

Session No.

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

13-4 THERMISTOR

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

13-4.1 Theory of Operation

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

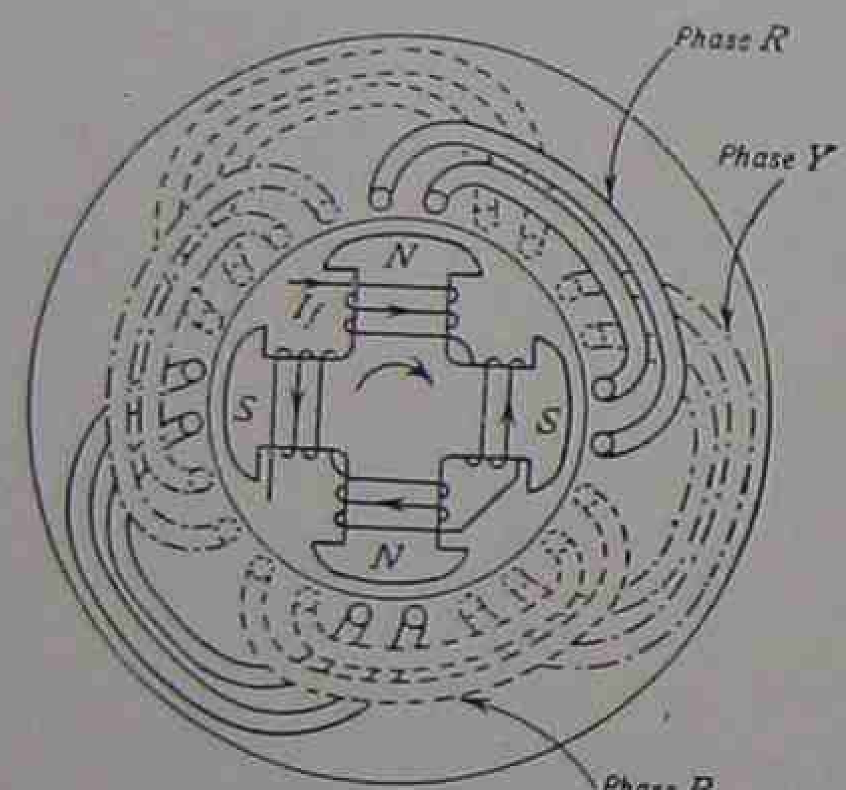


Fig. 188.—End connections of a three-phase single-layer winding.

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

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

13-4.2 Characteristics of the Thermistor

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

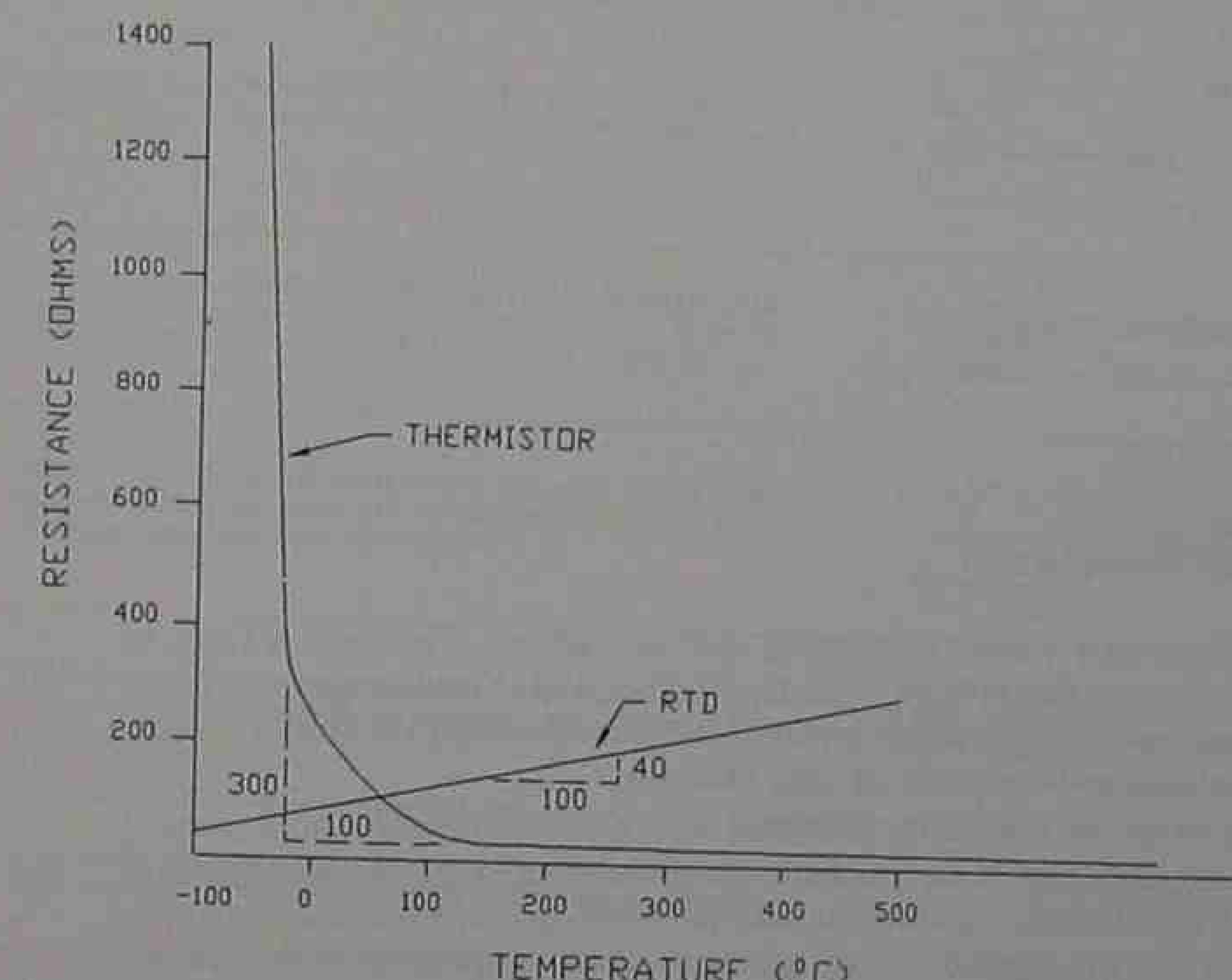


Figure 13-10 Temperature versus resistance curve for a thermistor.

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

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

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

Example 13-4

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

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

13-4.3 Typical Construction

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

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

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

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

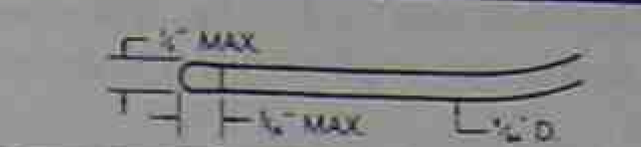
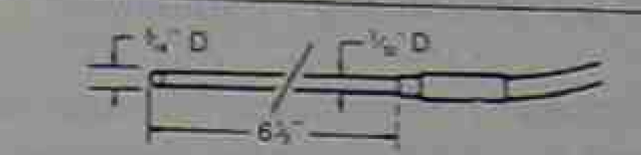
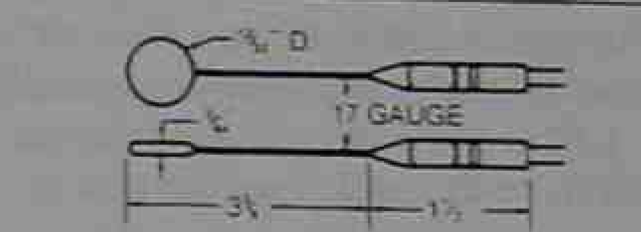
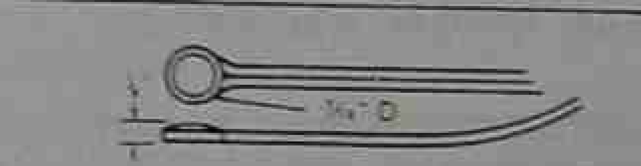
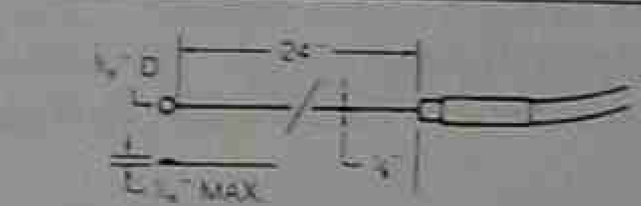
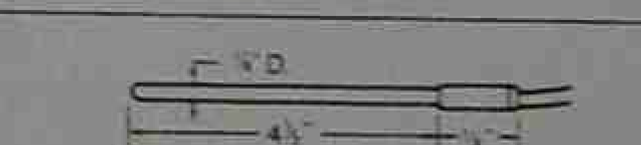
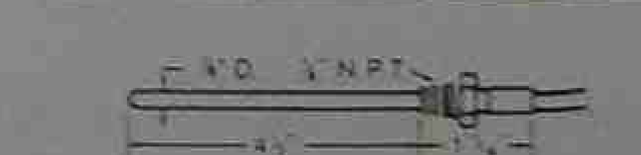
Description	Configuration
GENERAL PURPOSE. Vinyl tipped, most rugged probe. Used for short-term water and sub-soil readings. Waterproof construction now standard.	
SMALL FLEXIBLE. Vinyl sheath and tip. Cuvette temperatures. General purpose measurement.	
GENERAL PURPOSE. Non-immersible, epoxy tipped probe.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

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

13-4.4 Thermistor Self-Heat

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

13-4.5 Linearizing Thermistors

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

13-4.6 Practical Applications

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

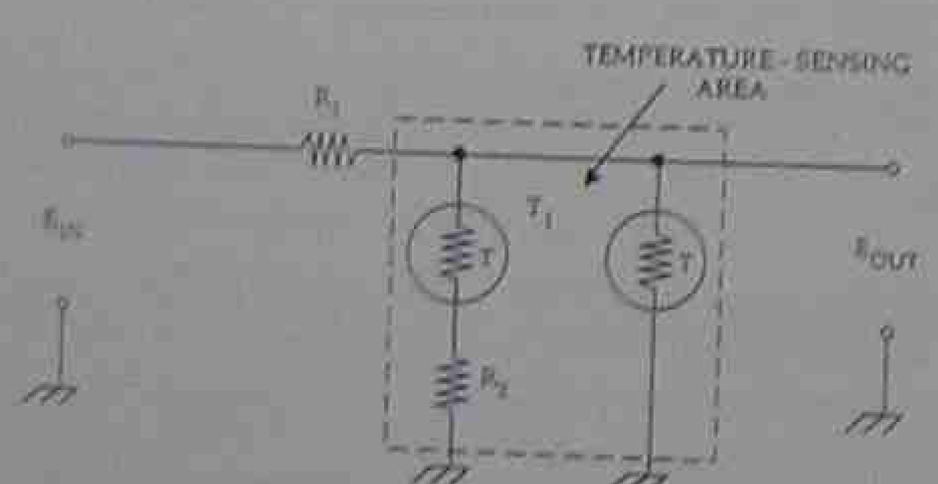


Figure 13-13 Linearizing circuit for a thermistor

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

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

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

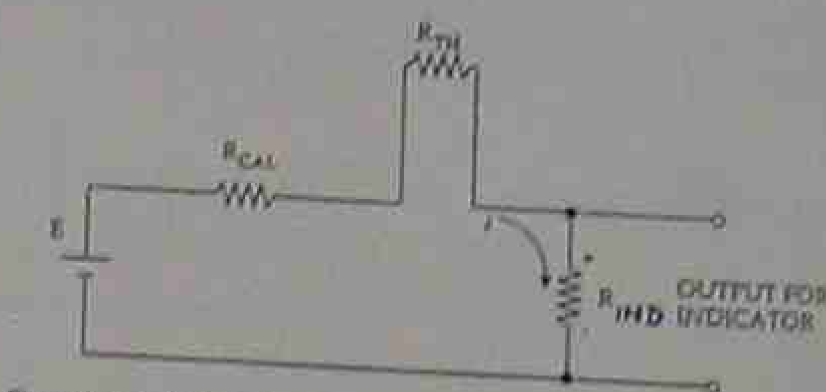


Figure 13-14 Using a simple ohmmeter circuit to read a thermistor's resistance

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

Example 13-5

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

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

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

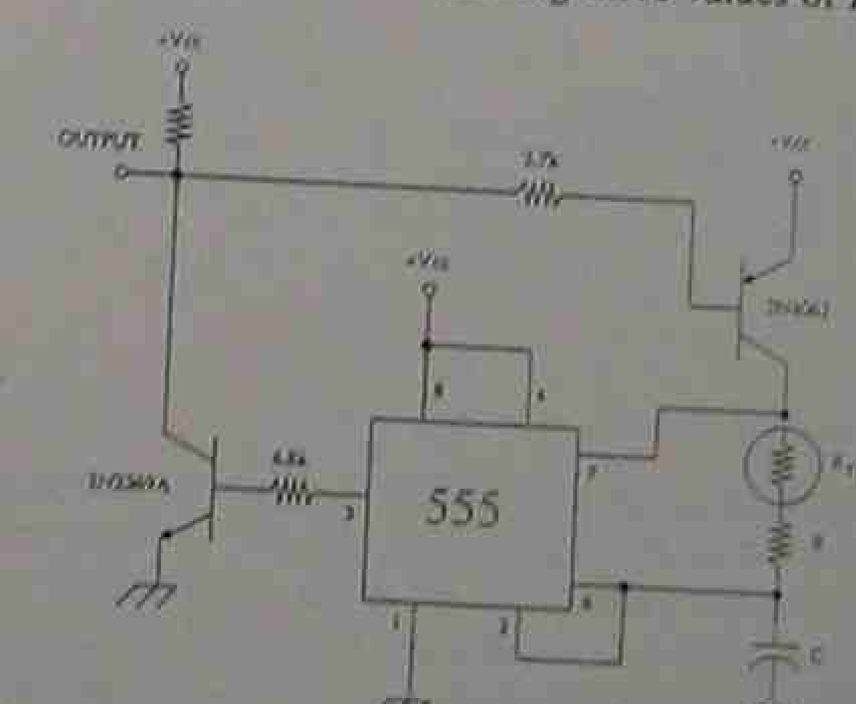


Figure 13-15 Using a thermistor to control the frequency output of an oscillator. (From Howard M. Berlin, *The 555 Timer Applications Sourcebook*, copyright 1976, p. 65. Reprinted by permission of the author.)

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

f_1

f_2

f_3

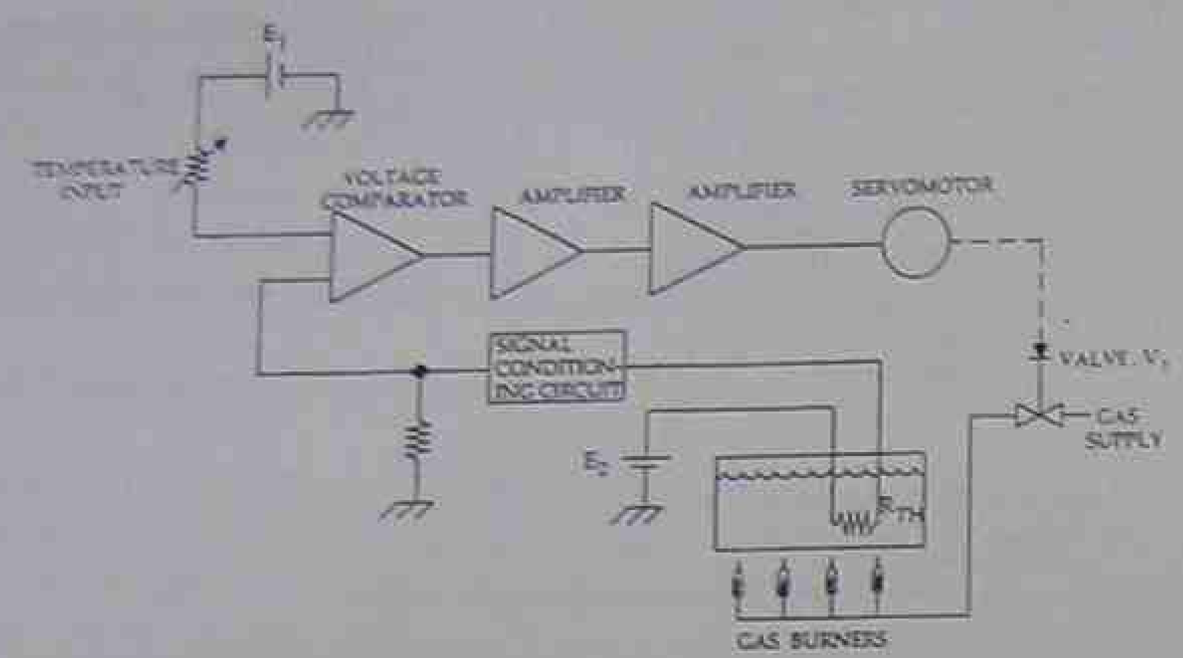


Figure 13-16 Thermistor being used in a process control system for controlling the temperature of a liquid in a heated container.

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

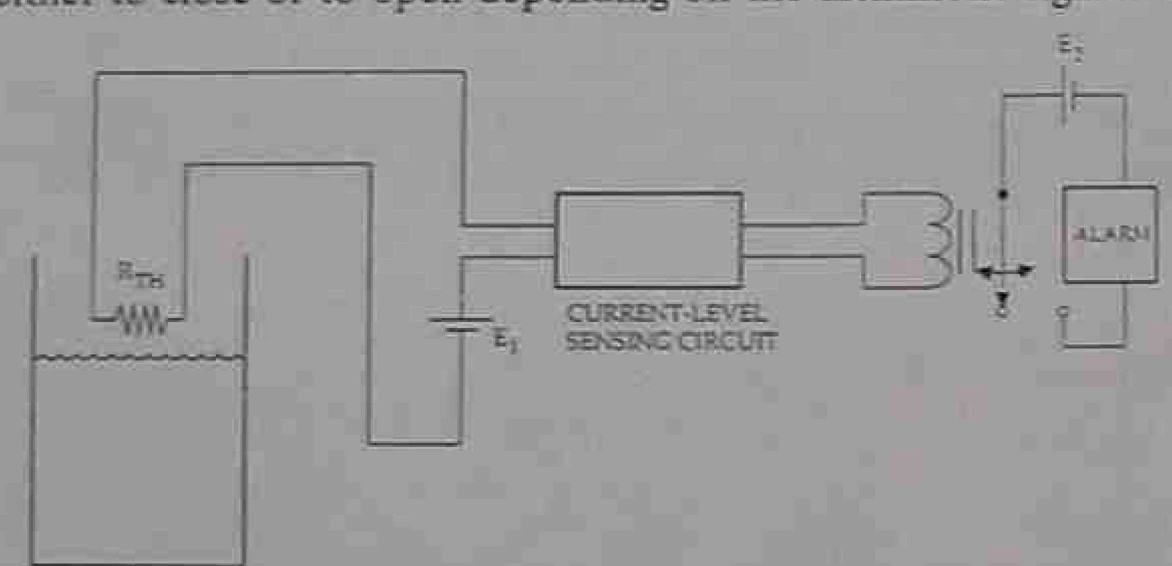


Figure 13-17 Thermistor being used as a level-indicating device.

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

Sydney Institute of Technology

School of Electrotechnology

Industrial Sensors

Theory Notes

Area ______________________

Topic ______________________

Session No.

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

Industrial Sensors

Topic -

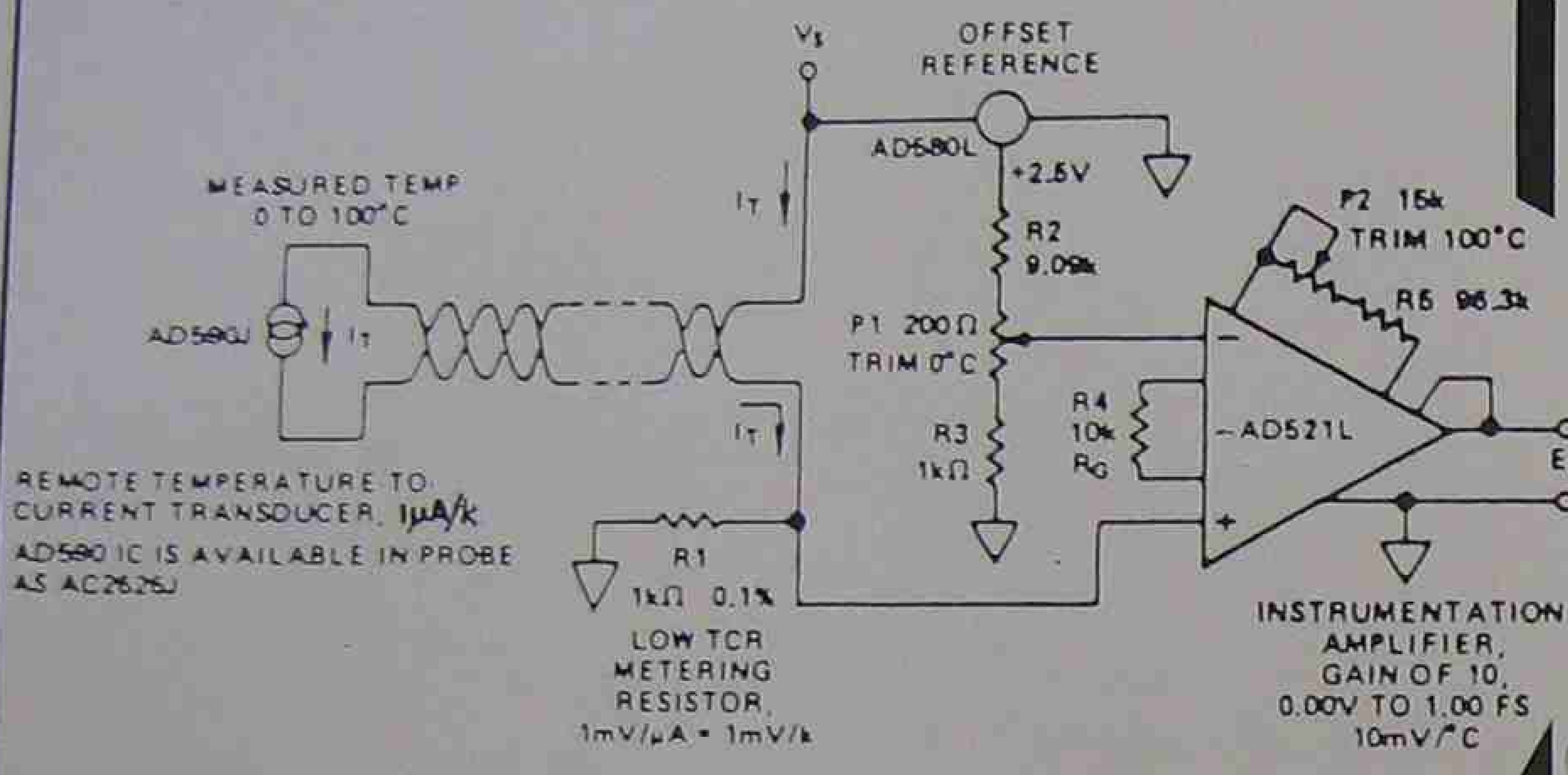


Figure 1. Thermometer Circuit

Orientation

Temperature Measurement Components

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

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

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

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

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

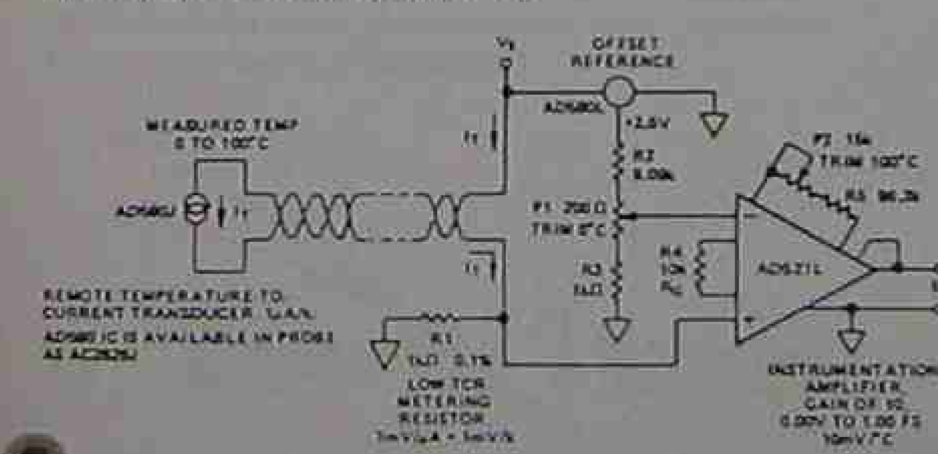


Figure 1. Thermometer Circuit

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

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

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

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

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

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

*Can be reduced by trimming the output offset.

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

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

8

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

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

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

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

AD521 bias-current tempco. The maximum bias-current change is 500pA/°C X 30° (range) = 15nA. The equivalent offset-voltage change is 15nA X 1kΩ = 15μV — 0.02°C

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

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

Total error (worst case) — 0.84°C

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

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

ADDITIONAL INFORMATION

COLD JUNTION COMPENSATION AMPLIFIER – AD594/AD595

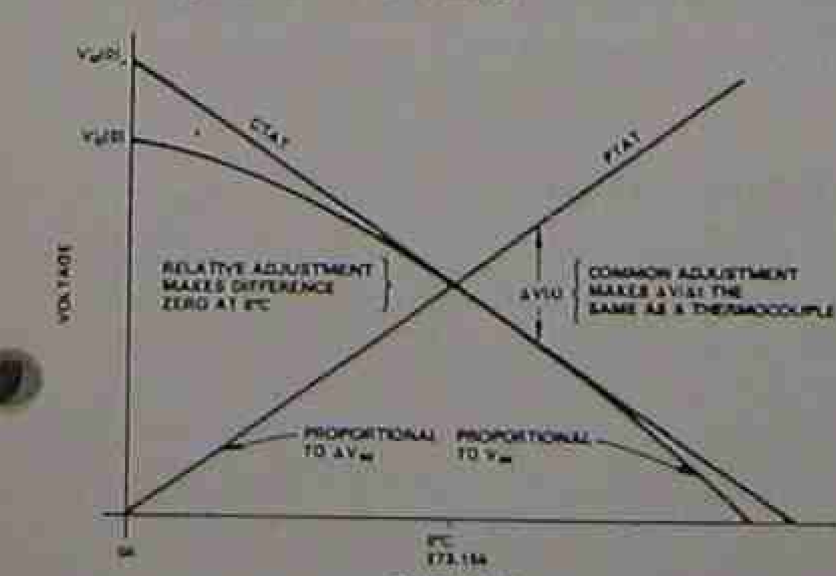
MONOLITHIC THERMOCOUPLE AMPLIFIER WITH COLD JUNCTION COMPENSATION

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

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

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

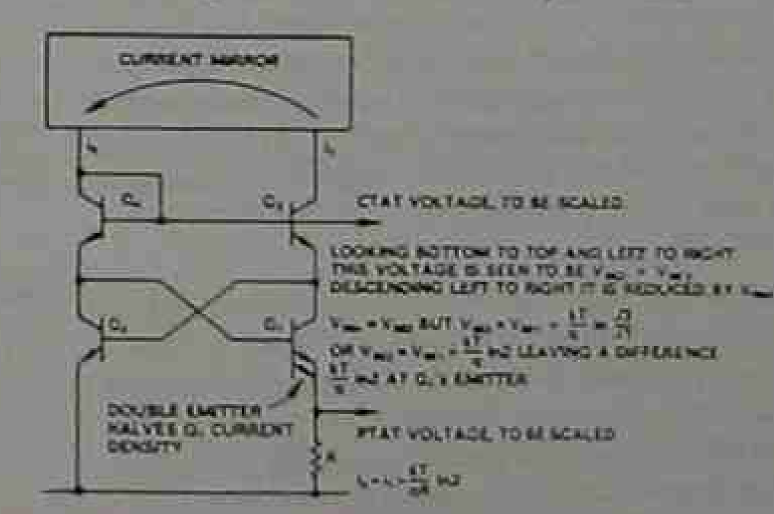
Although these two voltages are predictably related to absolute temperature, their difference can be related to Celsius temperature as shown here.



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

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

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



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

8

ANALOG DEVICES

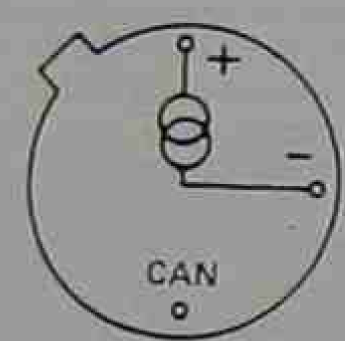
Two-Terminal IC Temperature Transducer

AD590*

FEATURES

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

AD590 FUNCTIONAL BLOCK DIAGRAM



TO-52
BOTTOM VIEW

8

PRODUCT DESCRIPTION

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

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

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

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

PRODUCT HIGHLIGHTS

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

*Covered by Patent No. 4,123,698

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

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

NOTES

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

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

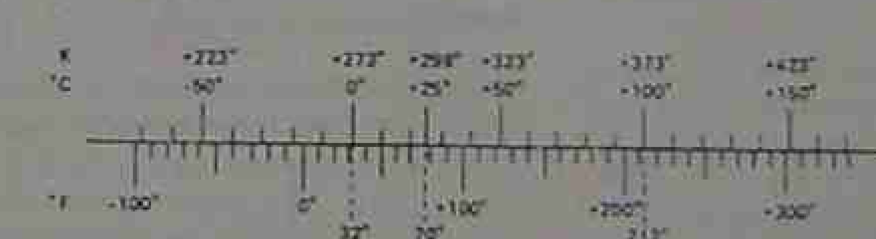
[3] Conditions: constant +5V, constant +125°C; guaranteed, not tested.

[4] Leakage current doubles every 10°C.

[5] See Section 19 for package outline information.

Specifications subject to change without notice.

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



TEMPERATURE SCALE CONVERSION EQUATIONS

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

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