SECTION II
PRECISION TRANSDUCER INTERFACES

- Bridge Circuit Applications
- Bridge Amplifiers
- Signal Conditioning for Temperature Measurement:
  Thermocouples, Resistance Temperature Devices (RTDs), Thermistors, Monolithic Thermocouple Amplifiers

SECTION II

PRECISION TRANSDUCER INTERFACES

JAMES WONG, ADOLFO GARCIA

PRECISION BRIDGE INTERFACES

JAMES WONG

Because of the many physical variables involved, real world signal processing requires a wide variety of sensing elements. Most are based on resistance, capacitance, or voltage output (thermocouple) elements, thereby simplifying the interface to electronic measuring and processing equipment.

Resistive elements are the most common transducers. They are cost effective 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), light signals. 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.

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

TYPICAL SENSING ELEMENTS FOR MEASURING REAL WORLD VARIABLES

- Resistance Output (RTDs, Strain Gauges)
- Capacitance or Charge Output (Piezoelectric)
- Voltage Output (Thermocouples)
- Current Output (Photodiode Detectors)
- ALL ARE LOW LEVEL SIGNALS REQUIRING PRECISION LOW-NOISE AMPLIFICATION

Figure 2.1

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. The following sections examine some of the important design tradeoffs involved.

BRIDGE AMPLIFIERS

A balanced bridge is one of the most useful circuits for sensing low level signals from a transducer. When applied properly, bridge amplifiers reproduce transducer signals accurately and reliably. Most bridges employ resistive elements, either for temperature sensing or for measuring the amount of strain from pressure or flex.

Figure 2.2 shows three commonly used bridge types. The single-element varying type is most suited for temperature sensing, such as an RTD (Resistance Temperature Device).

However, it produces the least amount of signal. The 2-element varying type produces twice the signal, which is significant for signals in the microvolt to millivolt range. It is the least popular of the three because the implementation of it in a transducer is more difficult. The all-element varying type produces the most signal for a given resistance change. Whatever the form, choosing the right bridge and amplifier configuration can minimize nonlinear behavior and simplify the design.

Bridge circuits are simple to use but can be difficult to optimize because the output signal is usually very small. The inherent high common-mode voltage of a balanced bridge also makes designing difficult. Designers must usually make tradeoffs between performance and cost. In circuits where microvolts of error are significant, the fewer components and the simpler the circuit, the better the design.

Bridge resistance varies over a wide range from as low as 100 Ω to 100's of kΩs or more. The resistance depends on the transducer design. Figure 2.3 shows some popular transducer sensors and their typical resistance ranges.

BASIC BRIDGE CONFIGURATIONS

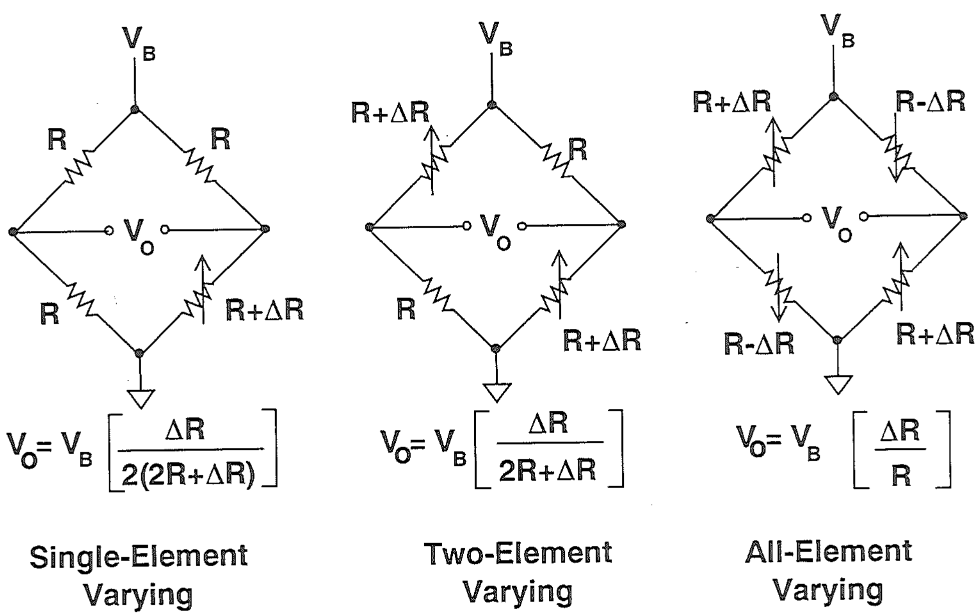


Figure 2.2

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

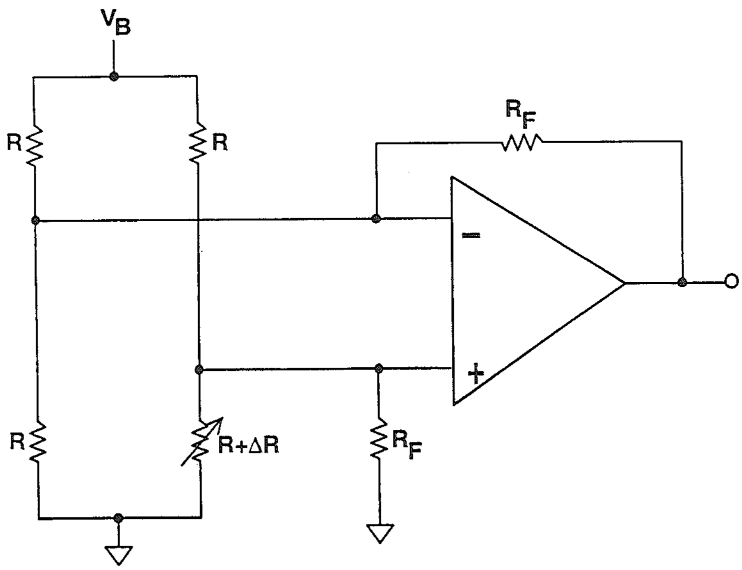
Figure 2.3

This section discusses the various bridge amplifier configurations, their relative linearity, and their advantages and disadvantages. The choice of the amplifier is also critical to the performance of the bridge.

SINGLE-ELEMENT VARYING SENSOR BRIDGES

Single-element sensors do not necessarily require a full bridge to produce usable signals. A variable resistance inside the feedback loop of an op amp produces usable signal. However, a bridge offers a convenient means to make more precise measurements of resistance change and provides a convenient means to level-shift. Figures 2.4 through 2.9 show different single-element varying bridge configurations, each having unique advantages and disadvantages.

SINGLE-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 1

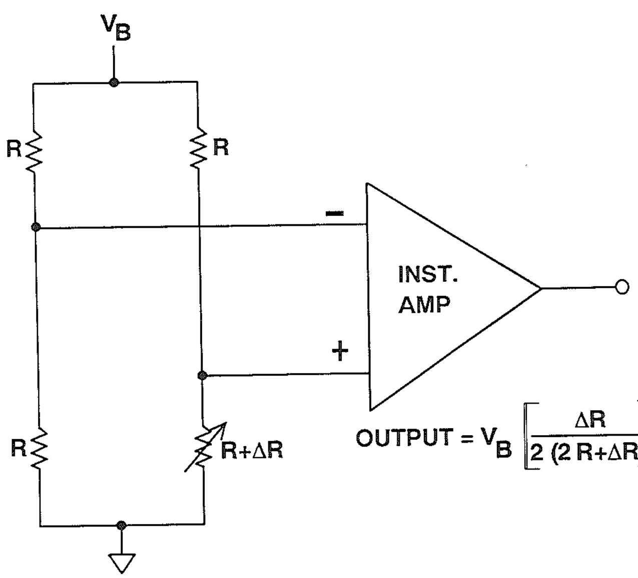


Advantages:
- No signal in = zero volts out
- Single supply operation
- Single op amp stage

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

Figure 2.4

SINGLE-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 2



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

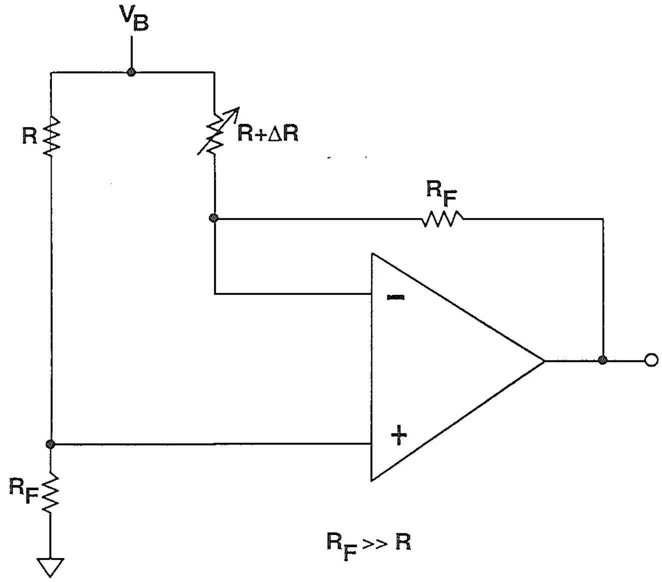
Advantages:
- Efficient design
- Only bridge R matching required
- Better gain accuracy

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

Figure 2.5

2

SINGLE-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 3

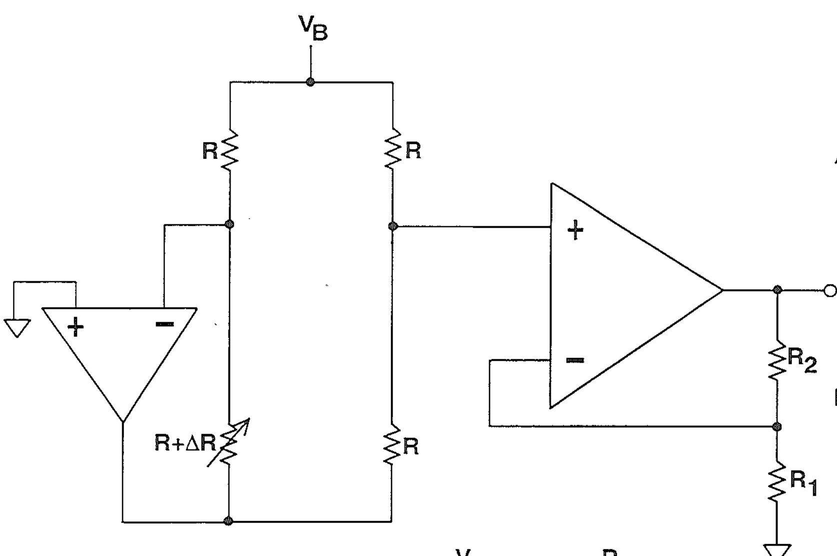


Advantages:
- Linear operation
- Low cost op amp
- No signal in = zero volts out
- Single supply

Disadvantages:
- Requires matching resistors
- Operates at high common-mode bridge voltage

Figure 2.6

SINGLE-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 4



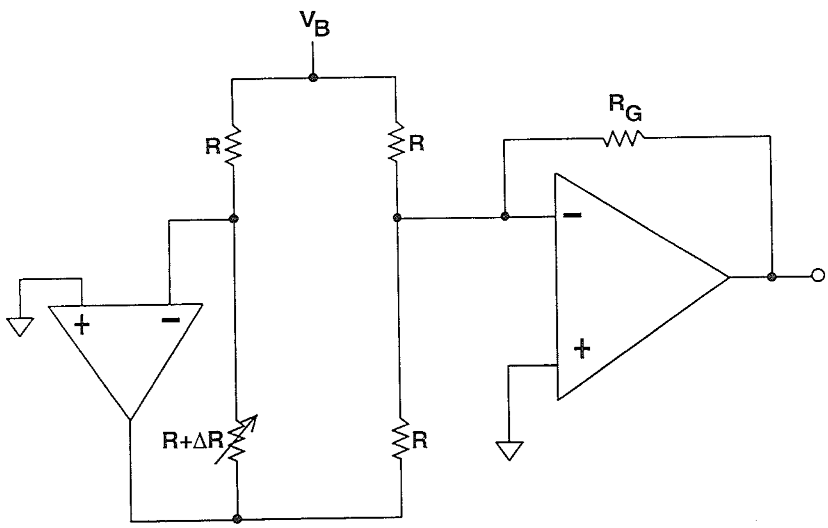
$$\text{OUTPUT} = \frac{V_B}{2}\left(\frac{\Delta R}{R}\right)\left(1 + \frac{R_2}{R_1}\right)$$

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

Figure 2.7

SINGLE-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 5



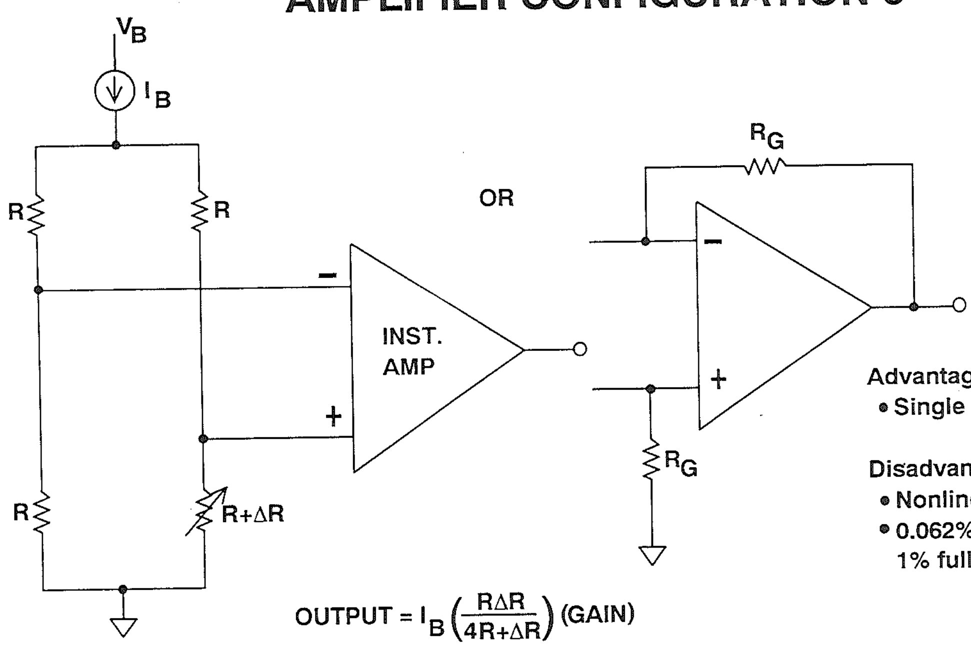
Advantages:
- Linear operation
- Single resistor sets gain
- Constant zero volts common-mode
- 2X signal resolution

Disadvantages:
- Requires 2 op amps
- Requires 2 supplies

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

Figure 2.8

SINGLE-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 6



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 2.9

A typical implementation of a single-element varying bridge is an RTD thermometer amplifier as shown in Figure 2.10. The bridge power is derived from a stable 2.500V reference. Each leg of the bridge normally carries only 600µA so as to minimize self-heating. The bridge is servoed to suppress the common-mode voltage to zero , thus simplifying the gain stage design as well as assuring linear performance.

At 600µA current, the RTD dissipates less than 0.1mW of power even at the maximum resistance . This is important because some RTDs have thermal resistance as high as 0.5 to 0.8 °C per mW. Consequently, even 1mW of dissipation can result in an appreciable error if the element is allowed to self heat. The design ensures self-heating contributes much less than 0.1 °C error over the full temperature range of the RTD.

If the RTD is remotely located some distance away from the measurement circuitry, the long line resistance will contribute measurement error. The optional zero adjustment circuit provides a method to balance the bridge. Calibration can be made at low temperatures or at the 0°C point. At 0°C the output should be adjusted to zero volts. The virtual ground summing node of the RTD is a convenient point to linearize the sensor by providing a small amount of positive feedback from the output. To calibrate, set the fullscale and linearity adjust pots to midpoint. Either apply a 500°C temperature to the sensor or substitute the equivalent 500°C RTD resistance. Then adjust the fullscale pot to 5.000V output. Apply half-scale temperature or the equivalent RTD resistance, then adjust the linearity pot to 2.500V output. Recheck the two settings and adjust as needed. Over the temperature range of -150° to +500°C, the circuit achieves better than ±0.5°C accuracy.

2

LOW POWER LINEARIZED RTD THERMOMETER BRIDGE AMPLIFIER

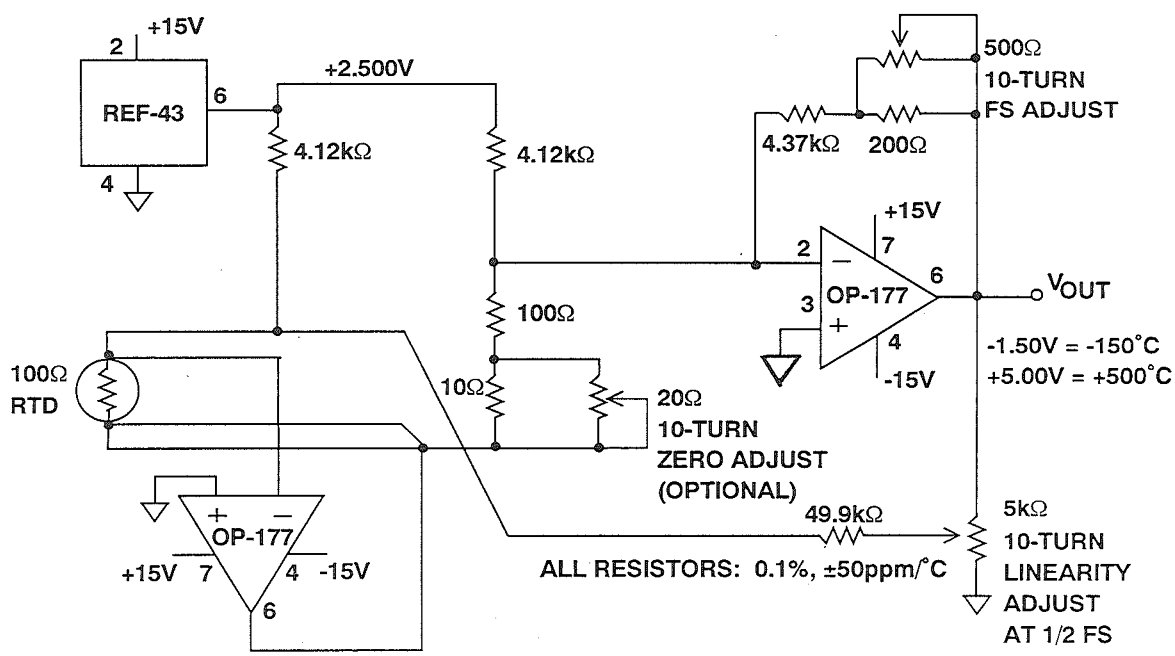
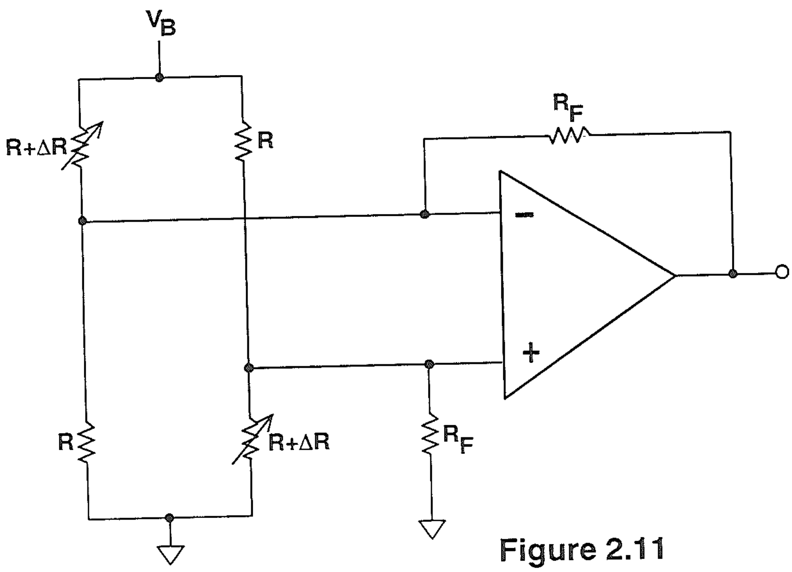


Figure 2.10

TWO-ELEMENT VARYING SENSOR BRIDGES

Two-element varying sensor bridges are commonly found in pressure transducers and flow meter systems. They have the benefit of producing twice the signal for a given input than the single-element bridges. Figures 2.11through 2.15 show several two-element bridge configurations and the relative merits of each.

TWO-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 1

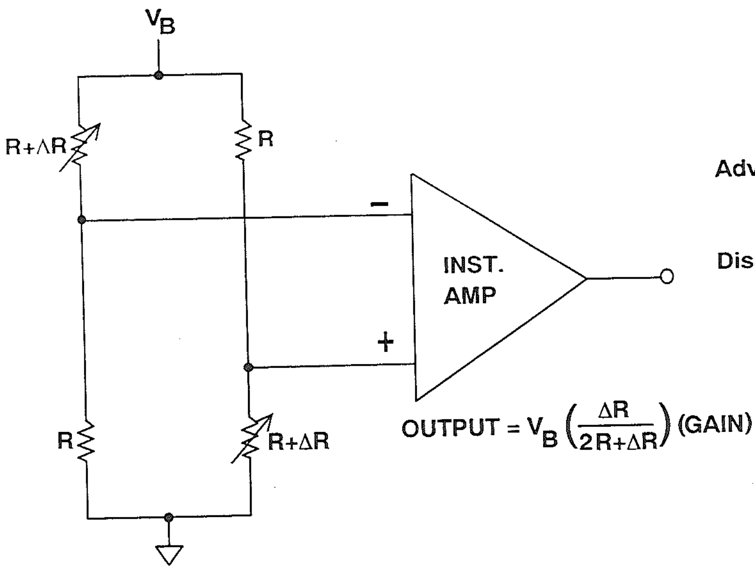


Advantages:
- Uses low cost op-amp
- Excellent gain-setting range
- No signal in = OV out
- Single supply operation

Disadvantages:
- Nonlinear operation 0.25% nonlinearity at 1% fullscale ΔR change
- Requires R matching

Figure 2.11

TWO-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 2

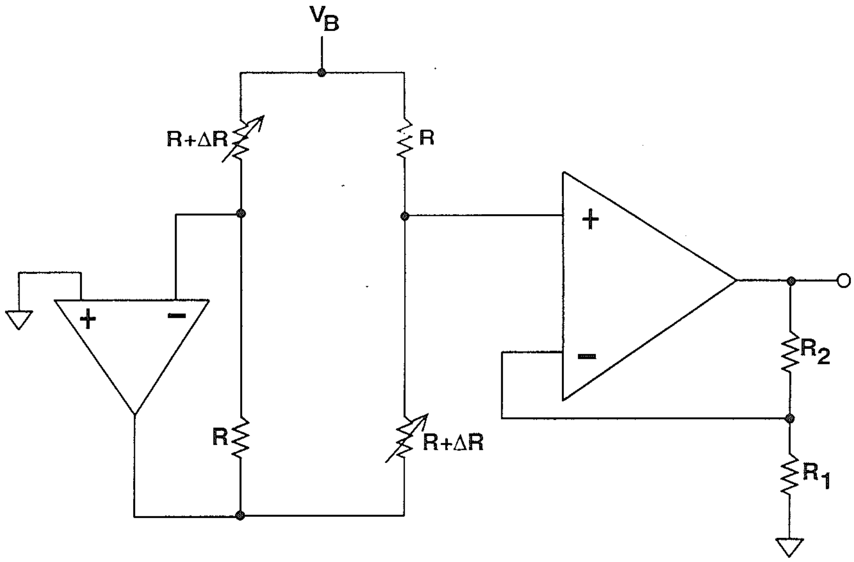


Advantages:
- Requires only bridge R matching

Disadvantages:
- Nonlinear operation 0.125% nonlinearity at 1% F.S. ΔR change

Figure 2.12

TWO-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 3



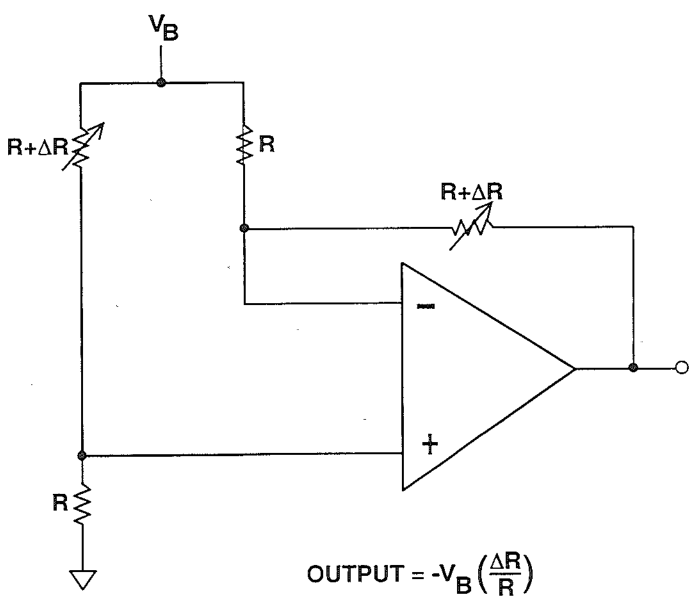
Advantage:
- 2X signal output

Disadvantages:
- Nonlinear operation 0.25% nonlinearity at 1% fullscale ΔR change
- Requires 2 op amps
- Requires 2 supplies
- R matching critical

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

Figure 2.13

TWO-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 4



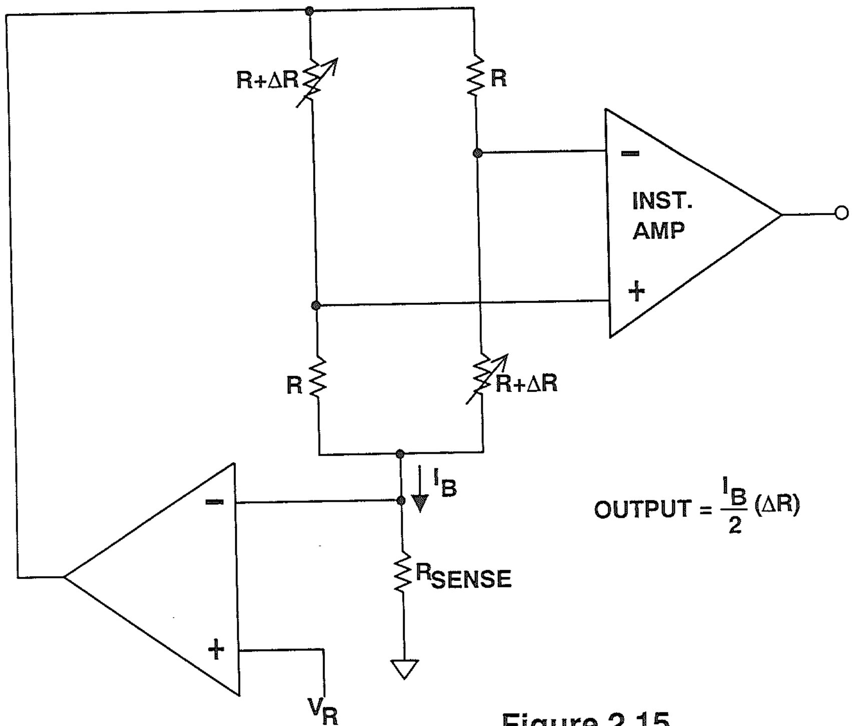
Advantage:
- Linear operation

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

$$\text{OUTPUT} = -V_B\left(\frac{\Delta R}{R}\right)$$

Figure 2.14

TWO-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 5



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

Disadvantages:
- Requires 2 amps

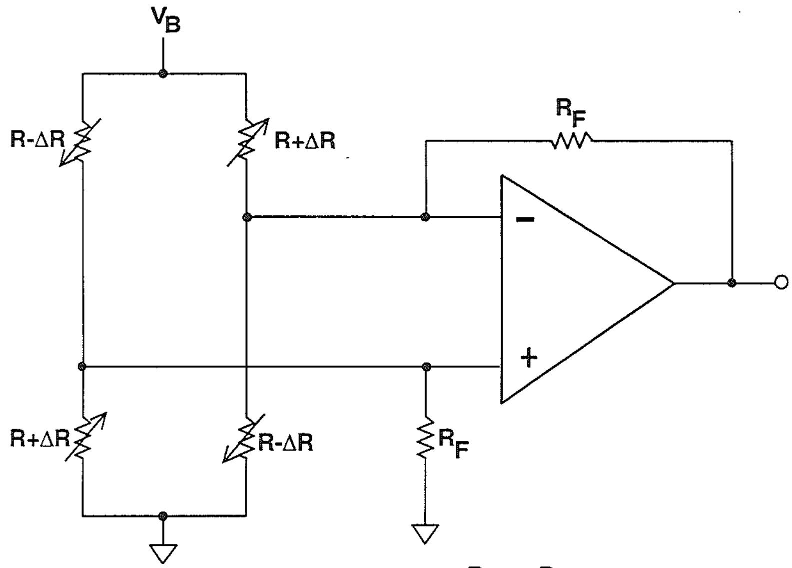
Figure 2.15

ALL-ELEMENT VARYING SENSOR BRIDGES

All-element varying bridge types are most popular among strain-gauge transducer applications such as weigh scales, pressure sensors, and strain measuring instruments. This bridge type produces twice the signal of the two-element bridge, and four times the signal of the single-element type. Figures 2.16 through 2.21 show several all-element varying configurations and the relative merits of each.

2

ALL-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 1



Advantages:

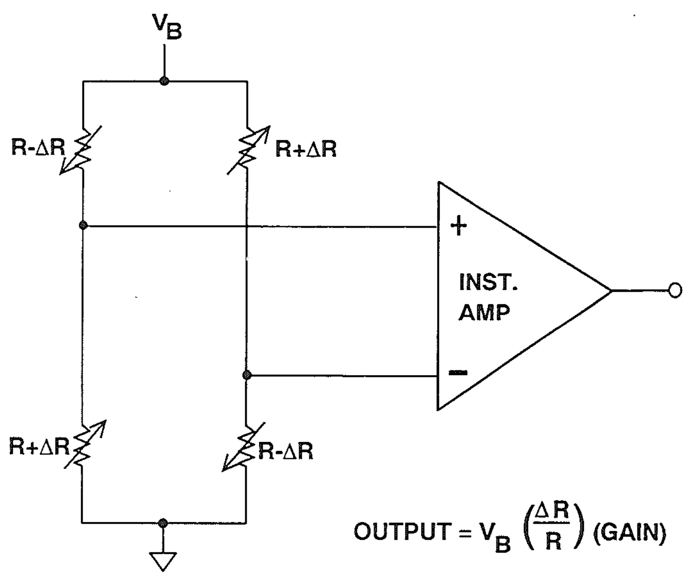
- Single supply operation
- No signal in = OV out
- Simple to set gain
- Uses low-cost op amp

Disadvantages:

- Slightly nonlinear 0.0037% nonlinearity at 1% F.S. ΔR change, at Gain = 100
- R matching critical

Figure 2.16

ALL-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 2



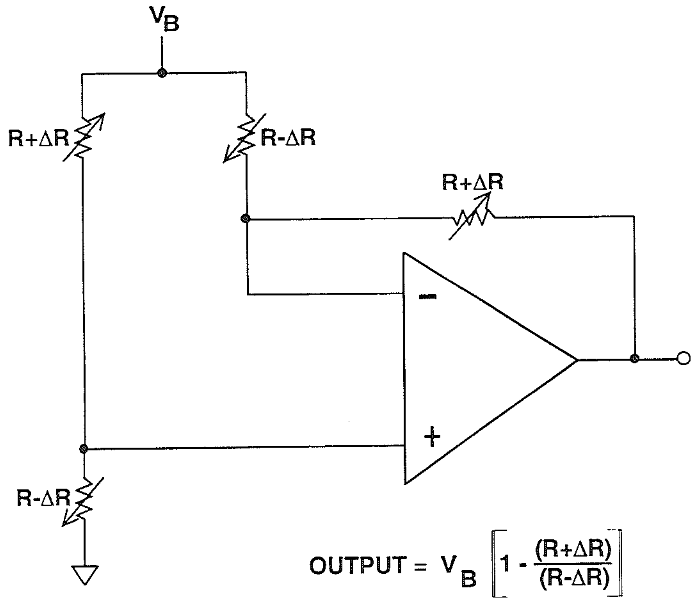
Advantages:

- Linear operation
- Single supply operation
- No signal in = OV out
- No R matching required

$$\text{OUTPUT} = V_B \left(\frac{\Delta R}{R}\right) (\text{GAIN})$$

Figure 2.17

ALL-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 3



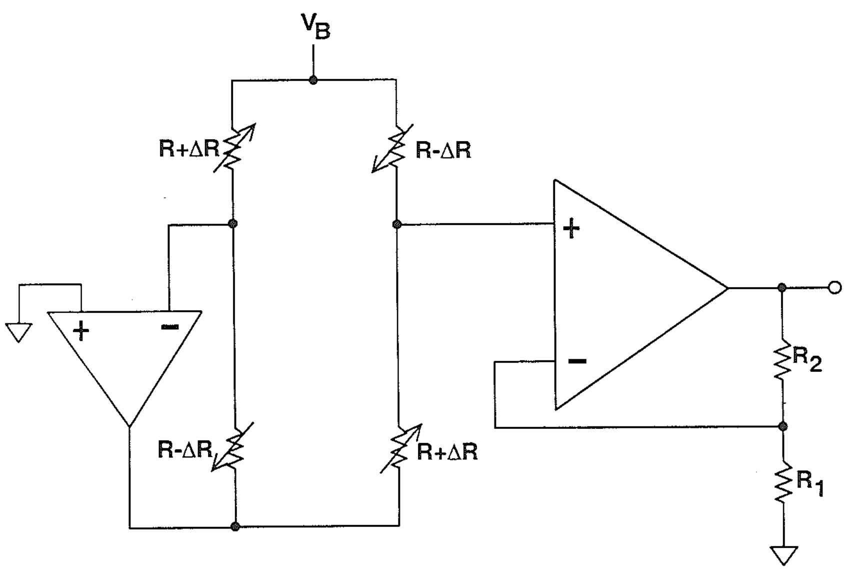
Advantages:
- None

Disadvantages:
- Nonlinear operation 0.25% nonlinearity for 1% fullscale ΔR change
- Low signal output
- Requires second gain stage

$$\text{OUTPUT} = V_B \left[1 - \frac{(R+\Delta R)}{(R-\Delta R)}\right]$$

Figure 2.18

ALL-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 4



Advantage:
- 2X signal output

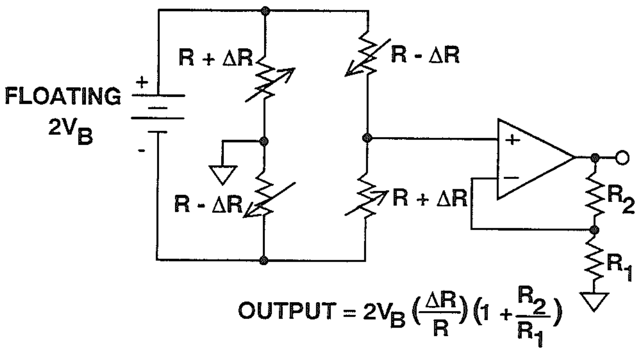
Disadvantages:
- Nonlinear operation 0.25% nonlinearity at 1% fullscale ΔR change
- Requires 2 op amps
- Requires 2 supplies
- More R matching

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

Figure 2.19

2

ALL-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 5

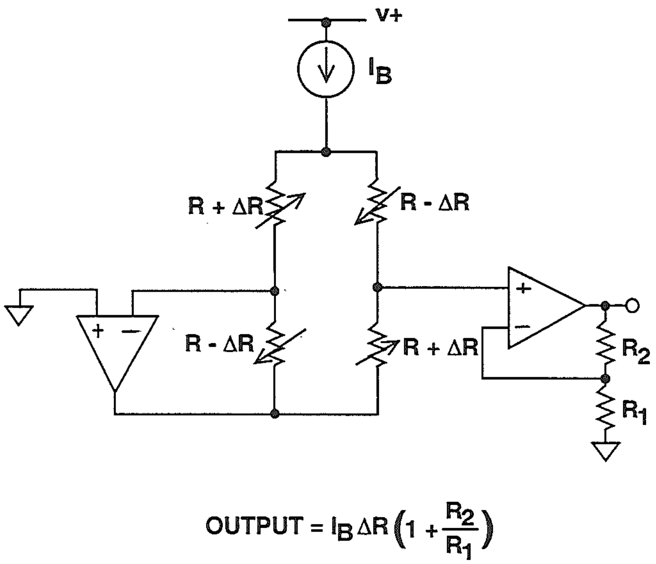


ADVANTAGES:
- Linear operation
- 2X signal output
- Uses low-cost op amp

DISADVANTAGES:
- Requires elaborate floating supply

Figure 2.20

ALL-ELEMENT VARYING BRIDGE, AMPLIFIER CONFIGURATION 6



ADVANTAGES:
- Linear operation
- 2X signal output
- Gain set resistors affect gain accuracy

DISADVANTAGES:
- Requires 2 op amps
- Requires a current source
- Requires 2 supplies

Figure 2.21

An example of an all-element varying bridge circuit is a fatigue monitoring strain sensing circuit as shown in Figure 2.22. 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 differentially amplified by the AD620 instrumentation amplifier. Full-scale strain voltage may be set by adjusting the 100 Ω 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

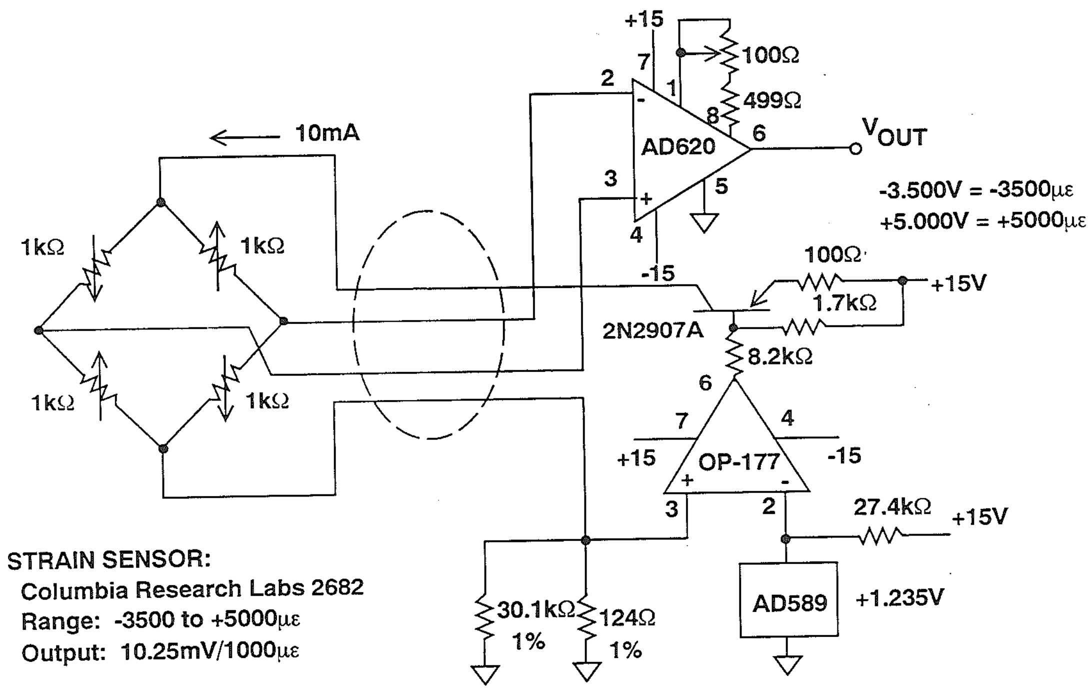


Figure 2.22

Another example is a weigh-scale amplifier circuit shown in Figure 2.23. A typical load cell has a bridge resistance of 350 Ω. 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 475 Ω resistor and the 100 Ω potentiometer have low temperature coefficients so the amplifier gain remains does not drift over temperature.

WEIGH-SCALE LOAD CELL AMPLIFIER

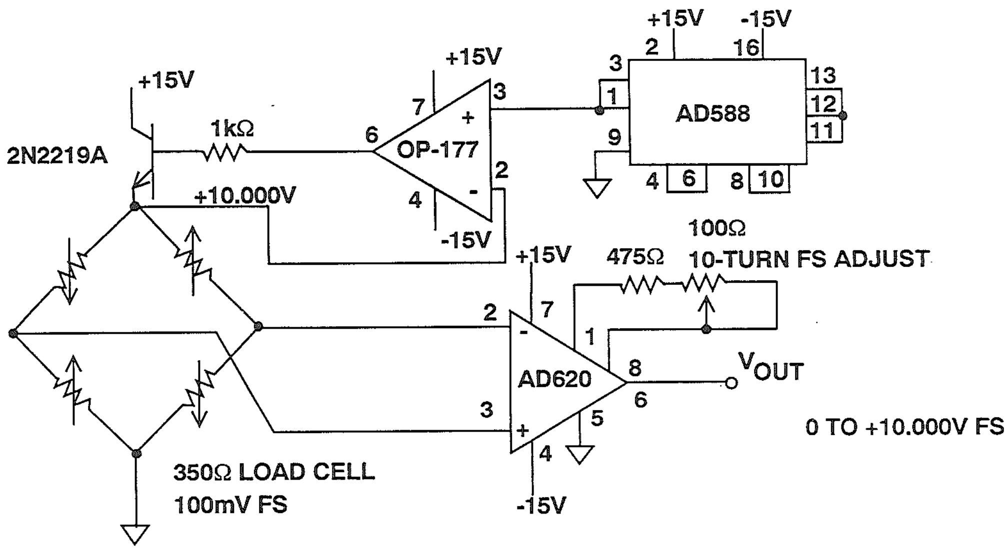


Figure 2.23

CHOOSING THE RIGHT AMPLIFIER FOR OPTIMUM PERFORMANCE

Once a bridge circuit configuration is chosen, it is of critical importance to choose the right amplifier for that circuit. There are several types of amplifiers from which to choose: precision op amps, instrumentation amplifiers, and chopper stabilized amplifiers. Each has unique attributes and capabilities that optimize a particular design feature. Other considerations that relate to system requirements are power consumption, bandwidth, supply voltages, and cost. The following sections discuss some of these tradeoffs.

Although many op amp parameters effect the accuracy of the circuit, it is important to identify the most important ones in the particular circuit application. For example, an op amp input offset current of 10 nA contributes only 1.75µV of error in a 350-ohm bridge. Clearly this is less critical than a 10µV input offset voltage error of a typical precision op amp. It is therefore important to focus on the parameters that have significant error impact on the circuit accuracy. Figure 2.24 lists some of the op amp critical parameters for high precision, high gain applications.

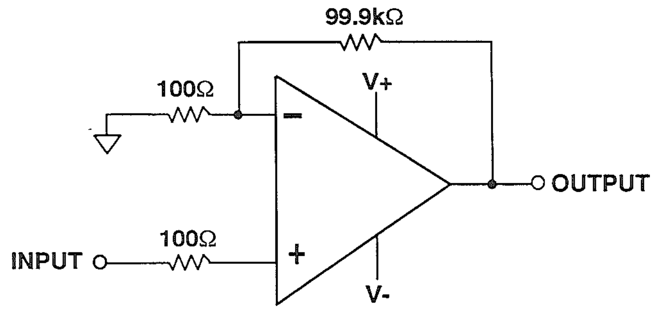
KEY OP AMP PARAMETERS FOR BRIDGE CIRCUITS

- Input Offset Voltage: $< 250\mu V$
- Input Offset Voltage Drift: $< 0.5\mu V/°C$
- Noise Voltage Density: $< 10nV/\sqrt{Hz}$
- Open Loop Gain: $> 1,000,000$

Figure 2.24

Let us compare some of the latest generation precision op amps and see how they perform under a complete DC error analysis. The circuit used is a simple gain of 1000 amplifier as shown in Figure 2.25. For the purpose of the analysis, only the error contribution due to the amplifier is considered. The external resistor tolerance is assumed to be zero. Since the bridge output has very little common-mode voltage change, no common-mode error is assumed. Total error in microvolts referred to the input for various op amps is shown in Figure 2.26.

AMPLIFIER ERROR BUDGET ANALYSIS



Assumptions:

- Zero Tolerance Resistors
- ±5% Power Supply Tolerance
- Negligible Common-Mode Voltage Change

Figure 2.25

TOTAL ERRORS IN MICROVOLTS REFERRED TO INPUT (RTI) FOR GAIN OF 1000 AMPLIFIER

	INDUSTRY STANDARD	ULTRA PRECISION	ULTRA PRECISION	LO POWER, LO I_B	LO POWER, LO I_B
PARAMETER	OP-07D	OP-177G	AD707J	OP-97F	AD705J
Max V_{OS}	150	60	90	75	90
Max V_{OS}/T	100	40	10	125	60
Max PSRR (3V)	153	9.5	9.5	12	12
Min A_{VOL} (FS)	100	5	1.2	66	66
Input Noise p-p	2	1.4	1.0	2.1	1.9
Max I_{OS}	0.8	0.5	0.15	0	0
Total Error RTI	506	116	112	280	230

Figure 2.26

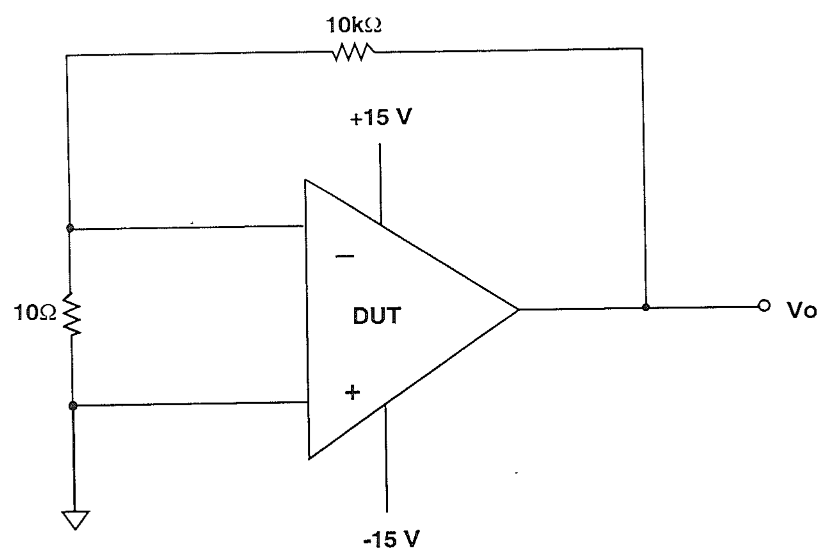
INPUT OFFSET VOLTAGE ERROR

Input offset voltage is usually one of the largest error sources for precision amplifier circuit designs. However, because it is a systemic error, it can be dealt with by using an offset null trim or by mathematically subtracting the output signal in a post-processor. Both solutions carry a cost penalty. In the first case, a potentiometer requires a manual adjustment step. Even if done automatically, additional circuitry is required. Similarly, additional circuitry is also needed in order for a processor to measure the offset during the calibration cycle.

Today's generation of precision op amps offers initial offset voltages as low as 10μV for bipolar devices, and far less for chopper stabilized amplifiers (more on chopper amps in the next sections). Most applications can tolerate a small amount of offset without resorting to null trimming or a costlier chopper amplifier. Only the most demanding applications will require a true zero offset error.

Measuring input offset voltages of a few microvolts requires that the test circuit does not introduce more error than the offset voltage itself. Figure 2.27 below shows a a precision input offset voltage test circuit.

INPUT OFFSET VOLTAGE TEST CIRCUIT



$$V_{IOS} = \frac{V_o}{1000}, \text{ Referred to Input}$$

Figure 2.27

As simple as it may look, this circuit may give inaccurate results unless the proper precautions are taken. The largest error comes from thermoelectric voltages caused by the temperature differential between the circuit component leads that form a pair of mismatched thermocouple junctions. Depending on the component or the IC package lead material, the thermoelectric voltage temperature coefficient may range anywhere from 2μV/ °C to more than 40μV/ °C when connected to a copper printed circuit board trace. Even a fraction of a degree difference at a single junction can be caused by air current and may induce large a enough error to render the offset voltage measurement invalid. Careful attention must also be given to proper layout, bypassing, and grounding techniques to avoid further error sources.

Measuring microvolts of offset reliably and repeatably may therefore depend more on the mechanical layout of the components and how they are placed on the PC board than on the circuit itself. Keep in mind that the two connections of a component such as a resistor create two equal, but opposite polarity, thermoelectric voltages (assuming they are connected to the same metal such as the copper trace on a PC board) which cancel each other — as long as they are at the exact same temperature. Clean connections and short lead lengths help to minimize temperature gradients and increase the accuracy of the measurement.

INPUT OFFSET VOLTAGE DRIFT

This temperature induced error is perhaps the most troublesome to deal with. One reason is that the drift characteristic of each op amp, even of the same type, is random and not necessarily linear. Predicting it accurately is impossible.

Today's precision op amps have reduced this error to satisfactory levels except for the most demanding applications. For example, ultra-precision op amps such as the OP-177 and the AD707 guarantee their input offset drifts to be less than 0.1μV/°C (for the top temperature grade devices).

These precision op amps achieve low drift by extremely careful chip layout. The photomicrograph of the OP-177 die as in Figure 2.28 illustrates a thermally symmetrical layout that minimizes drift. All offset and drift critical components lie symmetrically along a center line. Heat generated from the output stage and quiescent self-heating travels along this axis, thereby effecting all matched pairs equally.

There are other methods of minimizing offset voltage and offset voltage drift. One is a periodic auto-zero calibration under microprocessor control. Another is continuous zeroing as in a chopper stabilized amplifier. Chopper amplifiers do an excellent job of zeroing the offset and drift, however they tend to be much noisier. A way to achieve zero offset and drift and still have the low noise characteristics of a precision amplifier is shown in Figure 2.29 below. In addition to continuously nulling offset and drift, this circuit minimizes common-mode rejection error as well.

PHOTOMICROGRAPH OF OP-177 CHIP

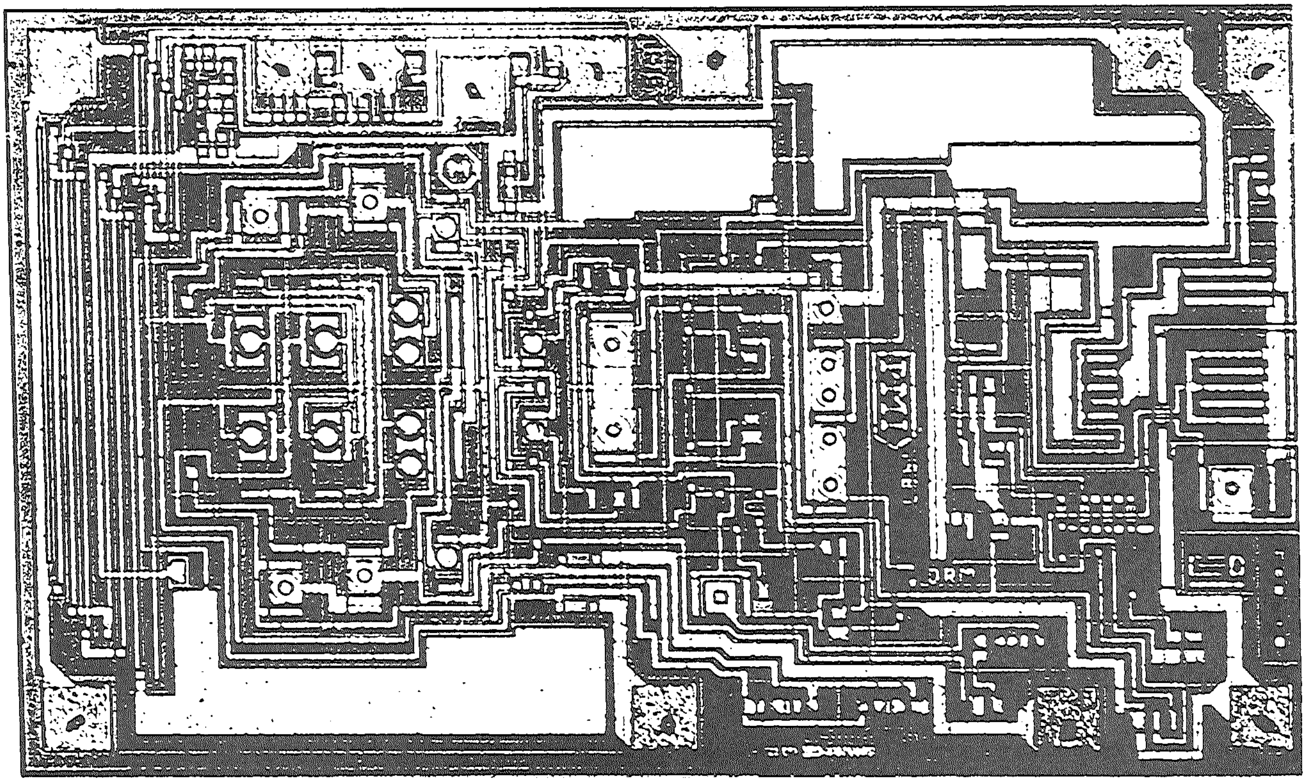


Figure 2.28

LOW-NOISE AUTO-ZERO CIRCUIT CORRECTS COMMON-MODE ERROR

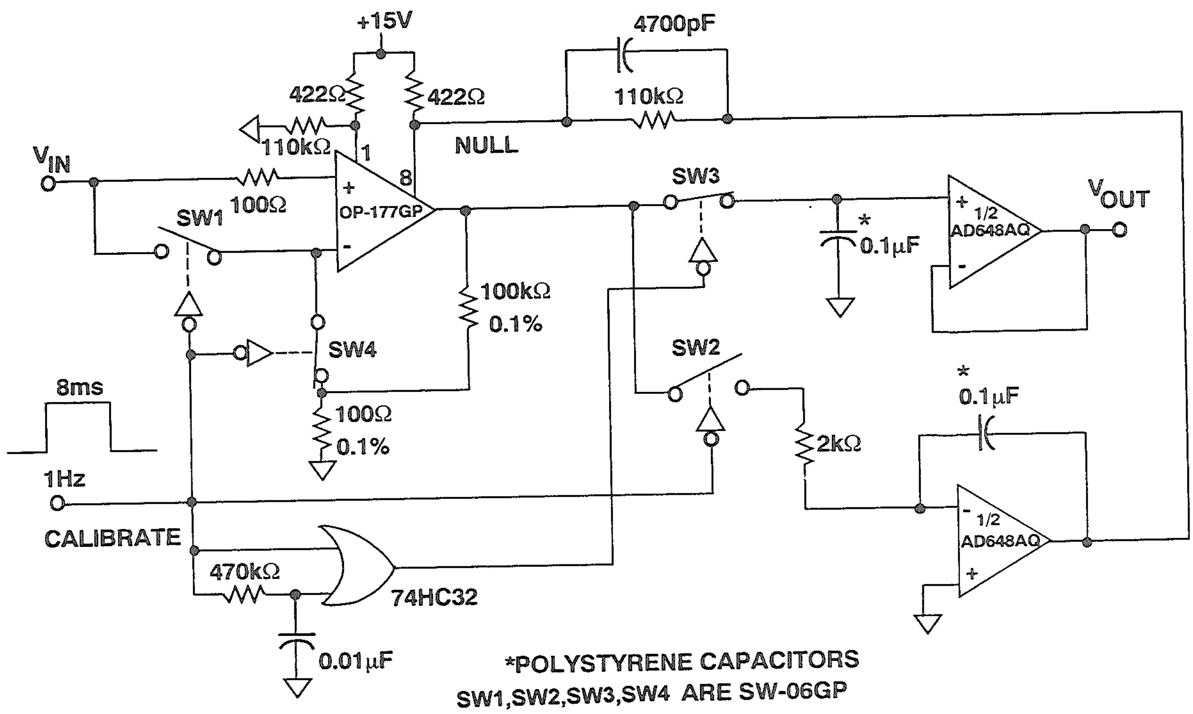
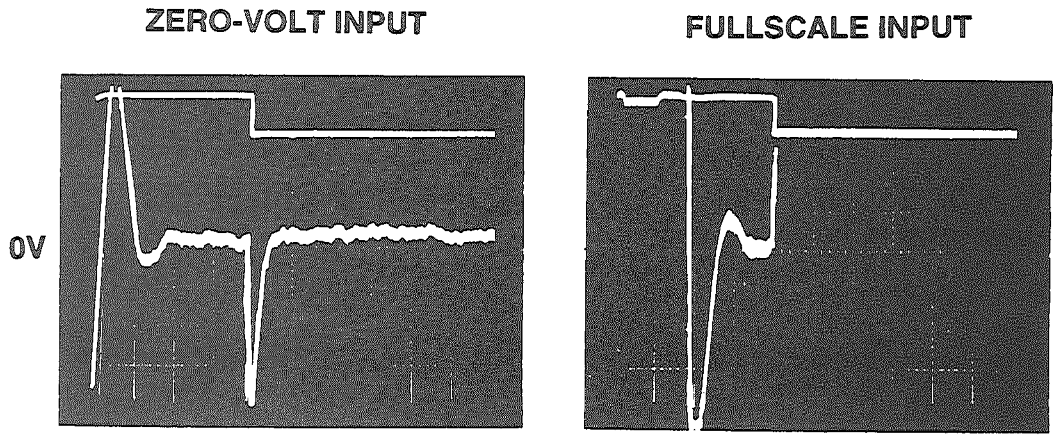


Figure 2.29

The amplifier operates at a gain of 1001 (although it can be set at any gain by adjusting the feedback resistors ratio). In normal operation, switches SW3 and SW4 closed, while SW1 and SW2 are open. The auto-zero cycle begins when a logic high is asserted at the CALIBRATE input for 8ms, toggling all 4 switches simultaneously. The opening of SW3 holds the most recent voltage at the output. At the same time the closing of SW1 shorts the two inputs of the OP-177 together, while SW4 opens the OP-177 loop. Nulling occurs as SW2 closes the OP-177's null circuit around the AD648 error amplifier. It servos the null pin (pin 8) until the OP-177 output goes to zero (plus or minus the input offset voltage of the AD648). The 4700pF shunting capacitor across the 110kΩ null bias resistor provides feedforward to speed up the OP-177 response. Otherwise the frequency response of the OP-177 will approach the cutoff frequency of the AD648 and cause a low frequency oscillation. 8ms is allowed for the servo loop to settle. The photos in Figure 2.30 depict this servo action at the OP-177 output.

AUTO-ZERO ACTION



TOP TRACES: CALIBRATE INPUT
BOTTOM TRACES: OP-177 OUTPUT

VERTICAL SCALE: 5mV / div. OR 5μV / div. (RTI)
HORIZONTAL SCALE: 2ms / div.

Figure 2.30

At the end of the 8ms auto-zero cycle, the CALIBRATE input logic goes low. All switches toggle except for SW3, thereby holding the nulling voltage at the AD648 error amplifier until the next auto-zero cycle. The closing of SW3 is delayed 5ms by the 470kΩ resistor and the 0.01μF capacitor to allow the OP-177 to settle after its loop is closed. This delayed switch closure minimizes disturbances at the sample and hold output.

The calibration cycle can be controlled by a microprocessor during its idle time and when the amplifier is not making a measurement. Alternatively, a simple timer circuit operating at 1 or 2 Hz can periodically calibrate the circuit. Any ambient temperature-induced drift is corrected by the periodic auto-zeroing. The circuit maintains a constant offset of less than 5μV referred to the input (RTI) regardless of temperature.

DESIGNING FOR LOW NOISE

Besides offset voltage error and drift, noise is a primary concern for precision amplifier applications. Highly accurate instruments require the noise floor to be as low as possible. In addition to choosing the lowest noise amplifier for a given circuit, there are other techniques which will keep noise to a minimum. A more detailed treatment of op amp noise is given in Section XI.

DESIGN FOR LOW NOISE

- **Choose A Low Noise Amp With Low 1 / f Corner Frequency and low peak-to-peak voltage noise in 0.1 to 10Hz bandwidth**
- **Use As Small Input Resistance As Practical To Reduce Johnson Noise And Effects Of Current Noise**
- **Limit Bandwidth**

Figure 2.31

SELECTING A LOW NOISE OP AMP

Before selecting the appropriate low noise op amp, the op amp source impedance and feedback resistances must be known. For high gain amplifiers that measure low level voltages, low voltage noise is important. For high source resistances, low current noise may become more critical. Figure 2.32 compares input offset voltage and 1kHz input voltage noise for a number of popular precision op amps.

OP AMP NOISE AND OFFSET VOLTAGE CHARACTERISTICS

OP AMP PART #	INPUT NOISE VOLTAGE SPECTRAL DENSITY @ 1kHz, nV/$\sqrt{\text{Hz}}$	INPUT OFFSET VOLTAGE, µV
AD797	0.9	100
SSM-2017	0.95	1200
AD743	2.9	1000
AD745	2.9	250
OP-27	3.0	25
OP-37	3.0	25
OP-270	3.2	75
OP-470	3.2	400
OP-50	4.5	25
OP-471	6.5	800
OP-07	7.0	25
OP-271	7.6	200
AD645	9.0	250
OP-177	9.6	10
AD707	9.6	15
OP-77	9.6	25

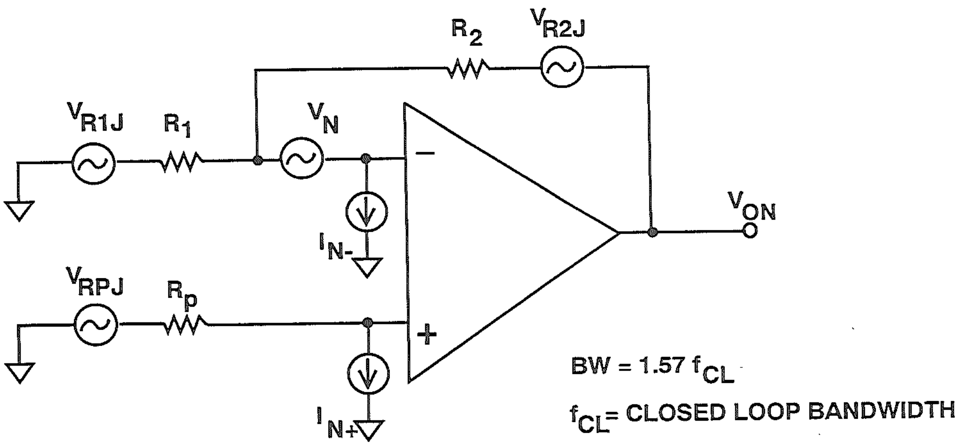
Figure 2.32

DETERMINE THE SOURCE RESISTANCE

While the amplifier's voltage noise is important, external resistances that are connected to the amplifier inputs may also generate additional noise at the op amp output. This noise is caused by the amplifier's input current noise as well as the individual resistors' thermal (Johnson) noise. All noise components are then root-sum-squared to calculate the total noise at the output of the op amp.

Virtually all amplifier circuits can be simplified to an equivalent circuit model that contains all noise sources as shown in Figure 2.33 .

OP AMP NOISE MODEL

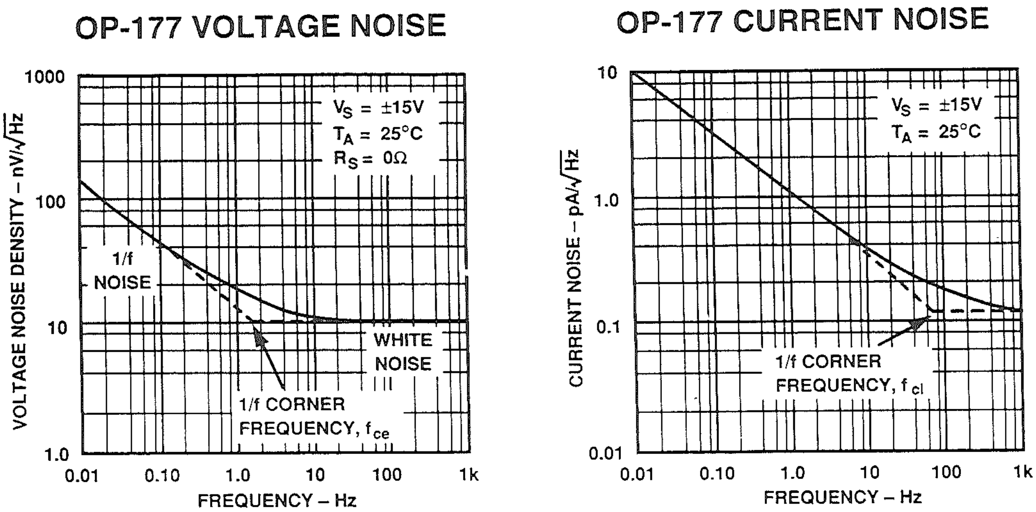


$$V_{ON} = \sqrt{BW}\sqrt{I_{N-}^{2} R_{2}^{2} + I_{N+}^{2} R_{p}^{2}\left[1+\frac{R_2}{R_1}\right]^2 + V_N^{2}\left[1+\frac{R_2}{R_1}\right]^2 + 4kTR_2 + 4kTR_1\left[\frac{R_2}{R_1}\right]^2 + 4kTR_p\left[1+\frac{R_2}{R_1}\right]^2}$$

Figure 2.33

Each noise component must be integrated over the appropriate bandwidth . The typical noise spectral density of an op amp is shown in Figure 2. 34 for the OP-177.

OP-177 INPUT NOISE SPECTRAL DENSITIES



$$E_n(f_H,f_L) = e_{nW}\sqrt{f_{ce}\cdot \ln\left(\frac{f_H}{f_L}\right) + (f_H - f_L)}$$

$$I_n(f_H,f_L) = i_{nW}\sqrt{f_{ci}\cdot \ln\left(\frac{f_H}{f_L}\right) + (f_H - f_L)}$$

Figure 2.34

The total rms output noise may be calculated using the equations shown in Figure 2.33 and Figure 2.34. A detailed description of this process is given in Section XI.

The above model may be used to estimate the RMS value of the output noise for virtually any amplifier circuit. Note that minimizing noise not only requires the use of low source resistances and low noise op amps, but also keeping the bandwidth as low as practical for the application. For very low bandwidth applications, choose an op amp that has a low voltage or current noise corner frequency, depending on which is the dominant contributor in the circuit.

It is important to note that JFET input op amps often have higher voltage noise and higher noise corner frequencies than bipolar op amps. For example, most JFET op amps' voltage noise is in the range of 30 to 40 nV/ √Hz with a noise corner frequency ranging from 100Hz to 500Hz. Bipolar op amps have a voltage noise ranging from a few nV/√Hz to 20 nV/√Hz, and their noise corner frequency is usually less than 100Hz.

New JFET input op amps have noise voltages and corner frequencies comparable with bipolar op amps. This trend is exemplified by the new AD745 ultra-low noise BiFET op amp. It achieves 2.9 nV/ √Hz voltage noise at 10kHz with a noise corner frequency of 50Hz. The combination of this low noise voltage and its 6.9 fA/√Hz noise current is unsurpassed for a FET-input monolithic op amp. Key specifications for the AD745 are given in Figure 2.35.

AD745 LOW NOISE, HIGH SPEED BiFET OP AMP

- Low Noise Voltage Density .... 2.9 nV/$\sqrt{Hz}$ @ 10kHz
- Low Voltage Noise .... 0.38 μVp-p, 0.1Hz to 10Hz
- Ultra-Low Current Noise .... 6.9 fA/$\sqrt{Hz}$ @ 1kHz
- High Gain-Bandwidth 20MHz
- High Slew Rate .... 12.5 V/μs

Figure 2.35

The benefits of these specifications become evident as the source resistance goes up, as in the case of a high impedance bridge or a high impedance transducer such as an ultrasound sensor. In these high impedance applications, the current noise may be a significant contributor to the output noise. This is where the AD745's ultra-low voltage and current noise combination can keep the noise floor at a minimum, in spite of the presence of a high source resistance. Figure 2.36 illustrates how the source resistance effects the input noise for both the bipolar OP-37 and the BIFET AD745. Note that for high values of source resistance, the AD745 is the best choice.

CURRENT NOISE IS IMPORTANT AS SOURCE RESISTANCE INCREASES. AD745 COMBINES LOW VOLTAGE AND CURRENT NOISE TO ACCOMMODATE A WIDE RANGE OF SOURCE RESISTANCES

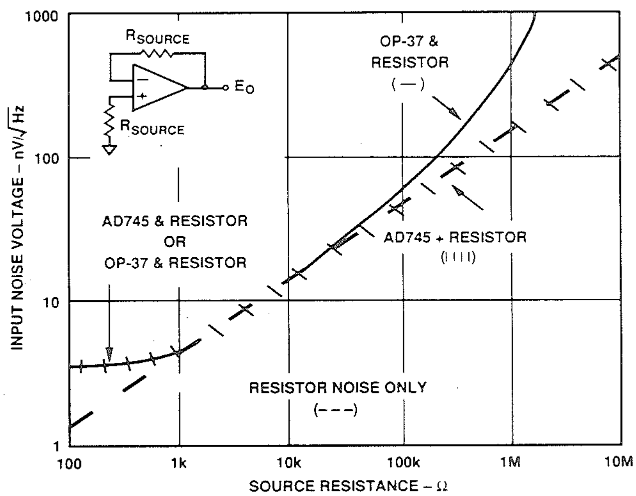


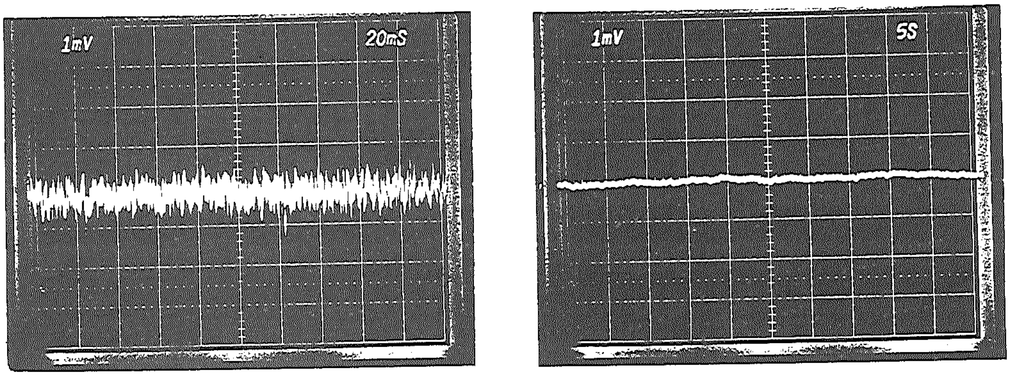
Figure 2.36

LIMIT BANDWIDTH TO REDUCE NOISE

As the noise equations suggest, noise is proportional to the square root of the circuit bandwidth. Therefore, noise can be minimized by limiting the bandwidth to that actually required by the signal. For example, if the measuring bandwidth of interest is only 1Hz, then signals above this frequency may be filtered to reduce the noise. Figure 2.37 shows the dramatic difference between the output noise level with a 1Hz lowpass filter and a 6kHz lowpass filter.

LIMITING BANDWIDTH TO REDUCE NOISE
OP-177 @ GAIN = 1000

LOWPASS FILTER = 6kHz | LOWPASS FILTER = 1Hz



VERTICAL SCALE: 1mV / div.
HORIZONTAL SCALE: 20ms / div.

VERTICAL SCALE: 1mV / div.
HORIZONTAL SCALE: 5s / div.

Figure 2.37

Because the gain-bandwidth product of a voltage feedback op amp is constant, high gains imply proportionally lower closed loop bandwidths. This means that the amplifier's own reduced bandwidth can serve as a low-pass filter at high gains. In most cases this effect helps to reduce the output noise.

CHOOSING INSTRUMENTATION AMPLIFIERS

When amplifying low level signal in the presence of high common-mode voltages, such as in the case of a bridge circuit, an instrumentation amplifier may provide an efficient high performance solution. Although an instrumentation amplifier can be built easily with the traditional two or three op amp configurations, monolithically integrated designs offer superior performance at low cost as shown in Figure 2.38 . Monolithic instrumentation amplifiers such as the AD620 do not require the costly external resistor matching. On chip thin film laser trimmed resistors allow the gain to be set accurately with a single external resistor.

CHOOSING INSTRUMENTATION AMPLIFIERS

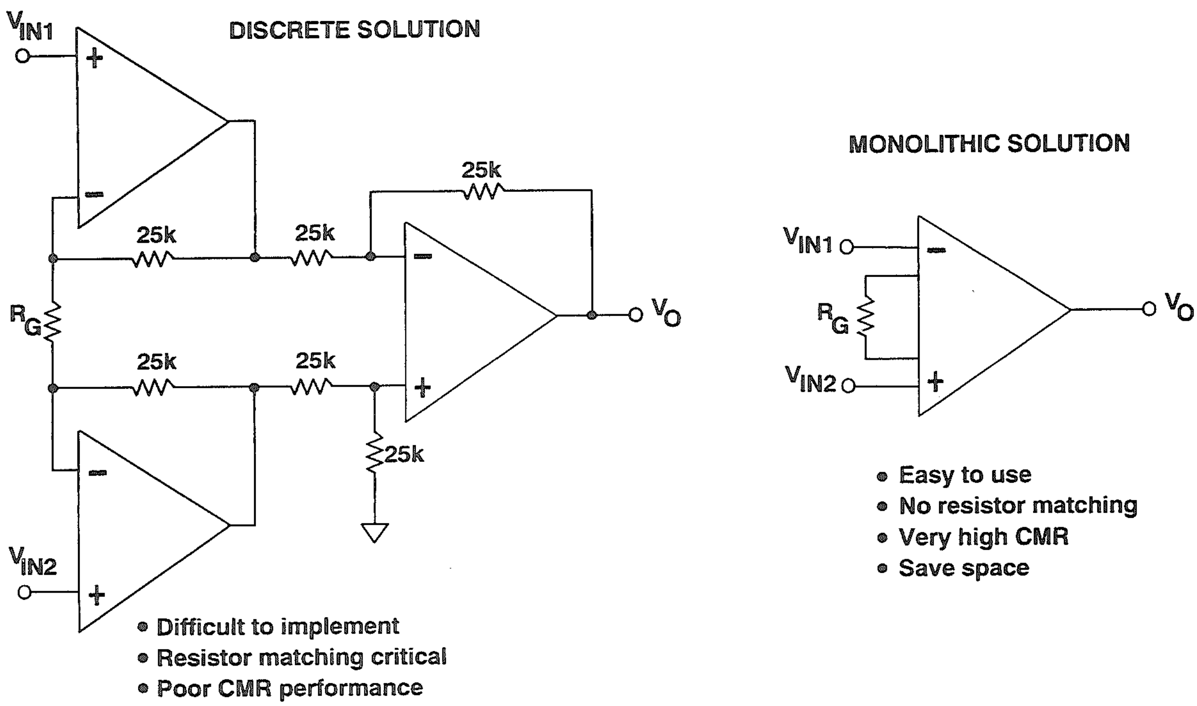


Figure 2.38

Instrumentation amplifiers may be implemented in discrete form if high performance levels are not required and printed circuit board area is available to accommodate the additional components required. For example, dual or quad op amps and 1% resistors may be used to build an instrumentation amplifier as shown in Figure 2.39.

This Super-β input instrumentation amplifier offers many performance benefits including low input bias currents, low input offset, and low offset voltage drift. It consumes only 1.0 mA quiescent current. For gains other than shown in the table, use the following equations:

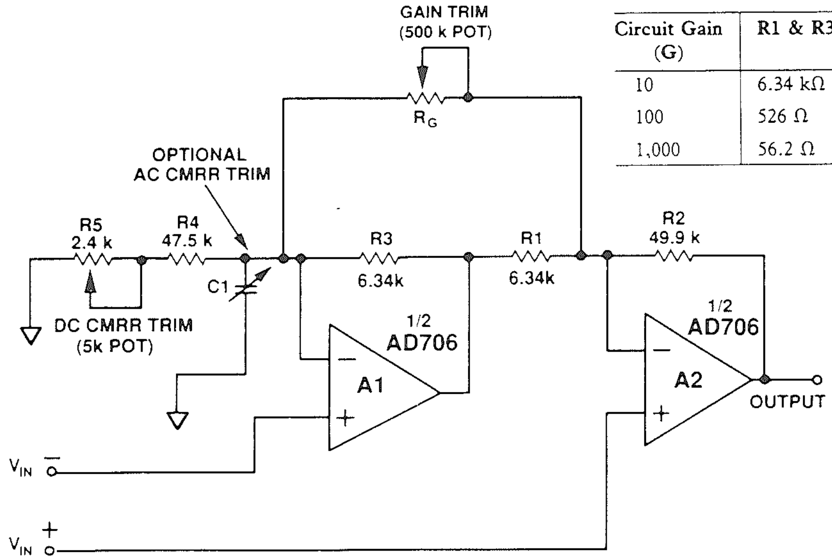
$$R2 = R4 + R5 = 49.9k\Omega$$

$$R1 = R3 = \frac{49.9k\Omega}{0.9G-1}$$

$$\text{Max Value of } R_G = \frac{99.8k\Omega}{0.06G}$$

$$C1 \sim \frac{1}{2\pi(R3)5 \times 10^5}$$

A DUAL OP AMP INSTRUMENTATION AMPLIFIER



Circuit Gain (G)	R1 & R3	R_G (Max Value of Potentiometer)	Bandwidth (−3 dB), Hz
10	6.34 kΩ	166 kΩ	50 k
100	526 Ω	16.6 kΩ	5 k
1,000	56.2 Ω	1.66 kΩ	0.5 k

$$G = 1 + \frac{R2}{R1} + \frac{2R2}{RG} \quad \text{(FOR R1 = R3, R2 = R4 + R5)}$$

ALL RESISTORS METAL FILM, 1%

Figure 2.39

The AC common mode rejection ratio maybe improved by adding the trim capacitor C1. The effects of adding this trim are shown in Figure 2.40.

AC COMMON-MODE REJECTION ENHANCEMENT USING A TRIMMING CAPACITOR

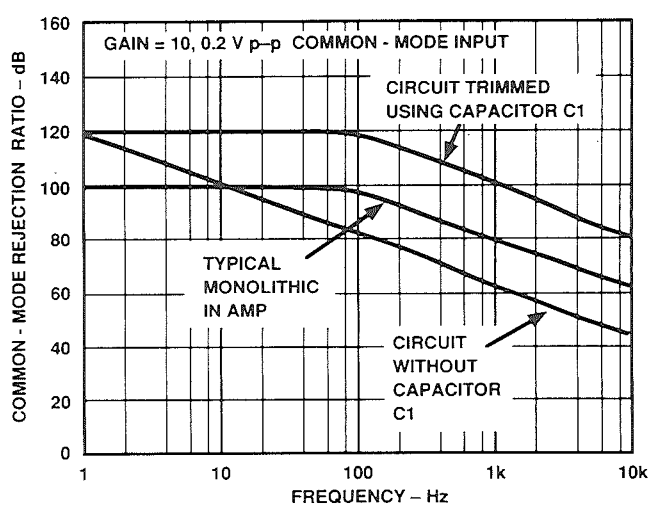
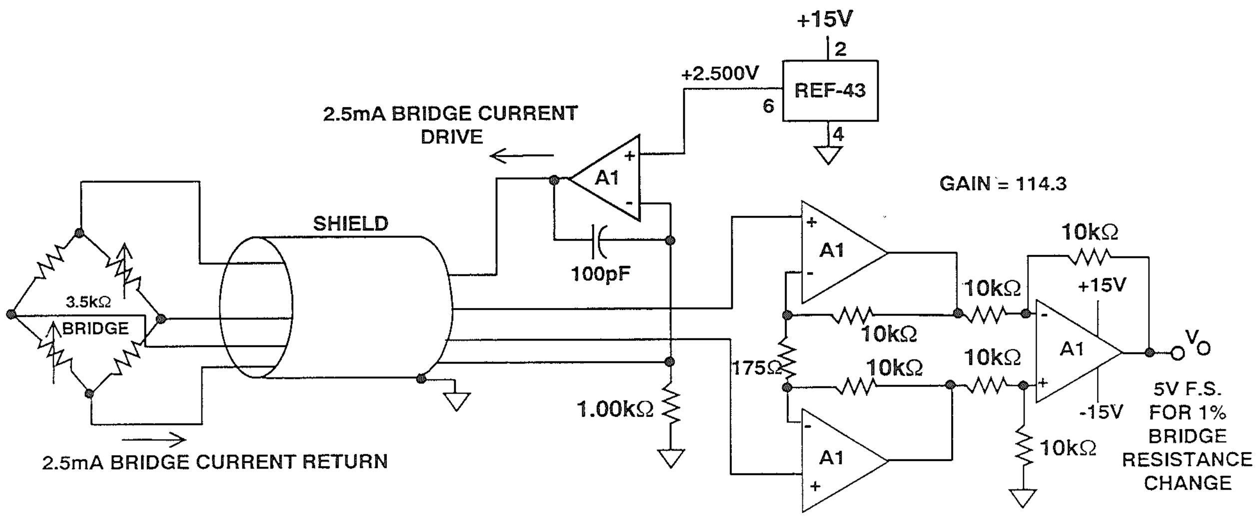


Figure 2.40

A three op amp instrumentation amplifier used in a remote bridge sensing application is shown in Figure 2.41.

3-OP AMP INSTRUMENTATION AMPLIFIER WITH REMOTE BRIDGE SENSOR



■ A1 = OP-497 OR AD704

■ ALL RESISTORS 1% OR BETTER

Figure 2.41

As precision and accuracy requirements increase, monolithic instrumentation amplifiers are much more cost effective and space efficient. Many designs have built-in gain setting resistors. Gains may be set either digitally by pin strapping. Some monolithic instrumentation amplifiers offer protection from input overvoltage. A comprehensive set of selection tables for instrumentation amplifiers is given in Figures 2.42 to 2.45.

2

INSTRUMENTATION AMPLIFIER SELECTION TABLE

HIGH PRECISION (G = 1000):

	GAIN ACCURACY %	V_{OS} µV	V_{OS} TC µV/°C	CMRR dB @ 60Hz	INPUT NOISE (0.1-10Hz) µV p-p
AD524	0.5	50	0.5	115	0.3
AD620	0.5	50	0.6	120	0.28
AD624	1.0	25	0.25	130	0.2
AD625	0.02	25	0.25	120	0.2
AMP-01	0.6	50	0.3	125	0.12
AMP-02	0.5	100	2	115	0.4

Figure 2.42

INSTRUMENTATION AMPLIFIER SELECTION TABLE -- CONTINUED

SINGLE SUPPLY

	SUPPLY RANGE	SUPPLY CURRENT	V_{OS}	CMRR @ 60Hz
AD626	+2.5→±7V	500µA	500µV	92dB
AMP-04	+4.5→±18	800µA	100µV	100dB

INPUT OVERVOLTAGE PROTECTED:

	Max Input Voltage
AD524	±36V
AD626	+54V
AMP-02	±60V

Figure 2.43

INSTRUMENTATION AMPLIFIER SELECTION TABLE -- CONTINUED

8-PIN PACKAGE:

	GAIN ACCURACY, %	V_{OS}, µV	CMRR, dB @60Hz
AD620	0.5	50	120
AD626	0.5	500	92
AMP-02	0.5	100	115
AMP-04	0.2	100	100

Figure 2.44

INSTRUMENTATION AMPLIFIER SELECTION TABLE -- CONTINUED

HIGH INPUT IMPEDANCE:

	INPUT RESISTANCE	INPUT BIAS CURRENT
AMP-05	$10^{12}\Omega$	50pA

DIGITALLY PROGRAMMABLE GAIN:

	GAIN RANGE	GAIN ACCURACY
AD365	1,10,100,500	0.1% Max @ G = 500

Figure 2.45

CMRR PERFORMANCE CRITICAL

Chief among parameters to look for in instrumentation amplifiers is the common-mode rejection performance, particularly over the rated operating temperature range.

Many monolithic instrumentation amplifiers employ the 3 op amp topology. The design is straight-forward, but depends heavily on stable on-chip laser trimmed resistors and tight resistor matching to achieve high common-mode rejection. Current feedback topology such as the AMP-01 as shown in Figure 2.46 solves some of these problems and yields exceptional CMRR performance over temperature.

AMP-01 USES CURRENT FEEDBACK TO ACHIEVE 125dB CMR @ GAIN = 1000

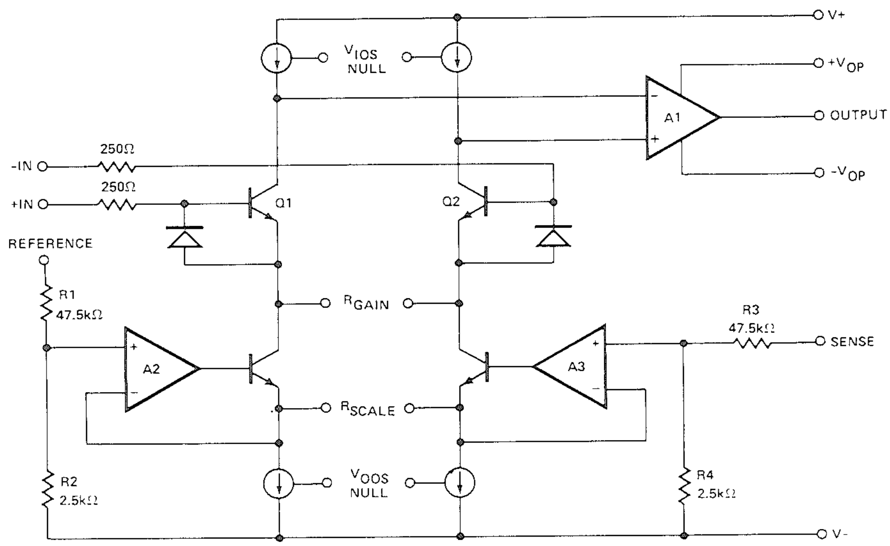


Figure 2.46

EASE OF USE IMPORTANT

Monolithic instrumentation amplifiers are quite simple to use. New designs such as the AD620 offered in 8-pin dual-in-line packages are even easier to use. Only a single external resistor is required to set the gain. The amplifier characteristics are highly predictable because the performance is fully specified as a complete unit, unlike discrete solutions. Key specifications for the AD620 instrumentation amplifier are summarized in Figure 2.47.

AD620
LOW POWER, PRECISION IN-AMP IN 8-PIN SO

- One Resistor Sets Gain From 1 to 1,000
- Low Offset Voltage . . . . .125μV Max
- Low Offset Voltage Drift . . . . . . 1.0μV/°C Max
- CMRR . . . . 93dB, G = 10
- Wide Specified Supply Range . . . ±5V to ±15V
- Low Power . . . . 900μA typ.
- Low Cost

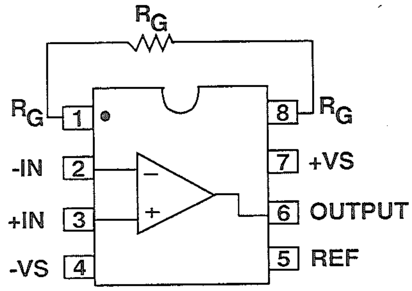


Figure 2.47

INPUT PROTECTION

The inputs of an instrumentation amplifier often come from a harsh environment. For example, unsafe voltages may be present in a factory and damage unprotected amplifier inputs if applied directly. If such a possibility exists, the inputs should be protected. The simplest method is to add resistor in series with each input to limit destructively high levels of current. One should always consult the Absolute Maximum Rating specification in the device data sheet for the maximum allowable input current. If not specified, 10mA is usually safe and may be applied under all overvoltage conditions. Figure 2.48 shows a protection scheme that will withstand +/-160V common-mode and differential overvoltage.

INPUT OVERVOLTAGE PROTECTION

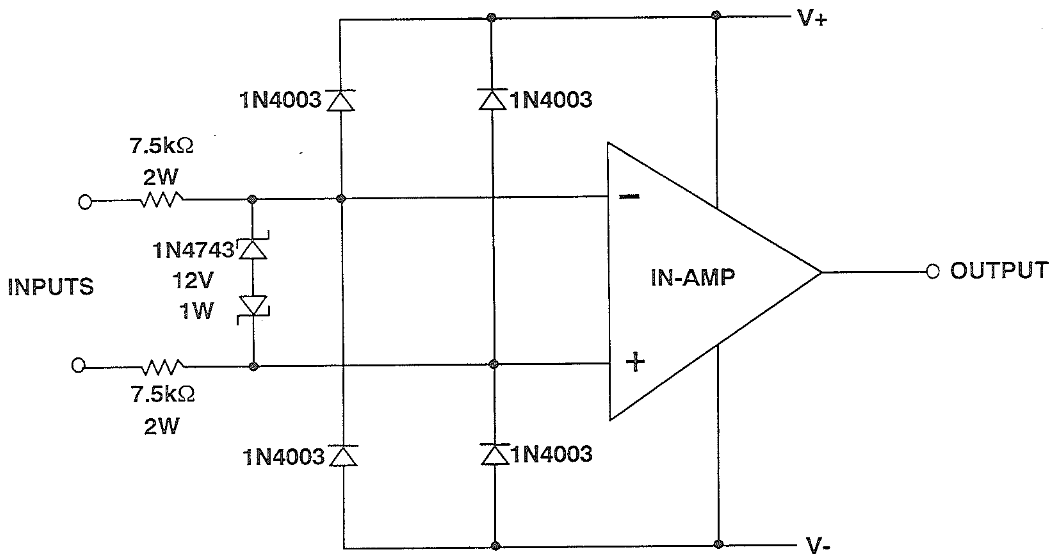


Figure 2.48

While the series input resistors provide overvoltage protection, they also increase the total output noise of the amplifier. Reducing the resistor values reduces the noise but the requires a higher wattage resistor to dissipate the power under full overvoltage conditions.

As mentioned before, some instrumentation amplifiers have built-in input protections. Figure 2.49 and Figure 2.50 show the two different input protection implementations of two different instrumentation amplifiers, the AD524 and the AMP-02, respectively. If additional protection is needed, a series resistor will help.

AD524's INTERNALLY PROTECTED INPUTS CAN WITHSTAND 36V BEYOND EITHER SUPPLY RAIL

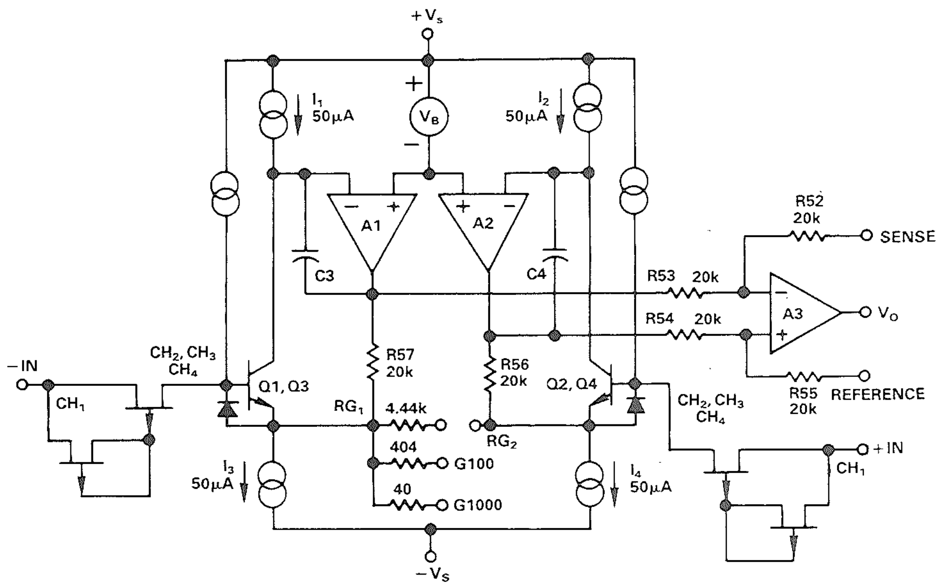


Figure 2.49

AMP-02's INTERNALLY PROTECTED INPUTS ALLOW ± 60V WITHOUT DAMAGE

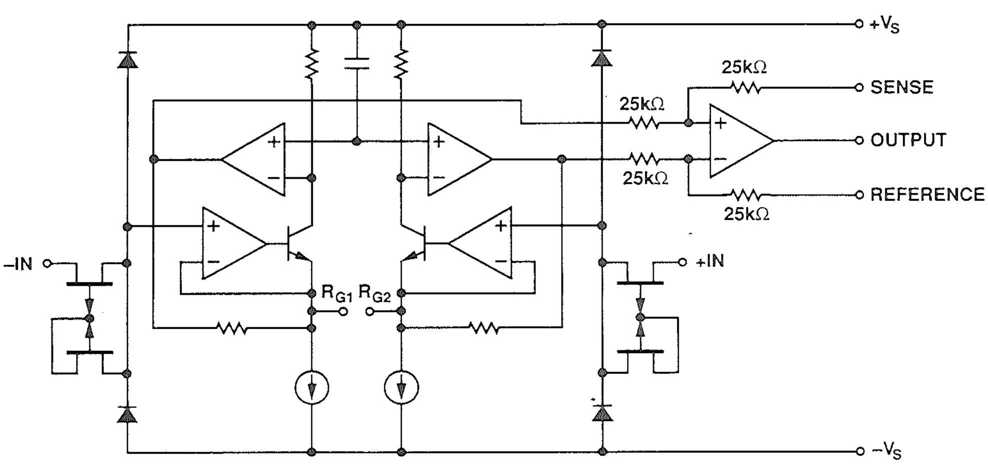


Figure 2.50

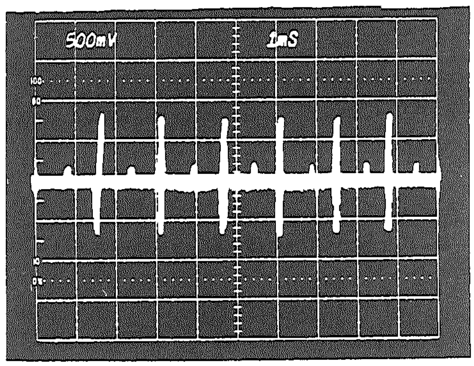
CHOPPER STABILIZED AMPLIFIERS

Chopper stabilized amplifiers offer virtually zero offset voltage and offset voltage drift by continuously nulling the offset. While this solves the problem of offset and drift, chopper stabilized amplifiers generate significantly more noise than their counterparts.

There are two sources of noise: glitch noise and intrinsic amplifier noise. The glitch noise amplitude is typically in the 100's of millivolts and is generated by the internal chopping action feeding through to the output. The chopping frequency is usually between a few hundred Hz to greater than 5kHz. Figure 2.51 shows the glitch noise that can be present at the output of a typical chopper amplifier. Adequate filtering is required to remove these glitches.

CHOPPER AMPLIFIERS HAVE SIGNIFICANT GLITCH NOISE AND REQUIRE OUTPUT FILTER

GAIN = 1000



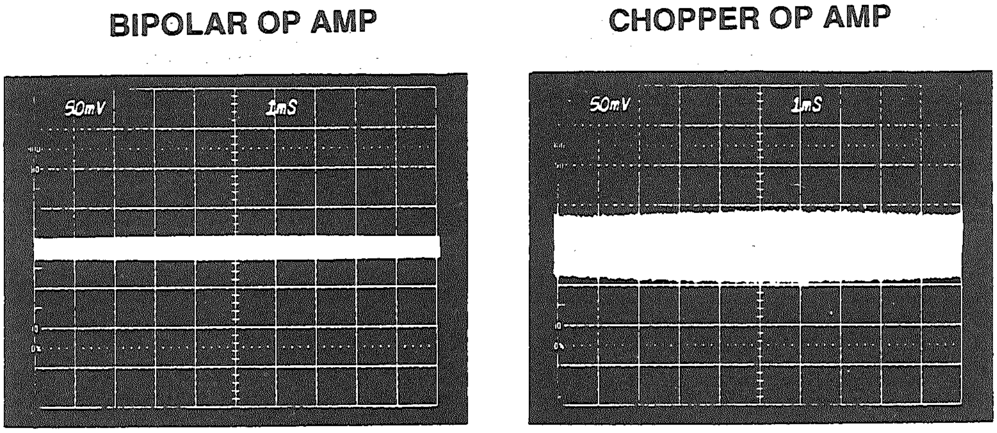
VERTICAL SCALE: 500mV / div.
HORIZONTAL SCALE: 1ms / div.

SCOPE SYNCHRONIZED
TO CHOPPER FREQUENCY

Figure 2.51

The other noise comes from the 1/f noise and the white noise of the amplifier, which is usually higher than low noise op amps. Figure 2.52 shows the unfiltered noise of a typical chopper stabilized amplifier. If not filtered, it significantly degrades the signal-to-noise performance of an amplifier circuit. Usually a two pole filter is required to reduce this noise to an acceptable level.

BIPOLAR OP AMP VERSUS CHOPPER WIDEBAND (UNFILTERED) NOISE G = 1000



VERTICAL SCALE: 50mV / div.
HORIZONTAL SCALE: 1ms / div.

SCOPE UNSYNCHRONIZED

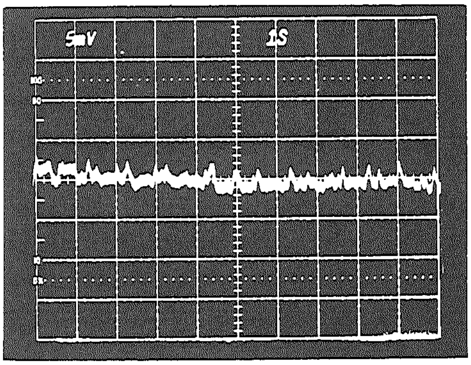
Figure 2.52

It is a common misconception that chopper stabilized amplifiers have inherently lower low-frequency noise because the low frequency noise behaves like an offset and therefore is nulled by the chopping action. This is incorrect because the switching action occurs at the inputs of the amplifier. The finite switch resistances and current noise spikes from the chopping action result in significantly higher noise below 10Hz. Figure 2.53 compares the 10Hz noise of a chopper with a bipolar precision op amp. The OP-177 has lower 10Hz noise than chopper stabilized amplifier.

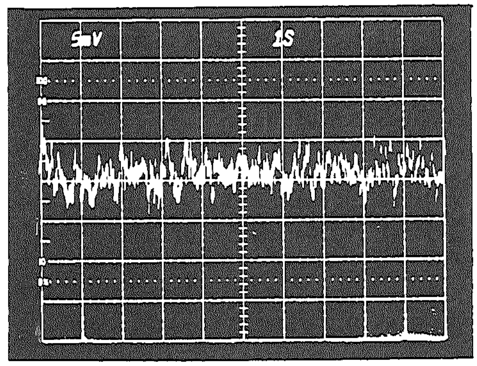
While chopper stabilized amplifiers do serve a useful solution in some applications where offset and drifts are critical, the designer should understand the tradeoffs before a selection is made. A useful comparison between chopper and a precision op amps is given in Figure 2.54.

BIPOLAR OP AMP VERSUS CHOPPER 0.1Hz TO 10Hz NOISE FOR GAIN = 1000

BIPOLAR OP AMP



CHOPPER OP AMP



VERTICAL SCALE: 5mV / div.
HORIZONTAL SCALE: 1s / div.

SCOPE UNSYNCHRONIZED

Figure 2.53

PRECISION OP AMP VERSUS CHOPPER: AN UNBIASED COMPARISON OF CRITICAL PARAMETERS

	BIPOLAR	CHOPPER
OFFSET VOLTAGE	10-50μV	<5μV
OFFSET DRIFT	0.1μV/°C	~0μV
OPEN LOOP GAIN	10 Million	10 Million
NOISE: HF GLITCH	None	>100mVp-p
NOISE: 0.1-10Hz	<0.2μVp-p	>1μVp-p
COST	Lower	Higher
EXTERNAL COMPONENTS	None	Most Require 2 Caps.
SATURATION RECOVERY	10-20μs	>100ms to seconds

Figure 2.54

SIGNAL CONDITIONING FOR TEMPERATURE MEASUREMENT

ADOLFO A. GARCIA

THERMOCOUPLE SIGNAL CONDITIONING

TEMPERATURE-VOLTAGE CHARACTERISTICS OF THERMOCOUPLES

Thermocouples are simple temperature sensors formed when any two dissimilar metals are joined together. They are very commonly used in hostile environments because their rugged construction can withstand wide temperature ranges without damage. These sensors are available in many varieties depending on temperature range and their low cost makes them one of the most commonly used devices for temperature measurement. When compared to other temperature sensors, thermocouples are quite linear. However, their low output voltages require very carefully designed, stable signal conditioning circuitry to keep measurement errors low. It is precisely this issue of signal conditioning for thermocouples that requires an understanding of the sensors themselves.

Illustrated in Figure 2.55 are voltage-temperature curves of four commonly used thermocouples. 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, constructed out of a noble metal alloy of platinum/platinum—rhodium, are the least sensitive to temperature changes. 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 stages.

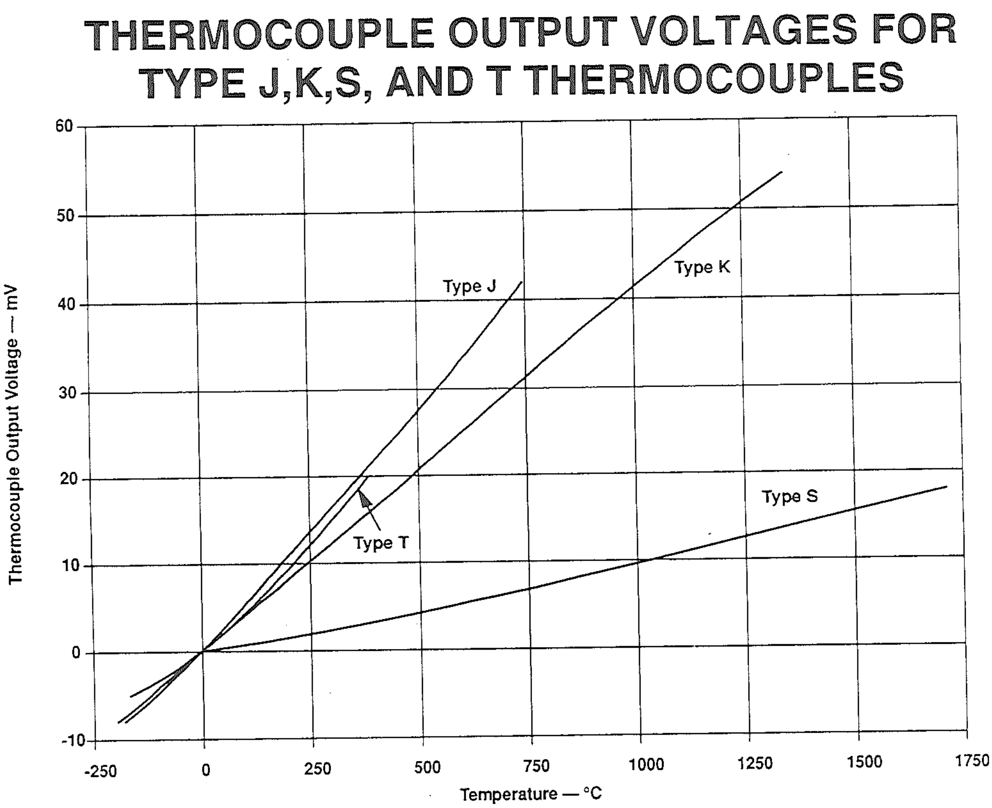


Figure 2.55

Most important in understanding thermocouple response is their non-linear temperature coefficient, or Seebeck coefficient. As was shown in Figure 2.55, each of the four thermocouple responses exhibit a non-linear output voltage versus temperature; however, the key to thermocouple signal conditioning is understanding the behavior of the coefficient of temperature and how it varies over temperature. Shown in Figure 2.56 are the Seebeck coefficients of the four thermocouples as a function of temperature. The important concept to remember when choosing a thermocouple for maximum linearity is selecting the temperature range where a thermocouple's Seebeck coefficient changes are small.

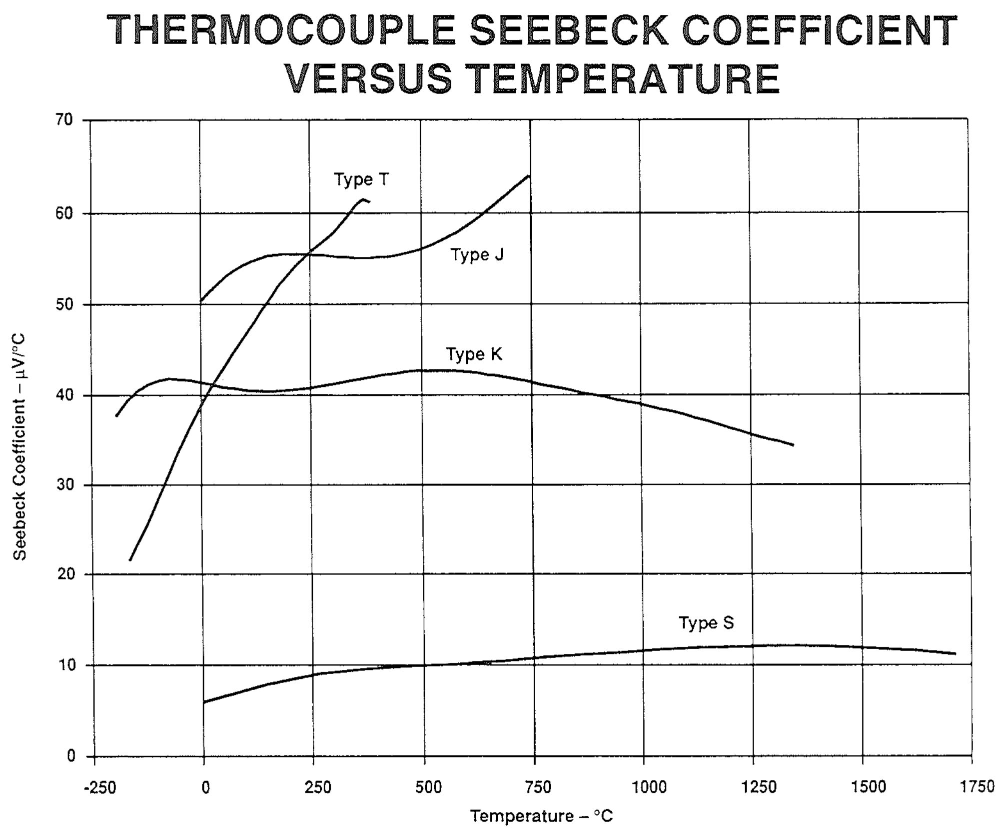


Figure 2.56

For example, a Type J thermocouple's Seebeck coefficient changes less than 1 μV/°C over any part of the range from 200°C to 500 °C. On the other hand, the Type T thermocouple's Seebeck coefficient increases non-linearly with increasing temperature; therefore, designing signal conditioning circuitry around the Type T thermocouple to cover accurately a wide temperature range would be a difficult task.

Presenting these data on thermocouples serves two purposes: Firstly, Figure 2.55 illustrates the range and sensitivity of the four thermocouple types. Thus, 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 to temperature changes. Secondly, the variations of Seebeck coefficients provide a quick guide to a thermocouple's linearity, both from a qualitative point of view and a quantitative one. For example, using Figure 2.56, 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 the variations of temperature are key instead of the absolute magnitude. These data also indicate what level of performance is required of the associated signal conditioning circuitry.

PRINCIPLE OF THERMOCOUPLE AND REFERENCE COLD-JUNCTION

Although external excitation is not required because of the inherent voltage output of the thermocouple, a stable reference is required to accurately measure the temperature of an unknown medium. Thus, in order to apply the Seebeck effect to temperature measurement, one junction must remain at a fixed reference temperature. Figure 2.57 illustrates one approach to maintaining a fixed reference temperature at the Type J thermocouple's terminating junction. This solution requires ice-point references which are very accurate and easy to build; however, they are very expensive and difficult to maintain. Another technique for maintaining the reference junction at 0 °C involves using a Peltier cooler to electronically simulate the ice-bath. However, this approach is very complex and bulky for most applications in that a platinum RTD (mated to the cooler) and a servo amplifier with a power stage to drive the cooler are required.

CLASSICAL COLD-JUNCTION COMPENSATION

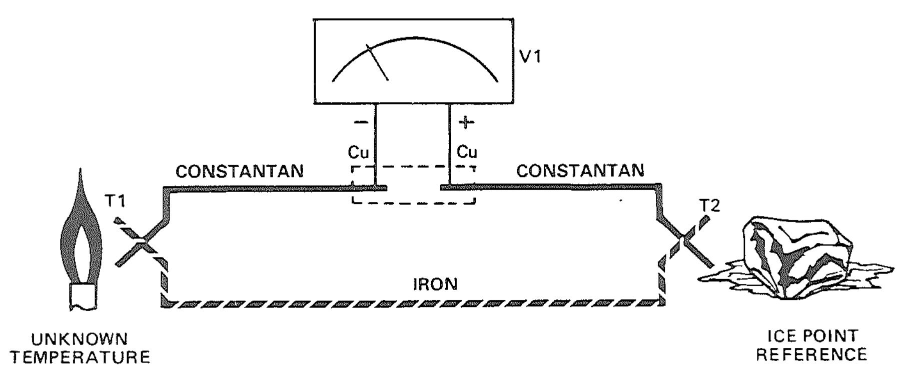


Figure 2.57

The third, and most commonly used approach, is an electronic compensation circuitry that provides an artificial reference level that is allowed to track ambient temperature variations of the reference junction. This approach is shown in Figure 2.58. A temperature sensor, T3, is placed in thermal proximity to the reference junction and generates a voltage, V3, equal to and opposite in polarity to that generated by the cold junction. This voltage is then added to the one produced by the thermocouple circuit. The result of this operation is that the voltage produced by the thermocouple at the measurement temperature. This has the same effect as maintaining the cold junction at a constant temperature, and it offers good accuracy and requires minimal maintenance.

USING A TEMPERATURE SENSOR FOR COLD-JUNCTION COMPENSATION

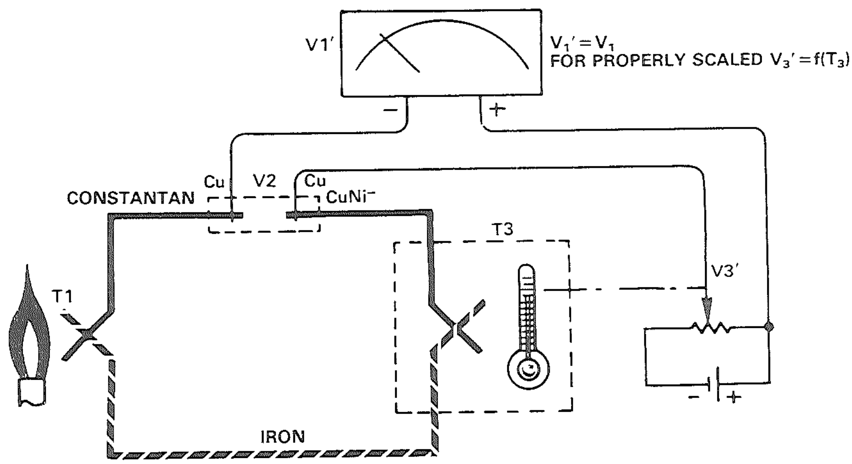


Figure 2.58

THERMOCOUPLE AMPLIFIER DESIGN CONSIDERATIONS

There are a number of ways to approach the design of thermocouple signal conditioning amplifiers. The appropriate thermocouple type for the temperature range should be chosen to reduce the accuracy requirements placed on the circuit design. In selecting the most linear thermocouple for a temperature range allows for straightforward designs with a minimum of components. Tight measurement accuracy over wide temperature ranges increases system costs by requiring more elaborate designs that linearize the thermocouple's inherently non-linear response. The keys to accurate measurements over narrow or wide temperature ranges are to use cold-junction compensation and a thermocouple whose Seebeck coefficient is linear over the temperature range of interest. It is not, however, just that straightforward — in many cases, a designer must use whatever thermocouple is at hand to do the job. To illustrate the issues of thermocouple signal conditioning, we'll begin by considering two approaches for a Type T thermocouple amplifier.

Once the measurement temperature range is known, the next step is to decide what type of an approximation will be applied to the thermocouple's non-linear response. As we have shown before, thermocouples do not have a unique and constant Seebeck coefficient over temperature. Hence, some type of linear approximation must be used. For the first example, a signal conditioning amplifier is required for a Type T thermocouple which measures temperatures over the range of 0 °C to 100 °C. As illustrated in Figure 2.59, a Type T thermocouple's response is shown with two linear approximations: end-point and least squares.

END-POINT AND LEAST-SQUARES APPROXIMATIONS TO A TYPE-T THERMOCOUPLE

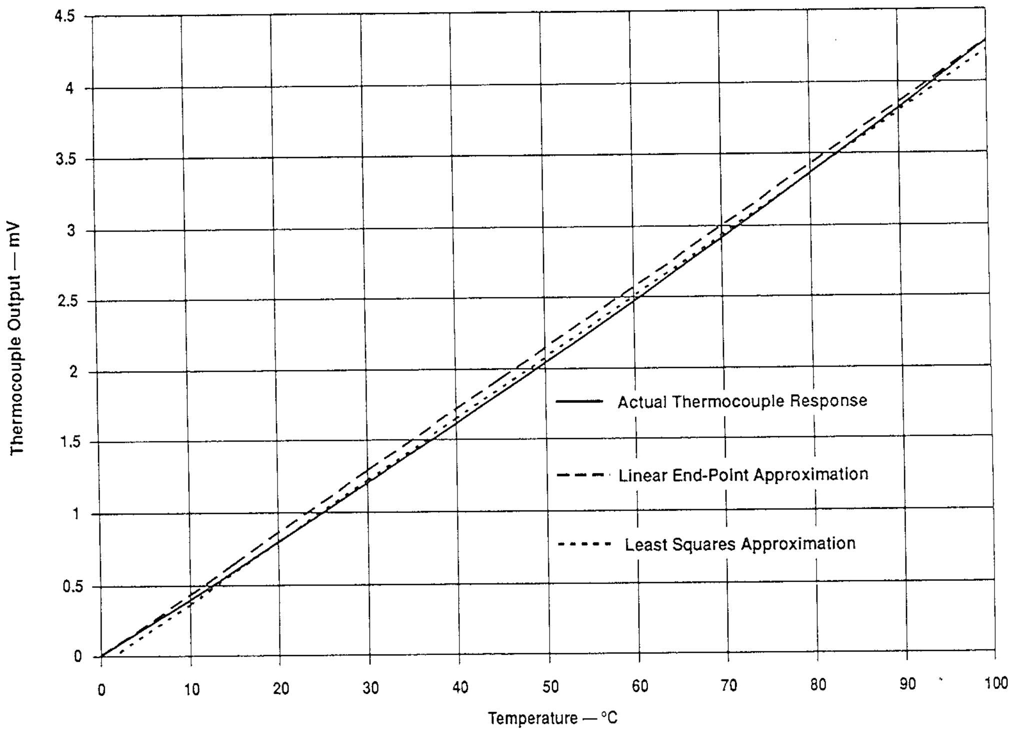


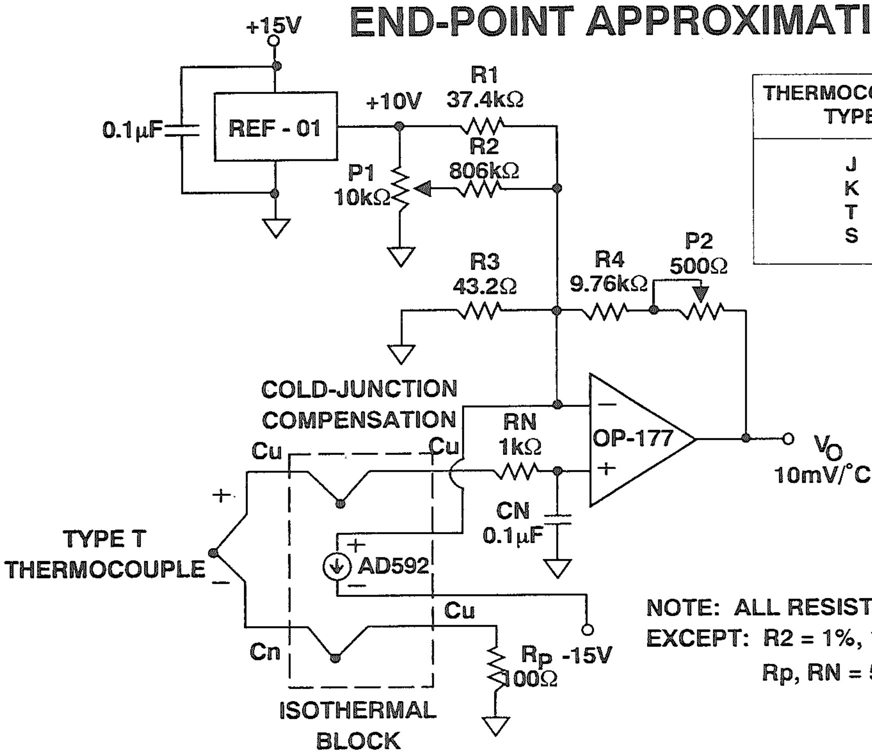
Figure 2.59

An end-point approximation applies a straight line on the thermocouple's response and intersects the two endpoints of the temperature range. The error in this approximation is zero at the endpoints and is a maximum at the midpoint of the temperature range, which is -2.5 °C at 50 °C in this example. The slope of the approximation is 42.8 μV/°C and is the Seebeck coefficient value 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 occurring at the midpoint and the endpoints. Typically what happens with this approach is that the error becomes too large at either temperature extreme, and the designer is forced to adjust the slope or the intercept of the line to reduce the overall error. Fortunately, most all scientific calculators make the job easier by providing a least squares curve fitting capability. The algorithm only works if enough data points are entered. In this example, data points were entered 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. This corresponded to a Seebeck coefficient of 42.8 μV/°C and with an intercept of -65 μV. Therefore, the only difference between the two approximations is an offset term. The only way to reduce the error in the approximation, be it end-point or least squares, is to narrow the measurement temperature range or use linearization techniques. Through this example, we have shown that linear approximations to a thermocouple's characteristic are more accurate for small deviations over temperature.

THERMOCOUPLE AMPLIFIER DESIGN #1

To demonstrate the design of a Type T thermocouple amplifier, a Seebeck coefficient of 42.8 μV/°C is used for a circuit which measures temperature over the range of 0 °C to 100 °C to an accuracy of ±0.4 °C. The amplifier, shown in Figure 2.60, is designed to provide a 10 mV/°C output and requires a gain of 233.8 to produce a 1 V output at 100 °C. Changing R3 as shown in the table allows the circuit to accommodate other thermocouple types.

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

NOTE: ALL RESISTORS ARE 1% 25 ppm/°C
EXCEPT: R2 = 1%, 100 ppm/°C
Rp, RN = 5% CARBON

Figure 2.60

A grounded thermocouple is used here because this configuration minimizes noise pickup from long leads. Two other components, R_N and C_N, serve as a noise filter to any ambient induced noise that the lead wires pick up. The filter's cutoff frequency of 1.6 kHz can be lowered by, preferably, increasing C_N. Although larger values for R_N can be used at the input of the OP-177, its value should be weighed against input bias-current induced offset and drift effects. Lastly, a resistor, R_P, is used in series with exposed thermocouples as protection in the event that it makes electrical contact to some high voltage potential. Otherwise, the thermocouple would be a dead short to ground and would certainly be destroyed without current limiting.

Cold-junction compensation is achieved using a AD592 monolithic temperature sensor located in thermal proximity to the thermocouple's terminating junction. The sensor continuously monitors the temperature of the cold-junction and provides a temperature-dependent current to the amplifier's summing junction. An offset current equal to the sensor's output current at 0 °C is generated by R1 and a REF-01, a precision 10 V reference. The offset current is summed and nulled with the sensor's initial current during calibration.

Calibration of the circuit is a two-step process. First, the circuit is allowed to warm-up and settle after 5 minutes or so to allow the temperature sensor, the reference, the resistors, and the op amp to stabilize. To protect the circuit from ambient thermal gradients or air currents that can affect calibration, it's a good idea to cover the circuit or enclose it in a box. For 0 °C calibration, the thermocouple is replaced by a short, and P1 is adjusted such that the output voltage equals the ambient temperature, according to the relationship 10 mV/°C • T_A. This part of the procedure trims out initial amplifier offset voltage and bias currents, resistor tolerances, and initial errors in the reference and the temperature sensor. For full-scale span adjust, the short is replaced by a precision dc voltage source with a value equal to the thermocouple's output voltage at 100 °C. For a Type T thermocouple, the voltage source is set to 4.277 mV. P2 is then adjusted such that the output voltage is 1V above the previously measured value; that is, V_{OUT} = 1V + 10 mV/°C • T_A. For example, if the ambient temperature of factory or a lab is 45 °C (whew!!), then P2 is adjusted so that V_{OUT} = 1.45 V.

The largest source of error over this temperature range comes from the approximation made to the thermocouple's characteristic. As we have shown before, a Type T thermocouple, although very sensitive to temperature changes, exhibits a Seebeck coefficient of 38.9 μV/°C at 0 °C that increases to 46.3 μV/°C at 100 °C. Therefore, using a constant 42.8 μV/°C over this range induces a 2.5 °C error in the measurements. For other thermocouple types, a Type K characteristic induces less than a 0.7 °C error and a Type J characteristic induces less than 1°C non-linearity error. This design is still quite very useful in applications where the output of the amplifier is digitized using software linearization with correction factors stored in a ROM lookup table.

The OP-177 is an excellent choice in 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 large static errors at the output and its low offset drift contributes less than 0.06 °C error for ambient temperatures that range from 20 °C to 50 °C; and (3) the amplifier's minimum open-loop gain of 5 million and excellent gain linearity keep the amplifier's gain error below an immeasurable 0.004 °C over the entire ambient temperature range.

THERMOCOUPLE LINEARIZATION TECHNIQUES

To improve measurement accuracy, linearization of the thermocouple's response can be applied. There are a number of techniques that can be used to linearize a thermocouple's characteristic. Some of the techniques are: offset addition, breakpoint correction, analog computation, and digital correction (Reference 1). Offset addition biases the non-linear characteristic with a constant offset. By adding an offset term, a least squares approximation is applied to the thermocouple response. Recall that the approximation's errors are high at the endpoints and low throughout the span. Offset correction can be implemented using the circuit in Figure 2.60 by adjusting the offset current and circuit gain during calibration so that the error in minimized at two points located roughly at one-third and at two-thirds of the measurement range.

Thermocouple non-linearity correction using analog computation techniques uses analog multipliers and operational amplifiers to synthesize a function that describes the thermocouple's voltage-temperature characteristic. For more detailed information on this subject, the interested reader should consult References 2 and 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 4). 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 Reference 5. Another advantage to digital techniques is the elimination of circuit calibration via 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 implementing this algorithm.

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 required by system accuracy to fit the thermocouple's characteristic. This approach is circuit intensive in that the number of linear segments determine the number of amplifiers the linearizer requires. In the next example, a design for such a linearizer will be described.

THERMOCOUPLE AMPLIFIER DESIGN #2

Shown in Figure 2.61 is a circuit that uses one breakpoint and two linear segments to approximate a Type T thermocouple's response from 0 °C to 100°C. The input amplifier is built around the AD707 and employs cold-junction compensation using an AD592 temperature sensor. The linearizing circuitry uses an AD706 dual to set the breakpoint of the approximation and changes the gain of the circuit as the signal increases. The output of the amplifier is designed to provide a 10 mV/°C output such that the output is 1 V when the thermocouple is at 100 °C.

COLD-JUNCTION COMPENSATED TYPE T THERMOCOUPLE USING A TWO-SEGMENT APPROXIMATION

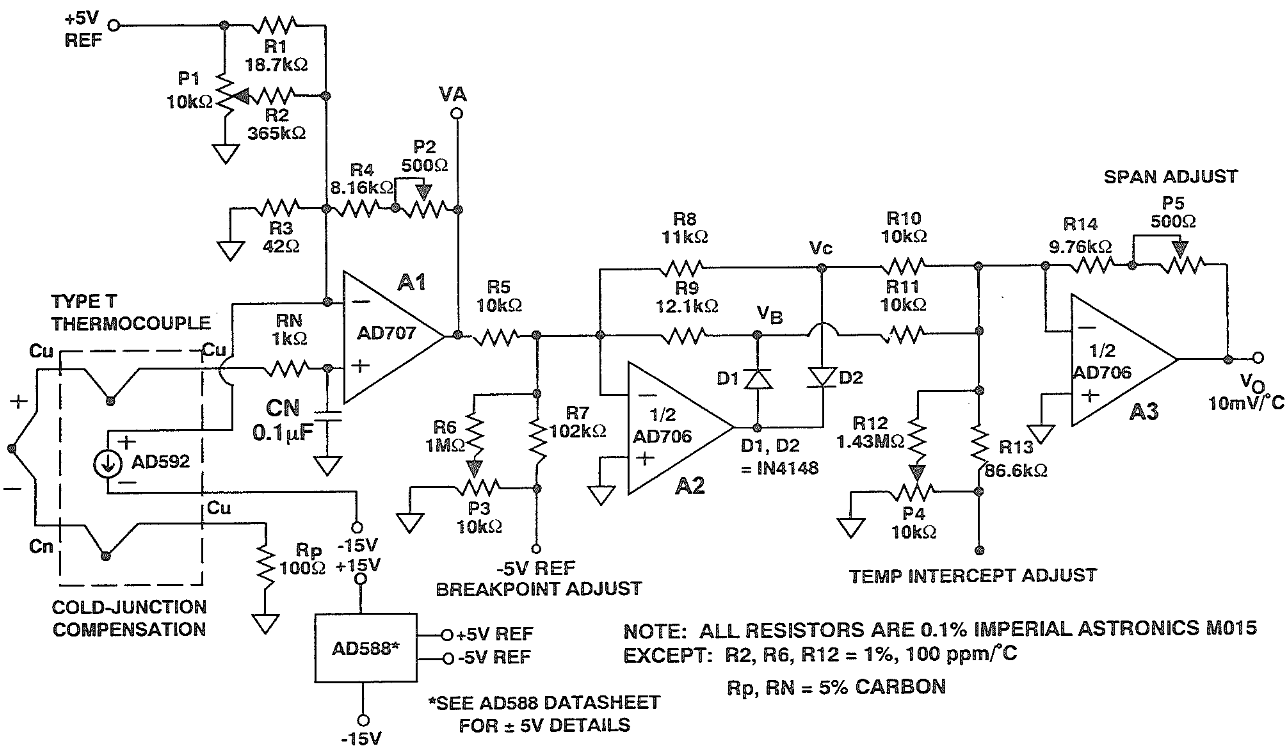


Figure 2.61

The input amplifier is virtually identical to the amplifier in Figure 2.60 except for one twist: although the gain that the thermocouple sees is 200, the cold-junction compensation circuitry is designed to track the terminating junction over a temperature range of 20 °C to 50 °C with a Seebeck coefficient of 41.7 μV/°C. The amplifier is then configured for a fixed gain of 200 to amplify the thermocouple's signal before being applied to the linearizer. An AD588 precision reference is configured for ± 5V and is used in the offset current, breakpoint adjust, and offset temperature adjust circuitry.

Designing the linearizing circuit is a straightforward process. On a piece of paper, a graph is drawn showing the thermocouple's voltage-temperature characteristic from 0 °C to 100 °C and is illustrated in Figure 2.62. On the ordinate is plotted the thermocouple's output voltage, and on the abscissa is plotted the temperature of the thermocouple.

LINEARIZING A TYPE-T THERMOCOUPLE'S RESPONSE USING TWO-SEGMENT APPROXIMATION

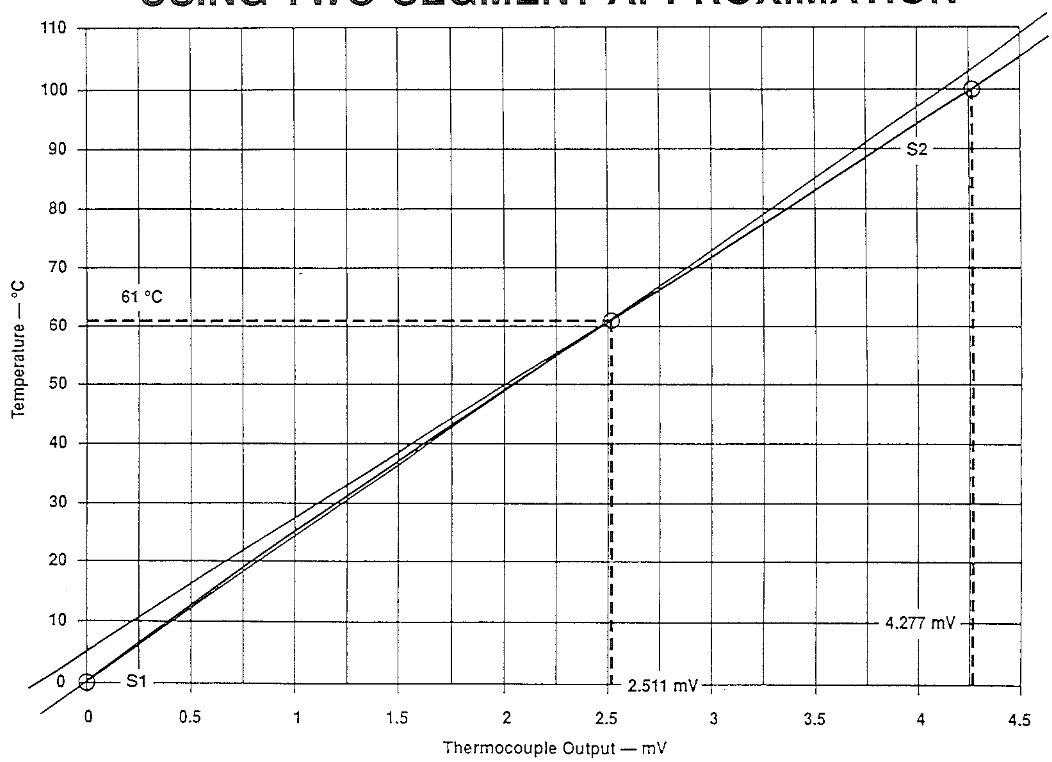


Figure 2.62

Two linear segments are drawn on the graph: S1, starting from the origin toward the full-scale output following the thermocouple's characteristic, and S2, second segment starting at 100 °C back toward the origin. The point of intersection sets the breakpoint for the approximation. The two segments intersect the thermocouple's temperature-voltage characteristic at about 61°C, which corresponds to a thermocouple output voltage of 2.511 mV.

To incorporate both segments into one stage, a high-accuracy dual-slope diode circuit built around the AD706 was used as the core of the linearizer. In order to change the gains in the circuit, there must be zero net current into A2 at the breakpoint. Thus, R7 and the AD588's -5V output are used together to provide that offset current. The breakpoint is set when the output of A1 is 502.2 mV (200 • 2.511 mV). Since the net input current into A2 is zero, both diodes D1 and D2 are turned off, and the net input current into A3 is also zero. At this point, the output of the signal conditioner should reflect a thermocouple temperature of 61 °C or 0.61 V. With A3's feedback network nominally set to 10 kΩ, R13 and the AD588's -5 V output supply another offset current to generate a 0.61 V output.

To set the gains of the linearizer, an equivalent Seebeck coefficient for each segment is calculated. From 0 °C to 61°C, the thermoelectric voltage of the thermocouple is less than 2.511 mV, making A2's net input current negative. Diode D1 turns on and the gain of the linearizer is given by: -(R9/R5). The slope of the first segment yields a Seebeck coefficient of 41.2 μV/°C and requires a gain of 242.9 to produce a 10 mV/°C output. Since the gain of the input stage is 200, the ratio of R9 to R5 is 1.214. The error in the approximation over this range of temperatures is less than 0.8 °C.

The second segment's slope yields 45.3 μV/°C from 61 °C to 100 °C and requires a gain of 220.8 to provide a 10 mV/°C output. In this case, the net current into A2 is positive and diode D2 turns on. The gain of the linearizer is now given by: -(R8/R5); hence, the ratio of R8 to R5 is 1.104. Using this Seebeck coefficient over this range produces an approximation error of 0.4 °C. The output of the two segments are then summed to A3 via R10 and R11.

The calibration procedure for this circuit is quite involved. First, the gain of the AD707 must be adjusted to 200 to avoid inducing a systematic gain error in the signal conditioner. To begin the calibration procedure, the AD592 is set to zero current by opening its lead to -15 V. Next, the AD588's + 5 V output is removed from P1, and then P1 is then shorted to circuit common. The next step is to replace the thermocouple with a precision dc voltage source set to 50.00 mV. P2 is adjusted so that the output of the AD707 is 10.00 V.

The calibration of the signal conditioner for 0 °C trim begins with reconnecting the AD592 and the AD588's + 5 V output back into the circuit. The dc source is set to 0.00 V, and P1 is adjusted so that the output of A1 is 8.338 mV/°C • T_A, the ambient temperature of the circuit. For example, if the ambient temperature were 25 °C, then P1 would be adjusted to set V_A equal to 208.45 mV.

The next step in the procedure is setting the breakpoint of the linearizer. The precision dc source is set to 2.511 mV to simulate the thermocouple's output at 61 °C. Since the AD707's gain was trimmed to 200, its output voltage is given by:

$$V_A = 2.511\ \text{mV} \bullet 200 + 8.338\ \text{mV/°C} \bullet T_A.$$

At this point, the net input current into A2 is positive and diode D2 is on. P3 is then adjusted so that V_C = -9.075 mV/°C • T_A. At this point, the offset temperature adjustment is made by adjusting P4 so that the output voltage is given by:

$$V_O = 9.075\ \text{mV/°C} \bullet T_A + 0.610\ \text{V}$$

The last step in the calibration requires trimming the full-scale output. The dc source is set to 4.277 mV to simulate the Type T's thermocouple output at 100 °C, and P5 is adjusted so that the output voltage of the signal conditioner is then by:

$$V_O = 1\ \text{V} + 9.075\ \text{mV/°C} \bullet T_A$$

Although high accuracy is achievable, one drawback to this approach is that the trims in the circuit are dependent, which means that the calibration procedure might require repeating to set accurately the endpoints of each segment. The simplicity of a digital approach to provide correction at any time is evident when compared to this analog approach.

Once all the adjustments are completed, the precision dc source is removed and the thermocouple is connected. Over an ambient temperature range of 20 °C to 50 °C, the accuracy of the circuit is ± 0.3°C. The largest source of error is in the matching of the resistors used in the linearizer. Tight tolerance, low-TCR resistors are required to minimize trim interactions and segment gain errors. In fact, a 10 kΩ thin-film resistor network can be used for R5, R10, R11, and R14 to help reduce circuit costs. Resistor networks can be acquired with tolerances as tight as 0.1 % and tracking temperature coefficients of 5 ppm/°C. The AD707 and the AD706 work very well in this application because their high open-loop gains, low input offset voltages and currents all combine to induce less than a 0.06 °C measurement error .

THERMOCOUPLE AMPLIFIER DESIGN #3

The design of the linearized Type T thermocouple amplifier can be extended to other thermocouples where in Figure 2.63 is illustrated a circuit that uses an OP-177 and an OP-297 dual operational amplifier to condition a Type S thermocouple's output. The circuit is designed to measure temperatures over the range of 200 °C to 1300 °C and does so to an accuracy of ± 0.5 °C. The measurement accuracy is lower in this circuit because of the larger gain required to scale the thermocouple's output to 10 mV/°C. The design procedure is identical to the Type T design where two segments were used to approximate the thermocouple's characteristic. In this design, the two segments intersect at 800°C where the thermocouple's output is 7.345 mV. From 200 °C to 800 °C, the approximation yields a Seebeck coefficient of 9.84 μV/°C with an error of 16 °C. Over the range of 800 °C to 1300 °C, the second segment uses a Seebeck coefficient of 11.7 μV/°C and exhibits an approximation error of 8 °C. To reduce the approximation errors, a narrower temperature range could have been chosen or more segments could have been used albeit at the price of increased circuit complexity and cost.

COLD-JUNCTION COMPENSATED TYPE S THERMOCOUPLE AMPLIFIER USING A TWO-SEGMENT APPROXIMATION OPERATING OVER 200° TO 1300°C

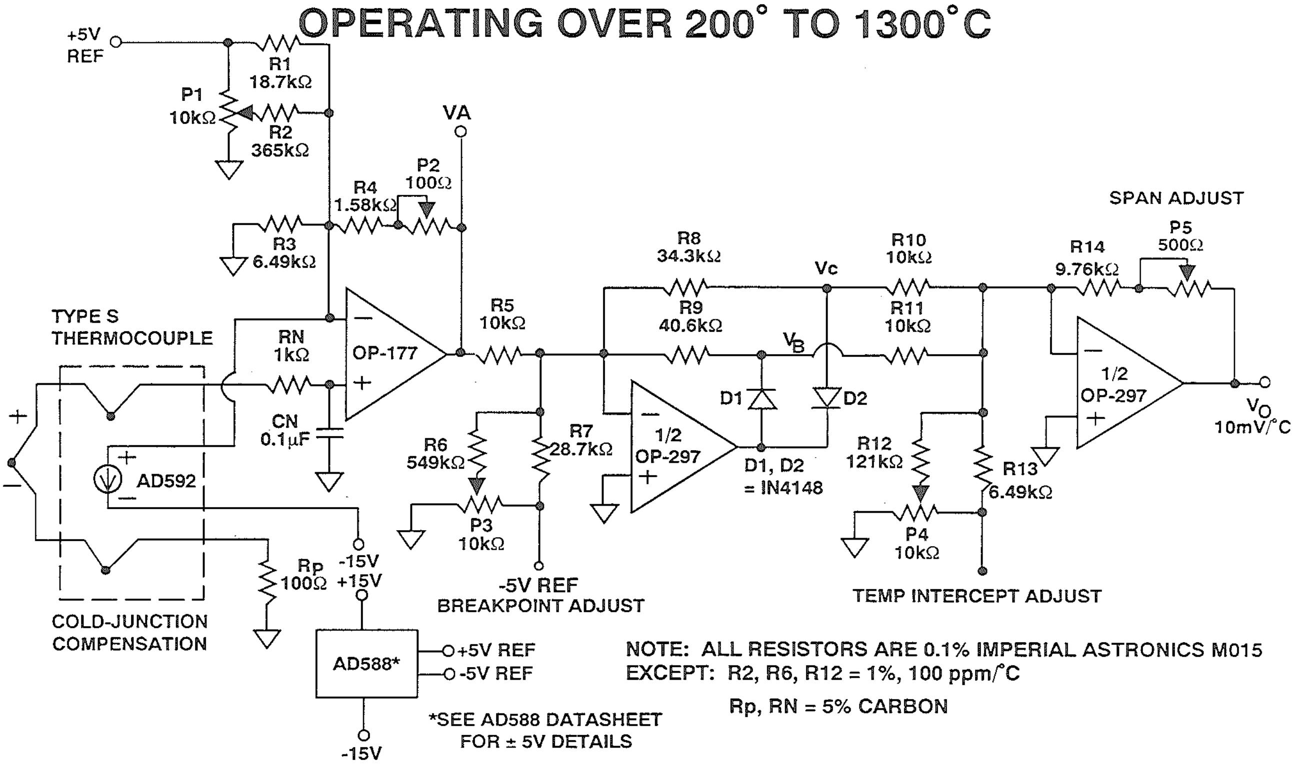


Figure 2.63

COMMON PITFALLS OF THERMOCOUPLE AMPLIFIER CIRCUITS

At this point, it is germane to discuss other circuit errors that doom thermocouple circuit accuracy. To achieve good accuracy in these systems requires diligence and care because, in general, achieving system accuracies better than 0.5 °C is very difficult. Thermocouple non-linearities notwithstanding, major (but common) pitfalls that affect accuracy include amplifier errors, choosing the right resistor tolerance and temperature coefficient, cold-junction considerations, and printed circuit board errors.

OPERATIONAL AMPLIFIER ERRORS

A critical design consideration in thermocouple signal conditioning is minimizing measurements errors in the recording equipment. The operational amplifier is a key element to accurate temperature measurement. Using the wrong one would ruin the measurement no matter how accurate the associated circuitry might be. The parameters that most affect the overall accuracy of the measurements are input offset voltage and drift, input bias current and drift, open-loop gain and temperature coefficient, and noise. Each of these parameters and their affect on measurement accuracy will discussed be individually.

As we have seen before, thermocouple Seebeck coefficients range from 6 to 15 μV/°C for Type S thermocouples and 40 to 60 μV/°C for Type J, K, and T thermocouples. Operational amplifiers used in these applications must exhibit very low input offset voltages to resolve minute changes in thermocouple temperature. In most applications where a Type J, K, or T thermocouples are used, the op amp's initial input offset voltage is trimmed during the calibration sequence. In the most critical applications (for example, a wide-range Type S thermocouple amplifier), the op amp's initial offset voltage is trimmed out using the amplifier's offset trim pins during the calibration sequence. Once the amplifier's initial offset error has been trimmed out, the op amp's offset voltage drift can contribute a significant measurement error. The offset voltage drift appears in series with the thermocouple signal and is amplified to the output. For example, in the Type T amplifier design, if an amplifier that exhibits a 1 μV/°C offset drift were used, the error at the output over a 25 °C change in ambient temperature is 0.6 °C (Drift error = 1 μV/°C • 25 °C • 233.8 ÷ 10 mV/°C). In a Type S thermocouple amplifier, this drift would induce an error of approximately 4 °C over the same ambient conditions. In these applications, halving in the input offset voltage drift halves the output measurement error. To achieve less than 0.1 °C error in Type J, K, and T applications requires that an amplifier's maximum input offset voltage drift be less than 0.2 μV/°C.

In the grounded thermocouple amplifier circuits shown, an amplifier's non-inverting input bias current flows through the thermocouple, and the inverting input bias current flows through the effective resistance of the feedback and offset current networks. Resistances in the circuit are low to minimize their noise contributions at the output so bias current errors can be neglected. In the case where the thermocouple is used in a high noise environment, an input filter is used; hence, the amplifier's non-inverting input bias current flows through the filter's resistor. In very noisy environments where 10 kΩ to 100 kΩ resistors are used, bias currents on the order of 10 nA generate 100 μV to 1 mV additional offset. Although these initial bias current errors might be trimmed at calibration, low input bias current drift is required to minimize measurement error over temperature. Hence, to achieve less than 0.1°C error over a 25°C change in ambient temperature requires an operational amplifier whose input bias current drift is less than 2 pA/°C when series resistors as high as 100 kΩ are used.

In each of the circuits illustrated thus far, the gain factors of the amplifiers were set to provide a 10 mV/°C outputs which allows the temperature to be converted into digital form using standard analog-to-digital converters or to be read directly off a 4 1/2- or a 5 1/2-digit precision digital voltmeter. Each of these circuits used a single high-gain stage to amplify the thermocouple's output. For the thermocouples discussed, the closed-loop gains range from 190 for Type J thermocouples to over 1500 for Type S thermocouples. If an output scaling of 100 mV/°C were used, all these gain factors would increase by an order of magnitude. High open-loop gain operational amplifiers are required for these applications to minimize gain-errors that produce false readings. To see how this works, recall from feedback theory that the exact expression for the closed-loop gain of an amplifier is given by:

$$A = \frac{V_{OUT}}{V_{IN}} = \frac{1}{\beta}\left(\frac{1}{1+\frac{1}{a\beta}}\right)$$

where A = the amplifier's closed-loop gain,

a = the amplifier's open-loop gain, and

β = the amplifier's feedback factor

Differentiating the closed-loop gain with respect to the open-loop gain and manipulating the algebra yields the sensitivity of the closed-loop gain (and, therefore, the output voltage) to the open-loop gain. This is given by:

$$\frac{dA}{A} = \frac{1}{1+a\beta} \bullet \frac{da}{a}$$

For example, let's say that a Type J thermocouple amplifier requires a closed-loop gain of 2000 to produce a 100 mV/°C output. If an op amp with an open-loop gain of 75,000 is used, the amplifier's initial gain error would be 2.6 % which can be trimmed during calibration. However, over temperature, that open-loop gain could change as much as 25 % which would then cause a gain-error of 3.5 % or 0.7 °C. If, on the other hand, an amplifier with an open-loop gain of 1 million were used, then the gain error over temperature would be less than 0.27 % or 0.06 °C. Therefore, to keep open-loop gain errors below 0.1 °C, the minimum recommended open-loop gain for Type J, K, and T thermocouples is 500,000. For Type S thermocouples amplifiers with 10 mV/°C outputs, the recommended open-loop gain of 500,000 is sufficient.

Another cause of measurement inaccuracy is operational amplifier noise. This error most greatly affects Type S thermocouple signal conditioning amplifiers which must resolve microvolts of thermocouple output. Wide-band circuit noise ultimately determines how small a signal the amplifier can resolve. To minimize an amplifier's noise, here are a few tips: (1) Design the circuit with low resistances – Using low resistances minimizes the effect of amplifier current noise flowing through the effective resistances appearing at the amplifier's input terminals. Using low-value resistors in the design also reduces their thermal noise and stray radio-frequency noise pick-up. For example, reconsider the first thermocouple amplifier design. Although the circuit required a gain of 233.8 to amplify the thermocouple's signal, the effective resistances at the amplifiers terminals were 43Ω (at IN-) and 1.1 kΩ (at IN+). These low resistances, coupled with the OP-177's low input noise voltage of 11 nV/√Hz and low input current noise of 0.17 pA/√Hz, only increased the total equivalent input referred noise to 11.8 nV/√Hz. The idea here is to let the amplifier's input voltage noise dominate. (2) Restrict the system bandwidth – Noise is generated outside the circuit and is all around it. External noise can best be dealt with by filtering, either at the thermocouple output or at the amplifier's output. The power supply leads are also a wonderful way high frequency supply noise can enter the circuit. In these applications (and most others that require high precision), we often recommend a stiff 10 μF tantalum electrolytic capacitor in parallel with a 0.1 μF or 0.01 μF ceramic disk capacitor right at the amplifier's supply pins. (3) Select a low-noise operational amplifier – Operational amplifiers that fall under the category of "high-accuracy" are specifically designed with input stages which use high currents and input bias current cancellation schemes to lower input voltage and current noise. For this reason, amplifiers used for thermocouple applications should exhibit less than 15 nV/√Hz equivalent input voltage noise.

Operational amplifiers such as the OP-177 and the AD707 are ideal amplifiers for these applications because they meet all the critical specifications that thermocouple signal conditioning requires. Both devices exhibit very high open-loop gains, excellent gain linearity, low input offset voltages and drift, low input bias currents, and low noise. For less critical applications, the AD705 and the OP-97 (and their duals, the AD706 and the OP-297) should be also considered as excellent thermocouple amplifiers for their high open-loop gains, low input offset voltages, very low input bias currents and drift (100 pA, 0.3 pA/°C), low noise, and low supply currents.

RESISTORS AND POTENTIOMETERS

Careful selection of resistors also greatly affects the accuracy of thermocouple signal conditioning amplifiers. Carbon composition resistors should not be used anywhere in the circuit except for input filtering and current-limiting protection for the thermocouple. In this role, their absolute value and temperature coefficient are unimportant. Referring to the circuit in Figure 2.60, R1, R3, and R4 should be precision wirewound or metal film resistors with 1 % tolerance and temperature coefficients of 25 ppm/°C. Measurement inaccuracies over temperature are dominated by these resistors and add a ± 0.2 °C error over a 25 °C change in ambient temperature. Using lower TCR resistors lowers this error but increases circuit cost. Loose tolerance resistors should be avoided because wide trim ranges would be required which would degrade circuit accuracy. R2's TCR, on the other hand, is not that critical in this application and a 1 %, 100 ppm/°C resistor is good enough.

In the examples of thermocouple signal conditioning circuits, trimming potentiometers are used for calibrating the circuit's response at two points, usually at zero and at full-scale. Although useful, trimming can lead to very large circuit errors if done improperly. In all the designs, potentiometers are placed where their absolute resistance is not important and their temperature coefficient is not critical. For example, in the first thermocouple amplifier design, the amplifier's nominal feedback network is 10 kΩ which is made up of a 9.76 kΩ resistor in series with a 500 Ω rheostat-connected potentiometer. The value of the potentiometer to achieve trim is roughly at 50 % of its value (i.e, trim range = 250 Ω ÷ 10 kΩ • 100 = 2.5 %).

What this means is that non-linearities of the potentiometer are avoided because mid-range values are more well-behaved than end-range values. Let's look at what happens to the feedback network over temperature. In these applications, R4 should be a 25 ppm/°C resistor, and cermet potentiometers typically exhibit a temperature coefficient of 100 ppm/°C. If the nominal feedback resistance were trimmed to 10 kΩ during calibration, then over a 25 °C change in ambient temperature the feedback network resistance would only increase by 7 Ω for an error of less than 0.02 °C. The same approach is used for trimming the offset current for cold-junction compensation. In this case, R2 is chosen so that the trim range is again 2 — 5 % of the total offset current required to bring the circuit in trim. So the moral of the story is: do your homework when it comes to applying trimming potentiometers to high accuracy circuits and use multi-turn stable cermet potentiometers for setting accuracy and low cost.

MAINTAINING PROPER COLD-JUNCTION COMPENSATION

One of the largest sources of error in thermocouple signal conditioning circuits is the failure of maintaining proper cold-junction compensation. Since a thermocouple's output voltage is a function of the temperature difference between its two junctions, any temperature delta between the terminating junction and the cold-junction temperature sensor will produce an error signal equal to their difference. It is therefore imperative that the layout of the temperature sensor and the thermocouple's terminating junction minimize any effects due to temperature gradients that might exist. Many times errors are inadvertently introduced when the cold junction is not properly transferred to the PC board. An example of transferring the cold junction to the pc board is shown in Figure 2.64.

TRANSFERRING THE COLD-JUNCTION FROM AN ENCLOSURE OR CONNECTOR TO THE PCB

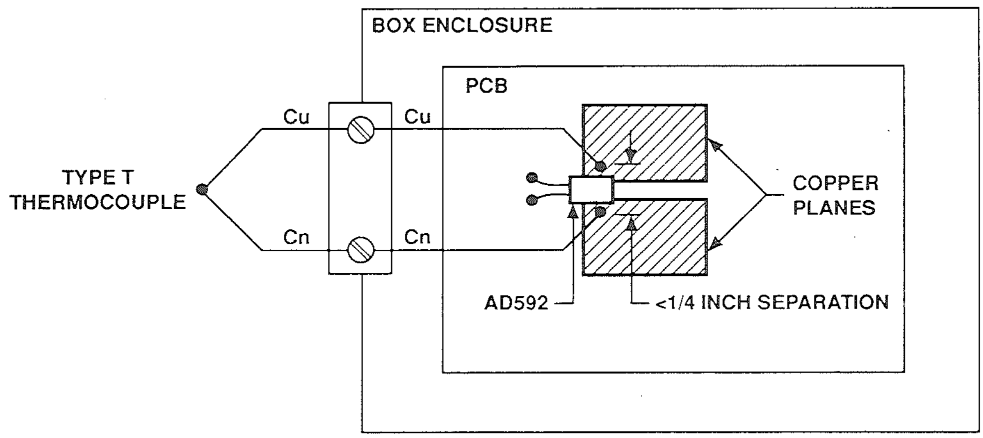


Figure 2.64

The layout illustrates how to extend properly the cold-junction reference plane from the terminal block (or connector) to the pc board, and how to locate the temperature sensor with reference to the thermocouple's terminating junction.

Bear in mind that the extension wires from the terminal block (or connector) to the pc board must be of the same wire type as that used for the thermocouple wires.

MINIMIZING PARASITIC PCB THERMOCOUPLE ERRORS

Thermocouple voltages are generated whenever dissimilar materials are joined. This includes the leads of IC packages, which may be kovar in TO-5 cans, alloy—42 or copper in dual-in-line packages, and a variety of other materials in plating finishes and solders. The net effect of these thermocouples is "zero" if all are at exactly the same temperature, but temperature gradients exist within IC packages and across printed circuit boards whenever power is dissipated. It is for this reason that extreme care must be used to ensure that no temperature gradients exist in the vicinity of thermocouple junctions, the temperature sensor, or the thermocouple amplifier. If a gradient cannot be eliminated, leads should be positioned isothermally, especially the temperature sensor, the amplifier inputs pins and the gain setting resistor leads. When the leads are positioned isothermally, parasitic thermocouple junctions cancel each other out because the leads are placed orthogonal to the temperature gradient. These critical components must also be placed as closely together as possible to prevent varying temperature gradients from affecting the components as well. To illustrate these two issues, Figure 2.65 is an example of a printed circuit board layout illustrating the isothermal placement and tight arrangement of the thermocouple terminating junction, the temperature sensor, the amplifier input pins, and the gain setting resistors.

An effect to watch for is amplifier offset voltage warm-up drift caused by mismatched thermocouple materials in the wire bond/lead system of the 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. Finally, it can be accommodated by calibrating and specifying the system after a five minute warm-up period.

PROPERLY LOCATING ALL COMPONENTS THAT ARE SENSITIVE TO THERMAL GRADIENTS

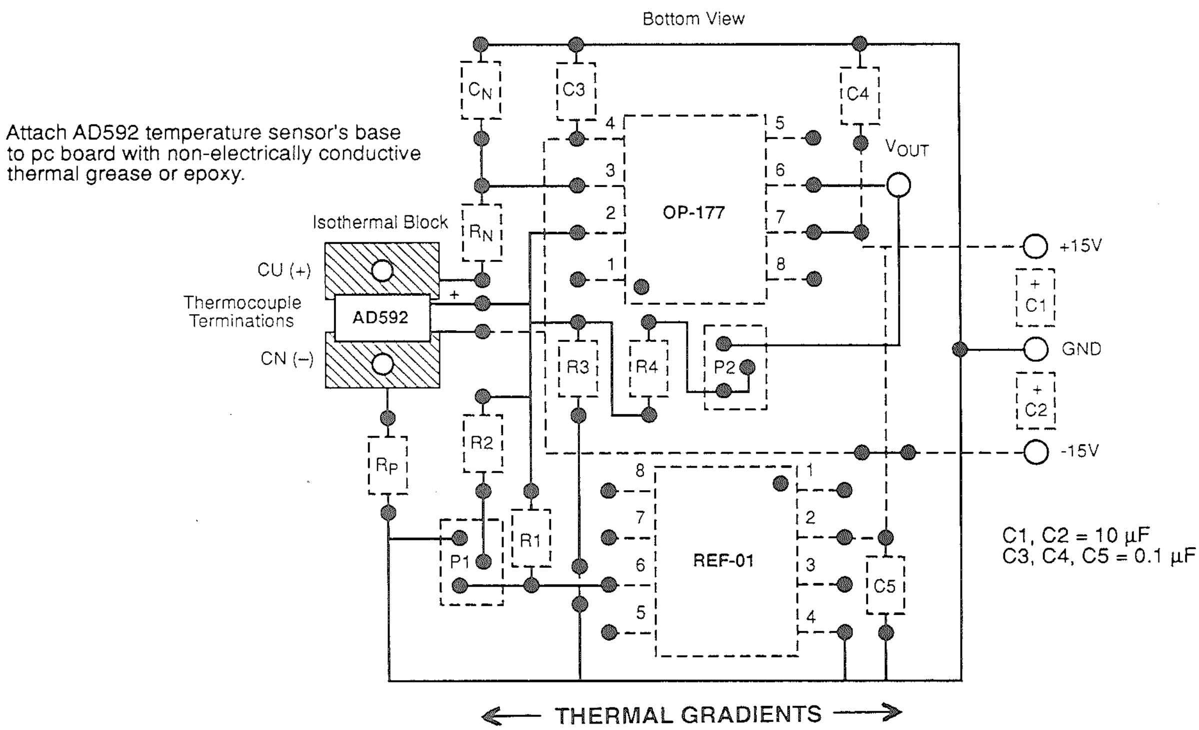


Figure 2.65

RESISTANCE TEMPERATURE DETECTOR (RTD) SIGNAL CONDITIONING

Of the commonly used temperature transducers, the resistance temperature detector, or the RTD, is one type of sensor whose resistance changes with temperature. Typically built of a platinum wire wrapped around a ceramic bobbin, the RTD exhibits a resistance versus temperature behavior which is more accurate and more linear over wide temperature ranges than a thermocouple. Figure 2.66 illustrates the resistance temperature coefficient of a 100-ohm RTD and the Seebeck coefficient of a Type S thermocouple. Over the entire temperature range, the RTD is a more linear temperature sensing device. Hence, linearizing an RTD's response requires less complexity than a thermocouple.

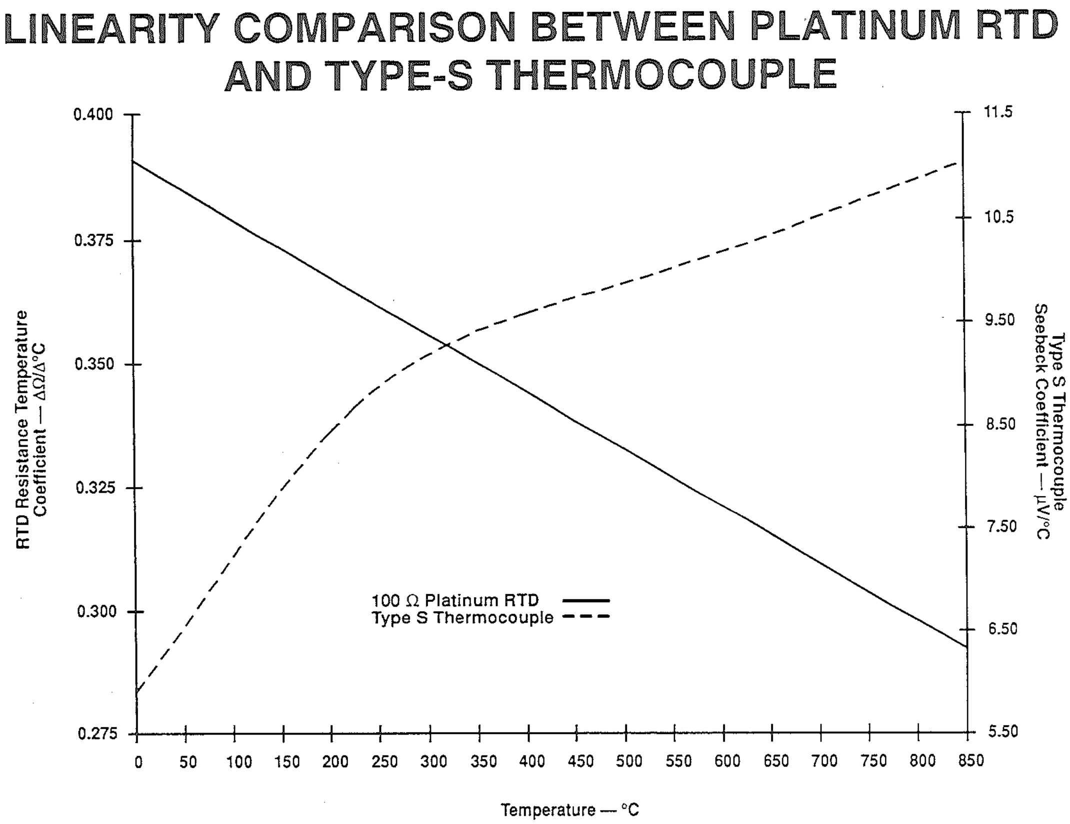


Figure 2.66

Unlike the thermocouple, however, the RTD is a passive temperature sensor and requires a current source excitation to produce an output voltage proportional to temperature. The RTD's low temperature coefficient of 0.385 Ω/°C (for a 100 ohm Pt RTD) requires the same high-performance signal conditioning circuitry as the thermocouple; however, the voltage drop across an RTD is a much larger signal than a thermocouple output voltage. A system designer may opt for large value RTDs given their low sensitivity to temperature; however, large-valued RTDs exhibit slow response times. Lastly, although the cost of RTDs is higher than thermocouples, they use ordinary copper extension wires; therefore, thermoelectric effects from terminating junctions do not affect measurement accuracy. And, because their resistance is a function of the absolute temperature, RTDs require no cold-junction compensation circuitry.

SECOND-ORDER EFFECTS

Caution must be exercised using current source excitation because the current through the RTD causes Joule, or I^2R, heating within the RTD. This self-heating effect changes the temperature of the RTD and appears as a measurement error. Hence, careful attention must be paid during the design of the signal conditioning circuitry so that self-heating is kept below a typical value of 0.5 °C/W in free air. Manufacturers typically 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 smallest amount of current through the RTD should be used for the required system resolution, and the largest RTD value should be chosen that results in an acceptable response time.

Another effect that can produce measurement error is that due to RTD lead wires. This is especially critical in low-valued, 2-wire RTDs because the temperature coefficient and the absolute value of the RTD are both small quantities. If the RTD is located a long distance from the signal conditioning circuitry, then the lead wires can be on the order of ohms or tens of ohms. Therefore, a small amount of lead resistance can contribute a significant error to the temperature measurement. To illustrate this point, let us assume that a 100-Ω platinum RTD with 30-gauge copper lead wires is used in a carpet manufacturing facility to sense the temperature of an adhesive and is located about 100 feet from a controller's display console. Since the resistance of 30-gauge copper wire is 0.105 Ω/ft, then the two leads of the RTD contribute a total 21 Ω to the network which is shown in Figure 2.67. This additional resistance will induce a 55 °C error in the measurement! Also not included in this example is the effect of the lead wires' temperature coefficient which can contribute an additional, and possibly significant, error to the measurement. To alleviate the effect of the lead wires, a 4-wire technique is used.

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

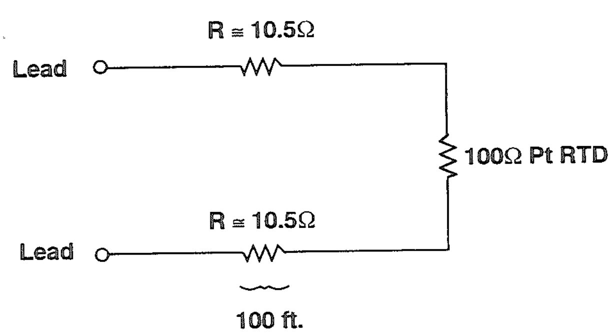


Figure 2.67

As shown in Figure 2.68, 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 drop can be 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 or low input bias currents. Since the SENSE leads do not carry current, this approach is insensitive to lead wire length. Sources of errors in this approach are the stability of the constant current source and the input impedance of the DVM or bias currents in the amplifier.

FOUR-WIRE OR KELVIN CONNECTION TO RTD

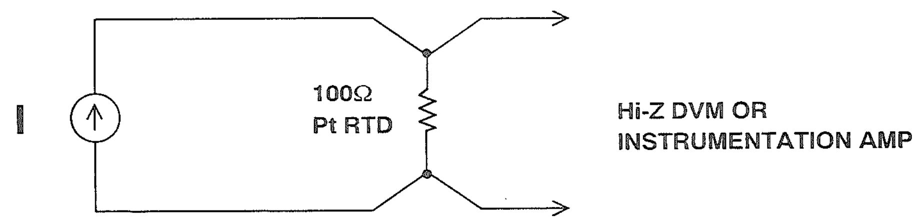


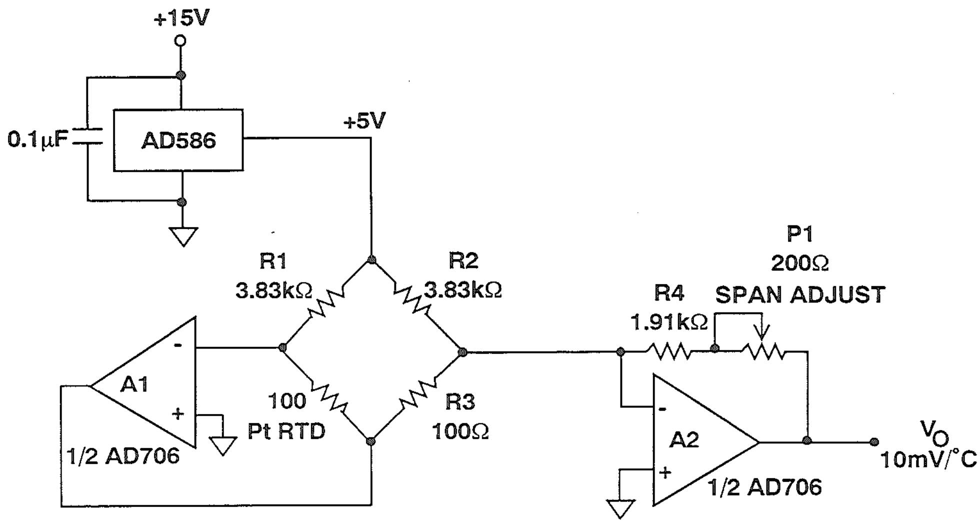
Figure 2.68

BRIDGE AMPLIFIER IMPLEMENTATION

Bridge circuits used in RTD signal conditioning do quite a nice job linearizing transducer outputs. They don't always, however, compensate for a major source of bridge-instrumentation error: an operational amplifier's input-offset-voltage drift. In many bridge circuit designs, an amplifier's input-offset-voltage thermal drift can be amplified four times greater than the sensor's signal because of the bridge's voltage-divider effects.

Without sacrificing linearity, an amplifier's input-offset-voltage drift can be minimized by applying feedback control to the bridge. A pair of op amps can be used in the circuit to place all bridge elements under fixed bias or feedback control. Shown in Figure 2.69 is a circuit that uses an AD706 dual operational amplifier and a 100 Ω 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 2.69

A1's inverting sum node forces a null at one bridge node such that a constant current, developed 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 over temperature generates a signal voltage at the output of A1 which is then developed into a signal current for A2 by R3. This signal current is then compared to the current generated by the reference's 5 V output and R2, with the result scaled by R4 to provide a 10 mV/°C output. Since the AD706's input bias currents are typically 30 pA, their error contribution can be neglected. Thus, a simplified equation for the output voltage is given by:

$$VO = \frac{R_4}{R_3} \bullet \frac{\Delta R}{R_1} V_{REF} + V_{OS2}\left(1 + \frac{R_4}{R_2}\right)$$
$$= \frac{R_4}{R_3}\left(V_{OS2} - V_{OS1}\right)$$

where ΔR is the change in the RTD's resistance over temperature from its 0 °C value. Substituting in values for the resistors yields:

$$V_O = 25.85E\text{-}3 \bullet \Delta R + 1.52 \bullet V_{OS2}$$
$$+ 19.8 \bullet (V_{OS2} - V_{OS1})$$

The circuit's response is now dominated by the RTD and the circuit achieves a much lower sensitivity to amplifier input offset voltage and drift. By placing the RTD in the feedback path of A1, A1's input offset voltage effects at the output are effectively suppressed and the circuit's response is very accurate. Although the output AD706's input offset voltage term 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 in ambient temperature.

Calibration of the circuit is a one trim procedure. Because of the AD706's low initial offset voltage of 10 μV and the use of precision wirewound resistors, the circuit's error at 0 °C is less than 0.1 °C. Calibration is then only required at the full-scale temperature of 100 °C. In this case, the RTD is replaced by a precision decade resistance box which is set to 138.5 Ω. P1 is adjusted such that V_O equals 1.00 V. With this single-trim 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 supply RTD non-linearity correction for very wide temperature ranges. For example, if this topology is used with the same RTD to measure the range of 0 °C to 200 °C, significant curvature of the RTD response would cause a measurement error of approximately 3 °C at full scale.

CURRENT DRIVE LINEAR AMPLIFIER IMPLEMENTATION

It was previously pointed out that the lead wires in a remotely located, low-valued RTD will introduce a measurement error. For example, a 2.6 °C error occurs for every ohm of lead wire resistance in a 100 Ω Pt RTD. In these applications, a 4-wire connection to the RTD should be used to eliminate the effect of lead wires. Figure 2.70 shows a circuit that uses a remotely located, 4-wire connection to the RTD and measures temperature from 0 °C to 100 °C with an accuracy of ± 0.8 °C.

FOUR-WIRE LINEAR AMPLIFIER WITH CONSTANT CURRENT SOURCE EXCITATION

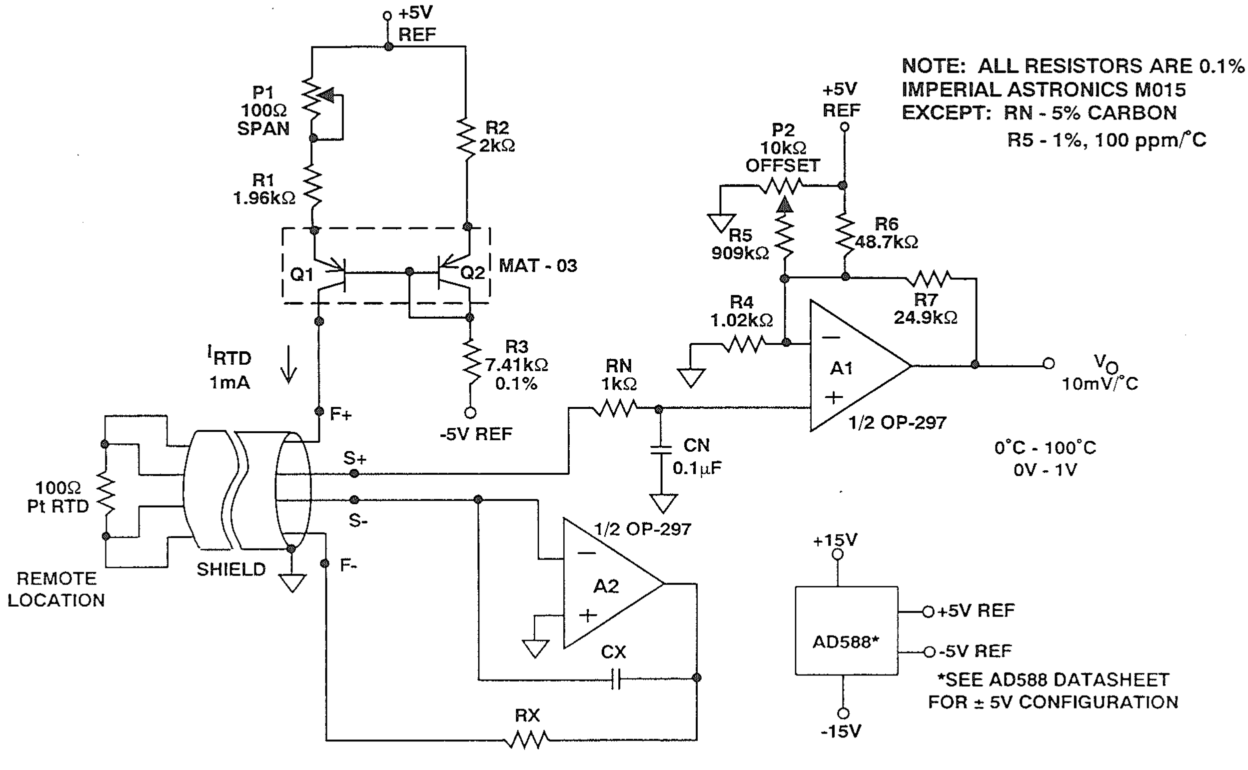


Figure 2.70

In this circuit, a 1 mA constant current is applied to the RTD, and the voltage developed across it is amplified by A1 to provide a 10 mV/°C output. What is unique about this circuit is that A2 servos the return end of the RTD to zero volts thereby eliminating the error induced by the lead wire. The only error the OP-297 introduces after trim is input offset voltage drift. For a 25 °C change in ambient temperature, the error is less than 0.04°C. For very remote locations where the parasitic capacitance of the cable might cause problems for the amplifier, R_X and C_X can compensate A2's response without loss of accuracy.

The 1 mA constant-current source consists of a matched pair of MAT-03 transistors and low-TCR precision resistors. In order to insure that the current source has a low sensitivity to its supply voltages, an AD588 precision voltage reference is configured for ± 5V outputs to drive the current mirror. The AD588's low 3 ppm/°C drift for ± 5 V outputs keeps its error contribution below ± 0.03 °C over a 25 °C change in ambient temperature. R2 and R3 set Q2's collector current accurately to 1 mA which is mirrored over to Q1. To adjust the RTD's current exactly for the full-scale span, a 100 Ω rheostat-connected potentiometer is provided in series with R1.

The gain of amplifier A1 is determined by the change in the RTD's resistance over the temperature range. From 0 °C to 100 °C, the change in the resistance of the RTD is 38.5 Ω. Using a 1 mA constant-current excitation, the full-scale voltage change is 38.5 mV; therefore, to produce a 1 V full-scale output requires a gain of 25.97 and can be adjusted at full-scale with P1. The AD588's 5 V output and R6 provide an offset current for the RTD's initial resistance of 100 Ω at 0 °C with adjustment for this and A1's input offset voltage supplied by P2 and R5. For very remotely located RTDs in industrial environments, R_N and C_N can be used as a filter for noise. Typical values of 1 kΩ for R_N and 0.1 μF for C_N effectively filter out induced noise above 1.6 kHz. Larger values can be used for R_N without introducing errors because of the OP-297's very low bias currents and drift.

To calibrate this circuit, a precision set of resistors or a precision decade box in place of the RTD work equally well. First, 100.00 Ω is substituted for the RTD at 0°C, and P2 is adjusted so that the output of A1 reads 0 V. Next, the decade box is set to 138.5 Ω for the RTD at 100 °C. The constant-current excitation is adjusted with P1 so that A1's output reads 1.0000 V. The trim sequence might need repeating to fix the two end points.

By far the largest source of error in the circuit is the temperature dependence of Q1's output current. Since the temperature coefficient of a base-emitter junction is approximately -2.1 mV/°C, a 25°C change in the ambient temperature will cause an error of 0.7 °C. To reduce the circuit's sensitivity to changes in the current mirror's base-emitter junctions over temperature, temperature independent biasing techniques can be applied in the design of the current source.

Although the circuit is initially designed for 100 Ω Pt RTDs, other RTD values can be used with only one modification to the amplifier's circuit. For example, 1 kΩ is another commonly used RTD value in many applications, and using it only requires changing R4 from 1.02 kΩ to 23.7 kΩ. Since the RTD's value at 100 °C is 1.385kΩ, the full-scale RTD voltage at 1 mA is 1.385 V. This large output voltage might be a source of measurement error due to Early voltage modulation of Q1's collector-base junction. However, because of the MAT-03's low output conductance of 10 μA/V at 1mA, the 1.25 V change in Q1's collector-base voltage induces less than a 2 % change in Q1's collector current. This delta can be easily trimmed at full-scale with P1.

THERMISTOR SIGNAL CONDITIONING

THERMISTOR BASICS

Similar in function to the RTD, thermistors are low-cost temperature-sensitive resistors and are constructed of solid semiconductor materials which exhibit a either positive or negative temperature coefficient. Although positive temperature coefficient devices are available, the most commonly used thermistors are those that exhibit a negative temperature coefficient. Figure 2.71 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 temperature sensitive.

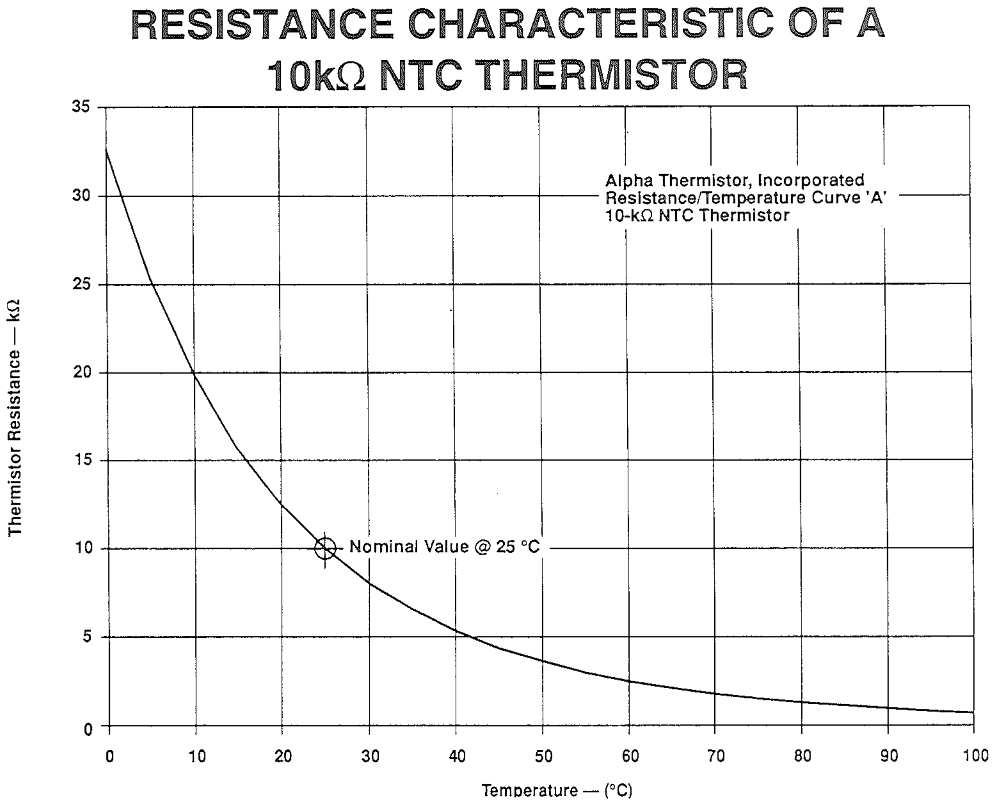


Figure 2.71

The thermistor's high sensitivity to temperature (typically, -44,000 ppm/°C at 25 °C), as shown in Figure 2.72, allows it to detect minute variations in temperature which would not be observed with an RTD or thermocouple. This high sensitivity to temperature 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 10 kΩ NTC thermistor, with a typical 25 °C temperature coefficient of -44,000 ppm/°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 the equivalent RTD error.

TEMPERATURE COEFFICIENT OF 10kΩ NTC THERMISTOR

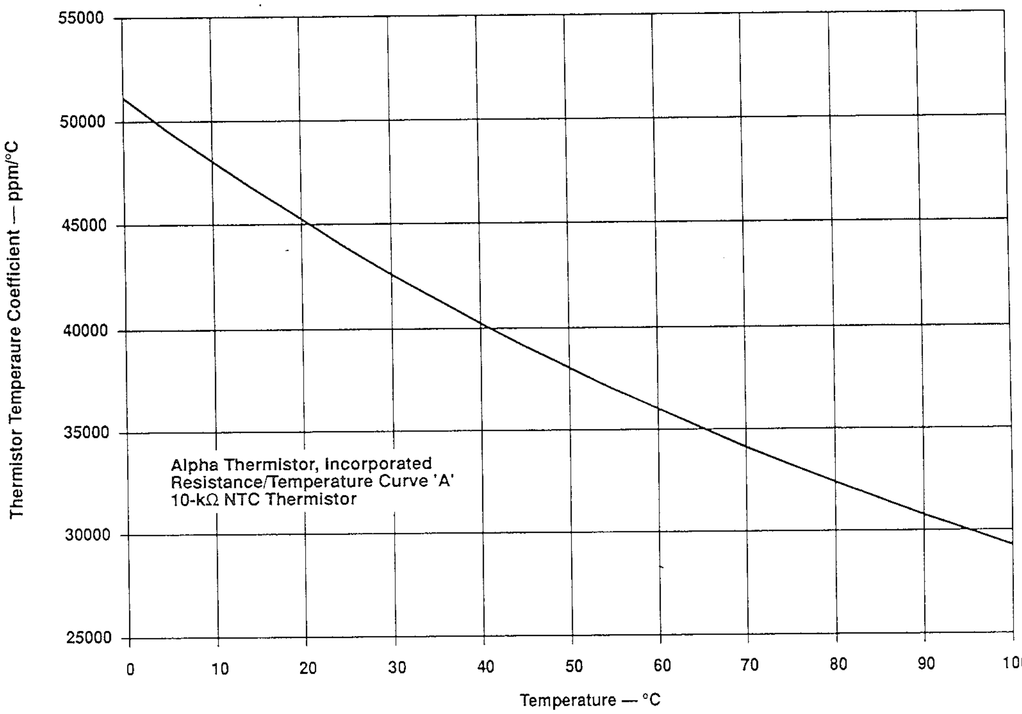


Figure 2.72

The thermistor's high sensitivity to temperature does not, however, come without a price. As was shown in Figure 2.72, the temperature coefficient of thermistors does not linearly decrease 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 permanent decalibration at high temperatures. Compared to thermocouples and RTDs, thermistors are quite fragile in construction and require careful mounting procedures to prevent crushing or bond separation. Although a thermistor's response time is quite short due to its small size, its small thermal mass makes it very sensitive to self-heating errors.

TYPICAL AMPLIFIER CONFIGURATION

As mentioned before, thermistors make for very inexpensive, highly sensitive temperature sensors. However, it was shown that a thermistor's temperature coefficient varies from -44,000 ppm/°C at 25 °C to -29,000 ppm/°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 wider temperature range only if the system designer can tolerate a lower sensitivity to achieve improved linearity. One approach to linearizing a thermistor is achieved by simply shunting it with a fixed resistor. Paralleling the thermistor with a fixed resistor increases the linearity significantly by desensitizing the network to temperature variations. As shown in Figure 2.73, the parallel combination exhibits a more linear variation to temperature when compared to the thermistor. Also, the sensitivity to temperature of the combination still remains high when compared to a thermocouple or an RTD. The primary disadvantage to this technique is that linearization can only be achieved within the measurement range of interest.

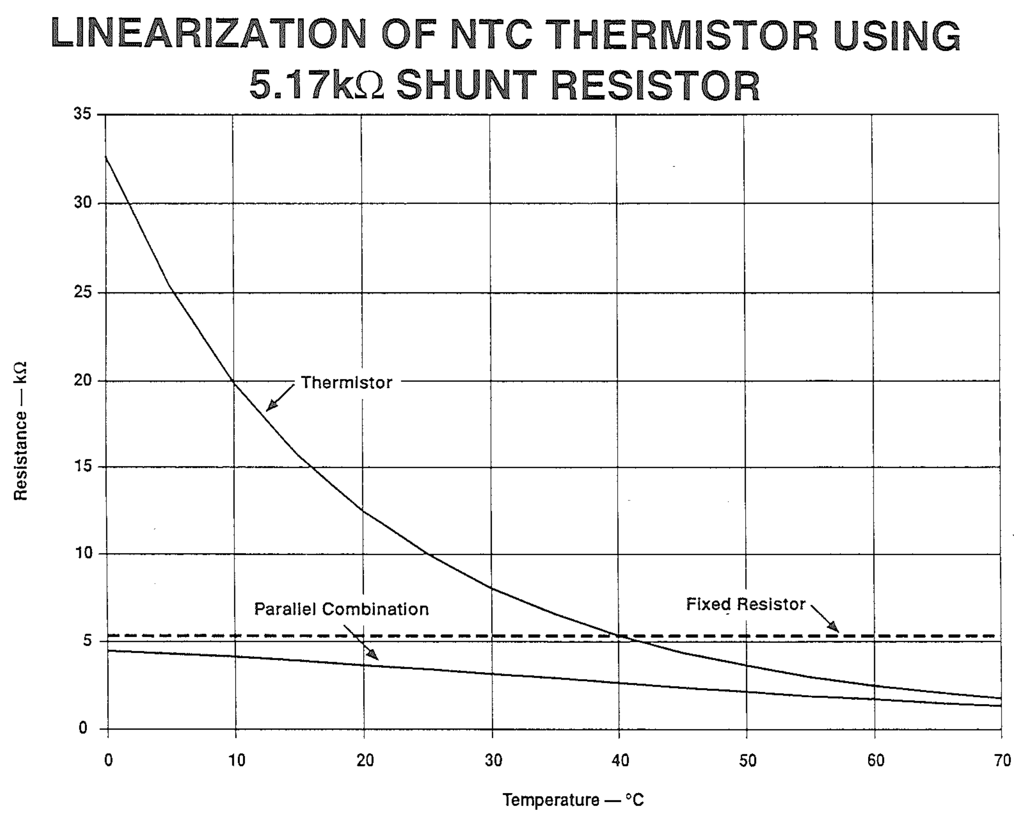


Figure 2.73

To implement this technique, a straight line is drawn through three equally spaced points along a thermistor's resistance-temperature characteristic. The three points are the midpoint and the two endpoints and must be within the measurement range. Thus, the value of the fixed resistor can be calculated from the following equation:

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

where RT1 = Thermistor resistance at T1, RT2 = Thermistor resistance at T2, RT3 = Thermistor resistance at T3.

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 5.17 kΩ for R. Of particular importance in linearizing a thermistor's response is the overall linearity deviation of the combined network. Accuracy requirements of the signal conditioning circuitry is dependent on the linearity deviation of the linearized network. For the example given above, Figure 2.74 illustrates the actual response of the combined network to the linear approximation and yields a departure from linearity of -2.3 °C/ + 2.0 °C.

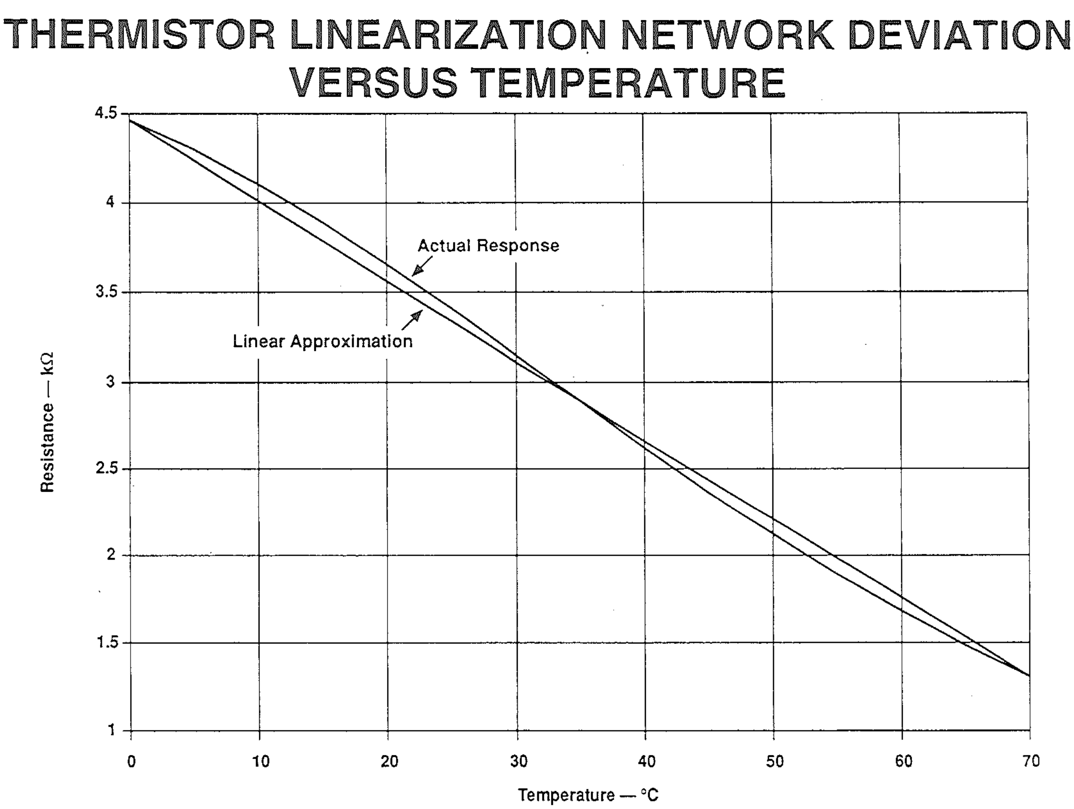


Figure 2.74

With this composite thermistor/resistor network, Figure 2.75 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. The 10 kΩ 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 deviation of the network under -2.3 °C/+ 2 °C. Much better linearity can be achieved with a narrower measurement temperature range. In many cases, the thermistor is placed some distance from the signal conditioning circuit. Under this condition, a 0.1 μF capacitor placed across R2 will help to suppress noise pickup.

In the circuit, R1 and the AD586 provide a constant current of 281 μA to the linearized 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 -10 mV/°C output. Lastly, R4 and the reference generate an offset current such that the output of the second amplifier is 0 V at 0 °C.

A calibration procedure is required for this circuit to trim resistor tolerance and the AD706s' initial offset voltages. To start, a precision decade box can be used in place of the thermistor. For 0°C trim, the decade box is set to 32.650 kΩ, and P1 is adjusted until the circuit's output reads 0 V. To trim the circuit at the full-scale temperature of 70 °C, the decade box in then set to 1.752 kΩ and P2 is adjusted until the circuit reads -0.70 V.

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

MONOLITHIC THERMOCOUPLE AMPLIFIERS WITH COLD JUNCTION COMPENSATION: THE AD594/AD595/AD596/AD597

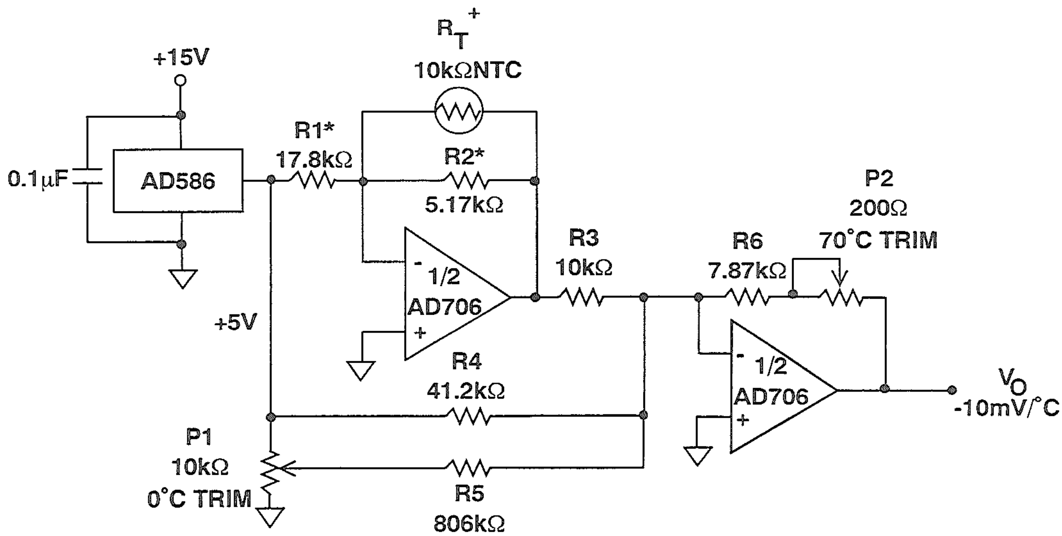
The AD594/AD595 is a complete instrumentation amplifier and thermocouple cold junction compensator on a monolithic chip. It combines an ice point reference with a precalibrated amplifier to produce a high level (10mV/°C) output directly from a thermocouple signal. Pin-strapping options allow it to be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

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

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

The AD594 is precalibrated by laser wafer trimming to match the characteristic of type J (iron-constantan) thermocouples, and the AD595 is laser trimmed for type K (chromel-alumel) inputs. The temperature transducer voltages and gain control resistors are available at the

LINEARIZED THERMISTOR AMPLIFIER



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

Figure 2.75

package pins so that the circuit can be recalibrated for other thermocouple types by the addition of two or three resistors. These terminals also allow more precise calibration for both thermocouple and thermometer applications. The AD594/AD595 is available in two performance grades. The C and the A versions have calibration accuracies of ±1°C and ±3°C, respectively. Both are designed to be used from 0 to +50°C, and are available in 14-pin, hermetically sealed, sidebrazed ceramic DIPs as well as low cost cerdip packages.

The AD596/AD597 is a monolithic temperature set-point controller which has been optimized for use at elevated temperatures such as those found in oven control applications. The device cold junction compensates and amplifies a type J or K thermocouple input to derive an internal signal proportional to temperature. The AD596/AD597 can be configured to provide a voltage output (10mV/°C) directly from a type J or K thermocouple signal. The device is packaged in a reliability qualified, cost effective 10-pin metal can and is trimmed to operate over an ambient temperature range from +25°C to +100°C.

Operation over an extended ambient temperature range is possible with slightly reduced accuracy. 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. The AD596/AD597 has 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.

REFERENCES

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