SECTION 9

TRANSDUCER INTERFACING

- Force Transducers
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- Driving Bridges
- Bridge Applications
- Temperature Transducers:
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- Photodiode Transducers:
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- Hall Effect Magnetic Sensors

SECTION 9

TRANSDUCER INTERFACING

Walt Kester, Adolfo Garcia, James Bryant

Because of the many physical variables involved, real world signal processing requires a wide variety of sensing elements. Most are based on resistance, capacitance, current, or voltage output elements, thereby simplifying the interface to electronic measuring and processing equipment. Most of these transducers produce low level outputs which require conditioning before further processing. Conditioning with precision while maintaining low noise performance presents a significant design challenge.

This section discusses the most popular sensors, how they operate, their outputs, and appropriate conditioning circuitry. Space does not permit a discussion of all sensors, but the general principles presented should still be applicable.

REAL WORLD PHYSICAL VARIABLES

VARIABLE	SENSOR*	OUTPUT
Force (Pressure)	Strain Gauge Piezo Transducer	Resistance Voltage
Temperature	Thermocouple Silicon Bandgap RTD Thermistor	Voltage Voltage or Current Resistance Resistance
Light Intensity	Photodiode	Current
Acceleration	Accelerometer	Capacitance
Magnetic Field	Hall Effect	Voltage
Displacement	LVDT Potentiometer	AC Voltage Voltage

- ALL GENERATE LOW LEVEL SIGNALS REQUIRING PRECISION LOW-NOISE CONDITIONING
- * Not all these sensors are considered in this section

Figure 9.1

FORCE TRANSDUCERS

The most popular electrical elements used in force measurements include the resistance strain gauge, the semiconductor strain gauge, and piezoelectric transducers. The strain gauge measures force indirectly by measuring the deflection it produces in a calibrated carrier.

The resistance strain gauge is a resistive element which changes in length, hence resistance, as the force applied to the base on which it is mounted causes stretching or compression. It is perhaps the most well known transducer for converting force into an electrical variable.

Unbonded strain gauges consist of a wire stretched between two points as shown in Figure 9.2. Force acting on the wire (area = A, length = L, resistivity = ρ) will cause the wire to elongate or shorten, which will cause the resistance to increase or decrease proportionally according to:

$$R = \rho L/A$$

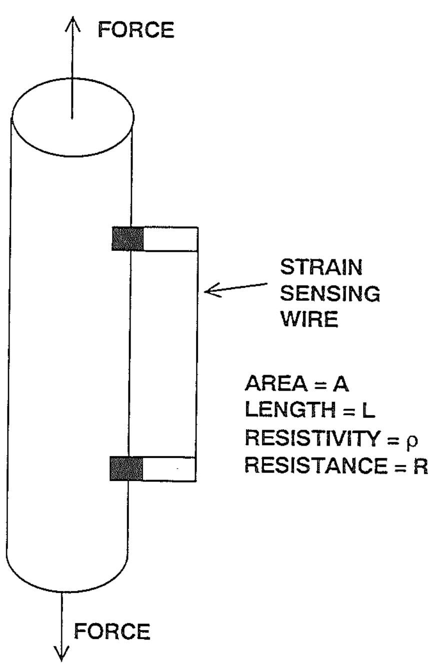
and

$$\Delta R/R = G \cdot \Delta L/L,$$

where G = Gauge factor (2.0 to 4.5 for metals, and more than 150 for semiconductors.

The dimensionless quantity $\Delta L/L$ is a measure of the force applied to the wire and is expressed in *microstrains* ($1\mu\varepsilon = 10^{-6}$ cm/cm).

UNBONDED WIRE STRAIN GAUGE



$$R = \frac{\rho L}{A}$$

$$\frac{\Delta R}{R} = \frac{G \cdot \Delta L}{L}$$

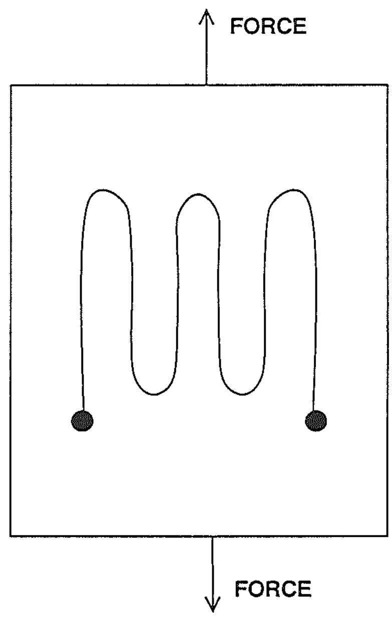
G = GAUGE FACTOR,
2.0 TO 4.5 FOR METALS
>150 FOR SEMICONDUCTORS

$$\frac{\Delta L}{L} = \text{MICROSTRAINS } (\mu\varepsilon) = 10^{-6} \text{ cm/cm}$$

Figure 9.2

Bonded strain gauges consist of a thin wire or conducting film arranged in a coplanar pattern and cemented to a base or carrier. The gauge is normally mounted so that as much as possible of the length of the conductor is aligned in the direction of the stress that is being measured. Lead wires are attached to the base and brought out for interconnection. Bonded devices are considerably more practical and are in much wider use than unbonded devices.

BONDED WIRE STRAIN GAUGE

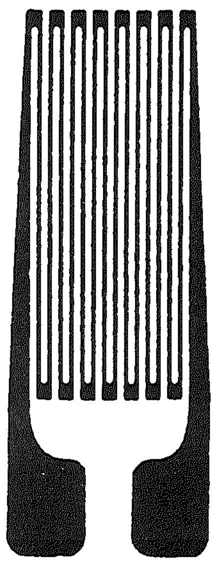


- SMALL SURFACE AREA
- LOW LEAKAGE
- HIGH ISOLATION

Figure 9.3

Perhaps the most popular version is the foil-type gauge, produced by photo-etching techniques, and using similar metals to the wire types (alloys of copper-nickel (Constantan), nickel-chromium (Nichrome), nickel-iron, platinum-tungsten, etc. (see Figure 9.4). Gauges having wire sensing elements present a small surface area to the specimen; this reduces leakage currents at high temperatures and permits higher isolation potentials between the sensing element and the specimen. Foil sensing elements, on the other hand, have a large ratio of surface area to cross-sectional area and are more stable under extremes of temperature and prolonged loading. The large surface area and thin cross section also permit the device to follow the specimen temperature and facilitate the dissipation of self-induced heat.

METAL FOIL STRAIN GAUGE

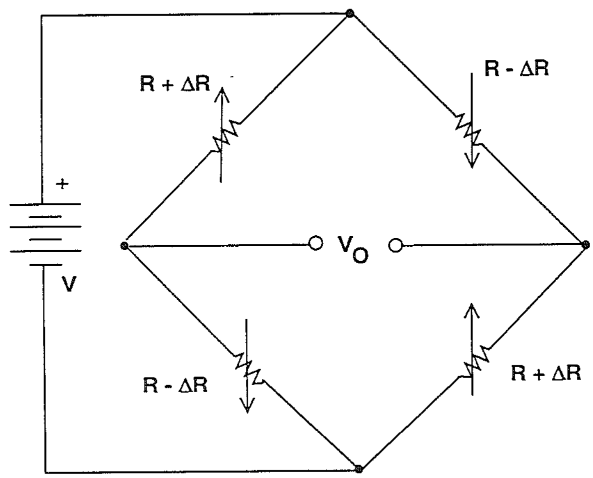


- PHOTO ETCHING TECHNIQUE
- LARGE AREA
- STABLE OVER TEMPERATURE
- THIN CROSS SECTION
- GOOD HEAT DISSIPATION

Figure 9.4

Strain gauges are low-impedance devices; they require significant excitation power to obtain reasonable levels of output voltage. A typical strain-gauge based load cell bridge will have (typically) a 350Ω impedance and is specified as having a sensitivity in terms of millivolts full scale per volt of excitation. The load cell is composed of four individual strain gauges arranged as a bridge as shown in Figure 9.5. For a 10V bridge excitation voltage with a rating of 3mV/V, 30 millivolts of signal will be available at full scale loading. The output can be increased by increasing the drive to the bridge, but self-heating effects are a significant limitation to this approach: they can cause erroneous readings or even device destruction. Further details regarding the accurate measurement of strain gauge resistance are covered in the section on bridges.

LOAD CELL COMPOSED OF FOUR STRAIN GAUGES



- REQUIRES EXCITATION: 2.5 TO 15V
- 350Ω TO 1000Ω IMPEDANCE
- LINEAR
- ΔR/R < 1%
- FOR V = 10V, FS OUTPUT 30 TO 100mV
- HIGH GAIN, PRECISION AMP REQUIRED

Figure 9.5

Semiconductor strain gauges make use of the resistance change in semiconductor materials in order to obtain greater sensitivity and higher-level output. Such bridges may have 30 times the sensitivity of bridges employing metal films, but are temperature sensitive and difficult to compensate. They are not in as widespread use as the more stable metal-film devices for precision work; however, where sensitivity is important and temperature variations are small, they may have some advantage. Instrumentation is similar to that for metal-film bridges but is less critical because of the higher signal levels and decreased transducer accuracy. A comparison between metal and semiconductor strain gauges is given in Figure 9.6 (Reference 1, page 51).

COMPARISON BETWEEN METAL AND SEMICONDUCTOR STRAIN GAUGES

PARAMETER	METAL STRAIN GAUGE	SEMICONDUCTOR STRAIN GAUGE
Measurement Range	0.1 to 40,000 $\mu\varepsilon$	0.001 to 3000$\mu\varepsilon$
Gauge Factor	2.0 to 4.5	50 to 200
Resistance, Ω	120, 350, 600, ..., 5000	1000 to 5000
Resistance Tolerance	0.1% to 0.2%	1% to 2%
Size, mm	0.4 to 150 Standard: 3 to 6	1 to 5

Figure 9.6

Piezoelectric force transducers are employed where the forces to be measured are dynamic (i.e., continually changing over the period of interest - usually of the order of milliseconds). These devices utilize the effect that changes in charge are produced in certain materials when they are subjected to physical stress. In fact, piezoelectric transducers are *displacement* transducers with quite large charge outputs for very small displacements, but they are invariably used as force transducers on the assumption that in an elastic material, displacement is proportional to force. Piezoelectric devices produce substantial output voltage in instruments such as accelerometers for vibration studies. Output impedance is high, and charge amplifier configurations, with low input capacitance, are required for signal conditioning.

The output of a piezoelectric transducer can be modeled as a voltage source in series with a small capacitor as shown in Figure 9.7. Step inputs of physical force result in an effective capacitance change. The op-amp summing voltage is held at zero, and the change in charge is transferred to the feedback capacitor, developing an output voltage at low impedance. The output of this circuit is inversely proportional to the value of the feedback capacitance. The op-amp must have extremely low bias current which is supplied by the very large feedback resistor.

CHARGE AMPLIFIER USED WITH PIEZOELECTRIC TRANSDUCER

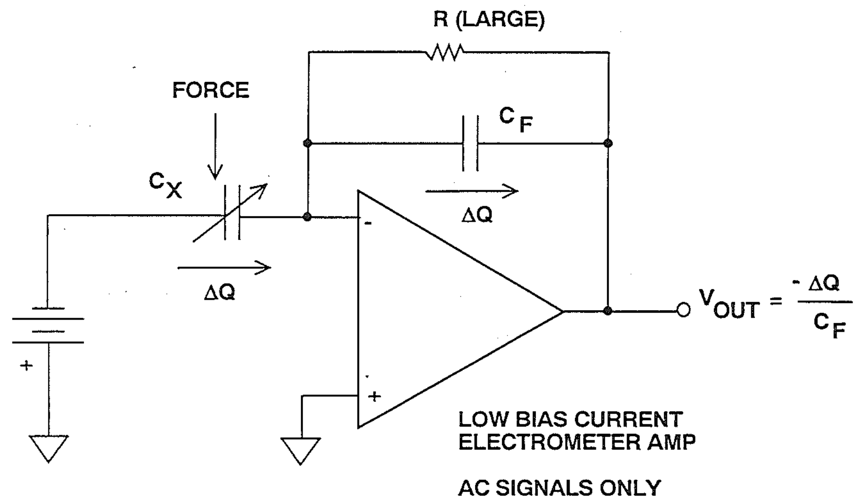


Figure 9.7

Pressures in liquids and gases are measured electrically by a variety of pressure transducers. A variety of mechanical converters (including diaphragms, capsules, bellows, manometer tubes, and Bourdon tubes) are used to measure pressure by measuring an associated length, distance, displacement, and to measure pressure changes by the motion produced. The output of this mechanical interface is then applied to an electrical converter such as a strain gauge or piezoelectric transducer. Unlike strain gauges, piezoelectric pressure transducers are typically used for high-frequency pressure measurements (such as sonar applications or crystal microphones).

PRESSURE TRANSDUCERS

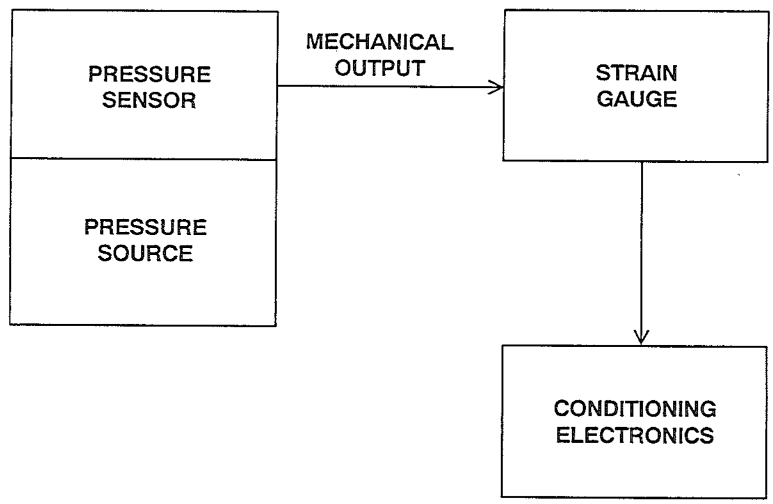


Figure 9.8

There are many ways of defining flow (mass flow, volume flow, laminar flow, turbulent flow). Usually the *amount* of a substance flowing (mass flow) is the most important, and if the fluid's density is constant, a volume flow measurement is a useful substitute that is generally easier to perform. One commonly used class of transducers, which measure flow rate indirectly, involves the measurement of pressure. Flow can be derived by taking the differential pressure across two points in a flowing medium - one at a static point and one in the flow stream. *Pitot tubes* are one form of device used to perform this function. The flow rate is obtained by measuring the differential pressure with standard pressure transducers as shown in Figure 9.9.

The cantilevered vane shown in Figure 9.10 is a simple way to measure flow and amenable to strain gauge instrumentation.

We have seen that the resistive strain gauge is a fundamental sensor used in a wide variety of force, pressure, and flow measurements. It is therefore important to understand bridge circuits, since they are by far the most common means of converting resistance changes into voltages.

PITOT TUBE USED TO MEASURE FLOW RATE

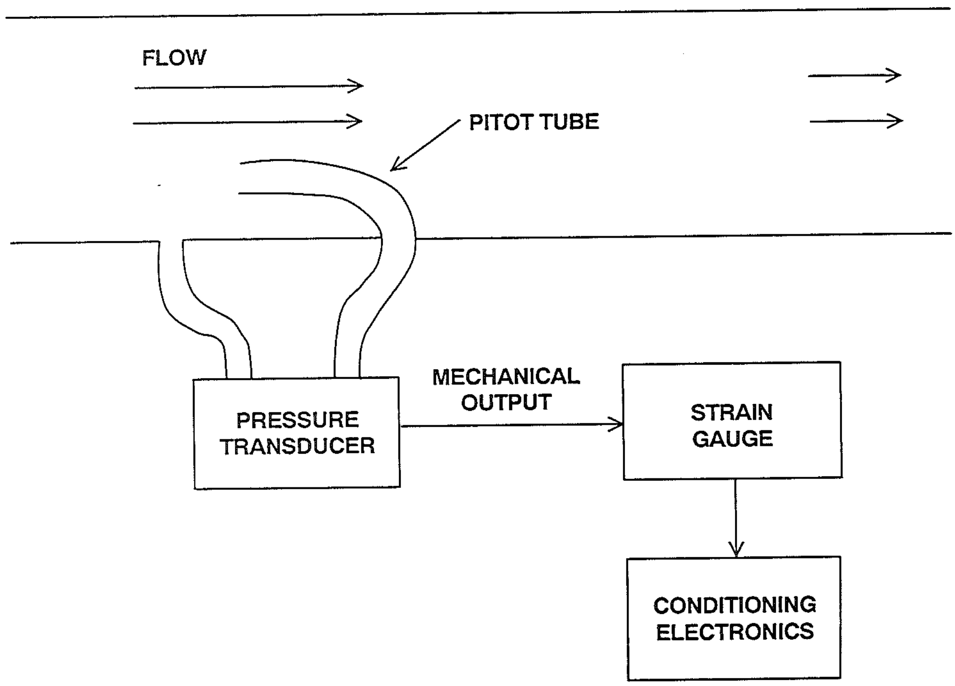


Figure 9.9

BENDING VANE (STRAIN GAUGE) USED TO MEASURE FLOW RATES

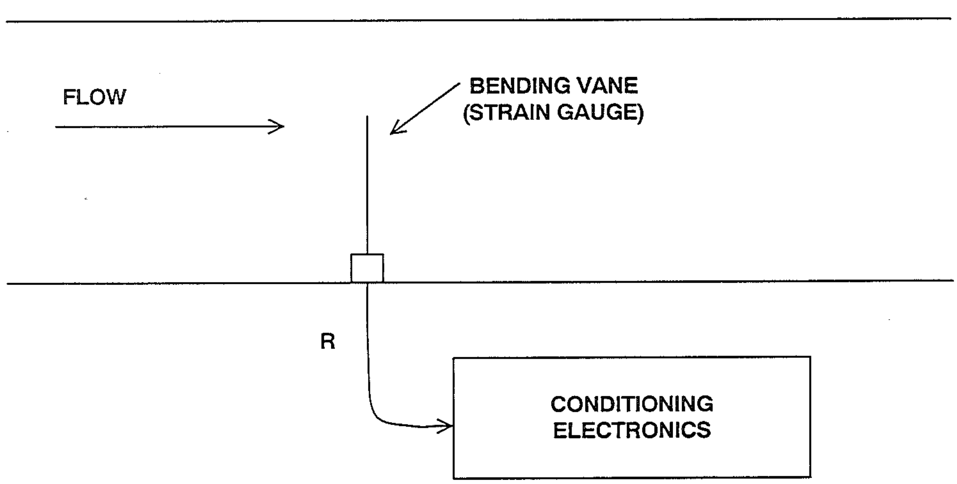


Figure 9.10

INTRODUCTION TO BRIDGES

Resistive elements are the most common transducers. They are inexpensive to manufacture and easy to interface with amplifier circuits. Furthermore, it is possible to integrate sensing elements directly on ICs. Resistive elements can be made sensitive to temperature, strain (by pressure or by flex), and light. Using these basic elements, many complex physical phenomena can be measured; such as fluid or mass flow (by sensing the temperature difference between two calibrated resistances) and dew-point humidity (by measuring two different temperature points), etc.

Transducer elements' resistances can range from less than 100Ω to several hundred kΩ, depending on the transducer design and the physical environment to be measured. For example, RTDs (Resistance Temperature Devices) are typically 100Ω. Thermistors are typically 3500Ω or higher. There are also high impedance piezoelectric or capacitance-modulating transducers whose impedances range from 10's to 100's of MΩ.

Most of these transducers produce very low level signals, and therefore high gain is necessary in order to get usable signal levels. Amplifying these signals with precision while maintaining low noise performance presents a significant design challenge.

RESISTANCE OF POPULAR TRANSDUCERS

■ RTD (Resistance Temperature Device)	100Ω
■ Pressure transducers	350 - 3500Ω
■ Strain Gauge	120, 350, 3500Ω
■ Weigh-Scale load cells	350 - 3500Ω
■ Thermistor	100 Ω - 10MΩ
■ Relative humidity	100kΩ - 10MΩ

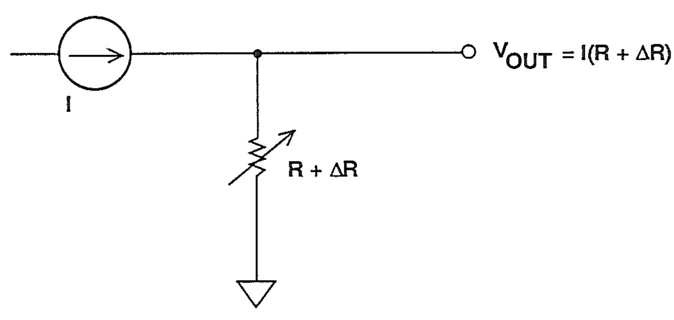
Figure 9.11

Resistive sensors such as RTDs and strain gauges produce small percentage changes in resistance in response to a change in a physical variable such as temperature or force. Platinum RTDs have a temperature coefficient of about 0.385%/°C. This corresponds to a 38.5% resistance change over a 0°C to 100°C.

Strain gauges present a significant measurement challenge because the typical change in resistance over the entire operating range of a strain gauge may only be 0.1%. Accurately measuring small resistance changes is therefore critical to applying resistive sensors.

One technique for measuring resistance (shown in Figure 9.12) is to force a constant current through the resistive sensor and accurately measure the voltage output. This requires an extremely stable current source because any change in the current will be misinterpreted as a resistance change. The power dissipation in the resistive sensor must be small so that self-heating does not produce errors, therefore the drive current must be small.

MEASURING RESISTANCE INDIRECTLY USING A CONSTANT CURRENT SOURCE

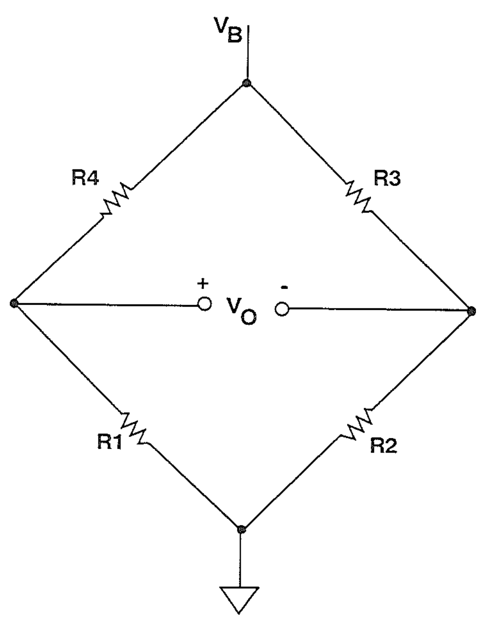


FULLSCALE $\frac{\Delta R}{R}$ MAY BE AS LOW AS 0.1%

Figure 9.12

Bridges offer an attractive alternative for measuring small resistance changes accurately. The basic Wheatstone bridge (actually developed by S. H. Christie in 1833) is shown in Figure 9.13. It consists of four resistors connected to form a quadrilateral, a source of excitation (voltage or current) connected across one of the diagonals, and a voltage detector connected across the other diagonal. The detector measures the difference between the outputs of two voltage dividers connected across the excitation.

THE WHEATSTONE BRIDGE



$$V_O = \frac{R1}{R1 + R4} V_B - \frac{R2}{R2 + R3} V_B$$

$$= \frac{\frac{R1}{R4} - \frac{R2}{R3}}{\left(1 + \frac{R1}{R4}\right)\left(1 + \frac{R2}{R3}\right)} V_B$$

AT BALANCE,

$$V_O = 0 \quad \text{IF} \quad \frac{R1}{R4} = \frac{R2}{R3}$$

Figure 9.13

A bridge measures resistance indirectly by comparison with a similar resistance. The two principal ways of operating a bridge are as a null detector or as a device that reads a difference directly as voltage or current.

When R1/R4 = R2/R3, the resistance bridge is at a *null*, irrespective of the mode of excitation (current or voltage, AC or DC), the magnitude of excitation, the mode of readout (current or voltage), or the impedance of the detector.

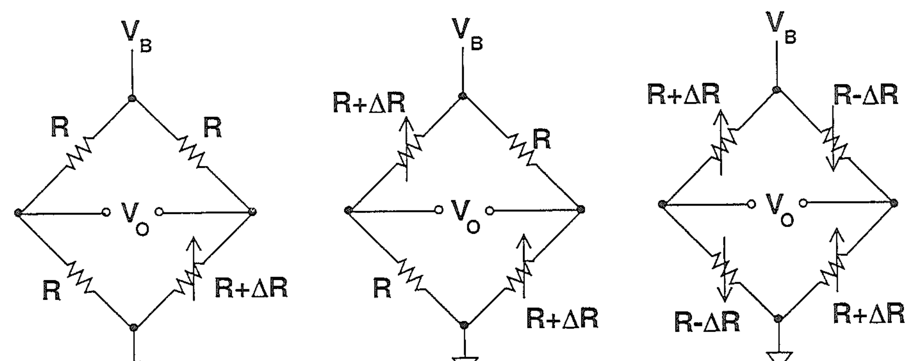
Therefore if the ratio of R2/R3 is fixed at K, a null is achieved when R1 = K·R4. If R1 is unknown and R4 is an accurately determined variable resistance, the magnitude of R1 can be found by adjusting R4 until null is achieved. Conversely, in transducer-type measurements, R4 may be a fixed reference, and a null occurs when the magnitude of the external variable (strain, temperature, etc.) is such that R1 = K·R4.

Null measurements are principally used in feedback systems involving electro-mechanical and/or human elements. Such systems seek to force the active element (strain gauge, RTD, thermistor, etc.) to balance the bridge by influencing the parameter being measured.

For the majority of transducer applications employing bridges, the deviation of one or more resistors in a bridge from an initial value is measured as an indication of the magnitude (or a change) in the measured variable.

Figure 9.14 shows the three commonly used bridges suitable for sensor applications and the corresponding equations which relate the bridge output voltage to the excitation voltage and the bridge resistance values.

BASIC BRIDGE CONFIGURATIONS



$$V_O = V_B\left[\frac{\Delta R}{2(2R+\Delta R)}\right]$$

Single-Element Varying

$$V_O = V_B\left[\frac{\Delta R}{2R+\Delta R}\right]$$

Two-Element Varying

$$V_O = V_B\left[\frac{\Delta R}{R}\right]$$

All-Element Varying

Figure 9.14

The *single-element varying* bridge is most suited for temperature sensing using RTDs or thermistors. All the resistances are nominally equal, but one of them (the sensor) is variable by an amount ΔR. As the equation indicates, the relationship between the bridge output and ΔR is not linear. For example, if R = 100Ω and ΔR = ±0.1Ω (±0.1% change in resistance), the output of the bridge is ±2.49875mV for V_B = 10V. This corresponds to an end-point linearity error of ±0.0125%. (Bridge end-point linearity error is calculated as the worst error in % FS from a straight line which connects the origin and the end points at ±FS, i.e. the FS gain error is not included). If ΔR = ±1Ω, (±1% change in resistance), the output of the bridge is ±24.8756mV, representing an end-point linearity error of approximately ±0.125%. The end-point linearity error of the single-element bridge can be expressed in equation form:

Single-Element Varying
Bridge End-Point Linearity Error ≈ % Change in Resistance ÷ 8

In some applications, this error may be acceptable, but there are methods available to linearize bridges which will be discussed shortly.

The *sensitivity* of a bridge is the ratio of the maximum expected change in the output voltage to the excitation voltage. For the examples given above, the sensitivities are 500µV/V (for ±0.1% resistance change) and 5mV/V (for ±1% resistance change).

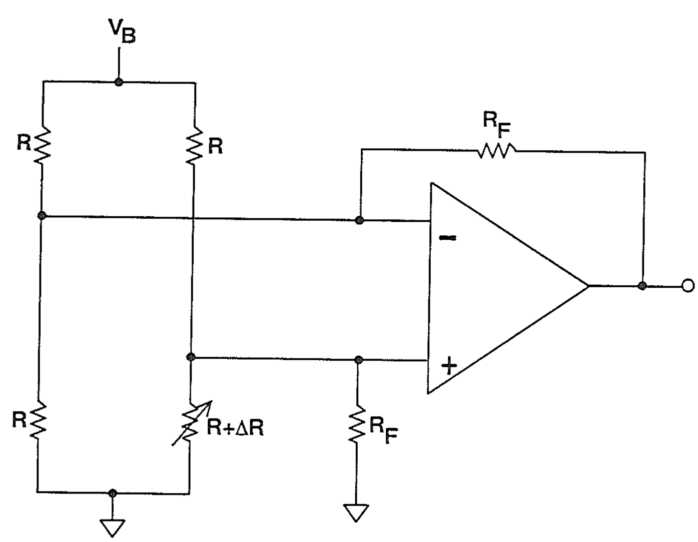
The *two-element varying* bridge produces twice the signal output but requires two identical sensors. The nonlinearity is the same as that of the single-element varying bridge. The two-element varying bridge is commonly found in pressure transducers and flow meter systems.

The *all-element varying* bridge produces the most signal for a given resistance change and is inherently linear. It is an industry-standard configuration for load cells which are constructed from four identical strain gauges.

The output of a single-element varying bridge may be amplified by an op-amp connected in the inverting mode as shown in Figure 9.15. This circuit has poor gain accuracy and also unbalances the bridge due to loading from RF and the op amp bias current. A much better approach is to use an instrumentation amplifier as shown in Figure 9.16. This efficient circuit provides better gain accuracy and does not unbalance the bridge.

Bridges can be driven with constant current sources rather than voltage sources. This eliminates errors due to DC voltage drops in the wiring to the bridge. The circuit shown in Figure 9.17 is an example and has the advantage of reduced nonlinearity (0.062% rather that 0.125% for a 1% FS resistance change).

SINGLE-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 1

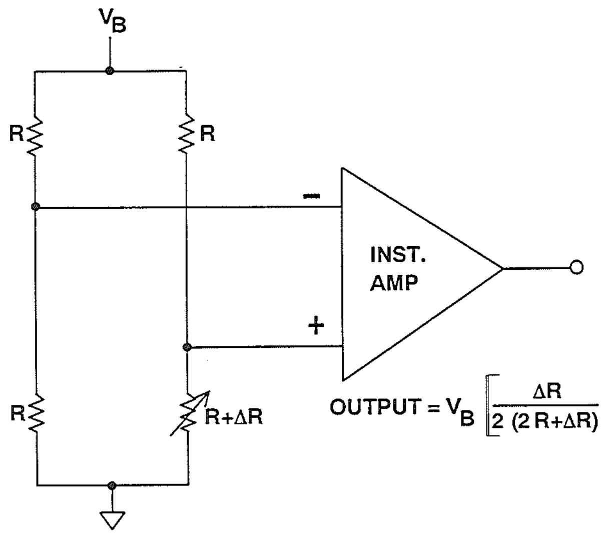


Advantages:
- No signal in = zero volts out
- Single supply operation
- Single op amp stage

Disadvantages:
- Nonlinear operation:
- 0.125% nonlinearity for 1% fullscale ΔR change
- Poor gain accuracy
- Unbalanced output R due to varying element

Figure 9.15

SINGLE-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 2



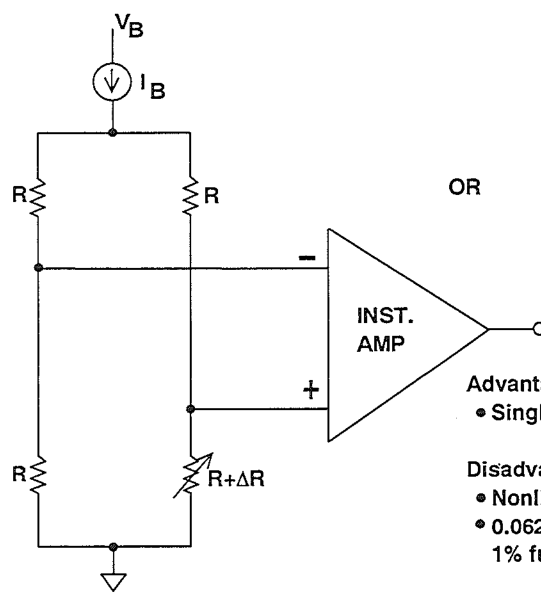
$$\text{OUTPUT} = V_B \left[\frac{\Delta R}{2\,(2R+\Delta R)}\right]\left[\text{GAIN}\right]$$

Advantages:
- Efficient design
- Only bridge R matching required
- Better gain accuracy

Disadvantages:
- Nonlinear operation
- 0.125% nonlinearity for 1% fullscale ΔR change

Figure 9.16

DRIVING A SINGLE-ELEMENT VARYING BRIDGE WITH A CONSTANT CURRENT SOURCE



OR

Advantages:
- Single supply operation

Disadvantages:
- Nonlinear operation
- 0.062% nonlinearity for 1% fullscale ΔR change

$$\text{OUTPUT} = I_B\left(\frac{R\Delta R}{4R+\Delta R}\right)(\text{GAIN})$$

Figure 9.17

Various techniques are available to linearize bridges, but it is important to distinguish between the linearity of the bridge equation and the linearity of the transducer response to the phenomenon being sensed. For example, if the active element is an RTD, the bridge used to implement the measurement might have adequate linearity; yet the output could still be nonlinear due to the RTD's nonlinearity. Manufacturers of transducers employing bridges address the nonlinearity issue in a variety of ways, including keeping the resistive swings in the bridge small, shaping complimentary nonlinear response into the active elements of the bridge, using resistive trims for first-order corrections, and others.

Figure 9.18 shows a single-element varying active bridge in which an op amp produces a null by adding a voltage in series with the variable arm. That voltage is equal in magnitude and opposite in polarity to the incremental voltage across the varying element and is linear with ΔR. Since it is an op-amp output, it can be used as a low impedance output point for the bridge measurement. This active bridge has a gain of two over the standard single-element varying bridge, and the output is linear, even for large values of ΔR. Because of the small output signal this bridge must usually be followed by an amplifier.

Two other circuits for linearizing a single-element varying bridge are shown in Figures 9.19 and 9.20. In Figure 9.19, the bottom of the bridge is driven by an op-amp, which maintains a constant current in the varying resistance. The output signal is taken from the right-hand leg of the bridge and amplified by a non-inverting op-amp. The op-amp in Figure 9.20 performs the same function, but the bridge output is amplified by a second op-amp configured in the inverting mode.

LINEARIZING A SINGLE-ELEMENT VARYING BRIDGE, METHOD 1

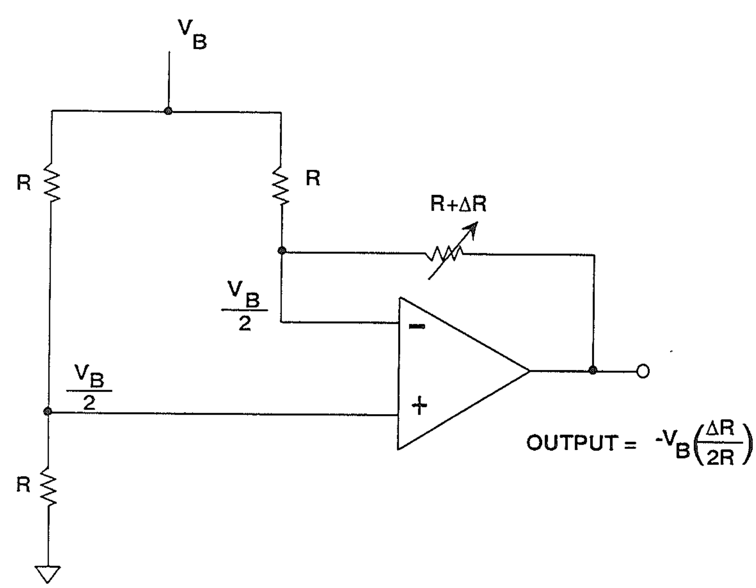
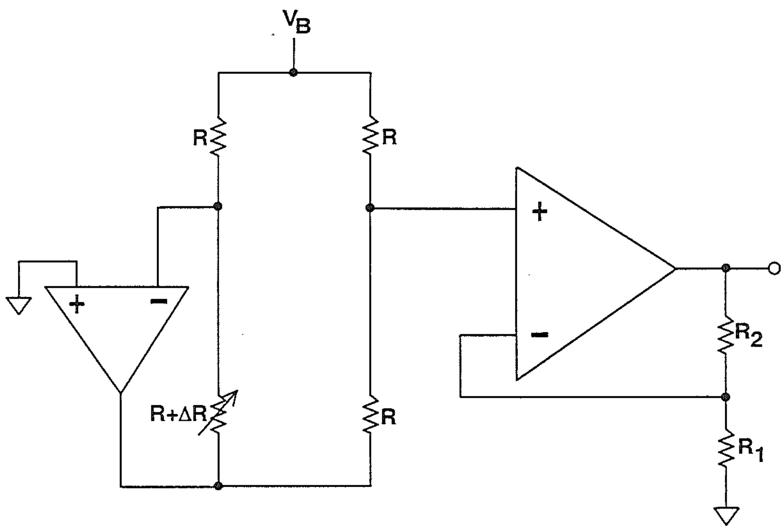


Figure 9.18

LINEARIZING A SINGLE-ELEMENT VARYING BRIDGE, METHOD 2: NON-INVERTING OUTPUT AMPLIFIER



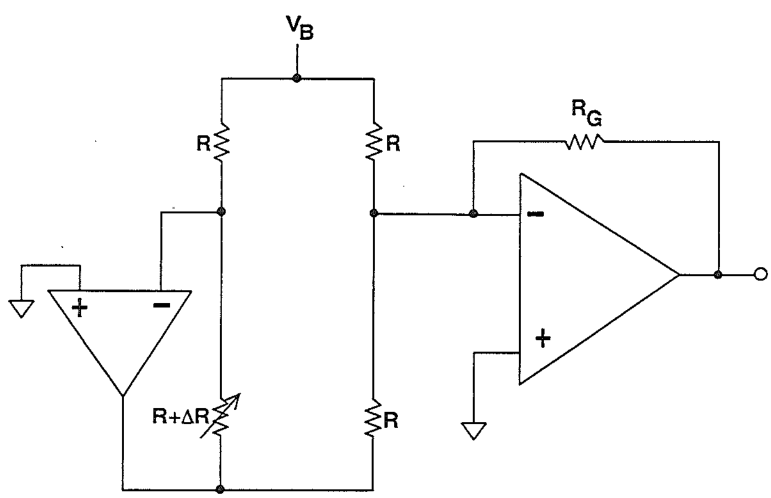
Advantages:
- Linear operation
- Uses low cost op amp
- No signal in = zero volts out
- Zero volts common-mode operation
- 2X signal resolution

Disadvantages:
- Requires 2 op amps
- Requires 2 supplies
- Requires R matching for bridge and gain

$$\text{OUTPUT} = \frac{V_B}{2}\left(\frac{\Delta R}{R}\right)\left(1 + \frac{R_2}{R_1}\right)$$

Figure 9.19

LINEARIZING A SINGLE-ELEMENT VARYING BRIDGE, METHOD 2: INVERTING OUTPUT AMPLIFIER



Advantages:
- Linear operation
- Single resistor sets gain
- Constant zero volts common-mode
- 2X signal resolution

Disadvantages:
- Requires 2 op amps
- Requires 2 supplies

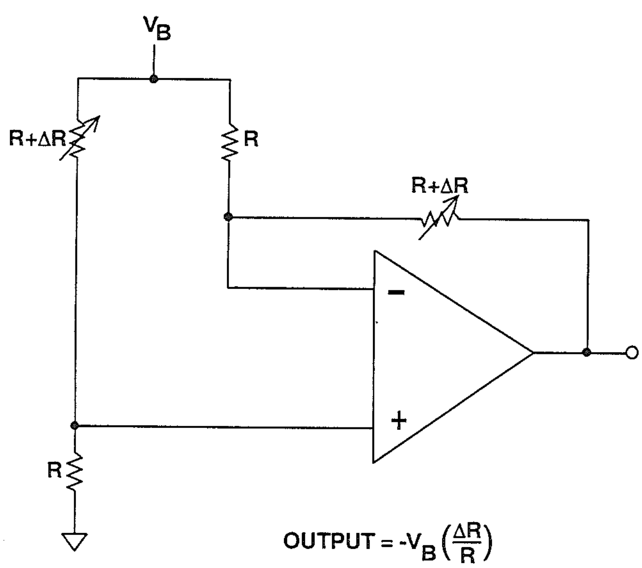
$$\text{OUTPUT} = V_B\left(\frac{\Delta R}{R^2}\right) R_G$$

Figure 9.20

A circuit for linearizing a two-element varying bridge is shown in Figure 9.21. This circuit is similar to Figure 9.18 and has twice the gain. Additional gain may be necessary.

The two-element varying bridge circuit in Figure 9.22 uses an op-amp to maintain a constant current through the bridge (V_R/R_{SENSE}). The current through each leg of the bridge remains constant ($I_B/2$) as the resistances change, therefore the output is a linear function of ΔR. An instrumentation amplifier provides the additional gain.

LINEARIZING A TWO-ELEMENT VARYING BRIDGE, METHOD 1

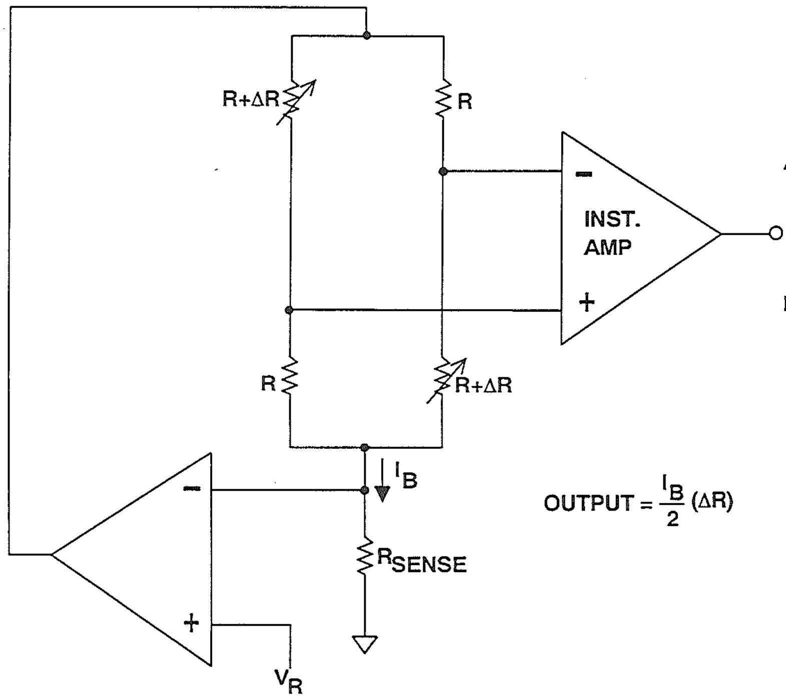


Advantage:
- Linear operation

Disadvantages:
- Low signal output implies larger errors
- Requires second amp for gain

Figure 9.21

LINEARIZING A TWO-ELEMENT VARYING BRIDGE, METHOD 2



Advantages:
- Linear operation
- No signal in = OV out
- Single supply operation

Disadvantages:
- Requires 2 amps

Figure 9.22

DRIVING BRIDGES

Wiring resistance and noise pickup are the biggest problems associated with remotely located bridges. Figure 9.23 shows how Kelvin sensing can be used to eliminate the effects of voltage dropped across the wiring resistance. The current in the sense lines is minimal, therefore the op-amps maintain a constant bridge voltage independent of the wiring resistance in the drive and return paths.

Another method often used to eliminate wiring resistance errors is to drive the bridge with a constant current source as shown in Figure 9.24.

9

DRIVING BRIDGES USING KELVIN SENSING

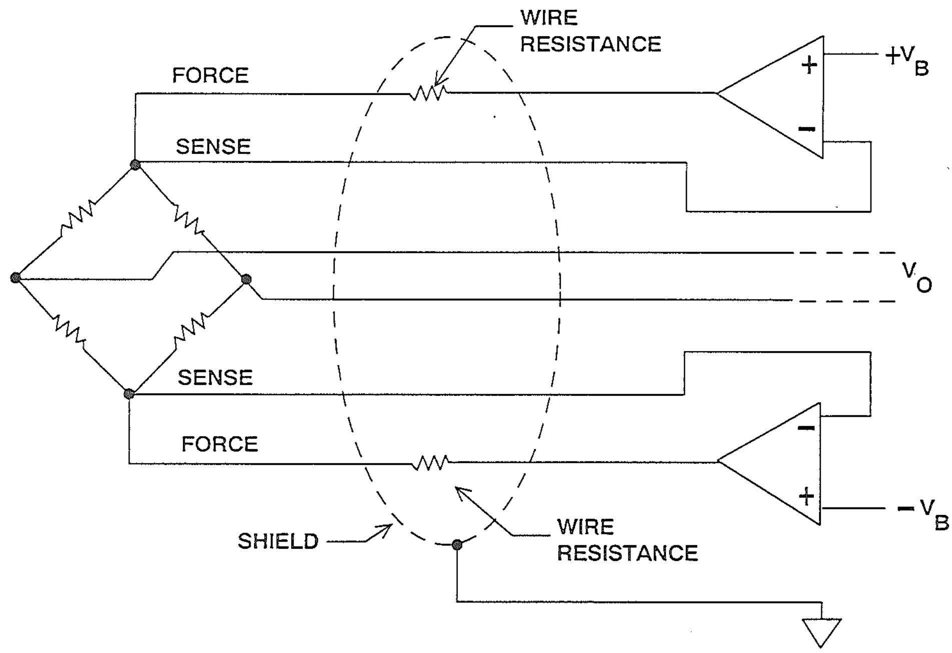


Figure 9.23

DRIVING BRIDGES USING A CONSTANT CURRENT SOURCE

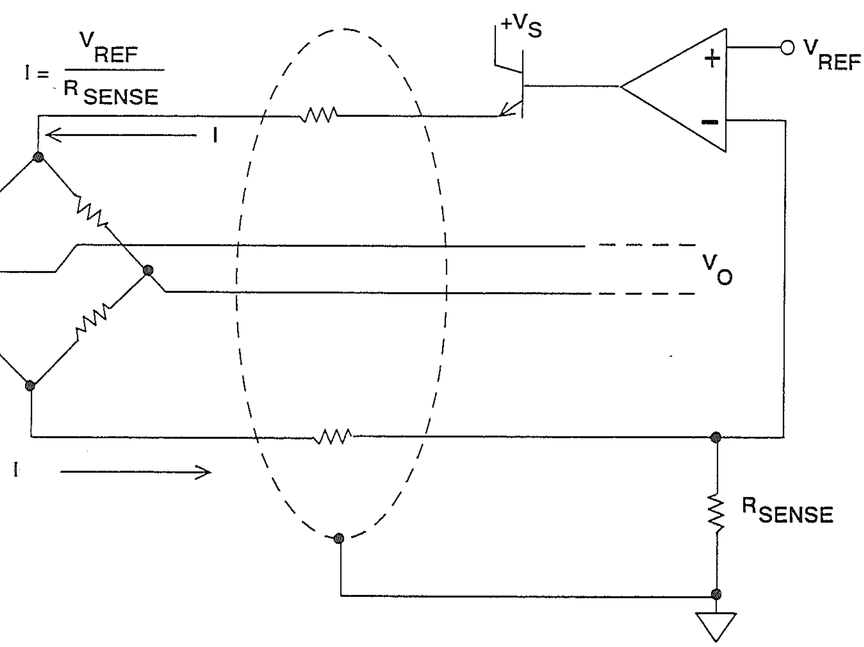


Figure 9.24

Remotely located transducers are a potential source of EMI/RFI pickup. It is recommended that shielded cable be used between the sensor and the instrumentation electronics. The optimum point to ground the shield is at the electronics, thereby shunting stray RF currents into a low impedance ground plane. Under no circumstances should the shield be grounded at both ends; this will cause ground-loop currents to flow in the shield and induce unwanted voltages on the conductors.

GROUNDING THE SHIELD AT THE RECEIVING END MINIMIZES RF PICKUP

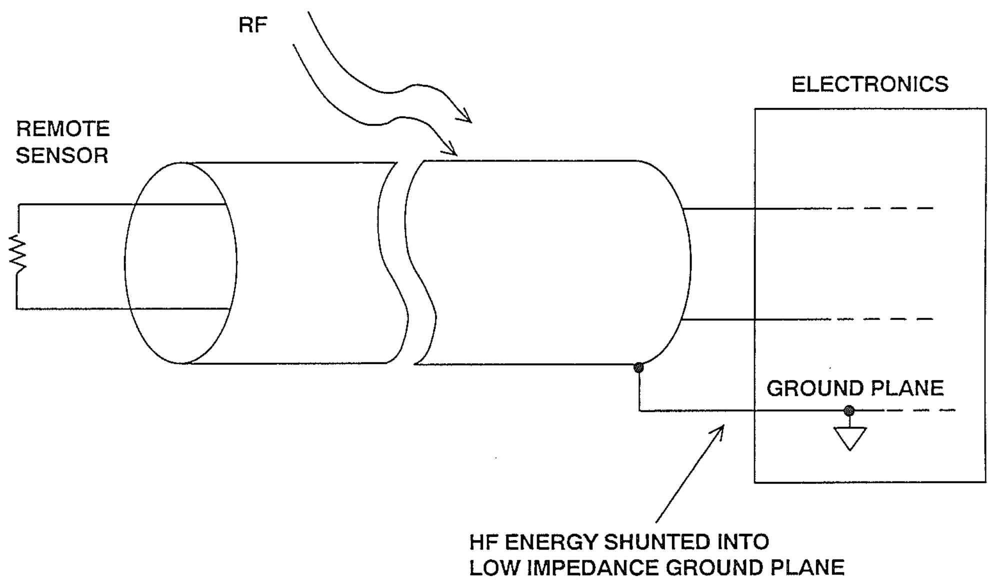


Figure 9.25

9

BRIDGE APPLICATIONS

An example of an all-element varying bridge circuit is a fatigue monitoring strain sensing circuit as shown in Figure 9.26. The full bridge is an integrated unit that can be attached to the surface on which the strain or flex is to be measured. In order to facilitate remote sensing, current excitation is used. The OP-177 servos the bridge current to 10mA around a reference voltage of 1.235V. The strain gauge produces an output of 10.25mV/1000με. The signal is amplified by the AD620 instrumentation amplifier. Full-scale strain voltage may be set by adjusting the 100Ω gain potentiometer such that, for a strain of –3500με, the output reads –3.500V; and for a strain of +5000με, the output registers a +5.000V. The measurement may then be displayed with a digital voltmeter.

FATIGUE LOAD STRAIN SENSOR AMPLIFIER

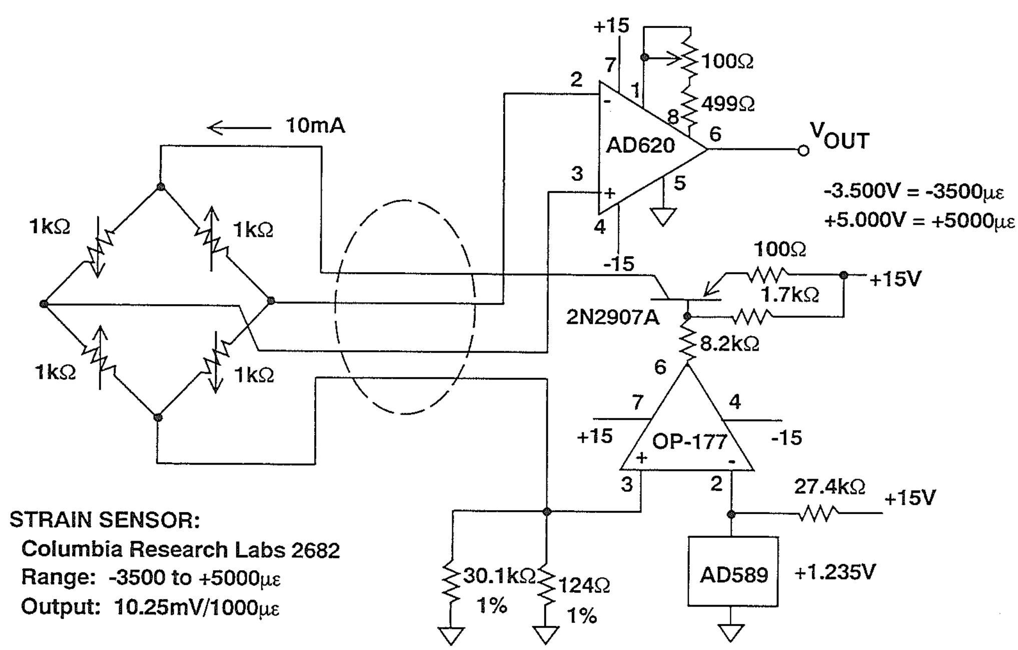


Figure 9.26

Another example is a weigh-scale amplifier circuit shown in Figure 9.27. A typical load cell has a bridge resistance of 350Ω. A 10.000V bridge excitation produces linear bridge behavior. To ensure this linearity is preserved, an instrumentation amplifier is used. This design has a minimum number of critical resistors and amplifiers, making the entire implementation accurate, stable, and cost effective. The only requirement is that the 475Ω resistor and the 100Ω potentiometer have low temperature coefficients so that the amplifier gain does not drift over temperature.

WEIGH-SCALE LOAD CELL AMPLIFIER

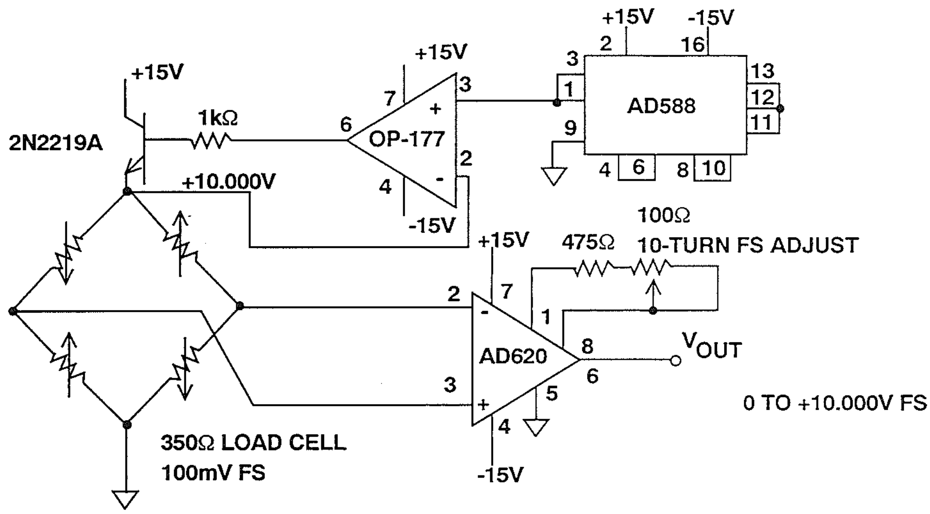


Figure 9.27

A precision weigh-scale transducer is usually configured as a 350Ω bridge. Figure 9.28 shows a load-cell amplifier that is powered from a single supply. The excitation voltage to the bridge must be precise and stable, otherwise it introduces an error in the measurement. In this circuit, a precision 5V reference is used as the bridge drive. The REF-195 reference can supply more than 30mA to a load, so it can drive the 350Ω bridge without the need of a buffer.

PRECISION SINGLE-SUPPLY LOAD-CELL AMPLIFIER

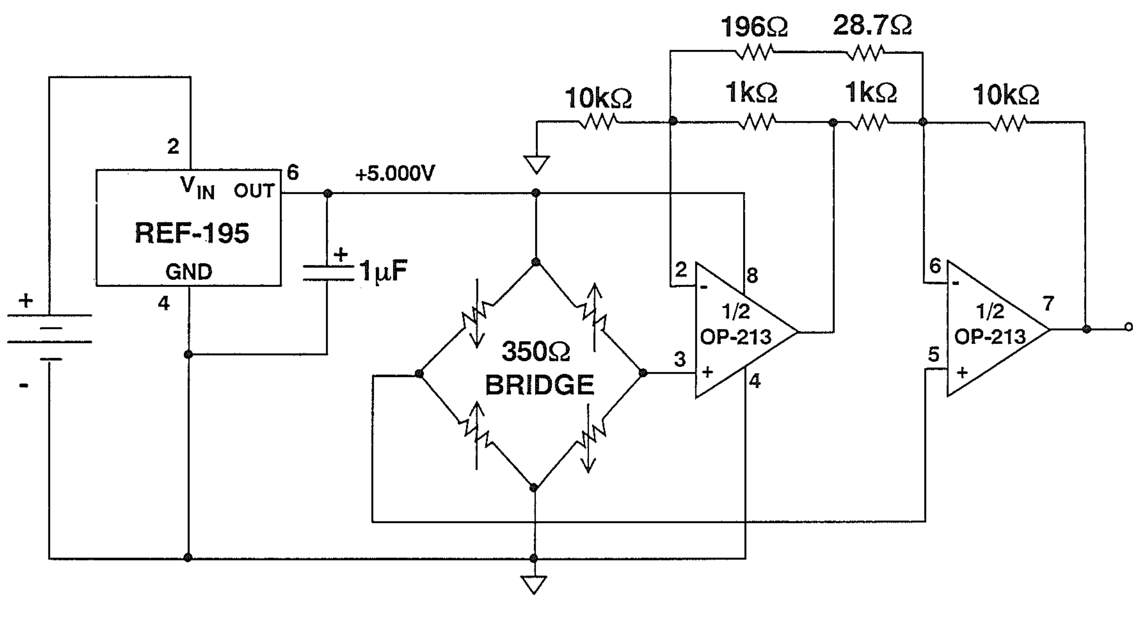


Figure 9.28

The bridge signal is amplified by the two OP-213 op amps connected as an instrumentation amplifier. The resistor network sets the gain according to the formula:

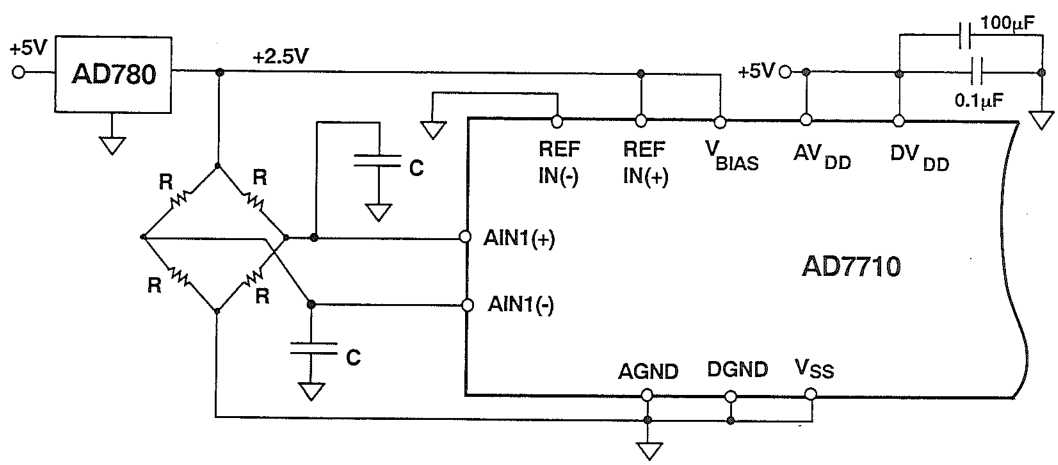
$$G = 1 + \frac{10k\Omega}{1k\Omega} + \frac{20k\Omega}{196\Omega + 28.7\Omega} = 100$$

For optimum common-mode rejection, the resistor ratios must be precise. High tolerance resistors (±0.5% or better) should be used.

For zero bridge signal, the amplifier will swing to within 2.5mV of 0V. This is the minimum output limit of the OP-213. Therefore, if an offset adjustment is required, one should start the adjustment from a positive voltage and adjust downward until the output stops changing. This is the point where the amplifier limits the swing. Because of the single supply design, the amplifier cannot sense signals which have negative polarity.

The AD7710/AD7711/AD7712/AD7713-family of 22-bit ADCs make a simple matter of interfacing bridges to a data converter. The simplified circuit in Figure 9.29 is an example. The bridge is connected directly to the differential inputs of the ADC. The internal PGA provides a gain of up to 128 to the bridge output (a 350Ω bridge requires 7.1mA @ 2.5V). The AD780 precision reference has sufficient drive current to supply both the bridge and the ADC reference input. With the internal PGA gain set for 64, a 40mV bridge output represents a full scale input to the ADC. The external reference in conjunction with the filter capacitors on the ADC input provide low noise so that an effective resolution (ENOB) of 17 bits is achieved at a 10Hz conversion rate. More details on the operation of the AD77xx-series of ADCs can be found in Reference 4.

WEIGH SCALE APPLICATION USING THE AD7710



- High impedance differential input interfaces directly to bridge
- External reference used for accuracy
- Wide range of impedances can be used for bridge
- Add capacitors to input to filter noise

Figure 9.29

TEMPERATURE TRANSDUCERS

Accurate measurement of temperature is critical to most measurement and process control applications. Thermocouples are capable of measuring extreme temperatures, but require cold-junction-compensation techniques, are non-linear, and have low-level outputs. Semiconductor temperature sensors provide integrated and accurate solutions and high-level outputs, but operate over a more limited range. Resistance Temperature Devices (RTDs) are accurate and linear, but require excitation current and bridge circuits. Thermistors have the most sensitivity but are the most non-linear. The characteristics of temperature sensors are summarized in Figure 9.30.

TEMPERATURE TRANSDUCERS

SEMICONDUCTOR	THERMOCOUPLE	RTD	THERMISTOR
Limited Range: -55 to +150°C	Widest Range: -184 to +2300°C	Range: -200 to +850°C	Range: 0 to +100°C
Linearity: 1°C Repeatability: 0-1°C Accuracy: 1°C	Repeatable Characteristics	Fair Linearity	Poor Linearity
Requires Excitation Source	Needs Cold Junction Compensation	Requires Excitation	Requires Excitation
10mV/K or 1µA/K Output Typical	Low-Voltage Output	Low Cost	High Sensitivity
	Small		

Figure 9.30

Modern semiconductor temperature sensors offer high accuracy and high linearity over an operating range of about –55°C to +150°C. Internal amplifiers can scale the output to convenient values, such as 10mV/°C. They are also useful in cold-junction-compensation circuits for wide temperature range thermocouples.

All semiconductor temperature sensors make use of the relationship between a bipolar junction transistor's (BJT) base-emitter voltage to its collector current:

$$V_{BE} = \frac{kT}{q}\ln\left(\frac{I_C}{I_S}\right)$$

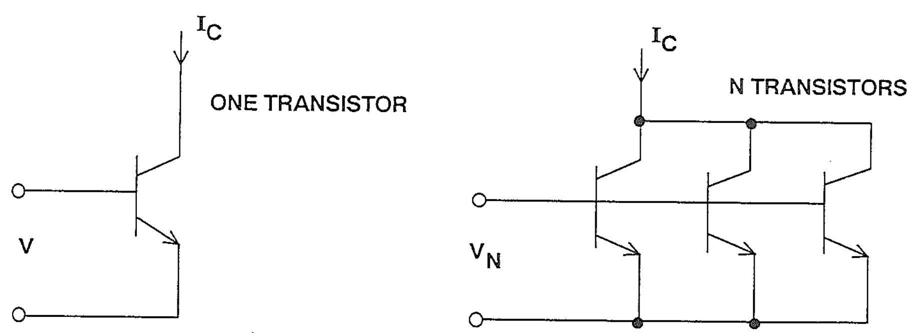
where k is Boltzmann's constant, T is the absolute temperature, q is the charge of an electron, and I_S is a current related to the geometry and the temperature of the transistors. (The equation assumes a voltage of at least a few hundred mV on the collector, and ignores Early effects.)

If we take N transistors identical to the first (see Figure 9.31) and allow the total current I_C to be shared equally among them, we find that the new base-emitter voltage is given by the equation

$$V_N = \frac{kT}{q}\ln\left(\frac{I_C}{N \cdot I_S}\right)$$

Neither of these circuits if much use by itself because of the strongly temperature dependent current I_S, but if we have equal currents in one BJT and N similar BJTs then the expression for the *difference* between the two base-emitter voltages is proportional to absolute temperature and does not contain I_S.

BASIC RELATIONSHIPS FOR SEMICONDUCTOR TEMPERATURE SENSORS



$$V = \frac{kT}{q} \ln\left(\frac{I_C}{I_S}\right) \qquad V_N = \frac{kT}{q} \ln\left(\frac{I_C}{N \cdot I_S}\right)$$

- $V - V_N = \frac{kT}{q} \ln(N),$
- INDEPENDENT OF I_C, I_S

Figure 9.31

$$V - V_N = \frac{kT}{q}\ln\left(\frac{I_c}{I_S}\right) - \frac{kT}{q}\ln\left(\frac{I_c}{N \cdot I_S}\right)$$

$$V - V_N = \frac{kT}{q}\left[\ln\left(\frac{I_c}{I_S}\right) - \ln\left(\frac{I_c}{N \cdot I_S}\right)\right]$$

$$V - V_N = \frac{kT}{q}\ln\left[\left(\frac{I_c}{I_S}\right) \Big/ \left(\frac{I_c}{N \cdot I_S}\right)\right] = \frac{kT}{q}\ln(N)$$

So if we arrange a circuit containing N+1 NPN BJTs, a temperature stable resistor, and a PNP current mirror as shown in Figure 9.32, the total current in the circuit will be proportional to its absolute temperature (PTAT), as will the voltage across the resistor (which is the difference between V and V_N in the equations above).

CLASSIC BANDGAP TEMPERATURE SENSOR

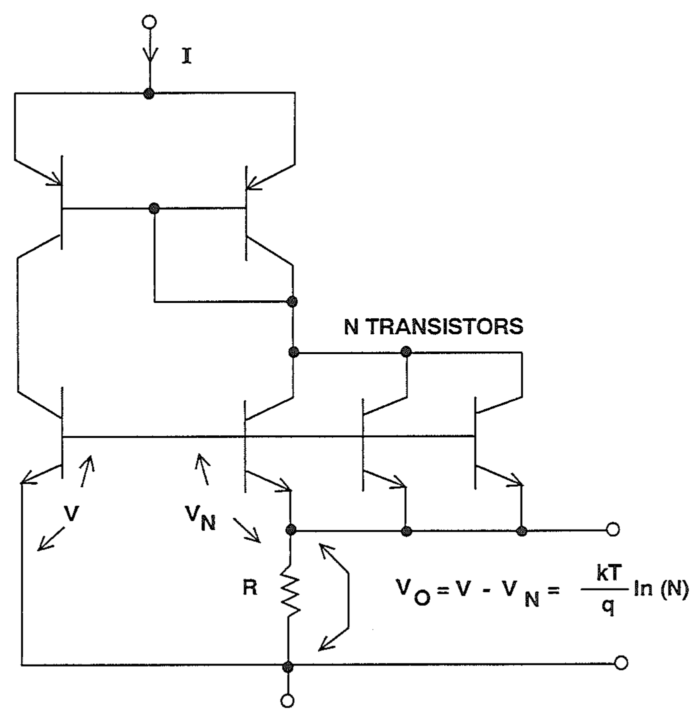


Figure 9.32

This circuit is the basic *band-gap* temperature sensor, and is widely used in semiconductor temperature sensors. The circuit in Figure 9.32 is incomplete, because it has two stable states - one with all the transistors biased on, and one with all the transistors turned off; therefore, turn-on circuitry is required to ensure that when a voltage is applied the transistors will turn on. This is done in different ways in different circuits.

The AD590, AD592, and AC2626 are two terminal temperature sensors with a current equal to 1μA/K (provided there is at least 4V applied to them). They are slightly vulnerable to RFI and should be decoupled at HF where there is a risk of this exposure.

The AD22100 is a low-cost sensor for automotive use. It uses a +5V supply, and its output is 22.5mV/°C from –50°C (at which temperature its output is 0.25V) to +150°C, but its output is proportional to the supply (i.e. if the supply is x% high, so is the output).

The TMP-01 is a sensor with an output of 5mV/K which also contains a voltage reference and two comparators with programmable hysteresis. It operates with supplies from 5 to 12V.

The AD537 is a voltage-to-frequency converter (VFC) which also contains a temperature sensor with 1mV/K output, so that it may be used as a temperature-to-frequency converter.

The AD594, AD595, AD596, and AD597 are special-purpose instrumentation amplifiers, intended for use with thermocouples, which contain temperature sensors to provide cold-junction compensation. (Thermocouples have two junctions and measure temperature *differences*, and to use one successfully it is necessary to compensate for the temperature of the cold junction. This series of devices work on the assumption that the cold junction is at the same temperature as the IC chip. It is therefore essential to ensure that the cold junction is close to the IC [not on the other side of the PC board] and that there are no local heat sources to disrupt it.

THERMOCOUPLES

Thermocouples are small, rugged, relatively inexpensive, and operate over the widest range of all temperature sensors. They are especially useful for making measurements at extremely high temperatures (up to +2300°C) in hostile environments. They produce only millivolts of output, however, and require precision amplification for further processing. They are more linear than many other sensors, and their non-linearity has been well characterized. Some common thermocouples are shown in Figure 9.33. The most common metals used are Iron, Platinum, Rhodium, Rhenium, Tungsten, Copper, Alumel (composed of Nickel and Aluminum), Chromel (composed of Nickel and Chromium) and Constantan (composed of Copper and Nickel).

SOME COMMON THERMOCOUPLES

JUNCTION MATERIALS	TYPICAL USEFUL RANGE (°C)	VOLTAGE SWING OVER RANGE (mV)	ANSI DESIGNATION
Platinum-6% Rhodium / Platinum-30% Rhodium	38 to 1800	13.6	B
Tungsten-5% Rhenium / Tungsten-26% Rhenium	0 to 2300	37.0	(C)
Chromel / Constantan	0 to 982	75.0	E
Iron / Constantan	–184 to 760	50.0	J
Chromel / Alumel	–184 to 1260	56.0	K
Platinum / Platinum-13% Rhodium	0 to 1593	18.7	R
Platinum / Platinum-10% Rhodium	0 to 1538	16.0	S
Copper / Constantan	–184 to 400	26.0	T

Figure 9.33

Figure 9.34 shows the voltage-temperature curves of four commonly used thermocouples, referred to a 0°C fixed-temperature reference junction. Of the thermocouples shown, Type J thermocouples are the most sensitive, producing the largest output voltage for a given temperature change. On the other hand, Type S thermocouples are the least sensitive. These characteristics are very important to consider when designing signal conditioning circuitry in that the thermocouples' relatively low output signals require low-noise, low-drift, high-gain amplifiers.

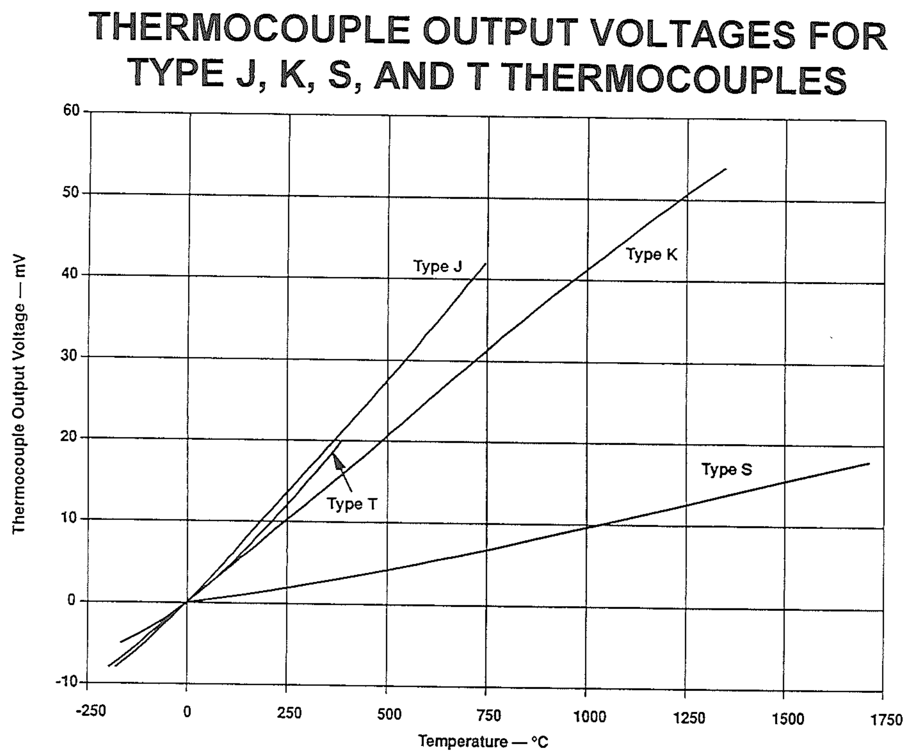


Figure 9.34

To understand thermocouple behavior, it is necessary to consider the non-linearities in their response to temperature differences. Figure 9.34 shows the relationships between sensing junction temperature and voltage output for a number of thermocouple types (in all cases, the reference *cold* junction is maintained at 0°C). It is evident that the responses are not quite linear, but the nature of the non-linearity is not so obvious.

Figure 9.35 shows how the Seebeck coefficient (the *change* of output voltage with *change* of sensor junction temperature - i.e., the first derivative of output with respect to temperature) varies with sensor junction temperature (we are still considering the case where the reference junction is maintained at 0°C).

When selecting a thermocouple for making measurements over a particular range of temperature, we should choose a thermocouple whose Seebeck coefficient varies as little as possible over that range.

THERMOCOUPLE SEEBECK COEFFICIENT VERSUS TEMPERATURE

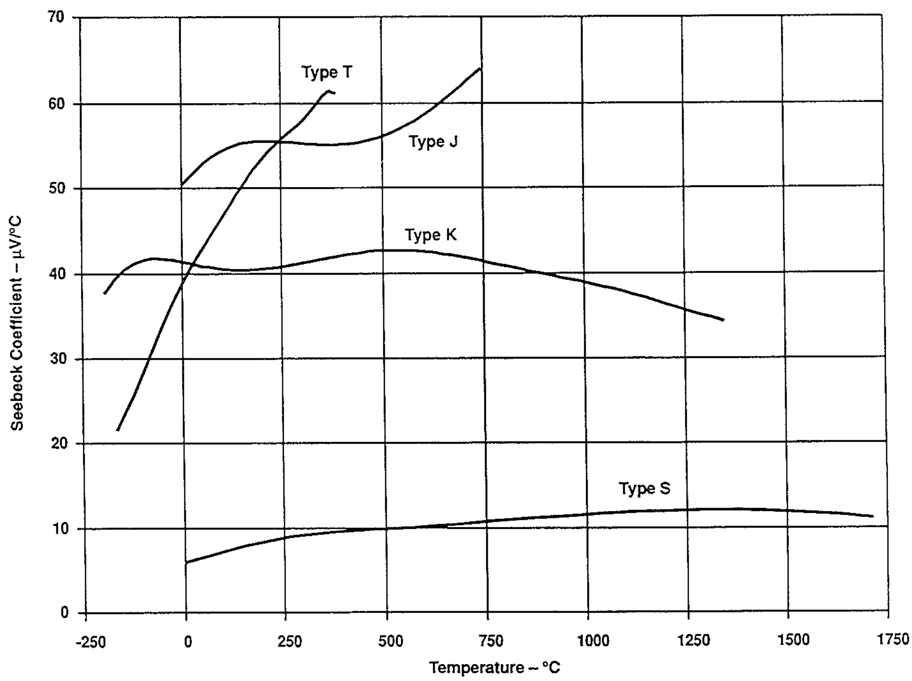


Figure 9.35

To consider two examples: a Type J thermocouple has a Seebeck coefficient which varies by less than 1μV/°C between 200 and 500°C, which makes it ideal for measurements in this range; a Type T thermocouple, on the other hand, has a Seebeck coefficient which increases steadily with increasing temperature, making its output non-linear (but predictably non-linear) over any range. A Type T (copper-constantan) device is useful, though, despite its non-linearity, because its materials are cheap, and one of them (copper) is widely used as an electrical conductor.

Presenting these data on thermocouples serves two purposes: First, Figure 9.34 illustrates the range and sensitivity of the four thermocouple types so that the system designer can, at a glance, determine that a Type S thermocouple has the widest useful temperature range, but a Type J thermocouple is more sensitive. Second, the Seebeck coefficients provide a quick guide to a thermocouple's linearity. Using Figure 9.35, the system designer can choose a Type K thermocouple for its linear Seebeck coefficient over the range of 400°C to 800°C or a Type S over the range of 900°C to 1700°C. The behavior of a thermocouple's Seebeck coefficient is important in applications where variations of temperature rather than absolute magnitude are important. These data also indicate what performance is required of the associated signal conditioning circuitry.

THERMOCOUPLE PRINCIPLES AND COLD-JUNCTION COMPENSATION

To use thermocouples successfully we must understand their basic principles. Consider the diagrams in Figure 9.36.

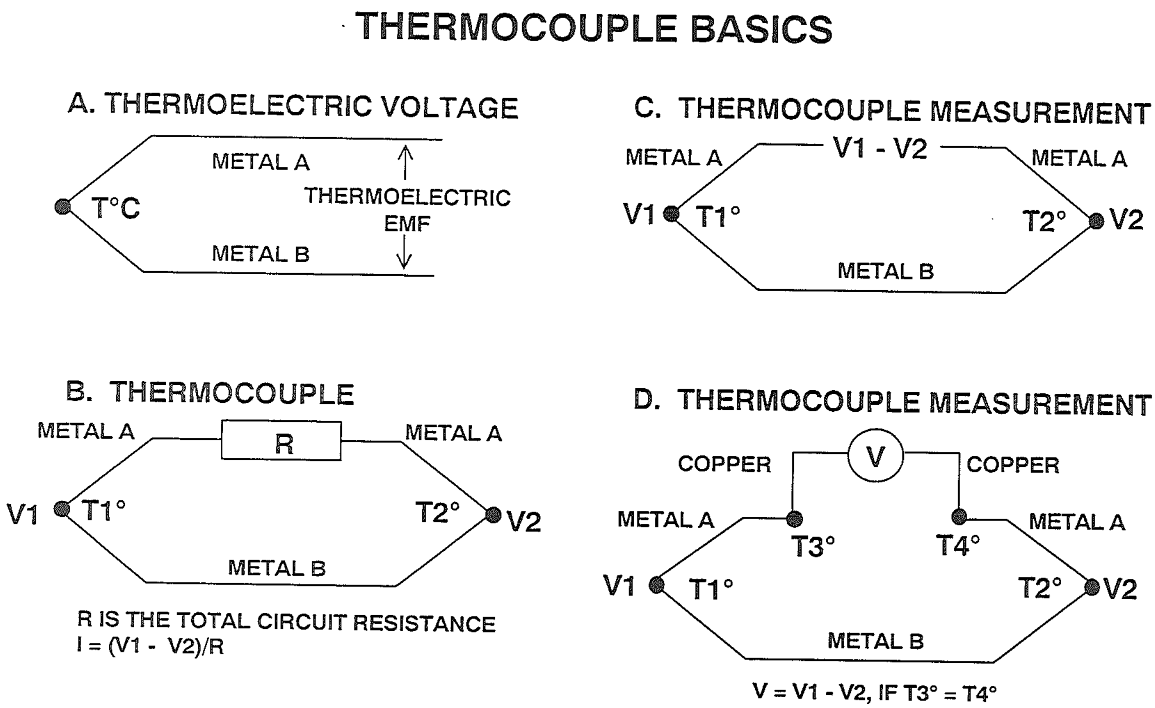


Figure 9.36

If we join two dissimilar metals at any temperature above absolute zero, there will be a potential difference between them (their "thermoelectric e.m.f." or "contact potential") which is a function of the temperature of the junction (Figure 9.36A). A thermocouple consists of two pieces of dissimilar metals (usually two wires) joined at two places. If the two junctions are at different temperatures, there will be a net e.m.f. in the circuit, and a current will flow determined by the e.m.f. and the total resistance in the circuit (Figure 9.36B). If we break one of the wires, the voltage across the break will be equal to the net thermoelectric e.m.f. of the circuit, and if we measure this voltage, we can use it to calculate the temperature difference between the two junctions (Figure 9.36C). *We must always remember that a thermocouple measures the temperature difference between two junctions, not the absolute temperature at one junction.* We can only measure the temperature at the measuring junction if we know the temperature of the other junction (often called the "reference" junction or the "cold" junction).

But it is not so easy to measure the voltage generated by a thermocouple. Suppose that we attach a voltmeter to the circuit in Figure 9.36C (Figure 9.36D). The wires with which we attach the voltmeter will form further thermojunctions where they are attached. If both these additional junctions are at the same temperature (it does not matter what temperature), then the

"Law of Intermediate Metals" states that they will make no net contribution to the total e.m.f. of the system. If they are at different temperatures, they will introduce errors. Since *every pair of dissimilar metals in contact generates a thermoelectric e.m.f.* (including copper/solder, kovar/copper [kovar is the alloy used for IC leadframes] and aluminum/kovar [at the bond inside the IC]), it is obvious that in practical circuits the problem is even more complex, and it is necessary to take extreme care to ensure that all the junctions in the circuitry around a thermocouple, except the measurement and reference junctions themselves, are at the same temperature.

Thermocouples generate a voltage, albeit a very small one, and do not require excitation. But they do require that the reference junction (often called the "cold" junction, even though, in some systems, it may be at a higher temperature than the measurement junction!) be at a well-defined temperature. A conceptually simple approach to this need is shown in Figure 9.37, but although an ice/water bath is easy to define, it is inconvenient to maintain.

CLASSICAL COLD-JUNCTION COMPENSATION USING AN ICE-POINT (0°C) REFERENCE JUNCTION

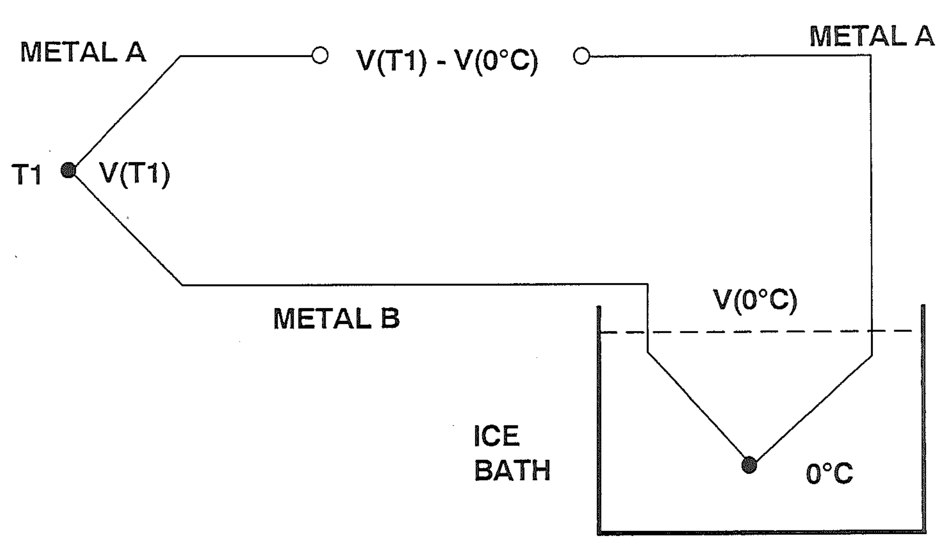
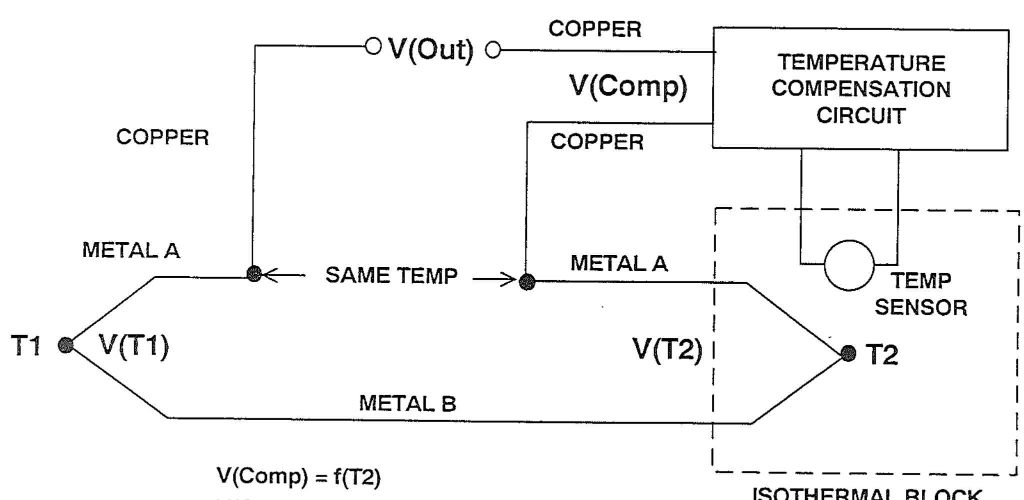


Figure 9.37

Today an ice-point reference, and its inconvenient ice/water bath, is generally replaced by electronics. A temperature sensor of another sort (often a semiconductor sensor, sometimes a thermistor) measures the temperature of the cold junction and is used to inject a voltage into the thermocouple circuit which compensates for the difference between the actual cold junction temperature and its ideal value (usually 0°C). Ideally, the compensation voltage should be an exact match for the difference voltage required, which is why the diagram gives the voltage as f(T2) (a *function* of T2) rather than KT2, where K is a simple constant. In practice, since the cold junction is rarely more than a few tens of degrees from 0°C, and generally varies by little more than ±10°C, a linear approximation (V=KT2) to the more complex reality is sufficiently accurate and is what is often used. (The expression for the output voltage of a thermocouple with its measuring junction at T°C and its reference at 0°C is a polynomial of the form $V = K_1T + K_2T^2 + K_3T^3 + ...$, but the values of the coefficients K_2, K_3, etc. are very small for most common types of thermocouple. References 8 and 17 give the values of these coefficients for a wide range of thermocouples.) Some thermocouple amplifiers (for example Analog Devices' AD594/5/6/7 series) incorporate cold junction compensation in the IC amplifier - with such devices it is very important that the IC chip is at the same temperature as the cold junction of the thermocouple, which is usually achieved by keeping the two in close proximity and isolated from any heat sources.

USING A TEMPERATURE SENSOR FOR COLD-JUNCTION COMPENSATION



V(Comp) = f(T2)
V(Out) = V(T1) - V(T2) + V(Comp)
V(Out) = V(T1) - V(0°C), if V(Comp) = V(T2) - V(0°C)

Figure 9.38

When electronic cold-junction compensation is used, it is common practice to eliminate the additional thermocouple wire and terminate the thermocouple leads in the isothermal block in the arrangement shown in Figure. 9.39.

TERMINATING THERMOCOUPLE LEADS DIRECTLY TO AN ISOTHERMAL BLOCK

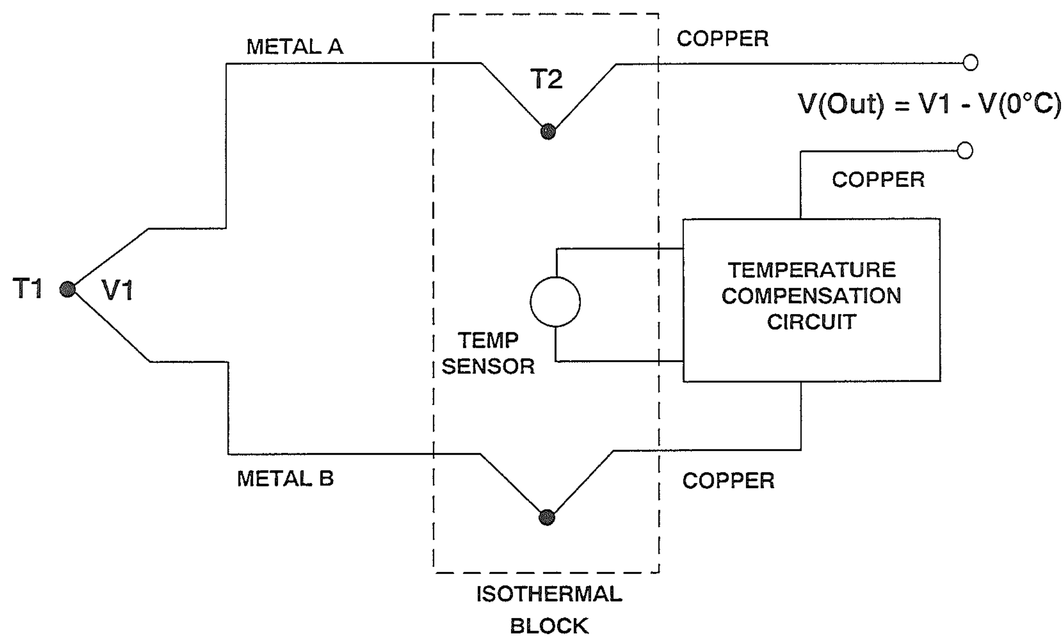


Figure 9.39

THERMOCOUPLE AMPLIFIER CONSIDERATIONS

There are a number of ways to design thermocouple signal conditioning amplifiers. The appropriate thermocouple for the temperature range is chosen to minimize the accuracy required of the circuit. The selection of the thermocouple which is most linear in the required temperature range permits straightforward designs with a minimum of components. The best way to achieve accurate measurements over narrow or wide temperature ranges is to use cold-junction compensation and a thermocouple whose Seebeck coefficient is linear over the temperature range of interest. It is not always that straightforward — in many cases, a designer must use whatever thermocouple is available. To illustrate the issues of thermocouple signal conditioning, let us begin by considering a Type T thermocouple amplifier.

Once the measurement temperature range is known, the next step is to decide what type of linearization must be applied to the thermocouple's nonlinear output. Thermocouples do not have a constant Seebeck coefficient over temperature, and some type of linear approximation must be used to linearize them. The first example is a signal conditioning amplifier for a Type T thermocouple which measures temperatures over the range of 0°C to 100°C. In Figure 9.40, a Type T thermocouple response is shown with two linear approximations: end-point and least squares.

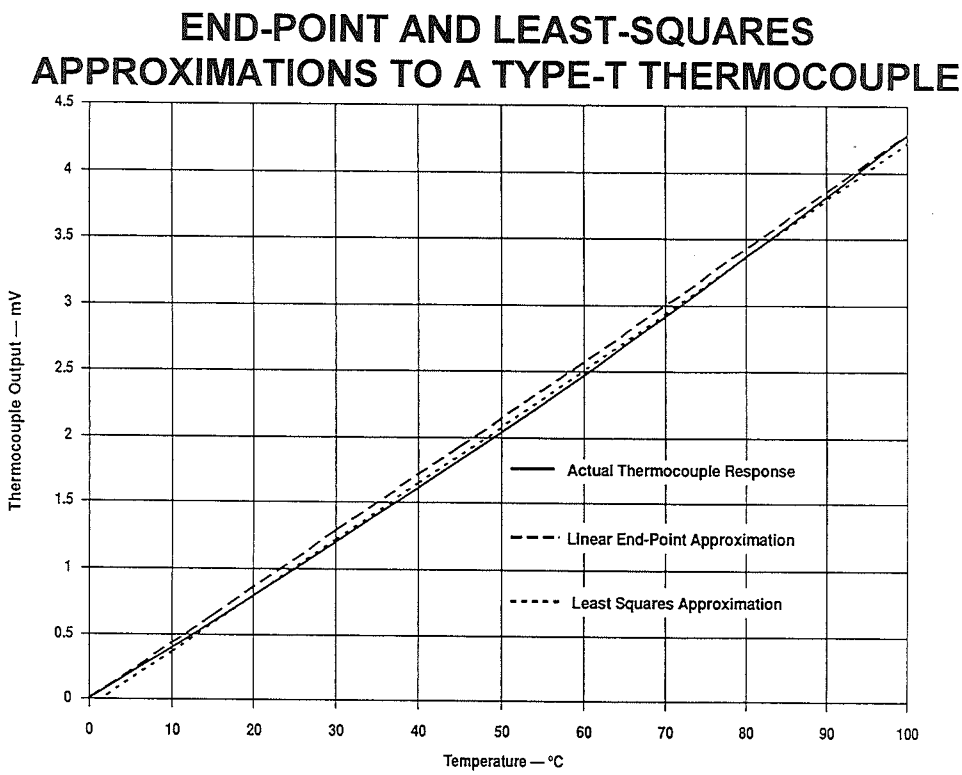
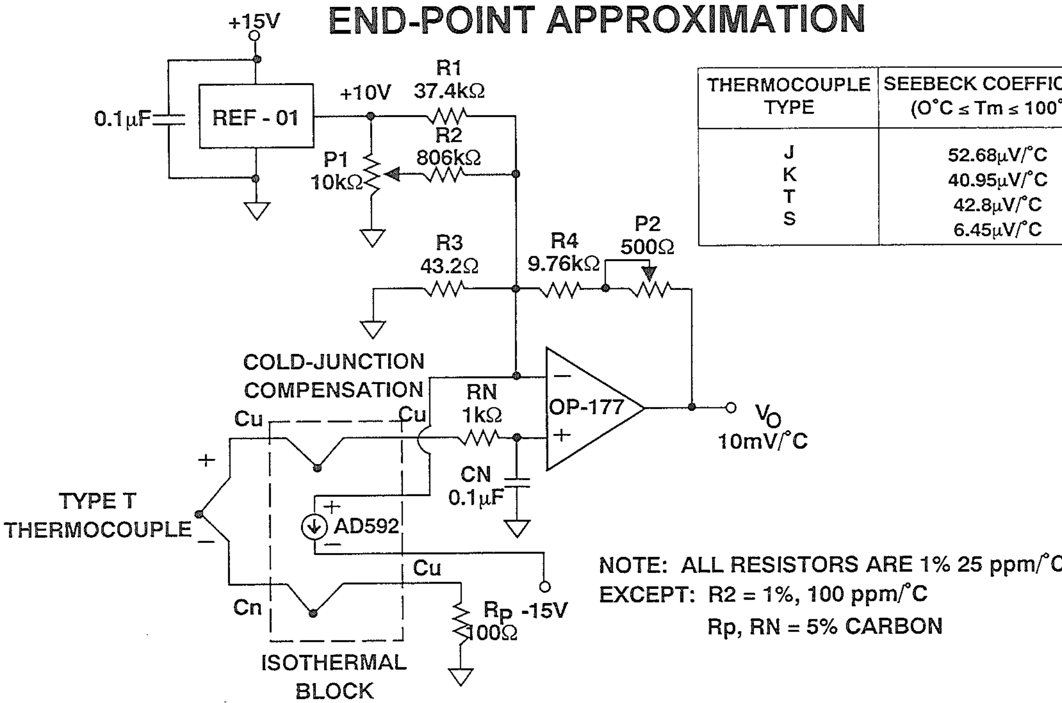


Figure 9.40

An end-point approximation consists of a straight line between the two endpoints of the temperature range. The error is zero at the endpoints and is a maximum at the midpoint of the range, –2.5°C at 50°C in this example. The slope of the line is 42.8μV/°C, and this is the Seebeck coefficient value used when the cold-junction compensator is designed. On the other hand, a least squares approximation to the thermocouple's response intersects the response at two points and exhibits a minimum mean-square error over the temperature range of interest. The error in this example is zero at 20°C and at 80°C, with the maximum error at the midpoint and the endpoints. Typically with this approach the error becomes too large at either temperature extreme, and the designer is forced to adjust the slope or the intercept to reduce the overall error. Fortunately, many scientific calculators provide a least squares curve fitting capability. The algorithm only works if there are enough data points. In this example, data points were in 5°C increments from 0°C to 100°C, and the approximation yielded an error of –1.5°C at the endpoints and + 0.9°C at midscale. The Seebeck coefficient is 42.8 μV/°C, and the intercept is –65 μV. The only difference between the two approximations is the offset. The only way to reduce the error in the approximation is to reduce the measurement temperature range or use linearization techniques. This example shows that linear approximations to a thermocouple's characteristic are accurate for small temperature ranges.

To illustrate the design of a Type T thermocouple amplifier, a Seebeck coefficient of 42.8μV/°C is used in a circuit which measures temperature over the range 0°C to 100°C to an accuracy of ± 0.4 °C. The amplifier, shown in Figure 9.41, is designed to provide a 10mV/°C output and requires a gain of 233.8. Changing R3 as shown in the table allows the circuit to work with other types of thermocouples.

COLD-JUNCTION COMPENSATED GROUNDED TYPE-T THERMOCOUPLE AMPLIFIER USING END-POINT APPROXIMATION



THERMOCOUPLE TYPE	SEEBECK COEFFICIENT (0°C ≤ Tm ≤ 100°C)	R3 (Ω)
J	52.68µV/°C	53.6Ω
K	40.95µV/°C	41.2Ω
T	42.8µV/°C	43.2Ω
S	6.45µV/°C	6.49Ω

Figure 9.41

A grounded thermocouple is used here because this minimizes noise pickup from long leads. R_N and C_N serve as a noise filter to any noise that the lead wires pick up. The cutoff frequency of 1.6kHz can be lowered by increasing C_N. Although larger values of R_N might be used at the input of the OP-177, they will increase input bias-current induced offset and drift effects. A resistor, R_P, is used in series with exposed thermocouples as protection against contact with some high potential.

Cold-junction compensation uses an AD592 monolithic temperature sensor at the thermocouple's reference (cold)

9

junction. The sensor monitors the temperature of the cold-junction and provides a temperature-dependent current to the amplifier's summing junction. An offset current nulling the sensor's output current at 0°C is produced by R1 and a precision 10V reference.

Calibration of the circuit is a two-step process. First, the circuit is allowed to warm-up for 5 minutes to allow the temperature sensor, reference, resistors, and op amp to stabilize. To protect the circuit from ambient thermal gradients or air currents, which can affect calibration, the circuit should be covered or enclosed in a box. For 0°C calibration, the thermocouple terminals are short-circuited, and P1 is adjusted such that the output voltage equals the temperature of the isothermal block, according to the relationship 10mV/°C • T_A. This trims out amplifier offset voltage and bias currents, resistor tolerances, and errors in the reference and the temperature sensor. For full-scale adjustment, the short is replaced by a precision dc voltage source set to the thermocouple's output voltage at 100°C. For a Type T thermocouple, the voltage is 4.277mV. P2 is then adjusted such that the output voltage is 1V above its previously measured value; that is, V_{OUT} = 1V + 10mV/°C • T_A. For example, if the ambient temperature of the factory or lab is 45°C (whew!!), then P2 is adjusted so that V_{OUT} = 1.45 V. The temperature of the isothermal block must not change during calibration.

The largest source of error over this temperature range comes from the approximation made to the thermocouple's characteristic. As we have shown, a Type T thermocouple exhibits a Seebeck coefficient of 38.9μV/°C at 0°C that increases to 46.3μV/°C at 100°C. Therefore, using a constant 42.8μV/°C over this range induces a 2.5°C error in the measurements. A Type K characteristic induces less than a 0.7°C error, and a Type J characteristic induces less than 1°C error. In applications where the output is digitized, it may be made more accurate by using software linearization with correction factors stored in a lookup table.

The OP-177 is an excellent choice for this application for a number of reasons: (1) its low input bias current allows the use of a filter and current limiting without generating large parasitic offset voltages and drift; (2) its low input offset voltage reduces static errors at the output, and its low offset drift contributes less than 0.06°C error for ambient temperatures between 20°C and 50°C; and (3) the amplifier's open-loop gain of greater than 5 million keeps the amplifier's gain error below 0.004°C over the entire ambient temperature range.

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 provide a high level (10mV/°C) output directly from the 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 becoming a stand-alone Celsius transducer with 10mV/°C 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 device can be powered from a single-ended supply (which may be as low as +5V), but by including a negative supply, temperatures below 0°C can be measured. To minimize self-heating, an unloaded AD594/AD595 will operate with a supply current of 160μA, but is also capable of delivering ±5mA to a load.

The AD594 is precalibrated by laser wafer trimming to match the characteristics of type J (iron/constantan) thermocouples, and the AD595 is laser trimmed for type K (chromel/alumel). 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 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 with cold junctions between 0 to +50°C. The circuit shown in Figure 9.42 will provide a direct output from a type J thermocouple (AD594) or a type K thermocouple (AD595) capable of measuring 0 to +300°C.

AD594/AD595 MONOLITHIC THERMOCOUPLE AMPLIFIERS WITH COLD-JUNCTION COMPENSATION

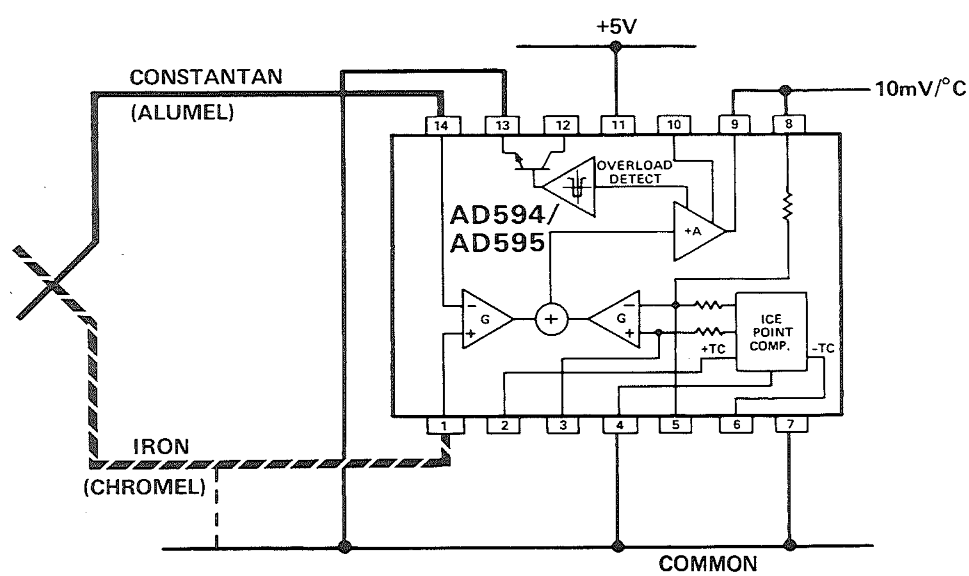


Figure 9.42

The AD596/AD597 are monolithic set-point controllers which have been optimized for use at elevated temperatures as are found in oven control applications. The device cold-junction compensates and amplifies a type J or K thermocouple to derive an internal signal proportional to temperature. They can be configured to provide a voltage output (10mV/°C) directly from type J or K thermocouple signals. The device is packaged in a 10-pin metal can and is trimmed to operate over an ambient range from +25°C to +100°C. The AD596 will amplify thermocouple signals covering the entire –200°C to +760°C temperature range recommended for type J thermocouples while the AD597 can accommodate –200°C to +1250°C type K inputs. They have a calibration accuracy of ±4°C at an ambient temperature of 60°C and an ambient temperature stability specification of 0.05°C/°C from +25°C to +100°C.

MAINTAINING PROPER COLD-JUNCTION COMPENSATION

One of the largest sources of error in thermocouple signal conditioning circuits is poor cold-junction compensation. Since a thermocouple output voltage is a function of the temperature difference between its two junctions, any temperature difference between the reference junction and the cold-junction temperature sensor will produce an error signal. It is therefore essential that the layout of the temperature sensor and the thermocouple reference junction minimizes any temperature gradients. Errors are often introduced when the cold junction is not properly transferred to the PC board. An example of correct transfer of the cold junction to the PC board is shown in Figure 9.43.

TRANSFERRING THE COLD-JUNCTION FROM AN ENCLOSURE OR CONNECTOR TO THE PC BOARD

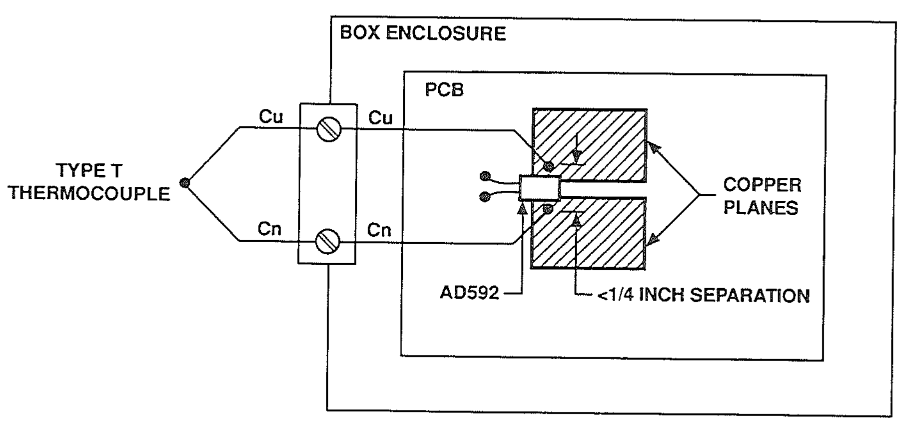


Figure 9.43

The layout illustrates how to extend the cold-junction reference from the terminal block (or connector) to the PC board, and how to locate the temperature sensor relative to the thermocouple reference junction. The wires from the terminal block (or connector) to the PC board must be of the same material as that used for the thermocouple wires.

MINIMIZING PARASITIC PCB THERMOCOUPLE ERRORS

Thermojunction voltages are generated whenever dissimilar conductors make contact. There are many such contacts in an electronic circuit board, where conductors include copper, kovar and alloy-42 (IC leads), aluminum (IC Bond wires), solder, plating (gold, nickel, silver and others), nichrome and cermets (in resistors), and brass and steel (in terminal blocks). Circuit boards are therefore full of unnoticed thermocouples. If they are all at the same temperature, their net effect is zero, but in reality all PC boards have innumerable temperature gradients, and thermocouple effects can seriously affect the accuracy of low-level DC measurements, including thermocouple measurements.

When designing such boards, the engineer should ensure that no temperature gradients exist which might induce thermoelectric voltages and so affect the accuracy of the system. If temperature gradients cannot be eliminated (the usual case), junctions and components should be placed isothermally and as close together as possible, especially the cold-junction sensor, the cold-junction itself, the amplifier input leads, and the gain resistor, so that parasitic thermoelectric voltages cancel out. Figure 9.44 is an example of a printed circuit board layout of the circuit in Figure 9.41 illustrating the isothermal placing and close location of the thermocouple terminating junction, the temperature sensor, the amplifier input pins, and the gain setting resistors.

Beware of amplifier offset voltage warm-up drift caused by mismatched materials in the wire bond/lead system of an IC package. This effect can be as high as tens of microvolts in TO-5 cans with Kovar leads. It has nothing to do with the actual offset drift specification and can occur in amplifiers with measured “zero” drift. Warm-up drift is directly proportional to amplifier power dissipation and can be minimized by avoiding TO-5 cans, using low-supply current amplifiers, and by using the lowest possible supply voltages. Its effects be minimized by calibrating and specifying the system after a warm-up period.

PROPERLY LOCATING ALL COMPONENTS THAT ARE SENSITIVE TO THERMAL GRADIENTS

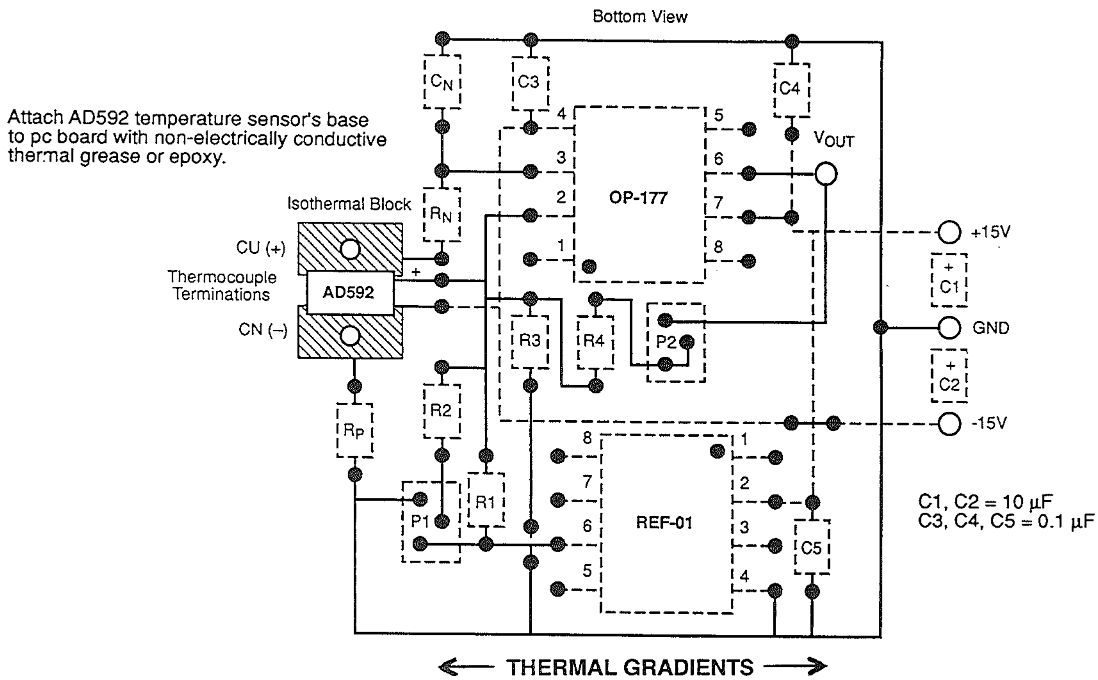


Figure 9.44

THERMOCOUPLE LINEARIZATION TECHNIQUES

To improve accuracy, the thermocouple's response can be linearized. There are a number of techniques that can be used to linearize a thermocouple, including breakpoint correction, analog computation, and digital correction (Reference 5).

Thermocouple non-linearity correction using analog computation makes use of analog multipliers and operational amplifiers to synthesize a function that models the thermocouple's voltage-temperature characteristic. For more detailed information on this subject, the reader should consult References 2 and 6.

Another technique for linearizing a thermocouple's response is to use breakpoints that change the gain of the circuit as the signal increases. Breakpoints use as many straight-line segments as are required by system accuracy to fit the thermocouple characteristic. This approach is circuit intensive: the number of linear segments determines the number of amplifiers the linearizer requires (see Section 2 of Reference 3).

An increasingly popular technique for thermocouple linearization is to digitize the output of an accurate thermocouple amplifier and apply continuous correction to thermocouple output with factors that are stored in ROM (Reference 7 and 18). One method uses a power series polynomial to implement the correction and has the form:

$$T = a_0 + a_1X + a_2X^2 + ... + a_nX^n$$

where T is the thermocouple temperature, X is the thermoelectric voltage, and a_0, a_1, a_2, ..., and a_n are the coefficients of the polynomial for each type of thermocouple. The coefficients for each thermocouple can be found in References 8 and 17. Another advantage of digital techniques is the elimination of trimming potentiometers. Computers and microprocessors can be programmed to execute a calibration sequence at any time and are far more efficient than analog linearization circuits in this application. Digital signal conditioners such as the AD1B60 (to be described shortly) provide complete signal conditioning and provide software linearization for a number of sensors.

RESISTANCE TEMPERATURE DETECTOR (RTD) SIGNAL CONDITIONING

The Resistance Temperature Detector, or the RTD, is a sensor whose resistance changes with temperature. Typically built of a platinum (Pt) wire wrapped around a ceramic bobbin, the RTD exhibits behavior which is more accurate and more linear over wide temperature ranges than a thermocouple. Figure 9.45 illustrates the temperature coefficient of a 100-ohm RTD and the Seebeck coefficient of a Type S thermocouple. Over the entire range (approximately –200°C to +850°C), the RTD is a more linear device. Hence, linearizing an RTD is less complex.

Unlike a thermocouple, however, an RTD is a passive sensor and requires current excitation to produce an output voltage. The RTD's low temperature coefficient of 0.385%/°C requires similar high-performance signal conditioning circuitry to that used by a thermocouple; however, the voltage drop across an RTD is much larger than a thermocouple output voltage. A system designer may opt for large value RTDs with higher output, but large-valued RTDs exhibit slow response times. Furthermore, although the cost of RTDs is higher than that of thermocouples, they use copper leads, and thermoelectric effects from terminating junctions do not affect their accuracy. And finally, because their resistance is a function of the absolute temperature, RTDs require no cold-junction compensation.

Caution must be exercised using current excitation because the current through the RTD causes heating. This self-heating changes the temperature of the RTD and appears as a measurement error. Hence, careful attention must be paid to the design of the signal conditioning circuitry so that self-heating is kept below 0.5°C. Manufacturers specify self-heating errors for various RTD values and sizes in still and in moving air. To reduce the error due to self-heating, the minimum current should be used for the required system resolution, and the largest RTD value chosen that results in acceptable response time.

LINEARITY COMPARISON BETWEEN PLATINUM RTD AND TYPE-S THERMOCOUPLE

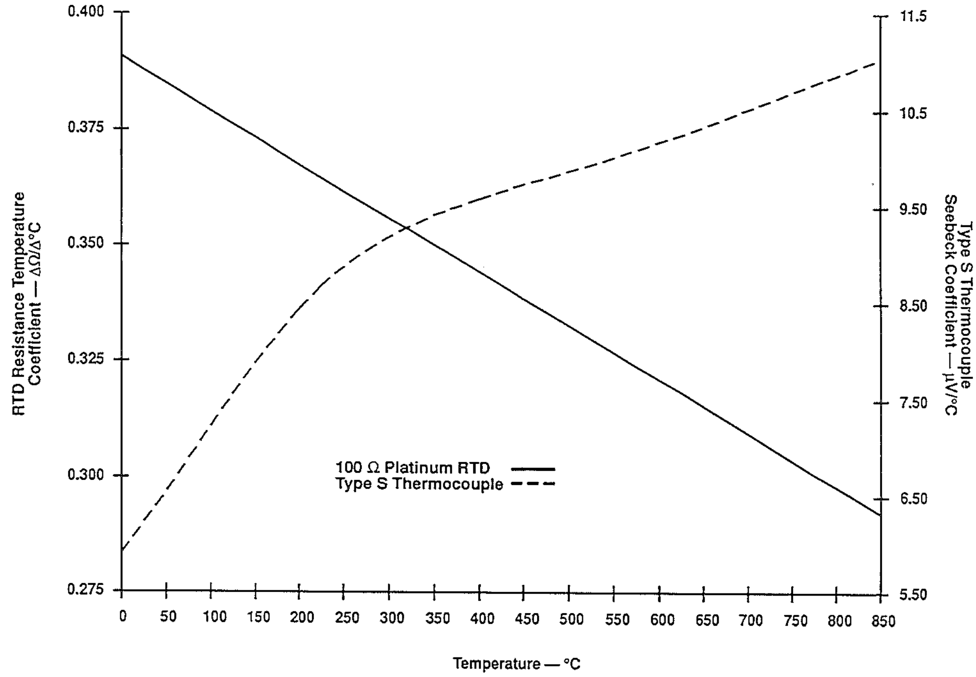
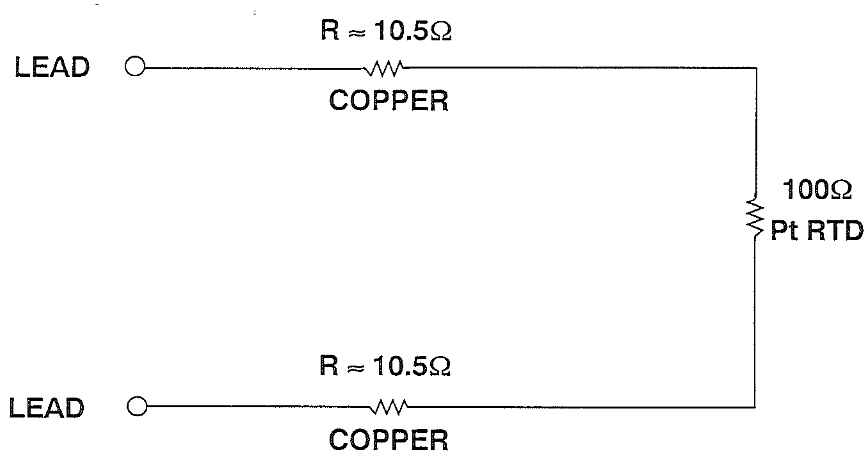


Figure 9.45

Another effect that can produce measurement error is voltage drop in RTD lead wires. This is especially critical with low-value 2-wire RTDs because the temperature coefficient and the absolute value of the RTD resistance are both small. If the RTD is located a long distance from the signal conditioning circuitry, then the lead resistance can be ohms or tens of ohms, and a small amount of lead resistance can contribute a significant error to the temperature measurement. To illustrate this point, let us assume that a 100Ω platinum RTD with 30-gauge copper leads is located about 100 feet from a controller's display console. The resistance of 30-gauge copper wire is 0.105Ω/ft, and the two leads of the RTD will contribute a total 21Ω to the network which is shown in Figure 9.46. This additional resistance will produce a 55°C error in the measurement! The leads' temperature coefficient can contribute an additional, and possibly significant, error to the measurement. To eliminate the effect of the lead resistance, a 4-wire technique is used.

A 100Ω PLATINUM RTD WITH 100 FEET OF 30-GAUGE LEAD WIRES



RESISTANCE TC OF COPPER ≈ 0.4%/°C @ 20°C

Figure 9.46

FOUR-WIRE OR KELVIN CONNECTION TO RTD

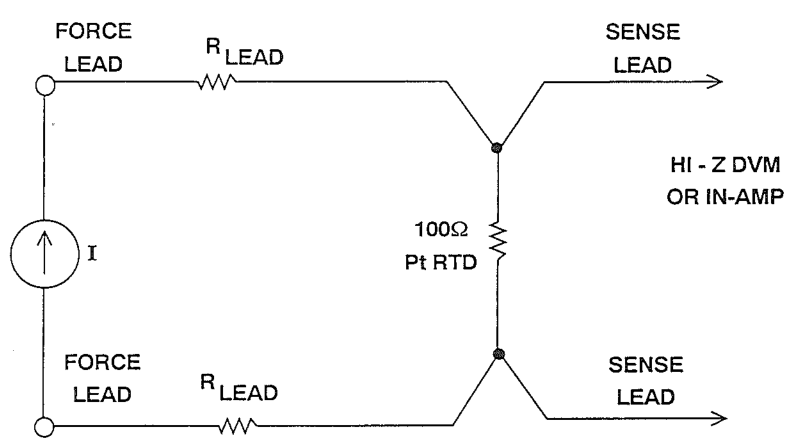


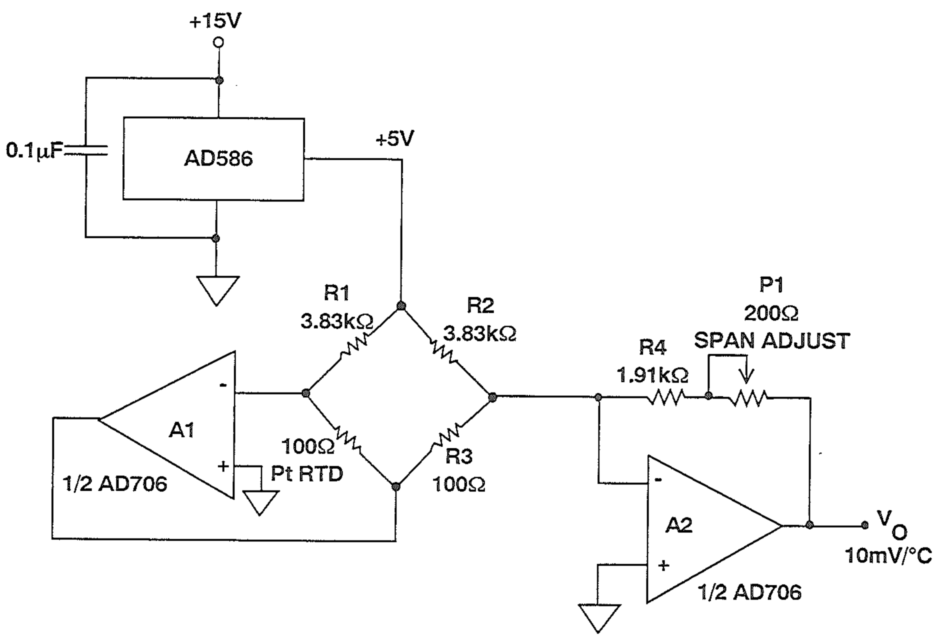
Figure 9.47

In Figure 9.47, a 4-wire, or Kelvin, connection is made to the RTD. A constant current is applied though the FORCE leads of the RTD and the voltage across the RTD itself is measured remotely via the SENSE leads. The measuring device can be a DVM or an instrumentation amplifier, and high accuracy can be achieved provided that the measuring device exhibits high input impedance and/or low input bias current. Since the SENSE leads do not carry appreciable current, this technique is insensitive to lead wire length. Sources of errors are the stability of the constant current source and the input impedance and/or bias currents in the amplifier or DVM.

Bridge circuits used in RTD signal conditioning effectively linearize transducer outputs. They do not compensate for a major source of bridge-instrumentation error: an operational amplifier's V_{OS} drift. In many bridge designs, amplifier ΔV_{OS} can see four times more gain than the sensor signal because of the bridge's voltage-divider effects.

Without sacrificing linearity, an amplifier's ΔV_{OS} can be minimized by applying feedback to the bridge. A pair of op amps can be used to place all bridge elements under fixed bias or feedback control. Shown in Figure 9.48 is a circuit which uses an AD706 dual operational amplifier and a 100Ω Pt RTD to measure temperature from 0°C to 100°C.

RTD BRIDGE AMPLIFIER



NOTE: ALL FIXED RESISTORS ARE 0.1% IMPERIAL ASTRONICS M015

Figure 9.48

A1 forces a null at one bridge node so that a constant current, defined by the AD586 and R1, flows through the RTD. Amplifier A2 is then used to establish a null at the other bridge node. Therefore, any variation of the RTD resistance generates a signal voltage at the output of A1 which is then converted to a signal current for A2 by R3. This signal current is then summed with the current generated by the reference's 5V output and R2, and the result scaled by R4 to provide a 10mV/°C output. Since the AD706 input bias current is typically 30pA, its error contributions can be neglected.

The circuit's response is dominated by the RTD, and the circuit has a low sensitivity to amplifier input offset voltage and drift. By placing the RTD in the feedback path of A1, A1's input offset voltage effects are effectively suppressed, and the circuit's response is very accurate. Although the output amplifier's V_{OS} appears at the output, it is not amplified by the same gain as in conventional approaches and introduces less than 0.002°C error at 0°C and <0.04°C error over a 20°C to 50°C range of amplifier temperature.

Calibration of the circuit involves one trim. Because of the AD706 low V_{OS} (10μV) and the use of precision wirewound resistors, the circuit error at 0°C is less than 0.1°C. Calibration is only required at the full-scale temperature of 100°C. The RTD is replaced by a precision decade resistance box which is set to 138.5Ω. P1 is adjusted such that V_O equals 1.00V. With this single calibration and 0.1% low-TCR resistors, the measurement error is better than ± 0.2 °C over a 20°C to 50°C ambient temperature range.

Although the servo loop controlled by A1 maintains a constant current in the RTD, this topology does not correct the non-linearity of the RTD. For example, if this topology were used with the same RTD to measure the range of 0°C to 200°C, curvature of the RTD response will cause a measurement error of approximately 3°C at full scale.

Linearization of RTD outputs can be achieved using op-amps and a small amount of positive feedback. Figure 9.49 shows a single supply circuit to correct the nonlinear behavior of an RTD. The RTD operates in one leg of a full bridge circuit that is excited by a constant current source established by 1/2 of an OP-295 dual op amp. The bridge current is regulated by servoing the current flowing into resistor R_{SENSE} and comparing with a 200mV reference voltage derived from the REF-43.

The temperature-dependent resistance change is amplified by an instrumentation amplifier, the AMP-04. Scaling is such that for 0°C the amplifier output is 0V, and at 400°C the amplifier output is +4.00V.

The RTD has an inherent non-linearity in its resistance-versus-temperature function. If uncorrected, the sensor nonlinearity would produce a 20°C error over the 400°C temperature range. This nonlinearity can be corrected by providing a small amount of positive feedback to the reference voltage, which increases the bridge current at high temperatures. The amount of positive feedback is so small as not to cause a stability problem.

Calibration is an interactive three-step procedure. The FULL-SCALE and LINEARITY potentiometers are first set to the middle of their adjustment ranges. The first calibration is made with the zero adjust, and must be made at a voltage other than zero, since zero volts is the negative voltage limit of the

PRECISION SINGLE SUPPLY RTD AMPLIFIER WITH LINEARIZATION

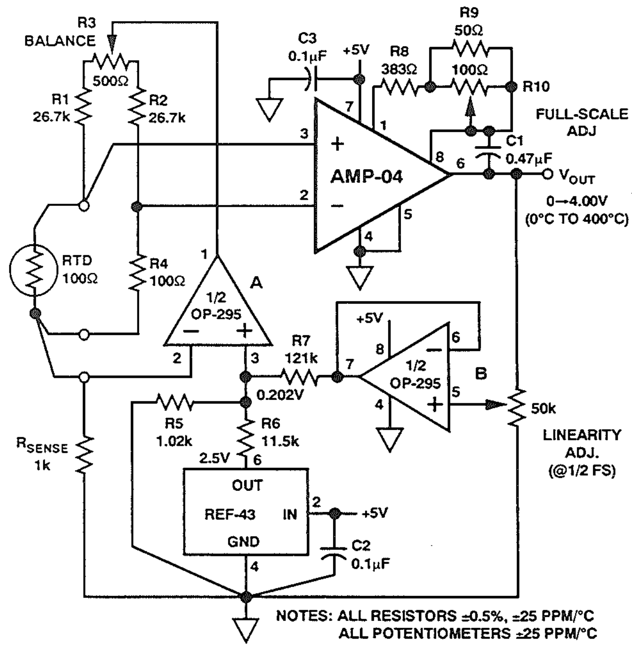


Figure 9.49

FOUR-WIRE RTD APPLICATION USING THE 22-BIT AD7711 SIGNAL CONDITIONING ADC

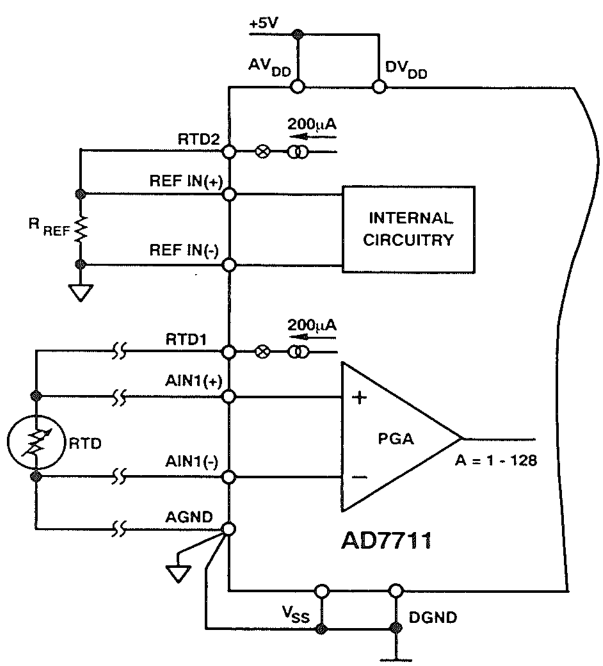


Figure 9.50

circuit. A convenient zero calibration point is +5°C. Substitute a stable 101.95Ω resistor for the RTD and then adjust the ZERO ADJUST potentiometer for 0.050V output.

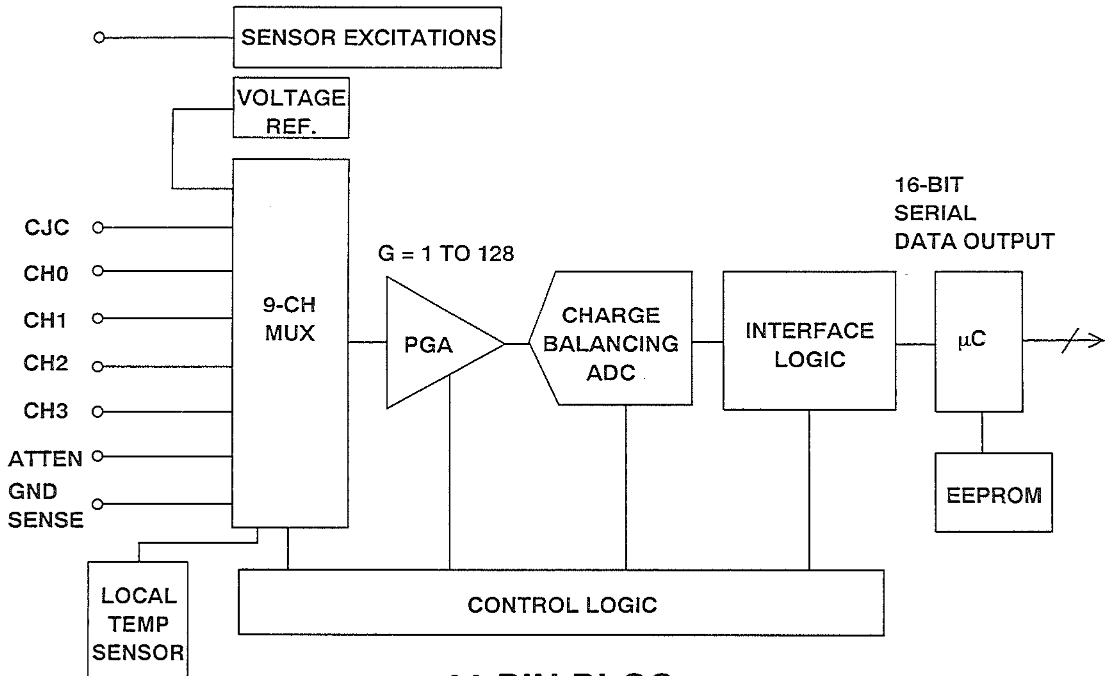
Next substitute the full-scale (400°C) equivalent RTD resistance of 247.04Ω. Adjust FULL-SCALE ADJUST for 4.000V output. Then substitute the resistance corresponding to half-scale (200°C), or 175.84Ω and adjust the LINEARITY ADJUST pot for a 2.000V output. Since the FULL-SCALE and LINEARITY adjustments are interactive, it is necessary to repeat the FULL-SCALE and LINEARITY calibration routine once or twice until no further adjustment is necessary.

Once calibrated, the amplifier is accurate to better than ±0.5°C over the 0°C to 400°C measurement range. If a higher supply voltage is available, the range can be increased.

Highly integrated signal conditioning ADCs, such as the AD7711, can be interfaced directly to RTDs as in Figure 9.50. The 22-bit AD7711 has two internal current sources of 200μA each. In this four-wire configuration, there are no errors associated with lead resistance as no current flows in the measurement leads, which are connected to AIN1(+) and AIN1(–). One of the current sources is used to provide the excitation current for the RTD. A 100Ω RTD will generate a 20mV signal which can be handled directly by the analog input of the AD7711. The second RTD excitation current is used to generate the reference voltage. This voltage is developed across R_{REF} and applied to the differential reference inputs. For a nominal reference voltage of +2.5V, R_{REF} is 12.5kΩ. This scheme ensures that the analog input voltage remains ratiometric to the reference voltage. Any errors in the analog input voltage due to temperature drift of the RTD current source are compensated by the variation in the reference voltage. The typical matching between the two RTD current sources varies by less than 3ppm/°C.

Figure 9.51 shows a simplified block diagram of a complete *intelligent digitizing signal conditioner*, the AD1B60. The device consists of a signal conditioning front end followed by a charge balance ADC and a mask-programmed microcontroller with EEPROM memory. The primary application of the AD1B60 is a user-configurable digitizing signal conditioner for RTDs, thermocouples, and low- and high-level voltage signals. (For a detailed description of the AD1B60, see Section 7 of Reference 4). The AD1B60 provides the algorithms required for scaling and linearizing many types of thermocouples and RTDs. For sensors not included in the built-in routines, a customized linearization algorithm can be created for *any* sensor. Besides its multiplexer, PGA, ADC, and microcontroller, the AD1B60 also provides digitally-controlled current sources for RTD and thermistor excitation and open thermocouple detection, and an internal 5:1 attenuator for high-level input signals (Figure 9.52).

AD1B60 INTELLIGENT, DIGITIZING SIGNAL CONDITIONER



44-PIN PLCC

Figure 9.51

AD1B60 BUILT-IN CONDITIONING CIRCUITS REDUCE OVERALL COMPONENTS COUNT

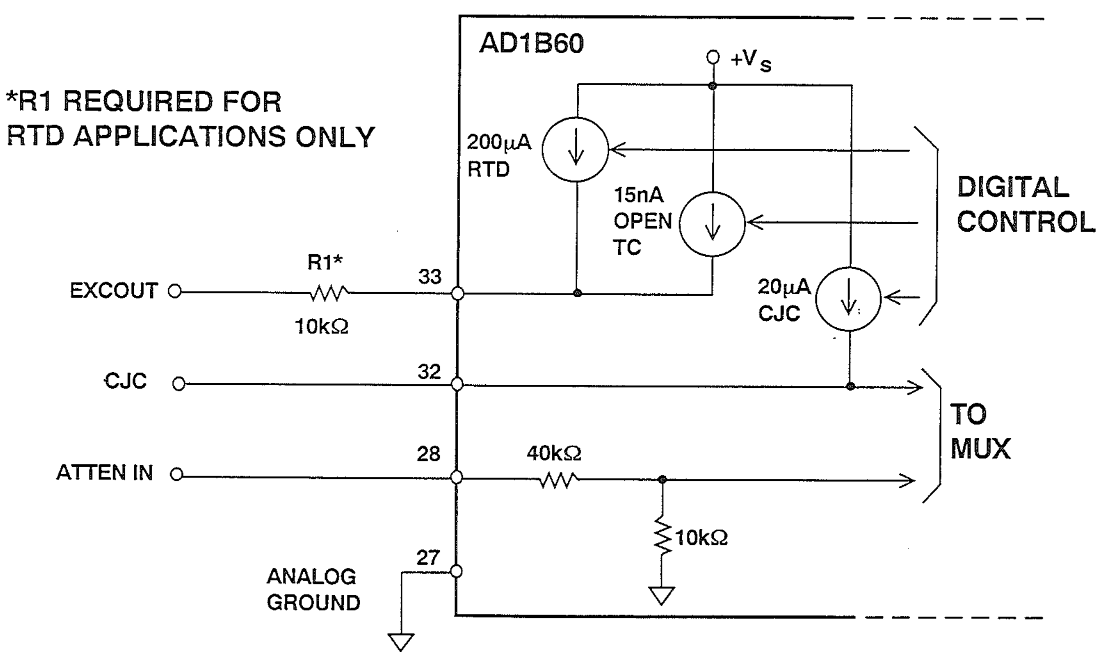
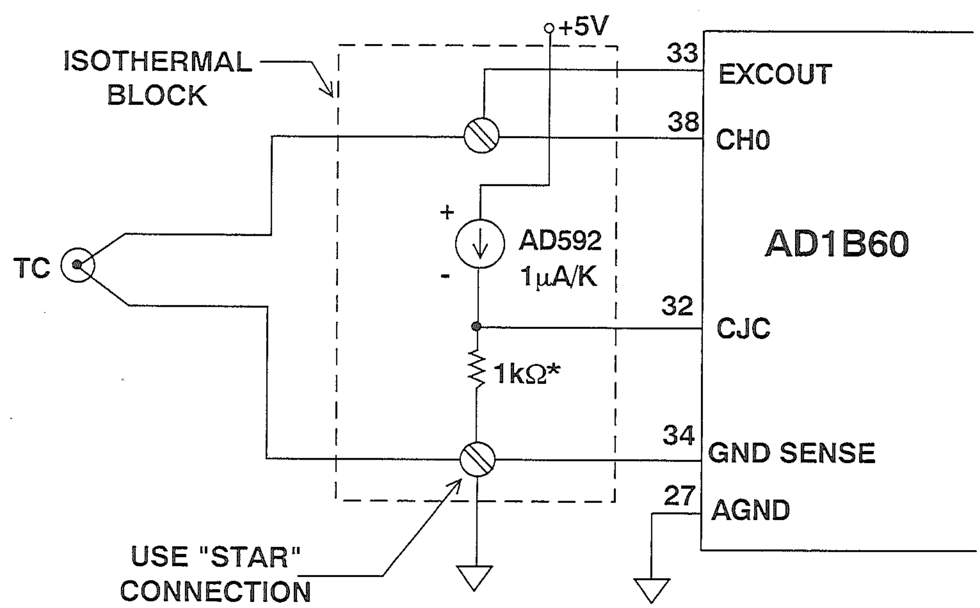


Figure 9.52

Figure 9.53 shows the AD1B60 thermocouple interface using an AD592 temperature sensor for cold junction compensation. The AD592 output current exhibits a temperature coefficient of 1μA/K which develops a voltage across the precision 1kΩ resistor to provide the cold junction compensation.

The AD1B60 can accommodate both 3- and 4-wire RTD applications, as shown in Figure 9.54. For three-wire RTD applications, the internal current excitation is connected directly to the RTD FORCE (+) lead wire through a 10kΩ resistor. This external resistor need not be a precision type — it serves only to reduce self-heating. Without this resistor, a gain error of approximately 0.2°C would be introduced. The RTD's FORCE (+) lead is directly connected to the AD1B60 CH0 input. Only in 3-wire applications is a "star" connection required to minimize additional errors generated by ground loops. The "star" connection is formed by the RTD's FORCE (–) lead wire and the AD1B60's GNDSENSE input connected to the analog ground. In four-wire applications the RTD FORCE(+) lead goes to the 10kΩ resistor, its SENSE(+) to CH1, and its SENSE(-) to CH3. There is no star connection, and FORCE(-) goes to GNDSENSE and AGND.

USING THE AD592 TEMPERATURE SENSOR WITH THE AD1B60 FOR COLD-JUNCTION COMPENSATION



*PRECISION RESISTOR: 0.01% OR BETTER
TCR ≤ 10ppm/°C

Figure 9.53

3- AND 4-WIRE RTD CONNECTIONS TO THE AD1B60

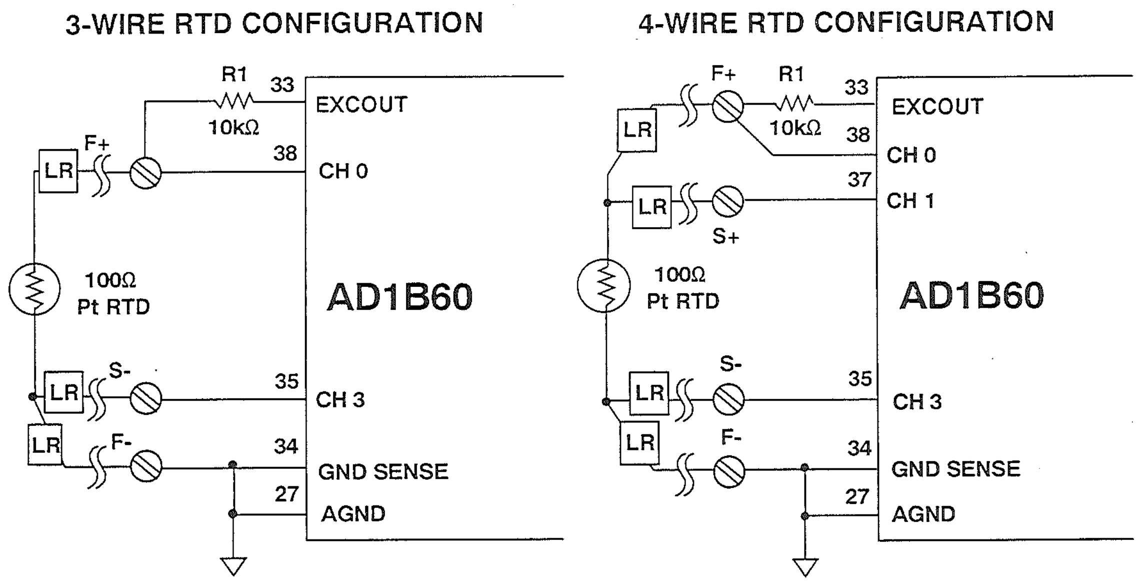


Figure 9.54

A summary of the performance of the AD1B60 (using the built-in linearization algorithms) is given in Figure 9.55.

THE AD1B60'S BUILT-IN ALGORITHMS LINEARIZE THERMOCOUPLE AND RTD OUTPUTS

Thermocouple Type	Temperature Range	Accuracy/Resolution (Typ)
J	0°C ≤ T ≤ 760°C	± 0.25°C / ± 0.15°C
K	0°C ≤ T ≤ 1000°C	± 0.55°C / ± 0.2°C
T	–100°C ≤ T ≤ 400°C	± 0.25°C / ± 0.15°C
E	0°C ≤ T ≤ 1000°C	± 0.2°C / ± 0.1°C
R	500°C ≤ T ≤ 1750°C	± 1.00°C / ± 0.55°C
S	500°C ≤ T ≤ 1750°C	± 1.15°C / ± 0.6°C
B	500°C ≤ T ≤ 1800°C	± 1.15°C / ± 0.7°C

100Ω RTD Type	Temperature Range	Accuracy/Resolution (Typ)
α = 3.85 mΩ/Ω/°C	–200°C ≤ T ≤ 800°C	± 0.2°C/± 0.15°C
α = 3.92 mΩ/Ω/°C	–200°C ≤ T ≤ 800°C	± 0.2°C/± 0.15°C

Figure 9.55

THERMISTOR SIGNAL CONDITIONING

Similar in function to the RTD, thermistors are low-cost temperature-sensitive resistors and are constructed of solid semiconductor materials which exhibit a positive or negative temperature coefficient. Although positive temperature coefficient devices are available, the most commonly used thermistors are those with a negative temperature coefficient. Figure 9.56 shows the resistance-temperature characteristic of a commonly used NTC (Negative Temperature Coefficient) thermistor. The thermistor is highly non-linear and, of the three temperature sensors discussed, is the most sensitive.

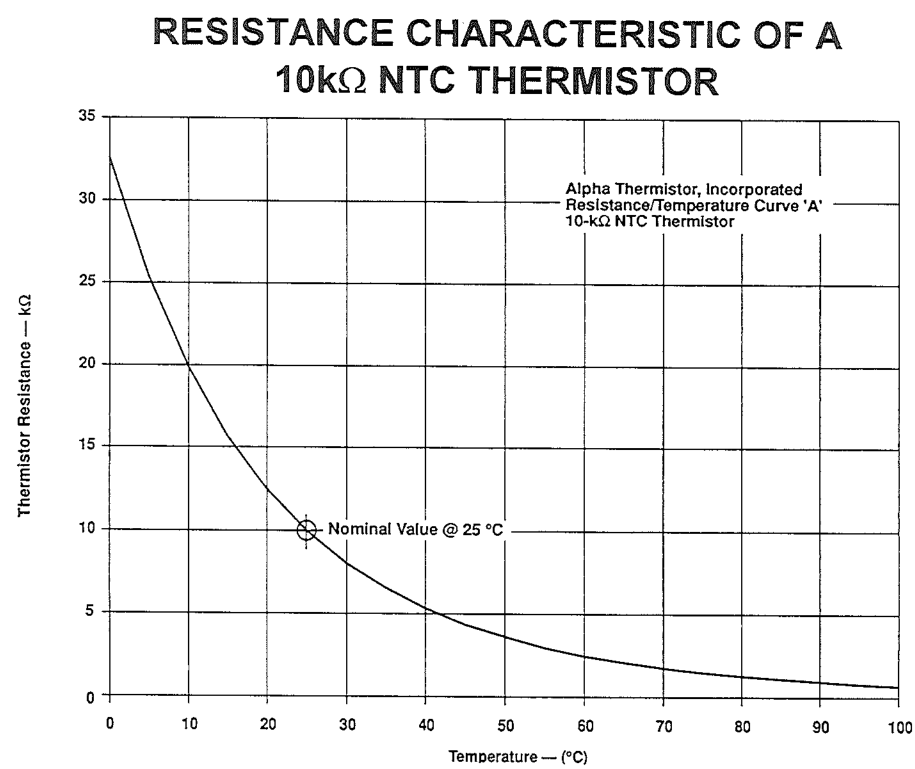


Figure 9.56

The thermistor's high sensitivity (typically, – 44,000ppm/°C at 25°C, as shown in Figure 9.57), allows it to detect minute variations in temperature which could not be observed with an RTD or thermocouple. This high sensitivity is a distinct advantage over the RTD in that 4-wire Kelvin connections to the thermistor are not needed to compensate for lead wire errors. To illustrate this point, suppose a 10kΩ NTC thermistor, with a typical 25°C temperature coefficient of –44,000ppm/°C, were substituted for the 100Ω Pt RTD in the example given earlier, then a total lead wire resistance of 21Ω would generate less than 0.05°C error in the measurement. This is roughly a factor of 500 improvement in error over an RTD.

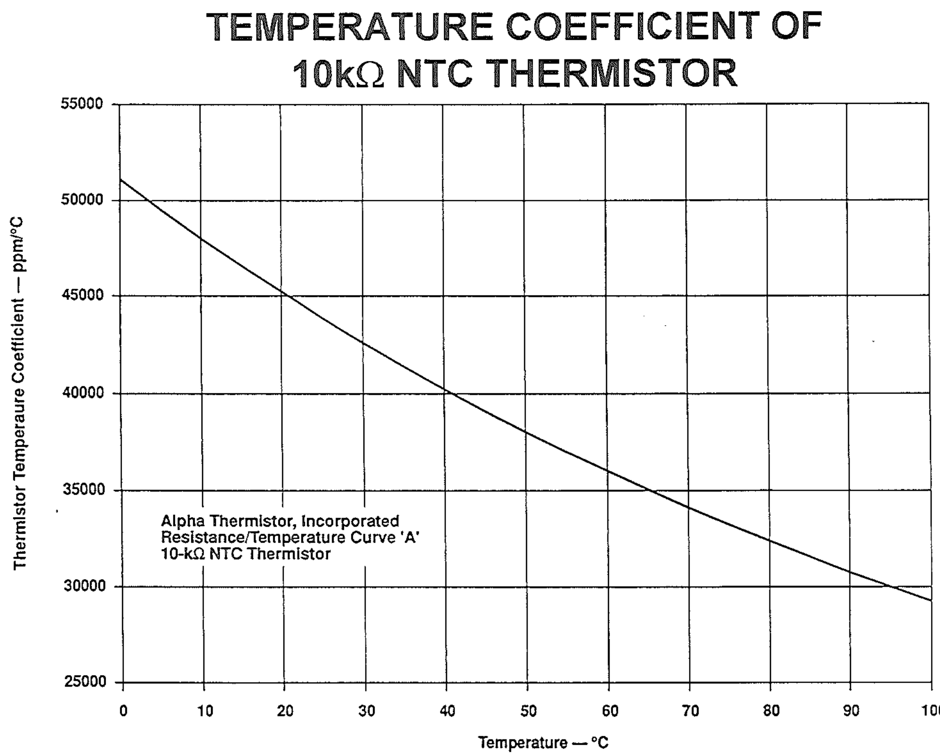


Figure 9.57

However, the thermistor's high sensitivity to temperature does not come without a price. As was shown in Figure 9.57, the temperature coefficient of thermistors does not decrease linearly with increasing temperature as it does with RTDs; therefore, linearization is required for all but the narrowest of temperature ranges. Thermistor applications are limited to a few hundred degrees at best because they are more susceptible to damage at high temperatures. Compared to thermocouples and RTDs, thermistors are fragile in construction and require careful mounting procedures to prevent crushing or bond separation. Although a thermistor's response time is short due to its small size, its small thermal mass makes it very sensitive to self-heating errors.

Thermistors are very inexpensive, highly sensitive temperature sensors. However, we have shown that a thermistor's temperature coefficient varies from –44,000 ppm/°C at 25°C to –29,000ppm/°C at 100°C. Not only is this non-linearity the largest source of error in a temperature measurement, it also limits useful applications to very narrow temperature ranges if linearization techniques are not used.

It is possible to use a thermistor over a wide temperature range only if the system designer can tolerate a lower sensitivity to achieve improved linearity. One approach to linearizing a thermistor is simply shunting it with a fixed resistor. Paralleling the thermistor with a fixed resistor increases the linearity significantly. As shown in Figure 9.58, the parallel combination exhibits a more linear variation with temperature compared to the thermistor itself. Also, the sensitivity of the combination still is high compared to a thermocouple or RTD. The primary disadvantage to this technique is that linearization can only be achieved within a narrow range.

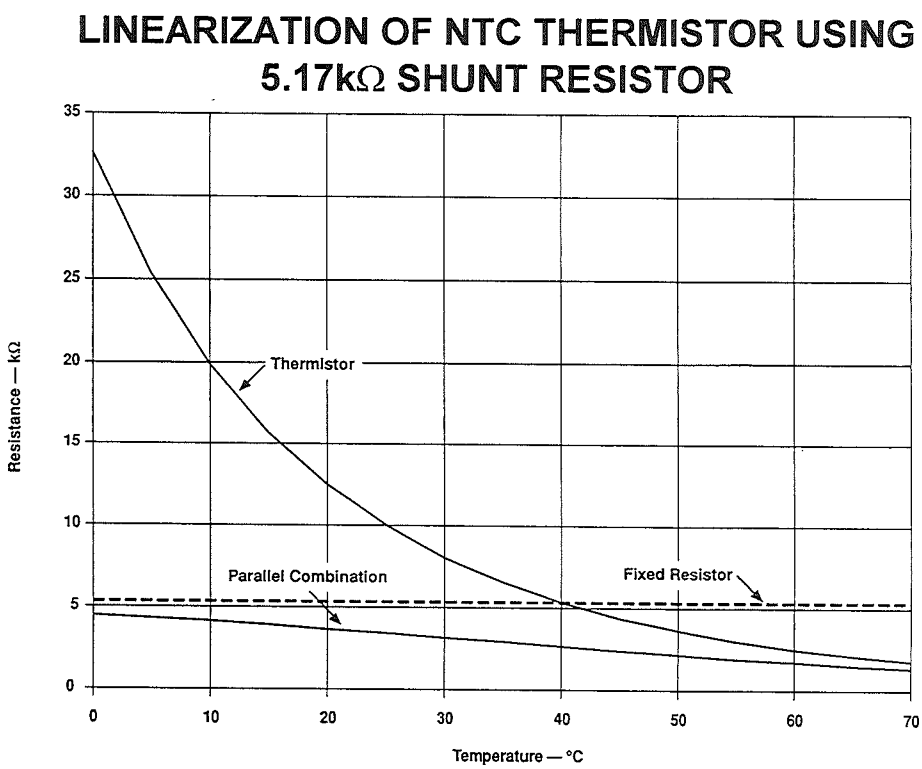


Figure 9.58

The value of the fixed resistor can be calculated from the following equation:

$$R = \frac{RT2 \cdot (RT1 + RT3) - 2 \cdot RT1 \cdot RT3}{RT1 + RT3 - 2 \cdot RT2},$$

where RT1 is the thermistor resistance at T1, the lowest temperature in the measurement range, RT3 is the thermistor resistance at T3, the highest temperature in the range, and RT2 is the thermistor resistance at T2, the midpoint, T2 = (T1 +T3)/2.

For a typical 10 kΩ NTC thermistor, RT1 = 32,650 Ω at 0°C, RT2 = 6,532 Ω at 35°C, and RT3 = 1,752 Ω at 70°C. This results in a value of 5.17kΩ for R. The accuracy needed in the signal conditioning circuitry depends on the linearity of the network. For the example given above, Figure 9.59 illustrates the response of the network and shows a non-linearity of – 2.3°C/ + 2.0°C.

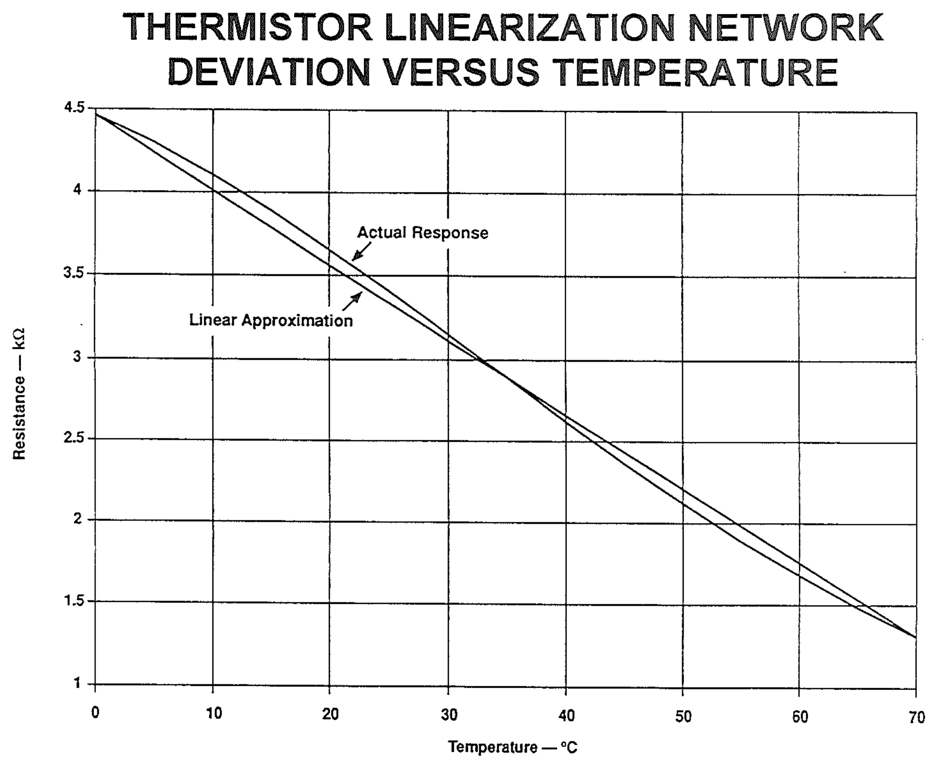
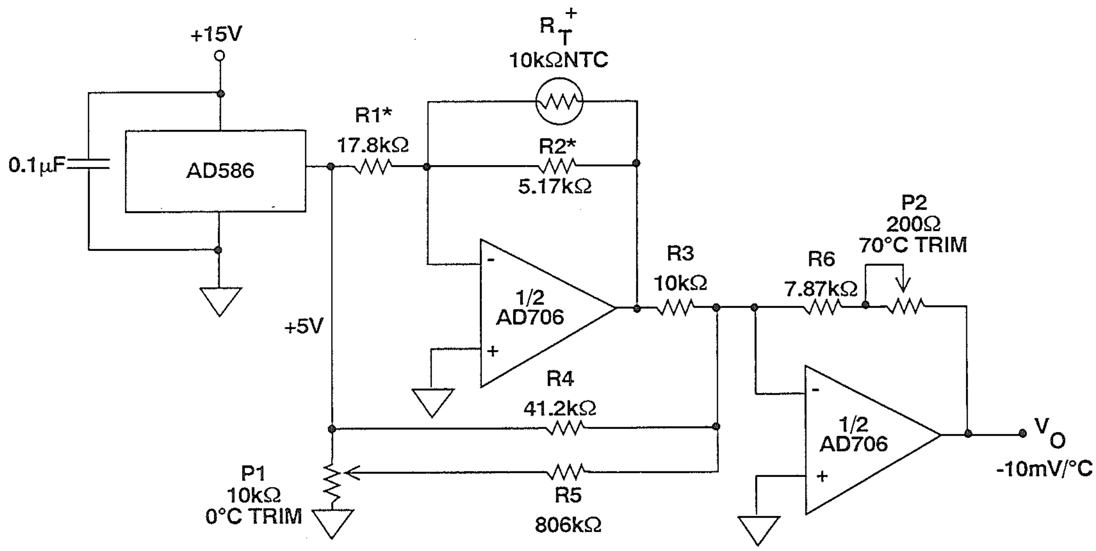


Figure 9.59

Figure 9.60 illustrates a linearized thermistor amplifier designed around the AD706 to measure temperatures over the range of 0°C to 70°C to an accuracy of ± 0.2°C using the network described above. The 10kΩ thermistor and R2 form the feedback network of an inverting amplifier whose gain increases with temperature. High absolute accuracy in R2 is required to maintain the linearity of the network to better than – 2.3°C/+ 2°C. Better linearity can be achieved over a narrower measurement temperature range, but R2 should then be recalculated. In many cases, the thermistor is placed some distance from the signal conditioning circuit. In this case, a 0.1μF capacitor placed across R2 will help to suppress noise.

LINEARIZED THERMISTOR AMPLIFIER



NOTES: + = ALPHA THERMISTOR 13A1002-C3
* = 0.1% IMPERIAL ASTRONICS M015
ALL RESISTORS ARE 1%, 25ppm/°C EXCEPT
R5 = 1%, 100ppm/°C

Figure 9.60

R1 and the AD586 provide a constant current of 281μA to the network such that the thermistor's self-heating error is kept below 0.1°C. Any variation in R1 changes the current through the thermistor network, so absolute accuracy in R1 is important. The second AD706 scales the output of the first amplifier to provide a –10mV/°C output. R4 and the reference generate an offset current such that the output of the second amplifier is 0V at 0°C.

A calibration procedure is required for this circuit to trim resistor errors and the AD706's V_{OS}. A precision decade box temporarily replaces the thermistor. For 0°C trim, the decade box is set to 32.650kΩ, and P1 is adjusted until the circuit output reads 0V. To trim the circuit at full-scale (70°C), the decade box is set to 1.752kΩ and P2 is adjusted until the circuit reads –0.70V.

Since the AD706 exhibits very low input bias currents, measurement errors due to them can be neglected in this application. The high open-loop gain and low ΔV_{OS} keep their error contribution below 0.003°C over a 25°C change in ambient temperature. To help reduce component costs, 50ppm/°C resistors can be substituted for R3, R4, and R6. With this modification, the circuit's accuracy is still better than ± 0.3°C over an ambient temperature range of 20°C to 50°C.

PHOTODIODE TRANSDUCERS

Photodiodes generate a small current which is proportional to the level of illumination. They have many applications ranging from precision light meters to high-speed fiber optic receivers.

PHOTODIODE APPLICATIONS

- Optical: Light Meters, Auto-Focus, Flash Controls
- Medical: CAT Scanners (X-Ray Detection), Blood Particle Analyzers
- Automotive: Headlight Dimmers, Twilight Detectors
- Communications: Fiber Optic Receivers
- Industrial: Bar Code Scanners, Position Sensors, Laser Printers

Figure 9.61

The equivalent circuit for a photodiode is shown in Figure 9.62. One of the standard methods for specifying the sensitivity of a photodiode is to state its short circuit photocurrent (I_{SC}) at a given light level from a well defined light source. The most commonly used source is an incandescent tungsten lamp running at a color temperature of 2850K. At 100 fc (foot-candles) illumination (approximately the light level on an overcast day), the short circuit photocurrent is usually in the picoamps to hundreds of microamps range for small area (less than $1mm^2$) diodes.

The short circuit photocurrent has a very linear relationship to illumination over 6 to 9 decades of light intensity, and is therefore often used as a measure of absolute light levels. The open circuit forward voltage across the photodiode varies logarithmically with light level, but because of its large temperature coefficient, the diode voltage is seldom used as an accurate measure of light intensity.

The shunt resistance is of order of 1000MΩ at room temperature, and decreases by a factor of 2 for every 10°C rise in temperature. Diode capacitance is a function of junction area and the diode bias voltage. A value of 50pF at zero bias is typical for small area diodes.

PHOTODIODE EQUIVALENT CIRCUIT

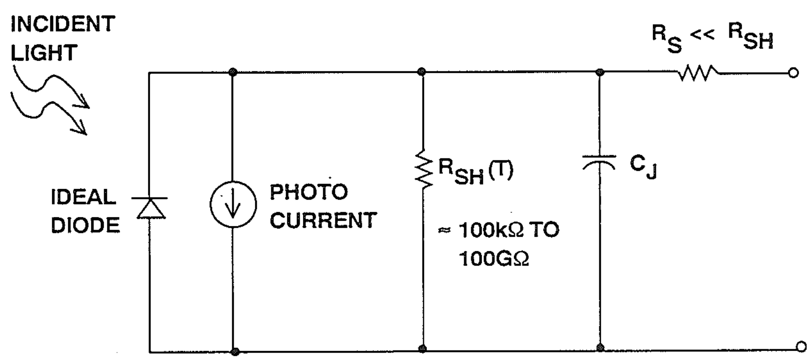
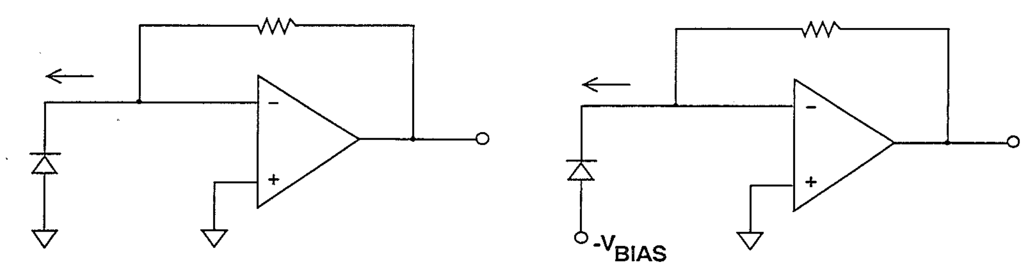


Figure 9.62

PHOTODIODE MODES OF OPERATION



PHOTOVOLTAIC	PHOTOCONDUCTIVE
· Zero Bias	· Reverse Bias
· No Dark Current	· Dark Current Exists
· Precision Applications	· High Speed Applications
· Low Noise (Johnson)	· Higher Noise (Johnson + Shot)

Figure 9.63

Photodiodes may be operated with zero bias (*photovoltaic* mode) or reverse bias (*photoconductive* mode) as shown in Figure 9.63. The most precise linear operation is obtained in the photovoltaic mode, while higher switching speeds are realizable when the diode is operated in the photoconductive mode. Under reverse bias conditions, a small leakage current called *dark current* will flow even when there is no illumination. There is no dark current in the photovoltaic mode. In the photovoltaic mode, the diode noise is basically the thermal noise of the shunt resistance. In the photoconductive mode, shot noise is an additional source of noise. Photodiodes are usually optimized during design for use in either the photovoltaic mode or the photoconductive mode, but not both.

Figure 9.64 shows the photosensitivity for a small photodiode (Silicon Detector Part Number SD-020-12-001). The specifications of the diode are summarized in Figure 9.65.

SHORT CIRCUIT CURRENT VERSUS LIGHT INTENSITY FOR PHOTODIODE (PHOTOVOLTAIC MODE)

ENVIRONMENT	ILLUMINATION (fc)	SHORT CIRCUIT CURRENT
Direct Sunlight	1000	30mA
Overcast Day	100	3mA
Twilight	1	0.03mA
Full Moonlit Night	0.1	3000pA
Clear Night / No Moon	0.001	30pA

Figure 9.64

PHOTODIODE SPECIFICATIONS
Silicon Detector Part Number SD-020-12-12-001

- Area: 0.2mm^2
- Capacitance: 50pF
- Shunt Resistance at 25°C: 1000 megohms
- Maximum Linear Output Current: 40mA
- Response Time: 12ns
- Photosensitivity: 0.03μA / fc

Figure 9.65

PHOTODIODE PREAMP CIRCUIT CONSIDERATIONS

A convenient way to convert the photodiode current into a usable voltage is to use an op amp as a current-to-voltage converter as shown in Figure 9.66. The diode bias is maintained at zero volts by the virtual ground of the op amp, and the short circuit current is converted into a voltage. If we wish to operate at maximum sensitivity, we must be able to detect a diode current of 30pA. This implies that the feedback resistor must be very large. For example, 1000MΩ will yield a voltage of 30mV for this current. Larger resistor values are impractical, so we shall use 1000MΩ for the most sensitive range. This will give an output voltage range from 10mV for 10pA of diode current to 10V for 10nA, a range of 60dB. For higher values of light intensity, the gain of the circuit must be reduced by using a smaller feedback resistor. With the maximum sensitivity, we should easily be able to distinguish between the light intensity on a clear moonless night (0.001fc) and that of a full moon (0.1fc)!

SIMPLIFIED OP-AMP CURRENT-TO-VOLTAGE CONVERTER

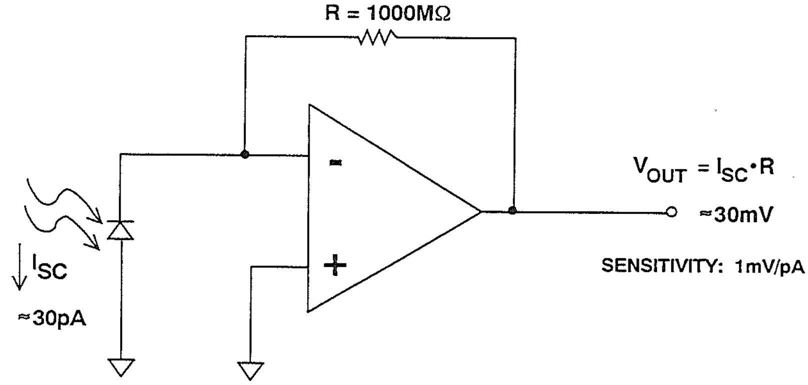


Figure 9.66

Notice that we have chosen to obtain as much gain as possible from one stage, rather than cascading two stages. This is in order to maximize the signal-to-noise ratio (SNR). If we halve the feedback resistor value, the signal level decreases by a factor of 2, while the noise due to the feedback resistor (Noise Voltage = $\sqrt{4kTR \cdot \text{Bandwidth}}$) decreases by $\sqrt{2}$, which reduces the SNR by 3dB, assuming the closed loop bandwidth remains constant. A detailed analysis (Reference 3, Section 3) shows that resistors are one of the largest contributors to the overall output noise.

PRECAUTIONS FOR PICOAMPERE CIRCUITS

Since the diode current is measured in picoamperes, extreme attention must be given to potential leakage paths in the actual circuit. Two parallel conductor stripes on a high-quality well-cleaned epoxy-glass PC board 0.05 inches apart running parallel for 1 inch have a leakage resistance of approximately 10^{11} ohms at +125°C (Reference 9, p.293). If there is 15 volts between these runs, there will be a current flow of 150pA.

The critical leakage paths for the photodiode circuit are enclosed by the dotted lines in Figure 9.67. The feedback resistor should be of thin-film on ceramic, or glass with glass insulation. The compensation capacitor across the feedback resistor should have a polypropylene or polystyrene dielectric. All connections to the summing junction should be kept short. If a cable is used to connect the photodiode to the preamp, it should be kept as short as possible and have Teflon insulation.

Guard rings (on both sides of the PC board) should be used around the inverting input pin of the op amp as shown in Figure 9.68. The case ground of the op amp (usually Pin 8) should also be connected to the grounded guard ring. Maintaining the guard ring at the same potential as the inverting input minimizes leakage current due to PC board resistance.

LEAKAGE CURRENT PATHS

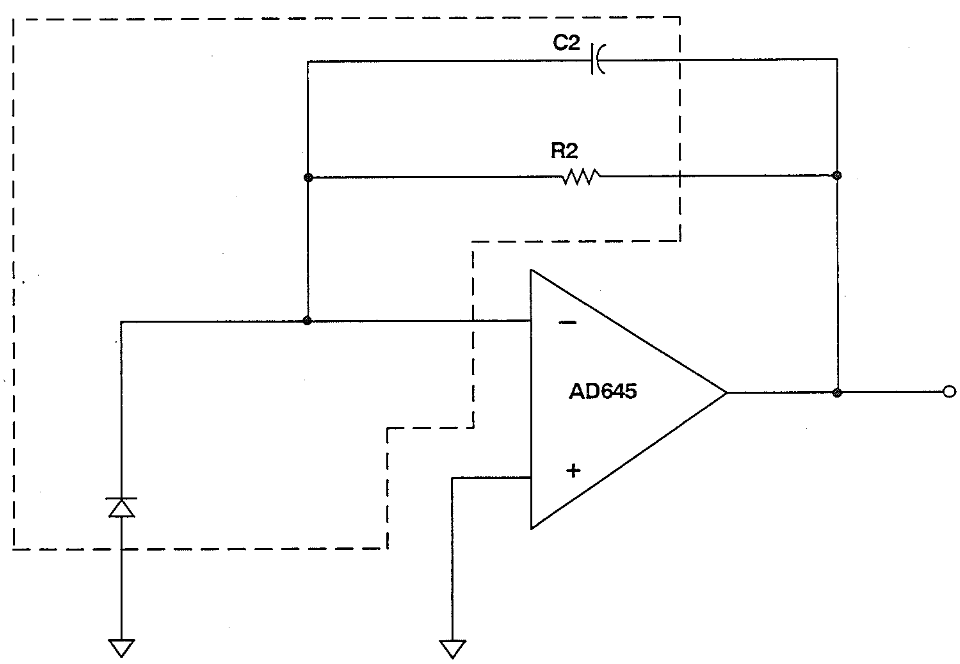


Figure 9.67

PC BOARD LAYOUT FOR GUARDING TO-99 PACKAGE

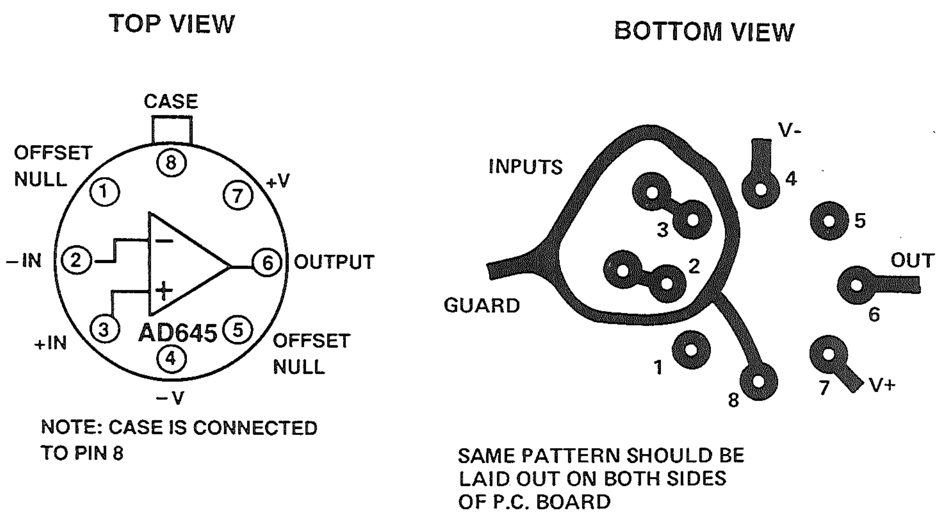


Figure 9.68

Ideally, all connections to the summing input of the op amp should be made to a virgin Teflon standoff insulator ("Virgin" Teflon is a solid piece of new Teflon material which has been machined to shape and has not been welded together from powder or grains). If mechanical and manufacturing considerations allow, the inverting input pin of the op amp should be soldered directly to the Teflon standoff (see Figure 9.69) rather than going through a hole in the PC board. The PC board itself must be cleaned carefully and then sealed against humidity and dirt using a high quality conformal coating material.

In addition to minimizing leakage currents, the entire circuit should be well shielded with a grounded metal shield to prevent stray signal pickup. Details regarding proper grounding, shielding, and noise reduction techniques are given in References 13 and 14 at the end of this section and in Section 11.

A VIRGIN TEFLON STANDOFF INSULATOR HAS MUCH LOWER LEAKAGE THAN A PCB TRACK

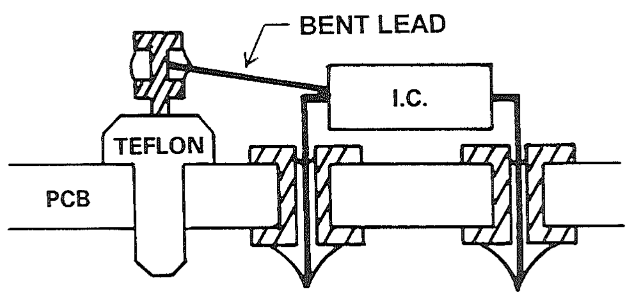


Figure 9.69

AMPLIFIER SELECTION

If we wish to measure accurately photodiode currents of tens of picoamps, the bias current of the op amp should be no more than a few picoamps. This narrows the choice considerably. The industry-standard OP-07 is an ultra-low V_{OS} (10µV) bipolar op amp , but its bias current is 4nA (4000pA!). Even super-beta bipolar op amps with bias current compensation (such as the OP-97) have bias currents of 100µA at room temperature. For this reason, an FET-input electometer-grade op amp such as the AD549 is required for the photodiode preamp. The bias current of the AD549 is *guaranteed* less than 60fA at T_A of 25°C (less than one electron every 3µs) but, as with all FET-input op amps, the bias current doubles with every 10°C increase of temperature, so by T_A = 125°C, I_b will be as much as 60pA.

There are many applications for photodiodes which require less sensitivity to illumination but more bandwidth (see Figure 9.70). The equivalent circuit shown in Figure 9.71 is useful in making the proper tradeoffs. The photodiode is operated in the photoconductive mode (reversed biased) in order to minimize the junction capacitance C_D. Because of the reverse bias, a small amount of *dark current* flows when there is no illumination. The op amp should be chosen for low bias current, high gain-bandwidth product (f_u), and low input capacitance C_{IN}. Notice that the total op amp input capacitance, C_1, is equal to the diode capacitance C_D plus the input capacitance of the op amp C_{IN}. The feedback capacitor C_2 is required in order to compensate for the op-amp input capacitance. It should be as small

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as possible (to provide the maximum signal bandwidth) and yet large enough to give adequate phase margin. It is usually optimized in the circuit for best pulse response. The equations are summarized in Figure 9.72 and are derived in Reference 3, Section 3. Finally, the effects of bias current and noise should be evaluated.

APPLICATIONS OF WIDE BANDWIDTH PHOTODIODE CIRCUITS

- Ring Laser Gyro Systems
- Bar Code Readers
- Fast Scanners
- Document Scanners
- Fax Machines
- Fiber Optic Receivers

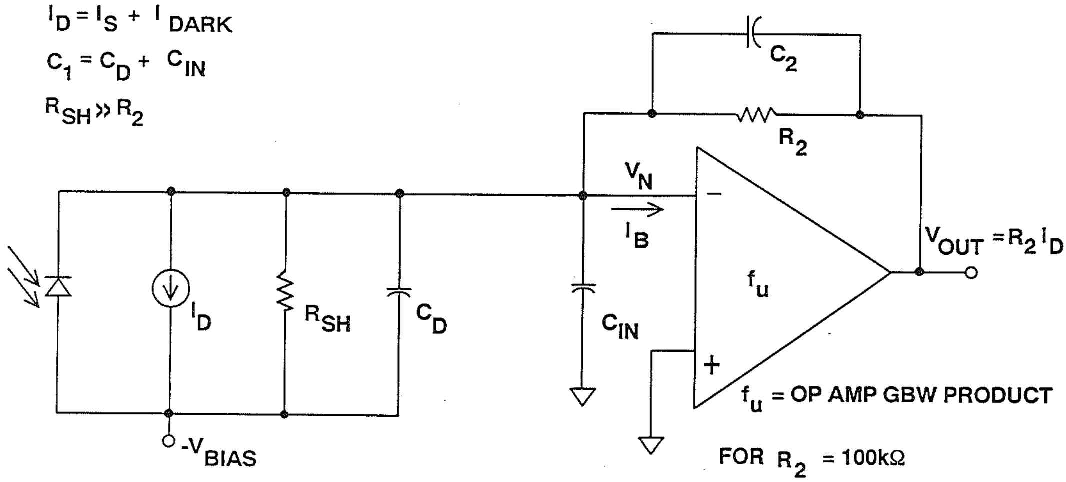
Figure 9.70

HIGH BANDWIDTH PHOTODIODE PREAMP EQUIVALENT CIRCUIT

$I_D = I_S + I_{DARK}$

$C_1 = C_D + C_{IN}$

$R_{SH} \gg R_2$



$$\text{SIGNAL BW} = \frac{1}{2\pi R_2 C_2}$$

FOR $R_2 = 100k\Omega$

$I_D = 100\mu A$

$V_{OUT} = 10V$

Figure 9.71

CONDITIONS FOR MAXIMALLY FLAT SIGNAL FREQUENCY RESPONSE WITH APPROXIMATELY 65° PHASE MARGIN AND 5% STEP FUNCTION OVERSHOOT

- $C_2 \approx 2\sqrt{\dfrac{C_1}{2\pi R_2 f_u}}$, and
- Signal Bandwidth $\approx \dfrac{1}{2}\sqrt{\dfrac{f_u}{2\pi R_2 C_1}}$.

Figure 9.72

GENERALIZED OP AMP SELECTION FLOW CHART FOR HIGH SPEED PHOTODIODE PREAMP

- SIGNAL BW ~ $\sqrt{\dfrac{f_u}{R_2 C_1}}$, f_u = Op Amp Unity Gain Bandwidth
  C_1 = Input Capacitance = $C_d + C_{in}$.
- Minimize Value of R_2 Based on Gain Requirements
- Try and make op amp input capacitance, $C_{in} \leq C_d$
- Maximize f_u/C_{in}, Minimize Voltage and Current Noise
- Calculate C_2 and Signal Bandwidth
- Examine Bias Current and Noise Effects on Circuit

Figure 9.73

9

Figure 9.74 shows a photodiode preamp circuit with a bandwidth just above 1MHz. The op-amp chosen is the AD843 FET-input type (gain-bandwidth product 34MHz, 6pF input capacitance). Using the circuit parameters $C_1 = C_D + C_{IN} = 4 + 6 = 10$pF, $f_u = 34$MHz, $R_2 = 100$kΩ, we find that $C_2 = 1.4$ pF using the equations in Figure 9.72. In practice, C_2 should be a 2pF low leakage variable capacitor so that the circuit may be optimized for the best compromise between frequency and pulse response. The 100kΩ resistors are made up of three 33.2kΩ film resistors in series to minimize stray capacitance.

1 MHz PHOTODIODE PREAMP USING AD843 OP AMP

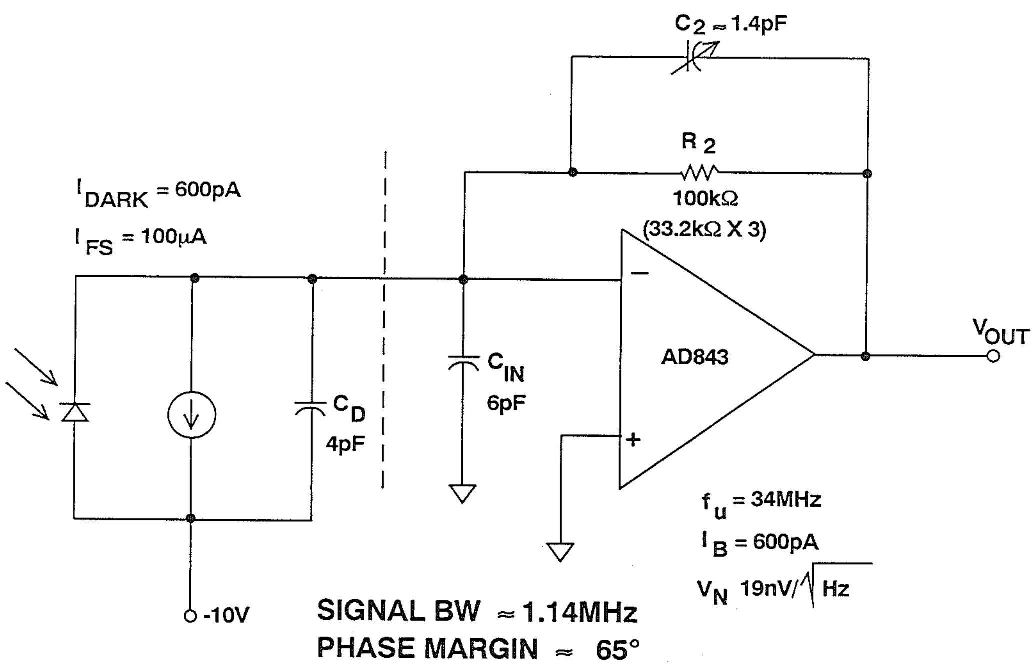


Figure 9.74

The noise analysis of the circuit is somewhat tedious and is given in Reference 2, Section 3. The total output noise of approximately 346µV rms is dominated by the input voltage noise integrated over the region of high frequency peaking which is typical of second-order circuits.

HIGH SPEED FIBER OPTIC RECEIVERS

When the primary function of the photodiode preamp is to amplify digital data from a fiber optic link, even more tradeoffs become possible to achieve higher speeds. In a fiber optic data transmission system, such as the one shown in Figure 9.75, the primary purpose of the preamp is to amplify the photodiode current to a voltage level which is sufficient to drive the input of a threshold comparator. In these applications, the data is coded in such a manner that the average duty cycle of the data is always 50% regardless of the actual bit pattern. (A Manchester coding scheme is one way to accomplish this.) Since the average duty cycle of the data stream is always 50%, ac coupling is possible, and the need for wideband precision op amps is eliminated. The preamp does not have to be a traditional dc-coupled op amp with feedback, and may be a low noise open loop GaAs "gain block". The photodiode, preamp, and the comparator are often fabricated on the same substrate in a hybrid package in order to minimize parasitics. Data transmission rates of much greater than 100MHz are possible using this approach.

HIGH SPEED FIBER OPTIC DATA RECOVERY

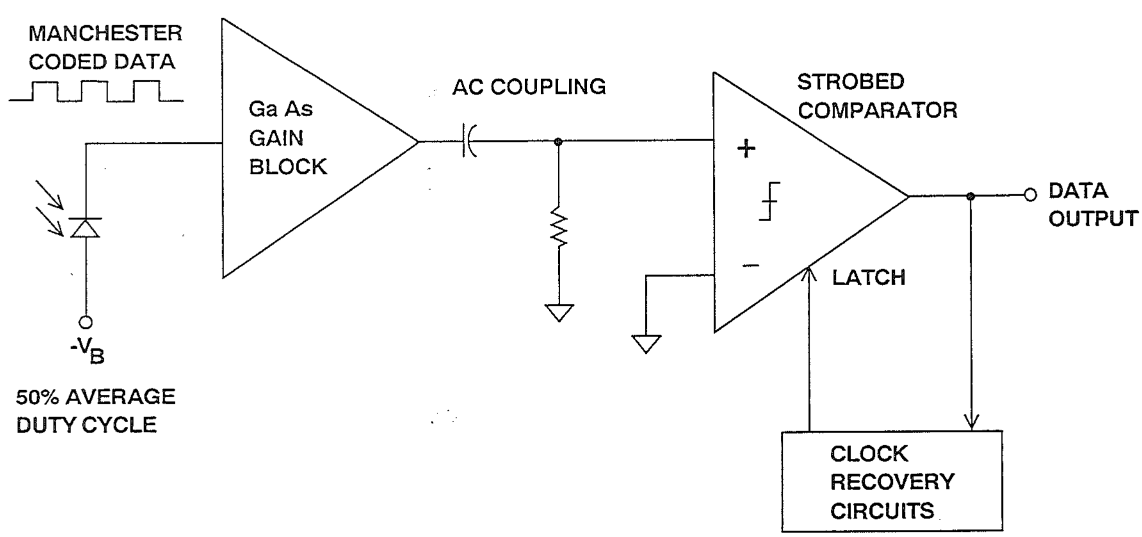
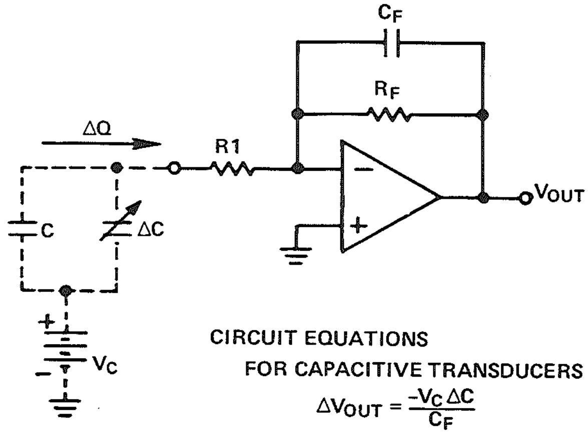


Figure 9.75

HIGH-IMPEDANCE CHARGE OUTPUT TRANSDUCERS

High impedance transducers such as hydrophones and some accelerometers require an amplifier which converts a transfer of charge into a change of voltage. Because of the high dc output impedance of these devices, appropriate buffers are required. The basic circuit of an inverting charge sensitive amplifier is shown in Figure 9.76. There are two types of charge transducers: capacitive and charge-emitting. In a capacitive transducer, the voltage across the capacitor (V_c) is held constant. The change in capacitance, ΔC, produces a change in charge, $\Delta Q = \Delta C V_c$. This charge is transferred to the op amp output as a voltage, $\Delta V_{out} = -\Delta Q/C_f = -\Delta C V_c/C_f$.

CHARGE-SENSITIVE AMPLIFIER



CIRCUIT EQUATIONS

FOR CAPACITIVE TRANSDUCERS

$$\Delta V_{OUT} = \frac{-V_C \Delta C}{C_F}$$

FOR CHARGE-EMITTING TRANSDUCERS

$$\Delta V_{OUT} = \frac{-\Delta Q}{C_F}$$

LOWER CUTOFF FREQUENCY (-3dB)

$$f_{o1} = \frac{1}{2\pi R_F C_F}$$

UPPER CUTOFF FREQUENCY (-3dB)

$$f_{o2} = \frac{1}{2\pi R_1 C}$$

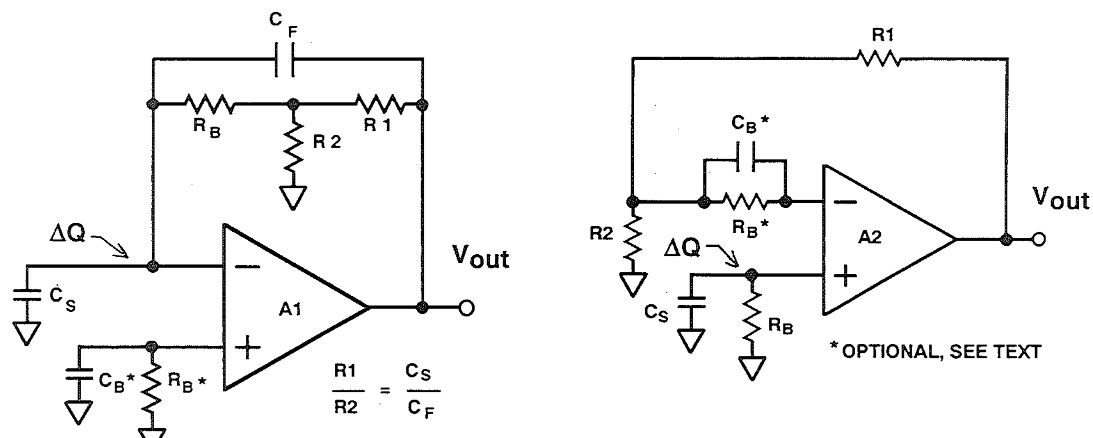
Figure 9.76

Charge-emitting transducers produce an output charge, ΔQ, and their output capacitance remains constant. This charge would normally produce an open-circuit output voltage at the transducer output equal to $\Delta Q/C$. However, since the voltage across the transducer is held constant by the virtual ground of the op amp (R1 is usually small), the charge is transferred to capacitor C_f, producing an output voltage $\Delta V_{out} = -\Delta Q/C_f$.

Figure 9.77 shows two ways to buffer and amplify the output of a charge output transducer. Both require using an amplifier which has a very high input impedance, such as the AD745. The AD745 provides both low voltage and low current noise. This combination makes this device particularly suitable in applications requiring very high charge sensitivity, such as capacitive accelerometers and hydrophones.

CHARGE AMPLIFIER CONFIGURATIONS

CHARGE OUTPUT MODE **VOLTAGE OUTPUT MODE**



*OPTIONAL, SEE TEXT

$$\Delta V_{out} = \frac{\Delta Q}{C_S}\left(1 + \frac{R1}{R2}\right)$$

$$\Delta V_{out} = \frac{-\Delta Q}{C_F}$$

Figure 9.77

The first circuit in Figure 9.77 uses the op amp in the inverting mode. Amplification depends on the principle of conservation of charge at the inverting input of amplifier A1. The charge on capacitor C_S is transferred to capacitor C_F, yielding an output voltage of $\Delta Q/C_f$. The amplifier's input voltage noise will appear at the output amplified by the ac noise gain of the circuit, $1 + C_S/C_f$.

The second circuit shown in Figure 9.77 is simply a high impedance follower with gain. Here the noise gain $(1 + R1/R2)$ is the same as the gain from the transducer to the output. Resistor R_B , in both circuits, is required as a dc bias current return.

To maximize dc performance over temperature, the source resistances should be balanced on each input of the amplifier. This is represented by the optional resistor R_B shown in Figure 9.77. For best noise performance, the source capacitance should also be balanced with the capacitor C_B. In general, it is good practice to balance the source *impedances* (both resistive and reactive) as seen by the inputs of a precision low noise BiFET amplifiers such as the AD743/AD745. Balancing the resistive components will optimize dc performance over temperature because balancing will mitigate the effects of any bias current errors. Balancing the input capacitance will minimize ac response errors due to the amplifier's non-linear common mode input capacitance, and, as shown in Figure 9.78, noise performance will be optimized. Figure 9.78 shows the required external components for both inverting and noninverting configurations. For values of C_B greater than 300pF, there is a diminishing impact on noise, and C_B can then be simply a large mylar bypass capacitor of 0.01µF or greater.

BALANCING SOURCE IMPEDANCES MINIMIZES EFFECTS OF BIAS CURRENTS AND REDUCES INPUT NOISE

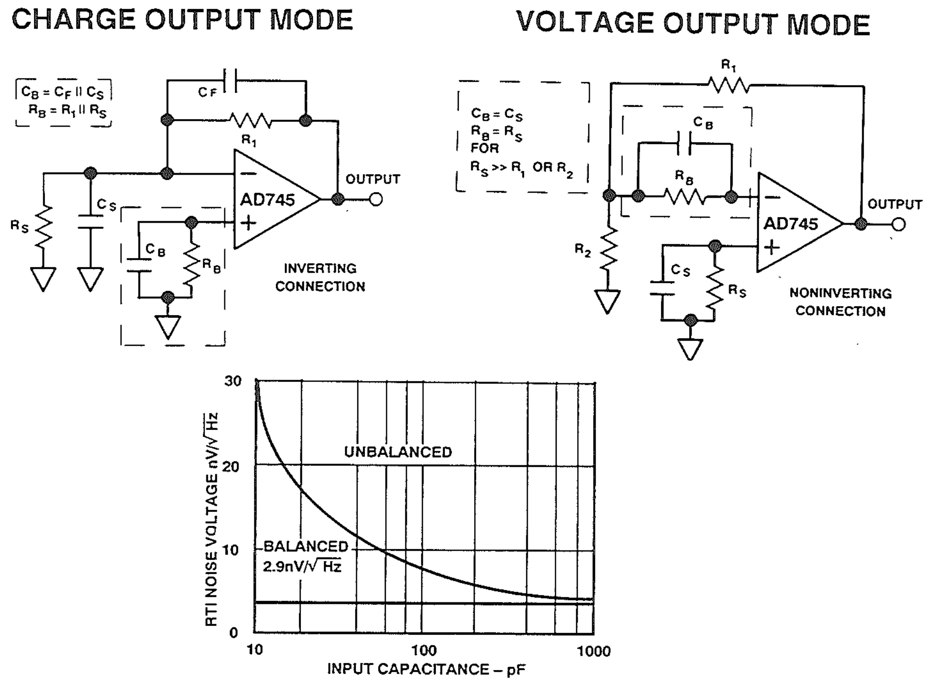
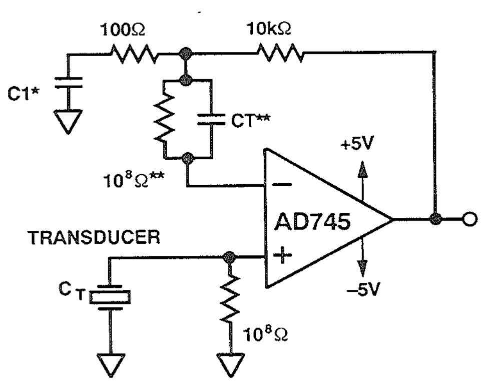


Figure 9.78

Figure 9.79 shows a piezoelectric transducer amplifier connected in the voltage-output mode. Reducing the power supplies to ±5V reduces the effects of bias current in two ways: first, by lowering the total power dissipation and, second, by reducing the basic gate-to-junction leakage current. The addition of a clip-on heat sink will further limit the internal junction temperature rise.

Without the ac coupling capacitor C1, the amplifier will operate over a range of –40°C to +85°C. If the optional ac coupling capacitor C1 is used, the circuit will operate over the entire –55°C to +125°C temperature range, but dc information is lost.

A GAIN OF 100 PIEZOELECTRIC TRANSDUCER AMPLIFIER



*OPTIONAL DC BLOCKING CAPACITOR
**OPTIONAL, SEE TEXT

- **±5V Power Supplies Reduce I_b for 0°C to +85°C Operation**
- **C1 Allows -55°C to +125°C Operation**

Figure 9.79

HYDROPHONES

Interfacing the outputs of highly capacitive transducers such as hydrophones, some accelerometers, and condenser microphones to the outside world presents many problems. Previously designers had to use costly hybrid amplifiers consisting of discrete low-noise JFETs in front of conventional op amps to achieve the low levels of voltage and current noise required by these applications. The AD743 and AD745 monolithic amplifiers allow designers to achieve new levels of system integration and performance.

In sonar applications, a piezo-ceramic cylinder is commonly used as the active element in the hydrophone as is shown in Figure 9.80. A typical cylinder has a nominal capacitance of 6,000pF with a series resistance of 10Ω. The output impedance is typically $10^8\Omega$ or 100MΩ.

Since the hydrophone signals are inherently ac in nature, noise is the overriding concern among sonar system designers. The noise floor of the hydrophone and the hydrophone preamplifier together limit the sensitivity of the system and therefore the overall usefulness of the hydrophone. Typical hydrophone bandwidths are in the range 1kHz to 10kHz. The AD743 and AD745 op amps, with their low noise figures of $2.9\text{nV}/\sqrt{\text{Hz}}$ and high input impedance of $10^{10}\Omega$ (or 10GΩ) are ideal for use as hydrophone amplifiers.

The AD743 and AD745 are companion amplifiers with different levels of internal compensation. The AD743 is internally compensated for unity gain. The AD745, stable for noise gains of 5 or greater, has a much higher bandwidth and slew rate. This makes the AD745 useful as a high-gain preamplifier where it provides both high gain and wide bandwidth. The AD743 and AD745 have extremely low levels of distortion: less than 0.0003% and 0.0002% (at 1kHz), respectively.

Hydrophone amplifiers are usually connected in the voltage-out mode rather than charge-out mode. Both modes are illustrated in Figure 9.80. The circuits shown in Figure 9.81 can be used to amplify the output of a typical hydrophone connected in the voltage-out mode.

HYDROPHONE AMPLIFIERS

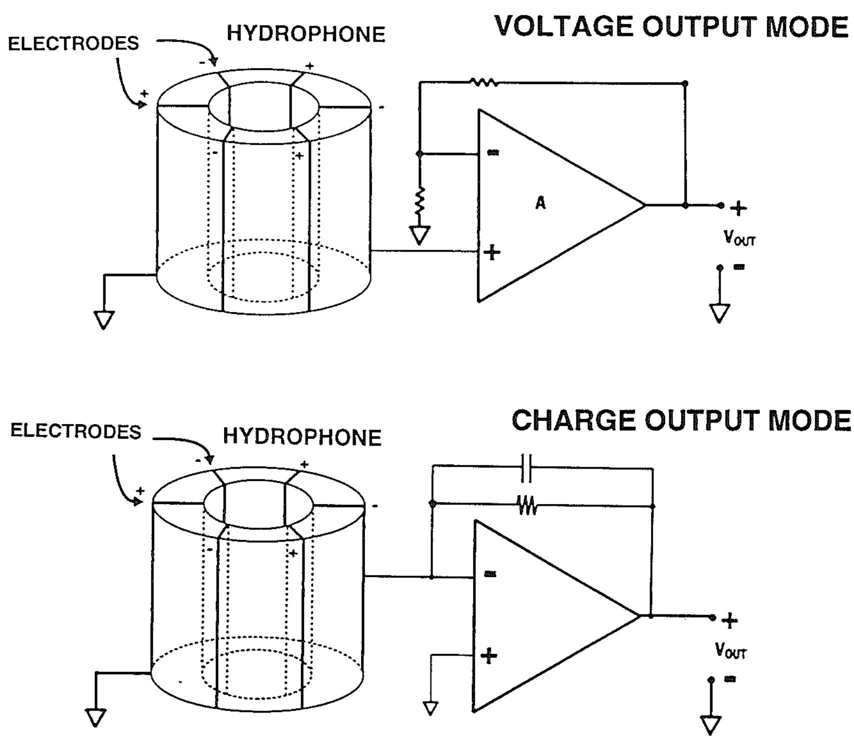


Figure 9.80

HYDROPHONE AMPLIFIER CONFIGURATIONS

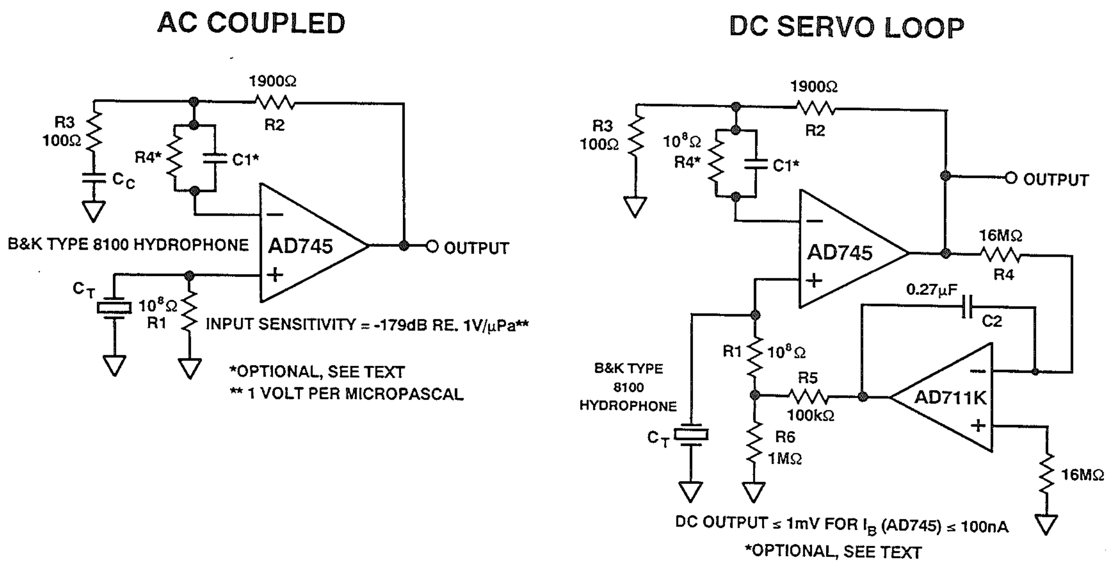


Figure 9.81

If the optional ac coupling capacitor, C_c, is used, the circuit on the left-hand side of Figure 9.81 will have a low frequency cutoff (F_L) :

$$F_L = \frac{1}{2\pi C_c 100\Omega}$$

which is determined by the time constant R_3C_c. With the ac coupling capacitor, the gain at dc is 1, and the gain above the low frequency cutoff will be a maximum of (1 + R2/R3) or 26dB.

A second type of hydrophone amplifier circuit is shown in the right-hand diagram in Figure 9.81. It uses a DC servo loop to force the dc output to 0V, within the input offset limits of the AD711 op amp, thereby maintaining full dynamic range. Power supply voltages should be reduced and heatsinking used to keep the input bias current of the AD745 below 100nA over the full military temperature range. For a smooth low frequency response, the time constant of R4 and C1 should be at least 10 times larger than that of R1 and C_T.

The transducer shown has a source capacitance of 7500pF. For smaller transducer capacitances (≤ 300pF), lowest noise can be achieved by adding a parallel RC network (R4 = R1, C_T = C1) in series with the inverting input of the AD745. As has been previously described, balancing the source impedances (both resistive and reactive) is good practice.

ACCELEROMETER AMPLIFIERS

Two of the most common charge-out transducers are hydrophones and accelerometers. Precision accelerometers are typically calibrated for a charge output measured in picocoulombs (pC) per g, where g is the Earth's gravitational constant ($9.81m/s^2$). Figure 9.82 shows two ways to configure the AD745 as a low noise charge amplifier for use with piezoelectric accelerometers. The output voltage, ΔV_{out}, of these circuits will be determined by the value of capacitor, C1, and the transducer charge output ,ΔQ, or

$$\Delta V_{out} = \frac{\Delta Q}{C1} .$$

HIGH PERFORMANCE ACCELEROMETER AMPLIFIERS

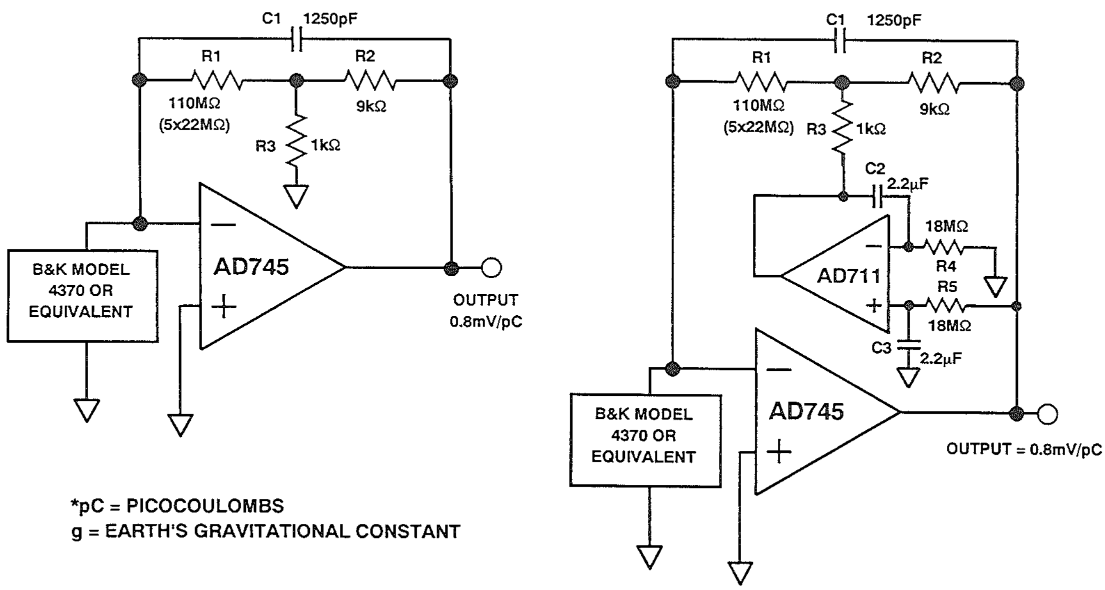


Figure 9.82

The ratio of capacitor C1 to the internal capacitance C_T of the transducer determines the ac noise gain of the circuit, $1 + C_T/C1$. The amplifier's voltage noise will appear at its output amplified by this amount. The low frequency bandwidth of these circuits will be dependent on the value of resistor R1. If a "T" network is used, the effective value is $R1(1+R2/R3)$.

The addition of the dc servo loop shown in Figure 9.82 can be used to assure a dc output less than 10mV without the need for a large compensating resistor when dealing with bias currents as large as 100nA. For low frequency performance, the time constant of the servo loop (R4C2 = R5C3) should be:

$$\text{Time Constant} = R4C2 = R5C3 \geq 10R1\left(1 + \frac{R2}{R3}\right)C1 .$$

MONOLITHIC ACCELEROMETERS

The ADXL50 is a complete acceleration measurement system on a single monolithic IC. It contains a polysilicon surface-micro-machined sensor and signal conditioning circuitry. The ADXL50 is capable of measuring both positive and negative acceleration to a maximum level of ±50g.

Figure 9.83 is a simplified view of the acceleration sensor at rest. The differential capacitor sensor consists of independent fixed plates and a movable "floating" central plate which deflects in response to acceleration. The two capacitors are connected in series, forming a capacitive divider with a common movable central plate. A force balance technique counters any deflection due to acceleration and servos the sensor back to its 0g position.

Figure 9.84 shows the sensor responding to acceleration. When this occurs, the common central plate or "beam" moves closer to one of the fixed plates while moving further from the other. The sensor's fixed capacitor plates are driven differentially by a 1MHz square wave: the two square wave amplitudes are equal but are 180° out of phase. When at rest, the values of the two capacitors are the same, and therefore the voltage output at their electrical center (i.e., at the center plate) is zero.

SIMPLIFIED DIAGRAM OF THE ADXL50 SENSOR AT REST

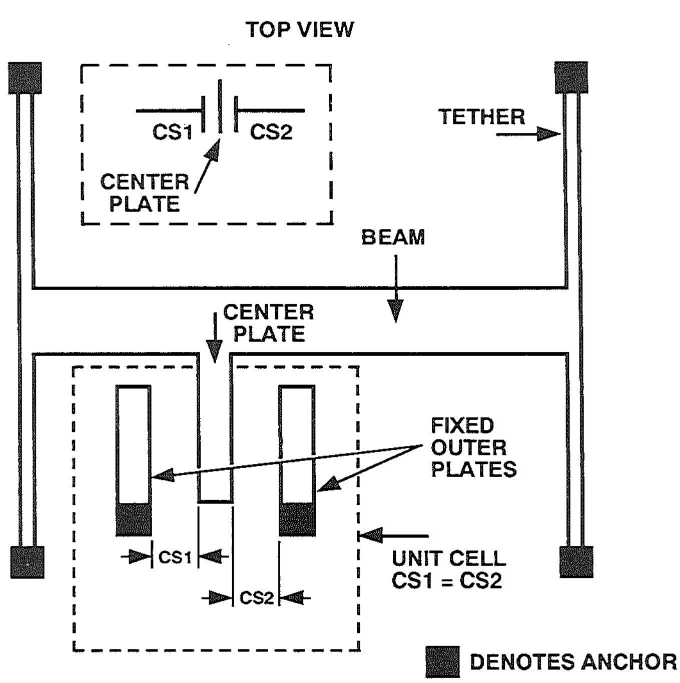


Figure 9.83

THE ADXL50 SENSOR MOMENTARILY RESPONDING TO AN EXTERNALLY APPLIED ACCELERATION

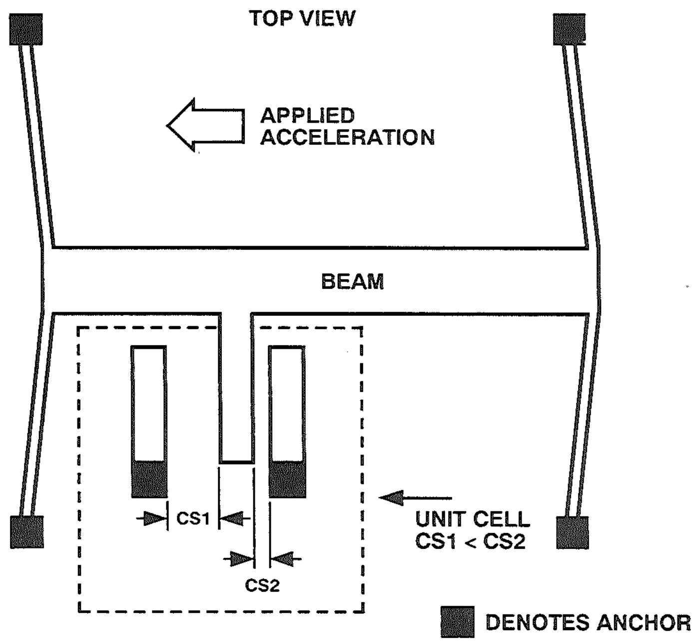


Figure 9.84

When the sensor begins to move, a mismatch in the capacitance produces an output signal at the central plate. The output amplitude will increase with the acceleration experienced by the sensor. The direction of beam motion affects the phase of the signal - synchronous demodulation is used to extract this information.

Figure 9.85 shows a simplified block diagram of the ADXL50. The voltage output from the central plate of the sensor is buffered and then applied to a synchronous demodulator which is also supplied with a 1MHz clock from the same oscillator which drives the fixed plates of the sensor. The demodulator will rectify any voltage which is synchronous with its clock signal. If the applied voltage is in sync and in phase with the clock, a positive output will result. If the applied voltage is in sync but 180° out of phase with the clock, then the demodulator's output will be negative. All other signals will be rejected. The bandwidth of the demodulator is set by an external capacitor C1.

ADXL50 ACCELEROMETER BLOCK DIAGRAM

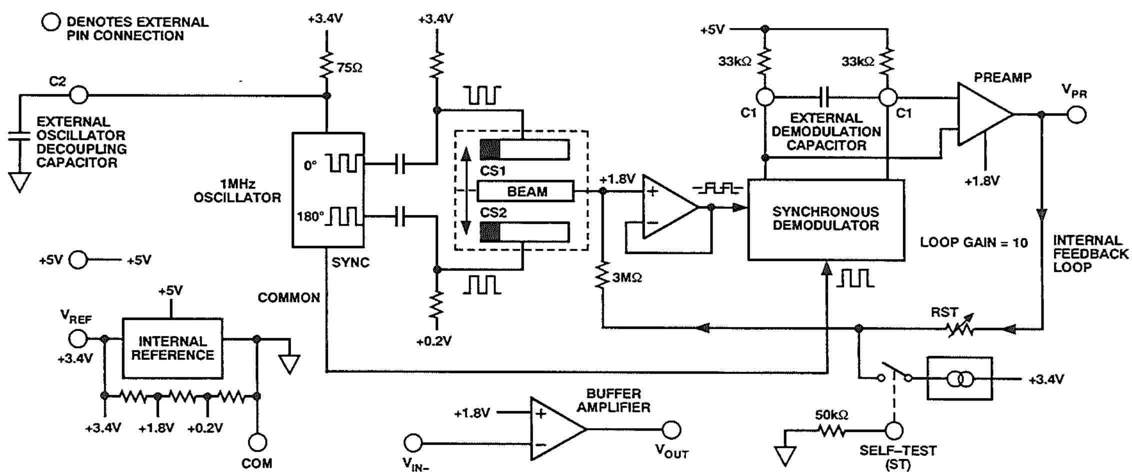


Figure 9.85

The output of the synchronous demodulator drives the preamp - an instrumentation amplifier which is referenced to +1.8V. The output of the preamp is fed back to the sensor through a 3MΩ isolation resistor and creates an electrostatic force which holds the sensor's center plate in the center position. The correction voltage is a direct measure of the applied acceleration and appears at the V_{PR} pin.

The loop bandwidth corresponds to the time required to apply feedback to the sensor and is set by external capacitor C1. The loop response is fast enough to follow changes in acceleration up to and exceeding 1kHz. The ADXL50's ability to maintain a flat response over this bandwidth keeps the sensor virtually motionless. This eliminates any nonlinearity or aging effects in the sensor beam's mechanical support spring, which is a problem with open-loop sensors.

The output of the preamp is 1.8V at zero acceleration with an output range of ±0.95V for a ±50g input, i.e., 19mV/g. An uncommitted buffer amplifier has been included on-chip to enhance the user's ability to offset the zero signal level and to amplify and filter the signal. The sensitivity may be increased up to ±10g (200mV/g) by using the internal buffer amplifier with the proper external gain-setting resistors.

9

CHARGE COUPLED DEVICES (CCDs)

Charge coupled devices (CCDs) contains a large number of small photocells called photosites or pixels which are arranged either in a single row (linear arrays) or in a matrix (area arrays). CCD area arrays are commonly used in video applications, while linear arrays are used in facsimile machines, graphics scanners, and pattern recognition equipment.

The linear CCD array consists of a row of image sensor elements (photosites, or pixels) which are illuminated by light from the object or document. During one exposure period each photosite acquires an amount of charge which is proportional to its illumination. These photosite charge packets are subsequently switched simultaneously via transfer gates to an analog shift register. The charge packets on this shift register are clocked serially to a charge detector (storage capacitor) and buffer amplifier (source follower) which convert them into a string of photo-dependent output voltage levels (see Figure 9.86). While the charge packets from one exposure are being clocked out to the charge detector, another exposure is underway. The analog shift register typically operates at frequencies between 1 and 10MHz.

LINEAR CCD ARRAY

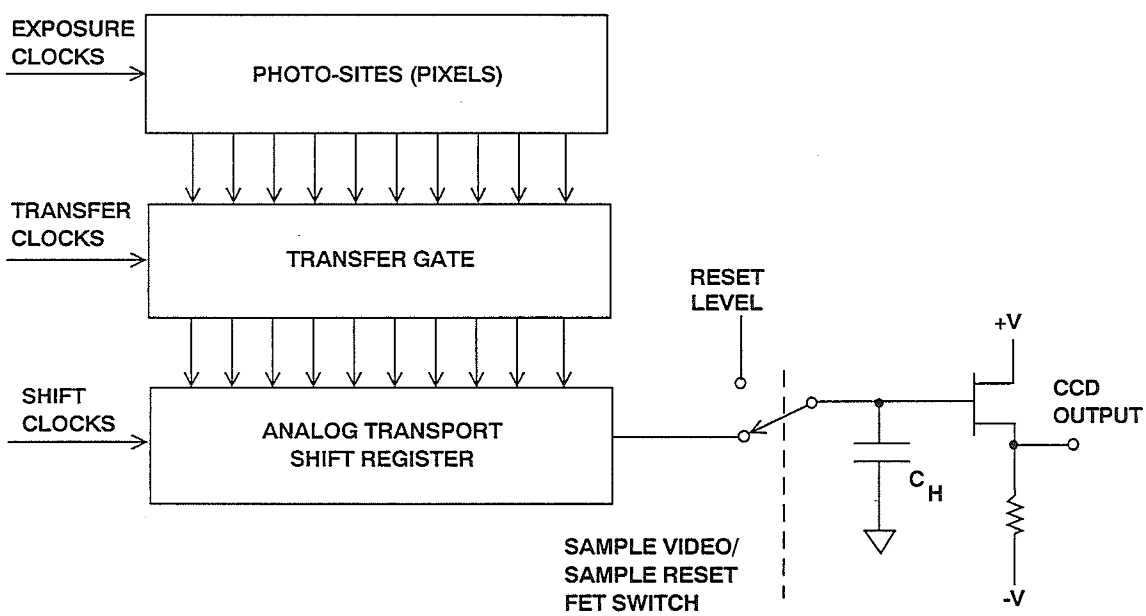


Figure 9.86

The charge detector readout cycle begins with a reset pulse which causes a FET switch to set the output storage capacitor to a known voltage. The switching FETs capacitive feedthrough causes a reset glitch at the output as shown in Figure 9.87. The switch is then opened, isolating the capacitor, and the charge from the last pixel is dumped onto the capacitor causing a voltage change. The difference between the reset voltage and the final voltage (video level) shown in Figure 9.87 represents the amount of charge in the pixel. CCD charges may be as low as 10 electrons, and a typical CCD output sensitivity is 0.6μV/electron. Most CCDs have a saturation output voltage of about 1V (see Reference 16).

CCD OUTPUT WAVEFORM

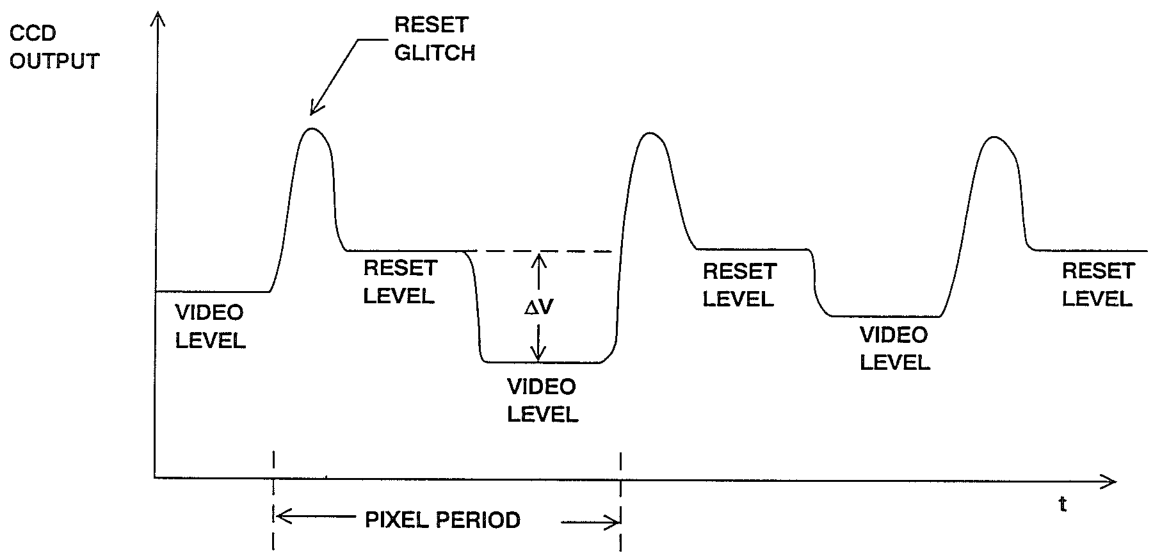


Figure 9.87

Since CCDs are generally fabricated on MOS processes, they have limited capability to perform on-chip signal conditioning. Therefore the CCD output is generally processed by external conditioning circuits.

CCD output voltages are small and quite often buried in noise. The largest source of noise is the thermal noise in the resistance of the FET reset switch. This noise may have a typical value of 100 to 300 electrons rms (approximately 60 to 180mV rms). This noise occurs as a *sample-to-sample* variation in the CCD output level and is common to both the reset level and the video level for a given pixel period. A technique called *correlated double sampling* (CDS)is often used to reduce the effect of this noise. Figure 9.88 shows two circuit implementations of the CDS scheme. In the top circuit, the CCD output drives both SHAs. At the end of the reset interval, SHA1 holds the reset voltage level. At the end of the video interval, SHA2 holds the video level. The SHA outputs are applied to a difference amplifier which subtracts one from the other. In this scheme, there is only a short interval during which both SHA outputs are stable, and their difference represents ΔV, so the difference amplifier must settle quickly.

Another arrangement is shown in the bottom half of Figure 9.88, which uses three SHAs and allows either for faster operation or more time for the difference amplifier to settle. In this circuit, SHA1 holds the reset level so that it occurs simultaneously with the video level at the input to SHA2 and SHA3. When the video clock is applied simultaneously to SHA2 and SHA3, the input to SHA2 is the reset level, and the input to SHA3 the video level. This arrangement allows the entire pixel period (less the acquisition time of SHA2 and SHA3) for the difference amplifier to settle.

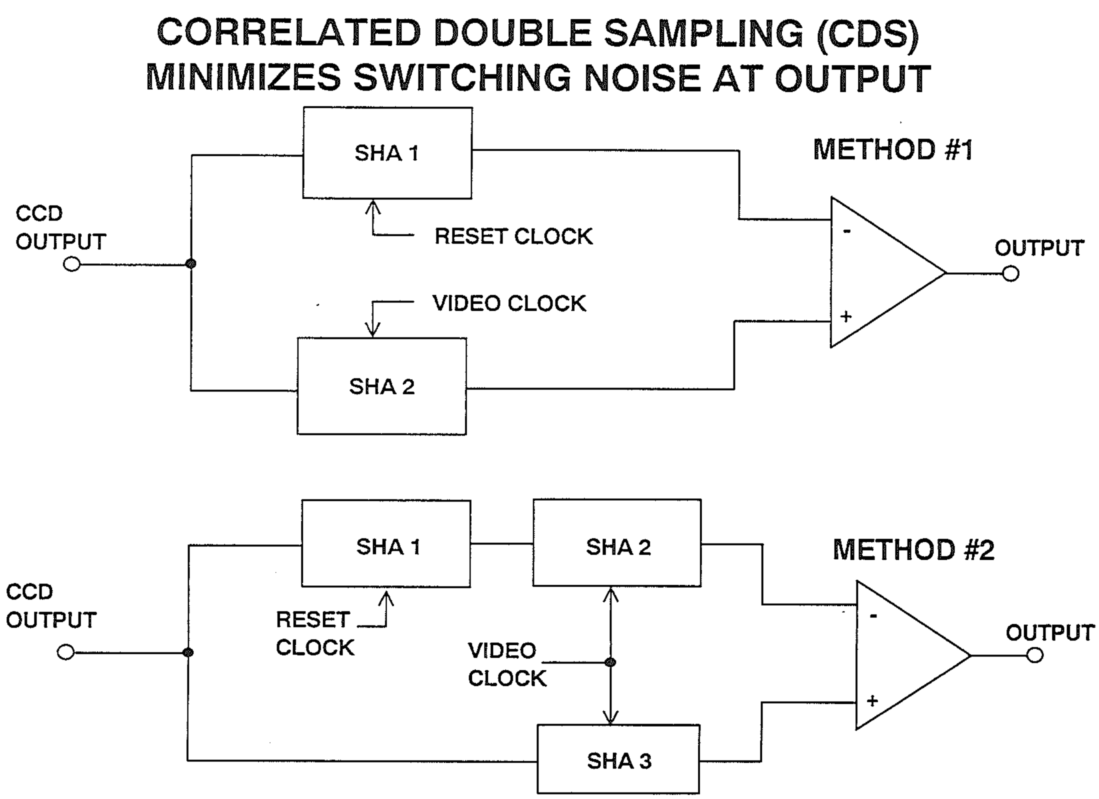


Figure 9.88

HALL EFFECT MAGNETIC SENSORS

If a current flows in a conductor (or semiconductor) and there is present a magnetic field perpendicular to the current flow, then the combination of current and magnetic field will generate a voltage perpendicular to both. This phenomenon is called the *Hall Effect*, and the voltage, V_H, is known as the *Hall Voltage*. V_H is a function of the current density, the magnetic field, and the charge density and carrier mobility of the conductor.

HALL EFFECT SENSORS

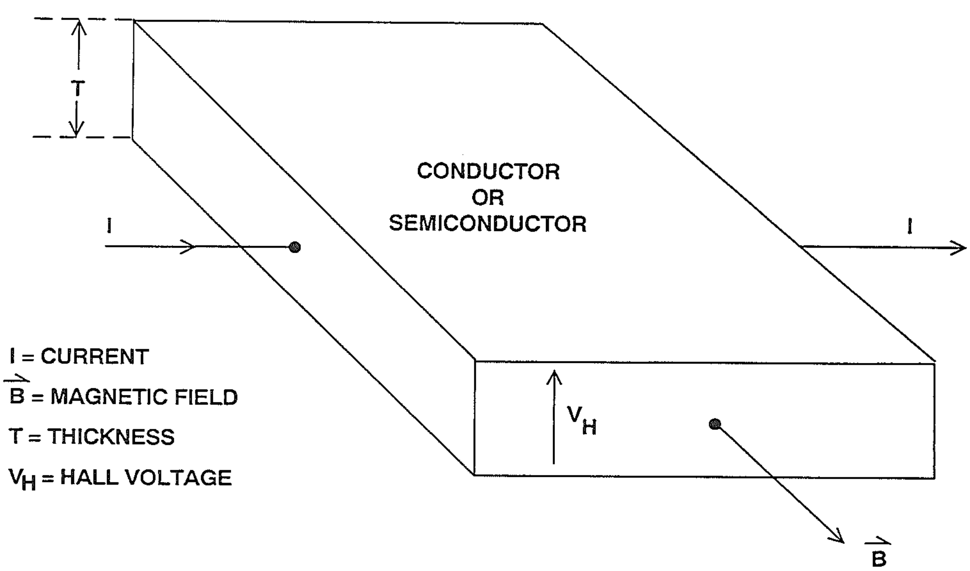


Figure 9.89

The Hall effect may be used to measure magnetic fields (and hence in contact-free current measurement), but its commonest application is in motion sensors where a fixed Hall sensor and a small magnet attached to a moving part can replace a cam and contacts with a great improvement in reliability. (Cams wear and contacts arc or become fouled, but magnets and Hall sensors are contact free and do neither.) Since V_H is proportional to magnetic field and not to rate of change of magnetic field like an inductive sensor, the Hall Effect provides a more reliable low speed sensor than an inductive pickup.

Although several materials can be used for Hall effect sensors, silicon has the advantage that signal-conditioning circuits can be integrated on the same chip. The AD22150 integrates a Hall effect sensor with a gain stage and a comparator as shown in Figure 9.90. The part is designed to detect rotation speed in automotive applications. The AD22150 responds to small changes in field (operate and release points at –12 and +17 Gauss, respectively). The comparator has built-in hysteresis, and the device operates up to 50kHz. A typical application is shown in Figure 9.91. The changing magnetic field caused by the rotating target wheel and the biasing magnet produces a square wave output from the AD22150.

AD22150 MONOLITHIC HALL EFFECT SENSOR WITH SIGNAL CONDITIONING

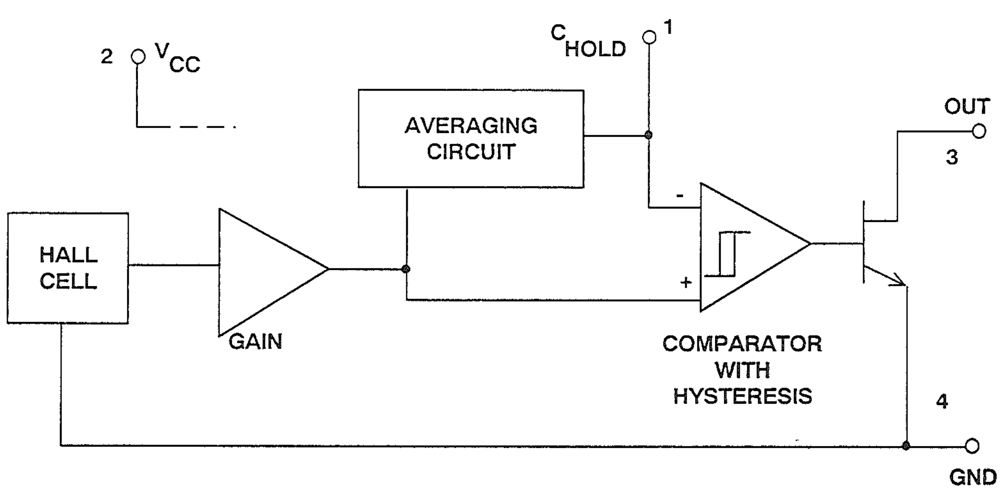


Figure 9.90

TYPICAL APPLICATION OF A HALL EFFECT SENSOR

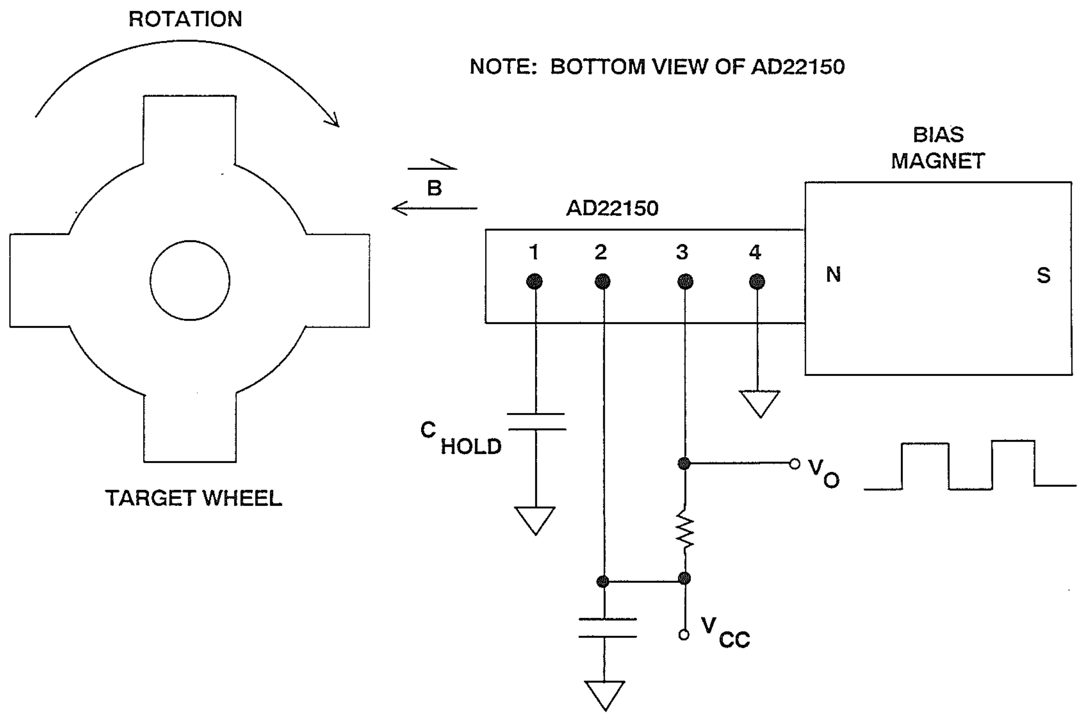


Figure 9.91

REFERENCES

1. Ramon Pallas-Areny and John G. Webster, **Sensors and Signal Conditioning**, John Wiley, New York, 1991.

2. Dan Sheingold, Editor, **Transducer Interfacing Handbook**, Analog Devices, Inc., 1980.

3. Walt Kester, Editor, **1992 Amplifier Applications Guide**, Section 2, 3, Analog Devices, Inc., 1992.

4. Walt Kester, Editor, **System Applications Guide**, Section 1, 6, Analog Devices, Inc., 1993.

5. Jim Williams, *Thermocouple Measurement,* **Linear Technology Application Note 28,** Linear Technology Corporation.

6. Dan Sheingold, **Nonlinear Circuits Handbook**, Analog Devices, Inc.

7. James Wong, *Temperature Measurements Gain from Advances in High-precision Op Amps*, **Electronic Design**, 15 May 1986.

8. **OMEGA Temperature Measurement Handbook**, Omega Instruments, Inc.

9. Thomas M. Fredrickson, **Intuitive Operational Amplifiers**, McGraw-Hill, 1988.

10. **Optoelectronics Data Book**, EG&G Vactec, St. Louis, MO, 1990.

11. Silicon Detector Corporation, Camarillo, CA, Part Number SD-020-12-001 Data Sheet.

12. **Photodiode 1991 Catalog**, Hamamatsu Photonics, Bridgewater, NJ

13. Ralph Morrison, **Grounding and Shielding Techniques in Instrumentation, Third Edition**, John Wiley, Inc., 1986.

14. Henry W. Ott, **Noise Reduction Techniques in Electronic Systems, Second Edition**, John Wiley, Inc., 1988.

15. Frank Goodenough, *Airbags Boom when IC Accelerometer Sees 50g*, **Electronic Design**, August 8, 1991.

16. *An Introduction to the Imaging CCD Array*, **Technical Note** 82W-4022, Tektronix, Inc., Beaverton, OR., 1987.

17. **Handbook of Chemistry and Physics**, Chemical Rubber Co.

18. James Bryant, *A Function Generator and Linearization Circuit Using the AD7569*, **Application Note AN-285**, Analog Devices, Inc.