SPECIAL PURPOSE AMPLIFIERS

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1. WHY SPECIAL PURPOSE AMPLIFIERS?

When working in the *real* world one must accept the fact that there will be deviations from the ideal. Zero output impedances and convenient output ranges are not characteristics of most practical transducers. Also included in the real world are environmental conditions which may complicate the process of acquisition.

In real world data acquisition the most simple scenario consists of a transducer which is connected directly to a data processing system. In most cases some amplification is required and under ideal conditions can be provided by a simple op-amp and a couple of resistors. However, electrical interference, voltage drops caused by current through the resistance of leads from remote locations, nonlinear transducers, requirements for galvanic isolation and fluctuating temperatures will often complicate the task of providing accurate amplification.

In this section we will discuss various means of signal conditioning in an environment hostile to precision measurements. Instrumentation amplifiers will often serve those applications where isolation is not required and where extremely high common-mode voltages are not encountered. Isolation amplifiers are intended for use under those latter conditions. We will also discuss some highly specialized signal conditioning circuits which simplify circuit design in specific applications.

WHY SPECIAL PURPOSE AMPLIFIERS ARE REQUIRED

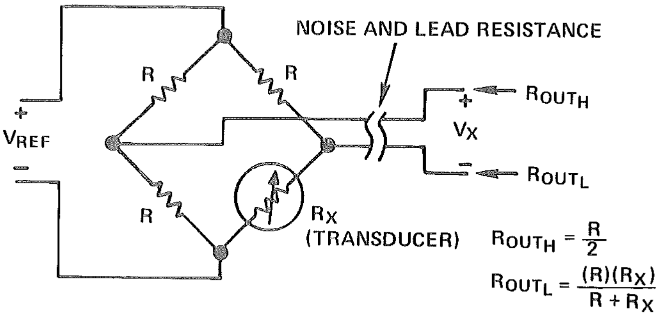
1) **Inconvenient Transducer Output Characteristics**
   - **Format of Output (Capacitance, Resistance, Current, Voltage, etc)**
   - **High Output Impedances**
   - **Inconvenient Voltage Ranges**
   - **Unbalanced Outputs**
2) **Hostile Environmental Conditions**
   - **Noise**
   - **High CMV**
   - **Remote Locations**
   - **Temperature Variances**
3) **Requirements for Isolation**
   - **Safety**
   - **Protection for Circuitry**
   - **Ground Loops**

2. LIMITATIONS OF OPERATIONAL AMPLIFIERS

IC operational amplifiers are without a doubt the most widely used analog building block in the electronics industry. Op amps are available with a wide variety of performance features at costs that are low for general purpose devices and slightly higher for devices which exhibit increased precision and/or speed. Reasons for op amp popularity are its extreme versatility (it can be configured to do more than simple amplification) and the fact that every analog circuit designer has at least a working knowledge of op amp techniques. However, in less than ideal situations, op amps have several serious shortcomings.

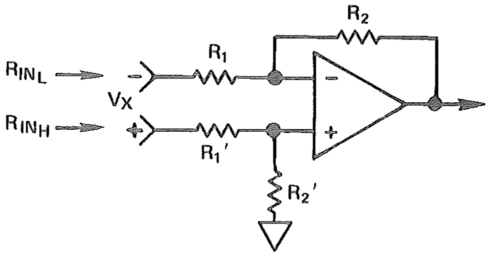
Practical transducer applications usually involve differential connections, nonzero source impedances and noise. A typical transducer bridge application is illustrated below:

PRACTICAL TRANSDUCER CIRCUIT
WITH DIFFERENTIAL OUTPUT



It can be seen that the output impedance of the above circuit is nonzero, unbalanced and varies with the parameter being measured. Also, lead resistance and noise pickup cannot be totally avoided. Due to the nature of the output, a differential amplifier is required, but a single op amp in a differential configuration is not well suited to such nonideal applications as shown below.

OP AMP IN DIFFERENTIAL CONNECTION



- Low Input Impedance
- Common Mode Rejection Depends on Resistor Ratio-Matching

To achieve balanced gain, G = (R2/R1) = (R2′/R1′)

For balanced input impedance, R1′ + R2′ = R1

To provide balanced impedance return paths for amplifier bias currents (to minimize offset voltage drift),

R1R2/(R1 + R2) = R1′R2′/(R1′ + R2′)

Furthermore, these mutually exclusive conditions are only valid for ideal transducers. Finite input impedances, even if balanced, can disturb unbalanced transducers with lead resistances further aggravating the situation.

Common-mode rejection for most op amps is typically between 60dB and 90dB (some types, like the AD517 and AD OP-07, may have up to 110dB). This may not be sufficient to reject common-mode noise.

Also, op amps do not exhibit galvanic isolation between input and output. They cannot handle signals superimposed upon common-mode voltages in excess of ± 10 volts and if the common-mode input voltage exceeds the supply voltage, they may be destroyed.

3. INSTRUMENTATION AMPLIFIERS

An instrumentation amplifier (IA) is a precision differential voltage amplifier that is intended for use when acquisition of a useful signal is difficult. As was stated earlier, real world signals from practical transducers are often plagued by problems such as noise, unbalanced output impedances, etc. IA's are designed to solve these signal acquisition problems. IA's have stable gain, high input impedances, low bias currents, high common-mode rejection and balanced differential inputs. Gain is normally determined via pin strapping, or a user-selectable resistor or resistor pair, and can be programmed in the range of 1 to 10,000. All other necessary precision components are internal which allows the manufacturer to guarantee a specified level of performance. The output is single ended, usually with sense and reference terminals.

INSTRUMENTATION AMPLIFIER CHARACTERISTICS

1) **Fixed, Stable Gain Determined by User-Selectable Resistor, Resistor Pair, or Pin Strapping (Internal Resistors)**
2) **High Input Impedance**
3) **Low Bias Currents**
4) **High Common-Mode Rejection**
5) **Balanced Differential Inputs**
6) **Stable, Well Characterized Specs**
7) **Single-Ended Output**

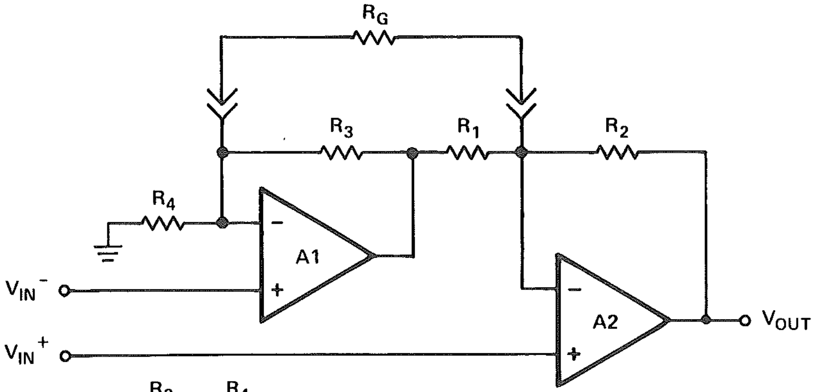
The performance of an IA, like that of an op amp, is described by its specifications. Unlike op amps, however, IA's do not have as wide a variety of performance levels. This is because IA's themselves are a specialized class of amplifier.

DESIGN TECHNIQUES

IA configurations are based on op amps. We have already analyzed the simplest design, using a single op amp, and found that it lacks the performance required for precision applications.

An IA can be designed using two amplifiers which overcomes some of the weaknesses exhibited by a single amplifier.

"TWO-AMPLIFIER" INSTRUMENTATION AMPLIFIER



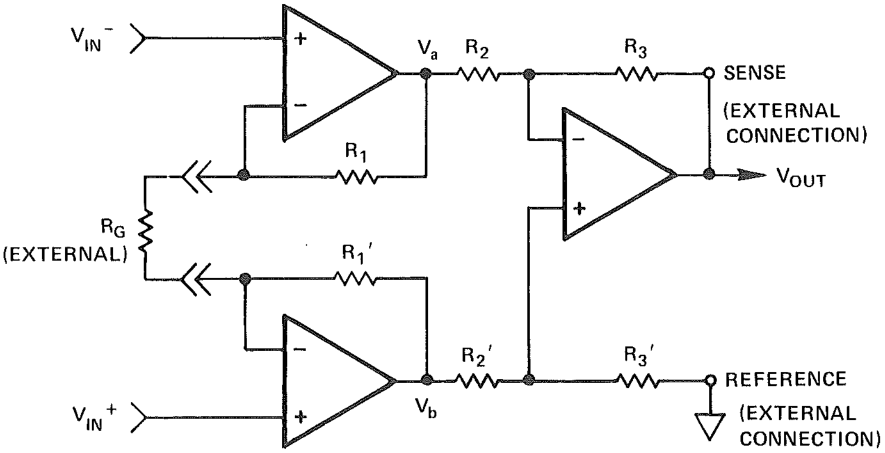
IF $\frac{R_2}{R_1} = \frac{R_4}{R_3}$ $\quad V_{IN} = V_{IN}^+ - V_{IN}^-$

$$\frac{V_{OUT}}{V_{IN}} = 1 + \frac{R_2}{R_1} + \frac{2R_2}{R_G}$$

The input impedance of this circuit is high, permitting the signal sources to have high and unbalanced output impedance. The major disadvantage of this design is that the common-mode voltage input range must be traded off against gain range. The amplifier A1 must amplify a common-mode signal by (R3 + R4)/R4. If R3>R4, saturation of A1 will occur if the common-mode signal is too high leaving no headroom to amplify the differential signal of interest and if R3<R4, low gains cannot be realized.

The most popular configuration for op amp based instrumentation amplifiers is shown below.

"CLASSIC" 3 OP AMP INSTRUMENTATION AMPLIFIER



The transfer function of this circuit is:

$$V_{OUT} = ((+V_{IN}) - (-V_{IN}))(2R1/RG + 1)(R3/R2)$$

where $R1 = R1'$, $R2 = R2'$, and $R3 = R3'$

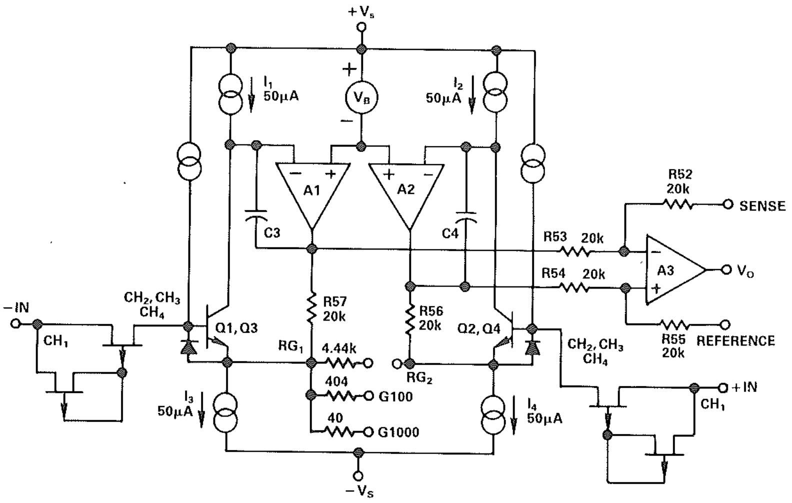
In this configuration, gain accuracy and CMR depend upon the ratio matching of R2, R2′, and R3 and R3′. Circuit analysis shows, however, that CMR does not depend on the matching of R1 and R1′.

Within limits, the user may take as much gain in the front end as he wishes (set by the value RG) without increasing the common-mode error signal. Thus, CMR will theoretically increase in direct proportion to gain. Furthermore, common-mode signals are only amplified by a factor of 1 regardless of gain (no common-mode voltage will appear across RG, hence, no common-mode current will flow in it because the input terminals of an op amp operating normally will have no significant potential difference between them). This means that the large common-mode signals (within the op amp limits) may be handled independent of gain.

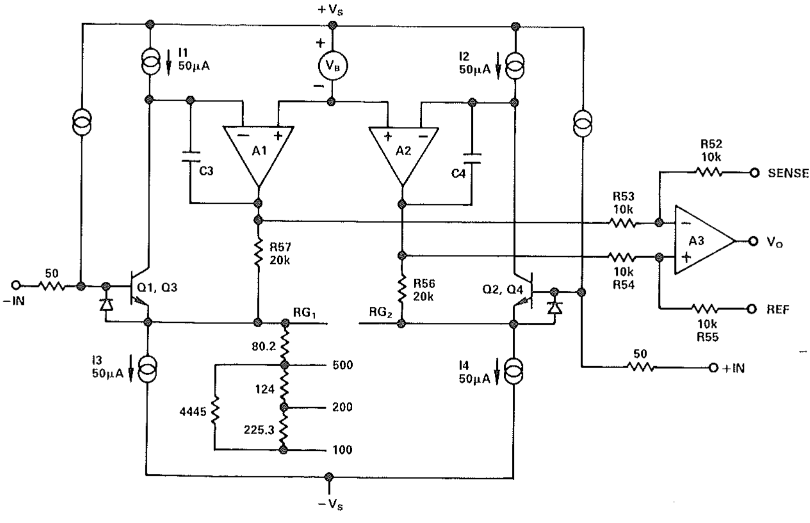
Finally, because of the symmetry of this configuration, first order common-mode error sources in the input amplifiers, if they track, tend to be cancelled out by the output stage subtractor. IA's of this type include the AD524, AD624 and AD625. They are characterized by extremely high precision. The input amplifiers may be either FET or Bipolar. FET input op amps have very low bias currents and are well suited for use with very high source impedances, but generally have poorer CMR than bipolar amplifiers because nongeometry-related mismatches usually cause larger input offset voltage drifts.

The AD524, AD624 and AD625 are monolithic instrumentation amplifiers based on the classic 3 op amp circuit. The advantage of monolithic fabrication is that the closely matched components required to construct the preamplifier are easily fabricated on a single chip. The preamplifier section develops the programmed gain by the use of feedback. The gain is programmed by varying the value of RG (smaller values increase the gain). Feedback forces the collector currents of Q1, Q2, Q3 and Q4 to be constant which impresses the input voltages across RG.

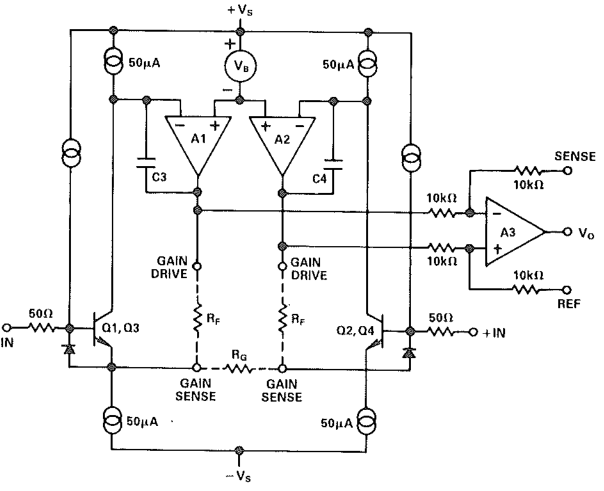
AD524 SIMPLIFIED SCHEMATIC



AD624 SIMPLIFIED SCHEMATIC



AD625 SIMPLIFIED SCHEMATIC



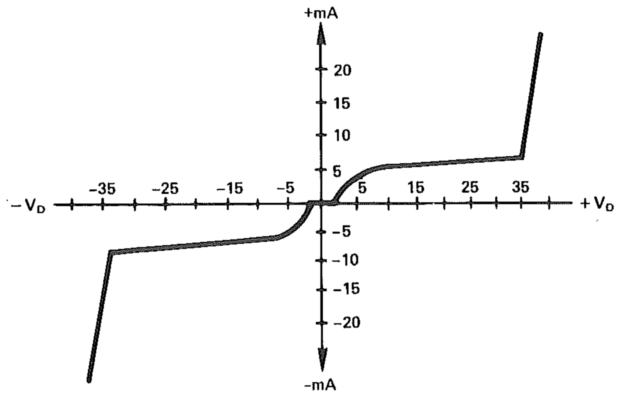
As RG is reduced, the transconductance of the input preamp increases to the transconductance of the input transistors. This has three important advantages. First, this approach allows the circuit to achieve a very high open loop gain of 3×10^8 at a programmed gain of 1000, reducing gain related errors to a negligible 30ppm. Second, the gain bandwidth product which is determined by C3, C4 and the input transconductance, reaches 25MHz. Third, the input voltage noise reduces to a value determined by the current of the input transistors for an RTI noise of $4\text{nV}/\sqrt{\text{Hz}}$, at $G = 1000$.

As interface amplifiers for data acquisition systems, instrumentation amplifiers are often subjected to input overloads, i.e., voltage levels in excess of the full scale for the selected gain range. At low gains, 10 or less, the gain resistor acts as a current limiting element in series with the inputs. At high gains the lower value of RG will not adequately protect the inputs from excessive currents. Standard practice would be to place series limiting resistors in each input, but to limit the current to below 5mA with a full differential overload (36V) would require over 7kΩ of resistance which would add $10\text{nV}/\sqrt{\text{Hz}}$ of noise. To provide both input protection and low noise the AD524 utilizes a special series protection FET.

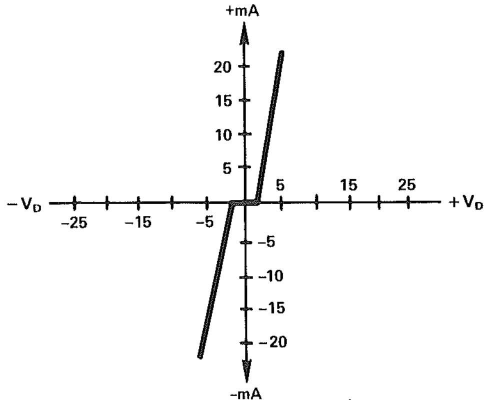
This unique FET design was used to provide a bidirectional current limit, protecting against both positive and negative overload conditions. Under nonovereloaded conditions, three channels CH_2, CH_3 and CH_4 act as a resistance (1kΩ) in series with the inputs as before. During an overload in the positive direction, a fourth channel, CH_1, acts as a small resistance (3kΩ) in series with the gate, which draws only the leakage current, and the FET limits to I_{DSS}. When the FET enhances under a negative overload, the gate current must go through the small FET formed by CH_1 and when this FET goes into saturation, the gate current is limited and the main FET will go into controlled enhancement. The bidirectional limiting holds the maximum input current to 3mA over the 36V range with power on or power off.

For applications where input overload is not expected, the AD624 and AD625 replace the protection FETs with 50Ω resistors. This reduces the input voltage noise to $4\text{nV}/\sqrt{\text{Hz}}$ while limiting input current to 10mA with a 2.5V differential input overload.

AD524
INPUT PROTECTION
CHARACTERISTIC



AD624
INPUT PROTECTION
CHARACTERISTIC

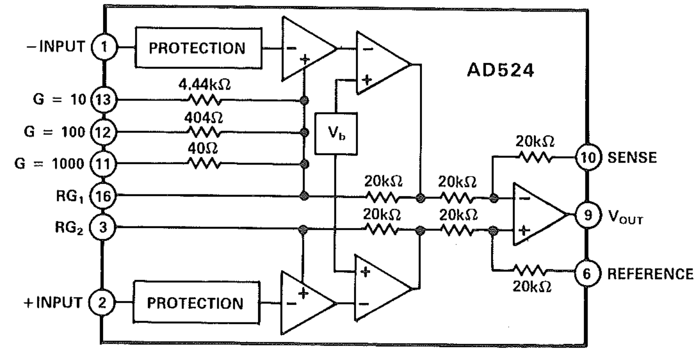


The AD524 and AD624 are high precision monolithic instrumentation amplifiers designed for data acquisition applications requiring high accuracy. As previously stated, the AD524 is fully protected from overloads with power on or off, making it suitable for operation under worst case conditions. The AD624 has been optimized for use with low level transducers, including load cells, strain gages and pressure transducers. Its combination of high linearity and low noise make the AD624 ideal for use in high resolution data acquisition systems.

AD524 FEATURES

Low Nonlinearity: 0.003% (G = 1)
High CMRR: 120dB (G = 1000)
Low Offset Voltage: 50μV
Low Offset Voltage Drift: 0.5μV/°C
Gain Bandwidth Product: 25MHz
Pin Programmable Gains of 1, 10, 100, 1000
Complete Input Protection Power On – Power Off
No External Components Required
Low Noise: 0.3μV p-p (0.1Hz to 10Hz)

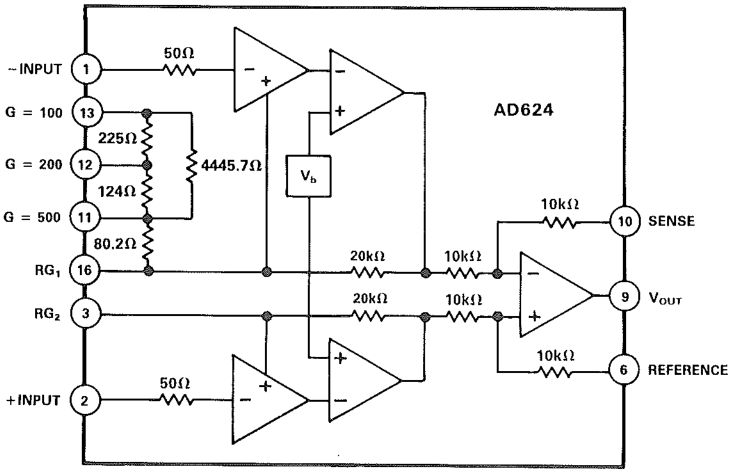
AD524 FUNCTIONAL BLOCK DIAGRAM



AD624 FEATURES

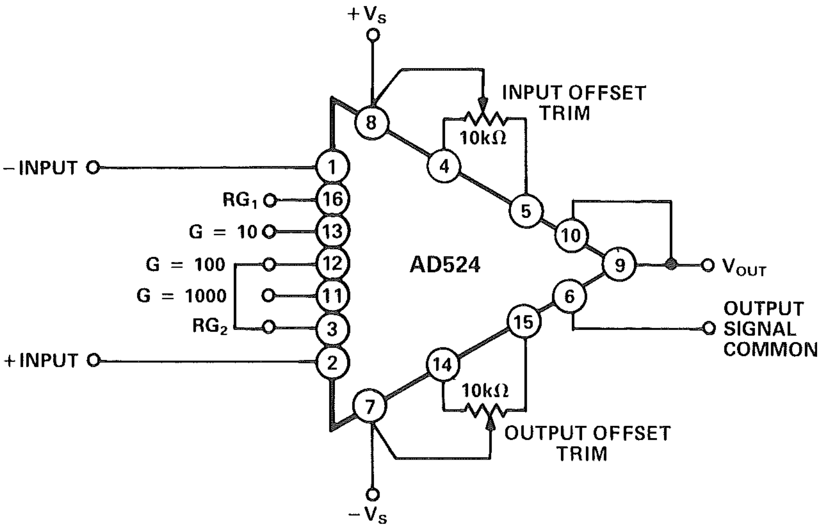
Low Nonlinearity: 0.001% (G = 1)
High CMRR: 115dB (G = 500)
Low Offset Voltage: 25μV
Low Offset Voltage Drift: 0.25μV/°C
Gain Bandwidth Product: 25MHz
Pin Programmable Gains of 1, 100, 200, 500, 1000, and More
No External Components Required
Internally Compensated
Low Noise: 0.2μV p-p (0.1Hz to 10Hz)

AD624 FUNCTIONAL BLOCK DIAGRAM



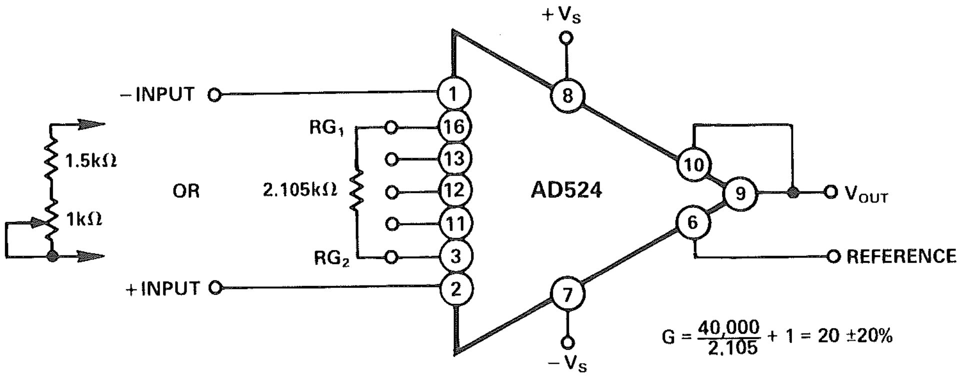
The AD524 and AD624 have internal high accuracy pretrimmed resistors for pin programmable gain setting (1, 10, 100 and 1000 for the AD524; 1, 100, 200, 500 and 1000 for the AD624). One of the present gains can be selected by pin strapping the appropriate gain terminal to RG_2.

OPERATING CONNECTIONS FOR G = 100



The AD524 and AD624 can be configured for other gains. An external resistor can be connected between pins 2 and 16 which programs the gain according to the formula G = 1 + 40kΩ/RG. The external RG should be a precision low TC resistor. An external RG effects both gain accuracy and gain drift due to the mismatch between it and the internal thin-film resistors. Gain accuracy is determined by the tolerance of the external RG and the absolute accuracy of the internal resistors (±20%). Gain drift is determined by the TC of RG compared to that of the internal resistors (−50ppm/°C typ).

OPERATING CONNECTIONS FOR G = 20



$$G = \frac{40,000}{2.105} + 1 = 20 \pm 20\%$$

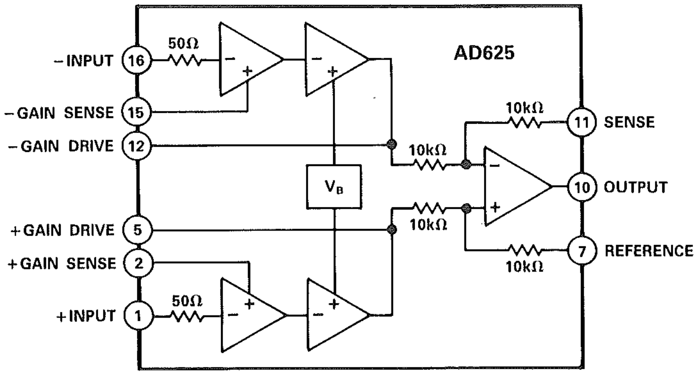
In the AD625, the appropriate circuit points have been pinned out so that the gain can be completely determined using external resistors. The absence of internal gain setting resistors alleviates the problem of gain accuracy and gain drift due to mismatch in internal/external resistors.

In the resistor-programmed mode of the AD625, only three external resistors are needed to select any gain from 1 to 10,000. Depending upon the application, discrete components or a pretrimmed network can be used. The gain accuracy and gain TC are primarily determined by the external resistors since the AD625 contributes less than 0.02% gain error and under 5ppm/°C gain TC. The gain sense current is insensitive to common-mode voltage, making the CMRR of the resistor programmed AD625 independent of the match of the two feedback resistors (RF).

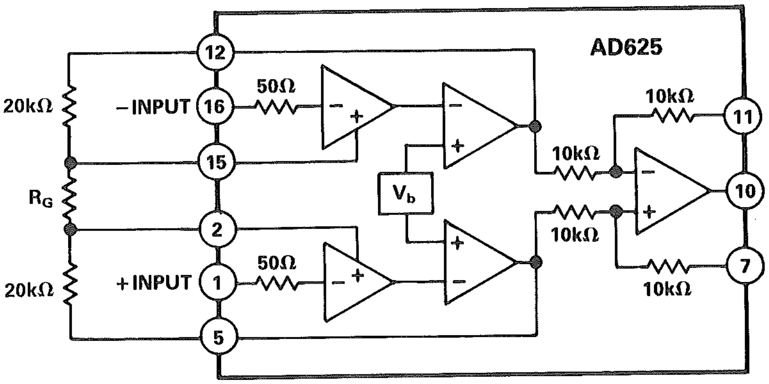
AD625 FEATURES

User Programmed Gains 1 to 10,000
Low Gain Error: 0.02% max
Low Gain T.C.: 5ppm/°C max
Low Nonlinearity: 0.001% max
Low Offset Voltage: 25μV
Low Noise: 4nV/Hz (at 1kHz) RTI
Gain Bandwidth Product: 25MHz

AD625 FUNCTIONAL BLOCK DIAGRAM



AD625 RESISTOR PROGRAMMED MODE



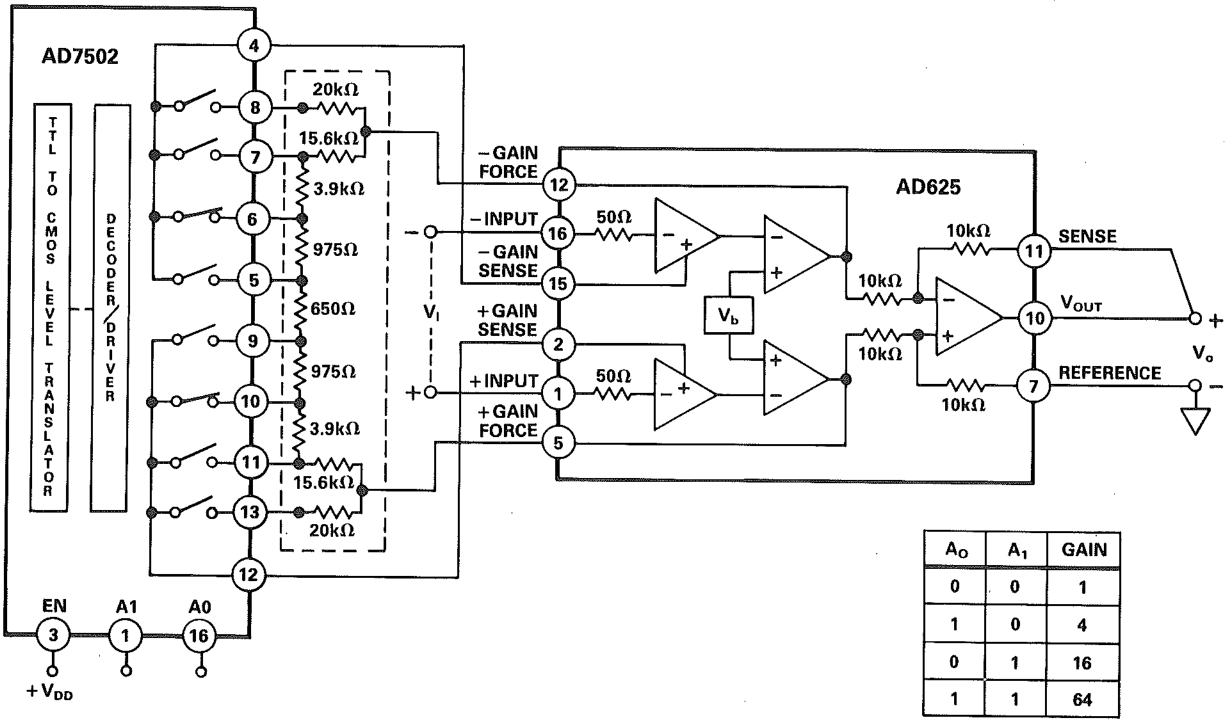
$$G = \frac{2R_F}{R_G} + 1$$

A software programmable gain amplifier (SPGA) provides the ability to program precision gains digitally. Historically, the problem in the systems requiring electronic switching of gains has been the ON resistance of the multiplexer, which appears in series with the gain setting resistor RG. This can result in substantial gain errors and gain drifts. The AD625 eliminates this problem by making the gain sense pins available (Pins 2, 15, 5 and 12). As a consequence, the multiplexers ON resistance is removed from the signal current path. This transforms the ON resistance error into a small, nullable offset error.

The figure below demonstrates an AD625-based SPGA with gain of 1, 4, 16 and 64. RG is the resistance between the gain sense lines (pins 2 and 15) of the AD625. With the multiplexer in the state shown, RG is the sum of two 975Ω resistors and the 650Ω resistor, or 2600Ω. RF is the resistance between the gain sense and the gain drive pins (pins 12 and 15, or pins 2 and 5), the 15.6kΩ resistor plus the 3.9kΩ resistor, or 19.5kΩ. The gain is therefore:

$$(2RF/RG) + = 9(19.5k\Omega)/(2.6k\Omega) + 1 = 16$$

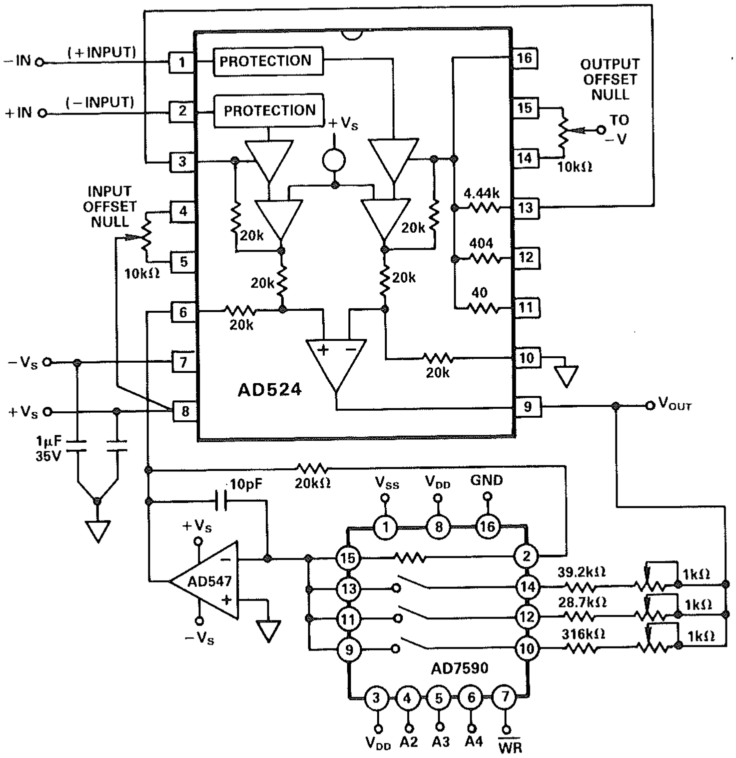
SOFTWARE PROGRAMMABLE GAIN AMPLIFIER USING THE AD625



A_O	A_1	GAIN
0	0	1
1	0	4
0	1	16
1	1	64

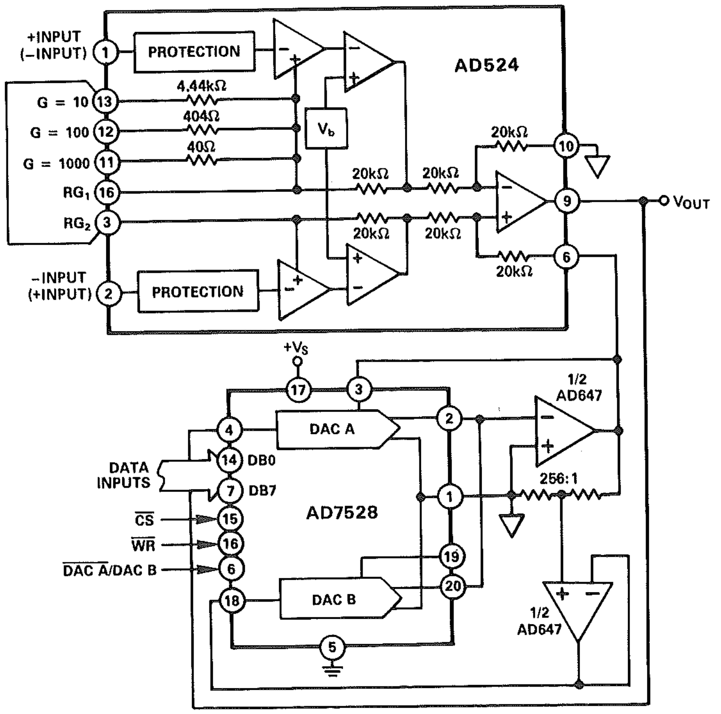
In many cases, an instrumentation amplifier can be connected for gain in the output stage. The figure below shows an AD547, a CMOS switch and resistors being used as a programmable active attenuation in the output amplifier's feedback loop. The active attenuation presents a very low impedance to the feedback resistors therefore minimizing CMRR degradation.

PROGRAMMABLE OUTPUT GAIN



Another method for developing gain in the output stage is to use a CMOS DAC. This method enables the user to select any arbitrary gain (Within limits). The AD7528 dual CMOS DAC acts as a pair of switched resistive attenuators having high analog linearity and symmetrical bipolar transmission. The multiplying DAC can handle inputs of either polarity without affecting the programmed gain. The circuit shown uses one half of the AD7528 (DAC A) to set the gain and the other half (DAC B) to perform a fine adjustment.

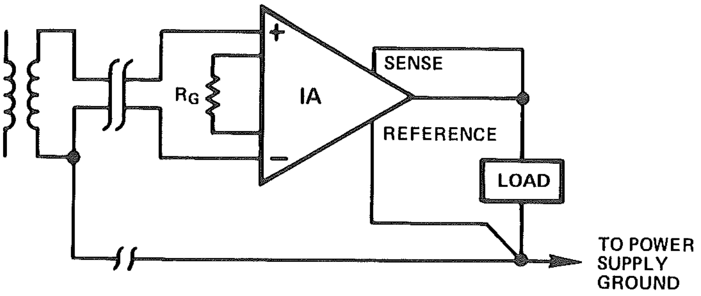
PROGRAMMABLE OUTPUT GAIN USING A DAC



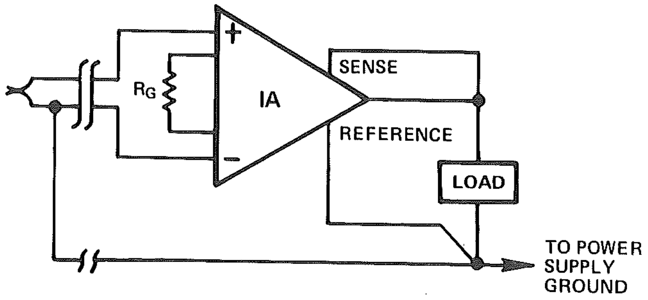
APPLICATION NOTES

Although instrumentation amplifiers have differential inputs, the bias currents do not flow from one to the other but to the external circuit. There must, therefore, be a return path for the bias currents. If this path is not provided, those currents will not flow, causing the output of the amplifier to drift uncontrollably or to saturate. Therefore, when amplifying floating input sources such as transformers and thermocouples, as well as ac-coupled sources, there must be a dc return path from each input to ground. Suitable connections are shown below. In applications where it is impossible to provide a dc path for bias currents, an isolation amplifier is necessary.

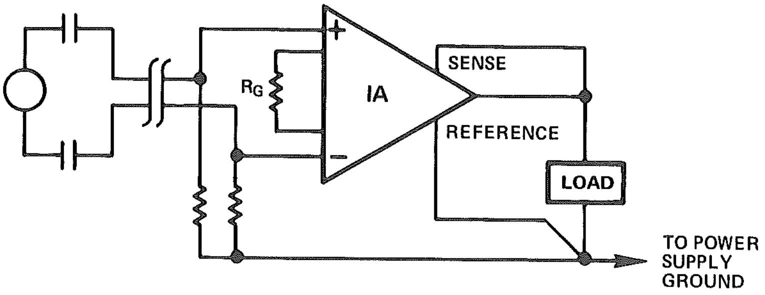
INDIRECT GROUND RETURNS FOR BIAS CURRENTS



a). TRANSFORMER COUPLED



b). THERMOCOUPLE

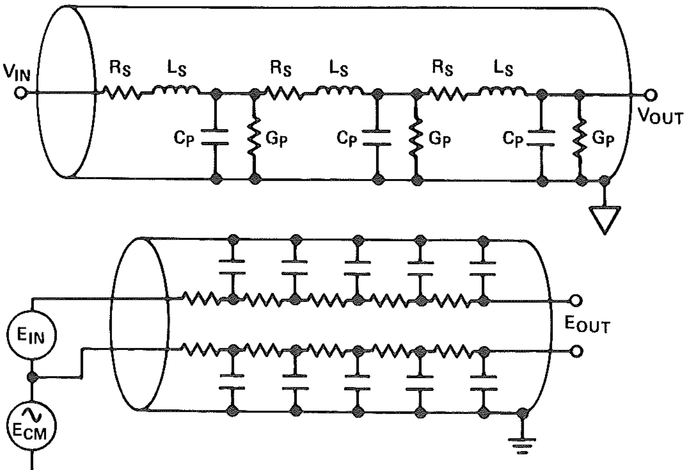


c). AC COUPLED

Data Guard

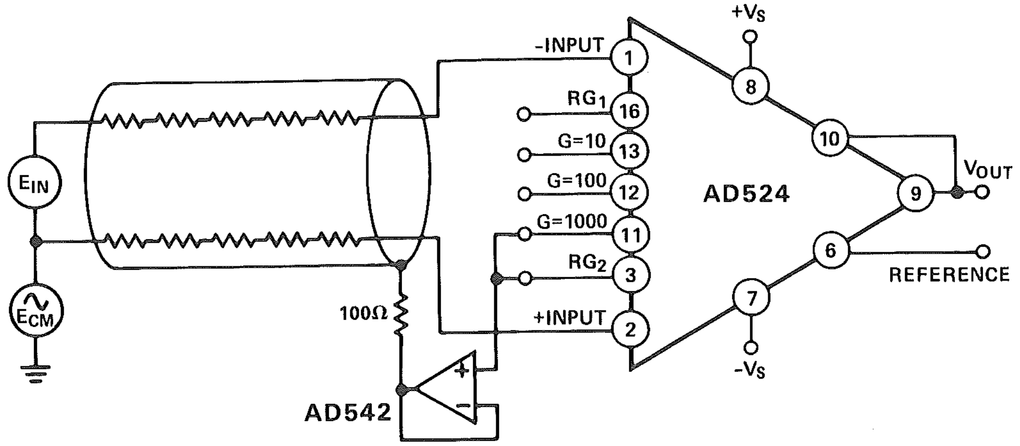
Signals from remote transducers are often transmitted to an IA through shielded cables. While this may reduce noise pickup, the distributed RCs in such cabling can cause differential phase shifts in those lines. When ac common-modes signals are present, these phase shifts will reduce common-mode rejection. If the shields are driven by the common-mode signal, the cable capacitance is "bootstrapped" thus making the capacitance effectively zero for common-mode signals.

RG58 CABLE

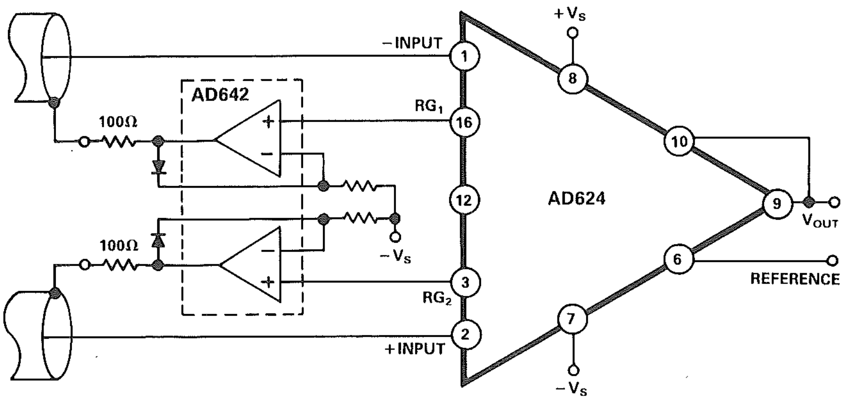


$E_{OUT} = E_{IN} + E_{CM}\ \omega(R\Delta C + \Delta RC)$

IC instrumentation amplifiers do not have a buffered data guard available externally. However, the common-mode voltage for amplifiers such as the AD525, AD624 and AD625 is available at RG_2 and should be buffered with an operational amplifier. The op amp should be of the FET input variety for low bias currents (AD542, AD548, etc). Low bias currents are required because bias currents flowing through the gain setting resistor will cause an offset error.



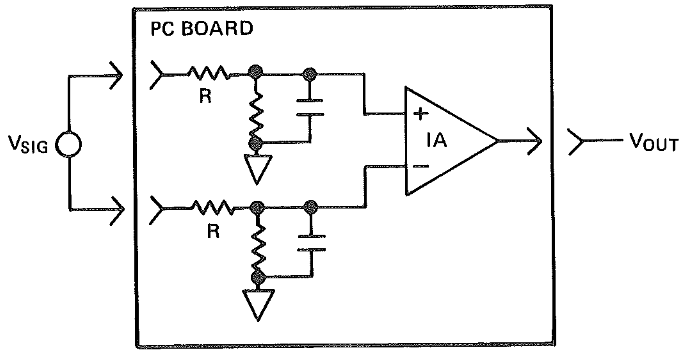
DIFFERENTIAL SHIELD DRIVER



Most circuits today are designed on printed circuit (PC) boards. Although the benefits of PC construction include reduction in circuit size and reduced noise pickup, the ability to apply shields and guards can be even more effective in improving circuit performance.

In some circuits, wiring capacitance and leakage resistance can degrade performance. In an instrumentation amplifier, ac common-mode rejection is directly limited by differential phase shift.

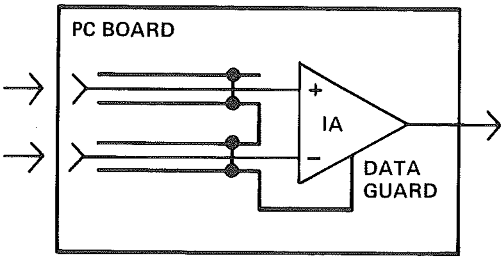
PARASITICS IN PRINTED CIRCUITS



Unequal drops across unbalanced PC board track resistances and differential phase shift due to varied stray capacities can cause a decrease in common-mode rejection similar to the effect of long input cables.

A data guard allows the common-mode rejection to be restored by guarding the inputs as shown in the diagram below.

DATA GUARD "BOOTSTRAPS" PARASITICS FOR COMMON-MODE SIGNALS

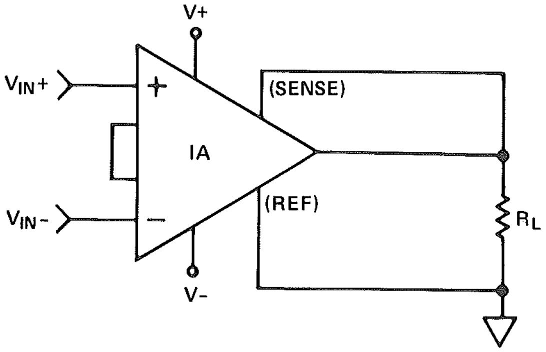


The parasitic components of the PC board are "bootstrapped" for common-mode signals. That is, if there is no voltage across the strays and leakages, no current will flow. Therefore, the effects of these parasitics are minimized.

Sense Terminal

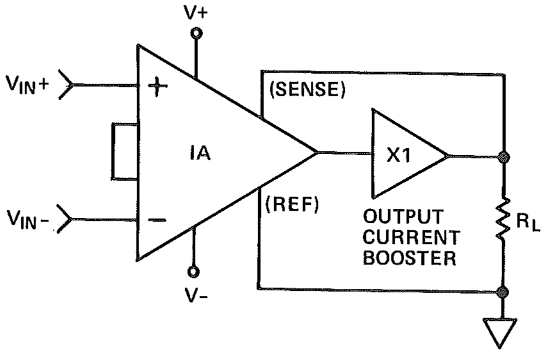
The sense terminal is the feedback point for the IA output amplifier. Normally it is connected directly to the IA output. If heavy load currents are to be drawn through long leads, IR drops can cause errors. Under these conditions, the sense terminal can be wired to the IA output at the load, putting the IR drops inside the loop and virtually eliminating this source of error.

INSTRUMENTATION AMPLIFIER OUTPUT CONNECTIONS



Typically, IC instrumentation amplifiers are rated for a full ± 10 volt output swing into a 2kΩ load. In some applications, however, more current is required. This cannot be provided on-chip because of the damaging effect of thermal gradients, due to differential heating, on the analog accuracy of an IA. A high-current booster, which can be an inexpensive audio amplifier in unity gain mode, may be connected inside the loop of an instrumentation amplifier to provide the required current boost without degrading overall performance. Nonlinearities, offset and gain inaccuracies of the buffer are minimized by the loop gain of the IA output amplifier, as is offset drift.

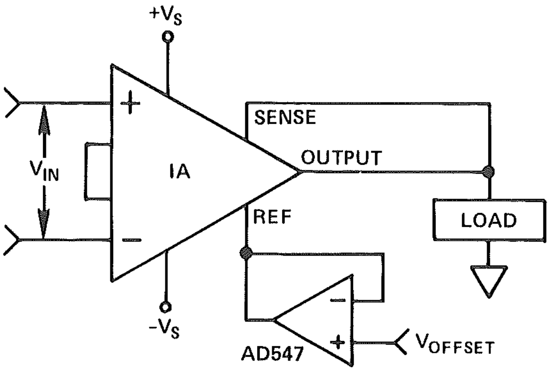
INSTRUMENTATION AMPLIFIER WITH OUTPUT CURRENT BOOSTER



Reference Terminal

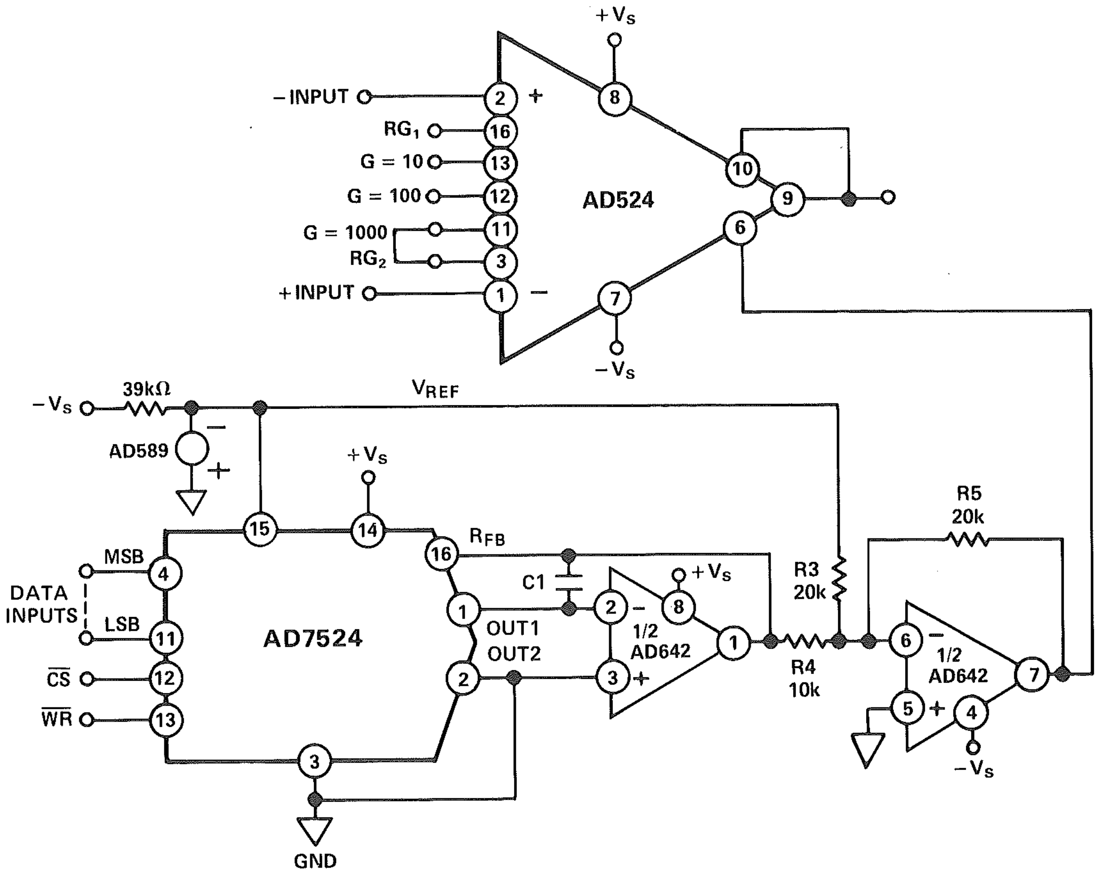
The reference terminal may be used to offset the output by up to ± 10 volts. This is useful when the load is floating or does not share a ground with the rest of the system. It also provides a means of injecting a precise offset. When injecting this offset, it is necessary that nearly zero impedance be presented to the reference terminal. Any significant resistance from the reference terminal to ground increases the gain of the noninverting signal path thereby upsetting the common-mode rejection of the IA. An operational amplifier is an ideal choice to provide this low impedance reference point.

USE OF REFERENCE THERMINAL TO PROVIDE OUTPUT OFFSET



In many applications it is necessary to provide very accurate data in high gain configurations. At room temperature, offset effects can be nulled by the use of trimpots. Over a wide operating temperature range, however, offset nulling becomes a problem. The circuit below uses a CMOS DAC connected to the reference terminal to provide software controllable offset adjustments.

SOFTWARE CONTROLLABLE OFFSET



Grounding and Decoupling

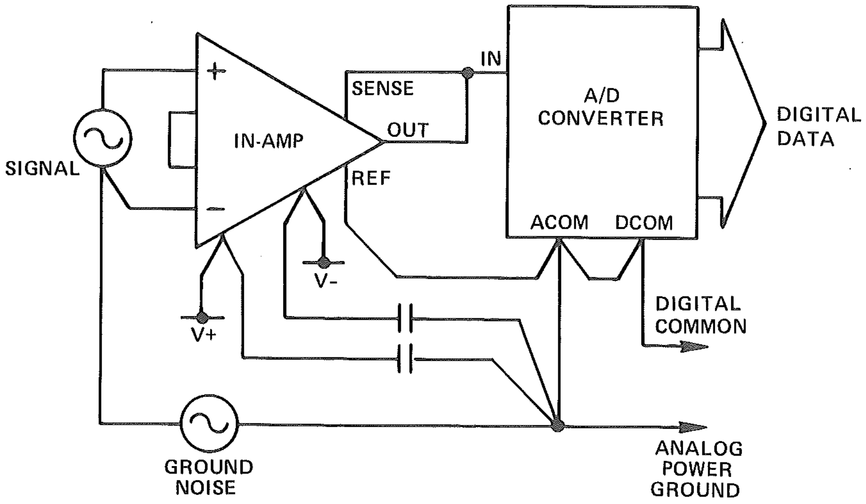
Data acquisition components usually have two or more ground pins which are not connected together within the device but have to be connected to a ground system externally. Ideally, a single solid ground would be desirable and typically that is what is used. The result is that a great amount of effort and many decoupling components are expended attempting to correct problems created by poor ground current management. In large systems and systems which combine high-level and low-level signals, "ground" (or common bus) management becomes an important aspect of design. Allowing low-level analog signals to share conductors with logic returns or power connections is an invitation to trouble.

In large systems it is often impractical to rely on a single common point for all analog signals. In these cases, some form of differential amplifier is required to translate signals between grounding systems. A simple subtractor or instrumentation amplifier can often be used for this purpose, translating a signal which is referred to one ground system into a similar or amplified signal which is referred to a different ground system. The common-mode rejection of the amplifier is used to eliminate the effects of voltage differences between the two ground or common points.

If an operational amplifier is used as a subtractor the op amp should be powered from the load power and/or decoupled with respect to load common. The reason for this can be deduced from the circuit architecture of the most common types of op amps. An op amp converts a differential input signal to a single-ended output signal. In many popular op amps the differential-to-single -ended conversion is done with respect to V−, and the resulting signal drives an integrator. The integrator characteristic is used to frequency compensate the amplifier, and its input is referred to the single-ended output, at V−. The integrator acts as a unity gain follower for fast signals applied to its noninverting (or reference) input. As a result, signals applied to the V− terminal have their high frequency components conveyed directly to the output. Signals having frequency components above the amplifier closed-loop bandwidth will be transmitted from V− to the output with little or no attenuation.

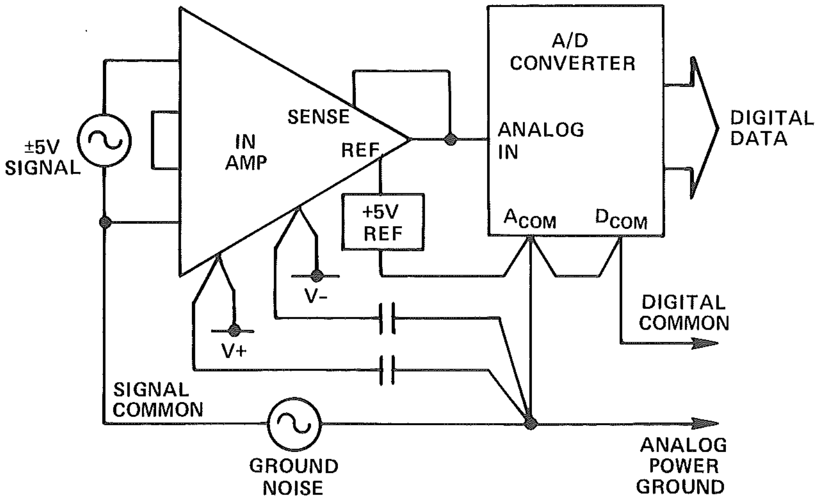
As discussed earlier, the noise rejection performance of the subtractor depends on carefully matched source and feedback resistance ratios and it cannot be used in all situations. Whenever the source impedance cannot be controlled or is exceptionally high, the subtractor (or dynamic bridge) becomes impractical. In this situation, ground noise and other remote grounding difficulties can often be avoided by the use of an instrumentation amplifier. The IA accepts differential input signals at its high impedance input. It provides a fixed gain without introducing overall feedback joining the input and output circuitry. The output signal is developed with respect to a reference terminal, which may be connected to the input common of a remote load-circuit.

INSTRUMENTATION AMPLIFIER SIMPLIFIES SIGNAL CONDITIONING FOR A/D CONVERTER



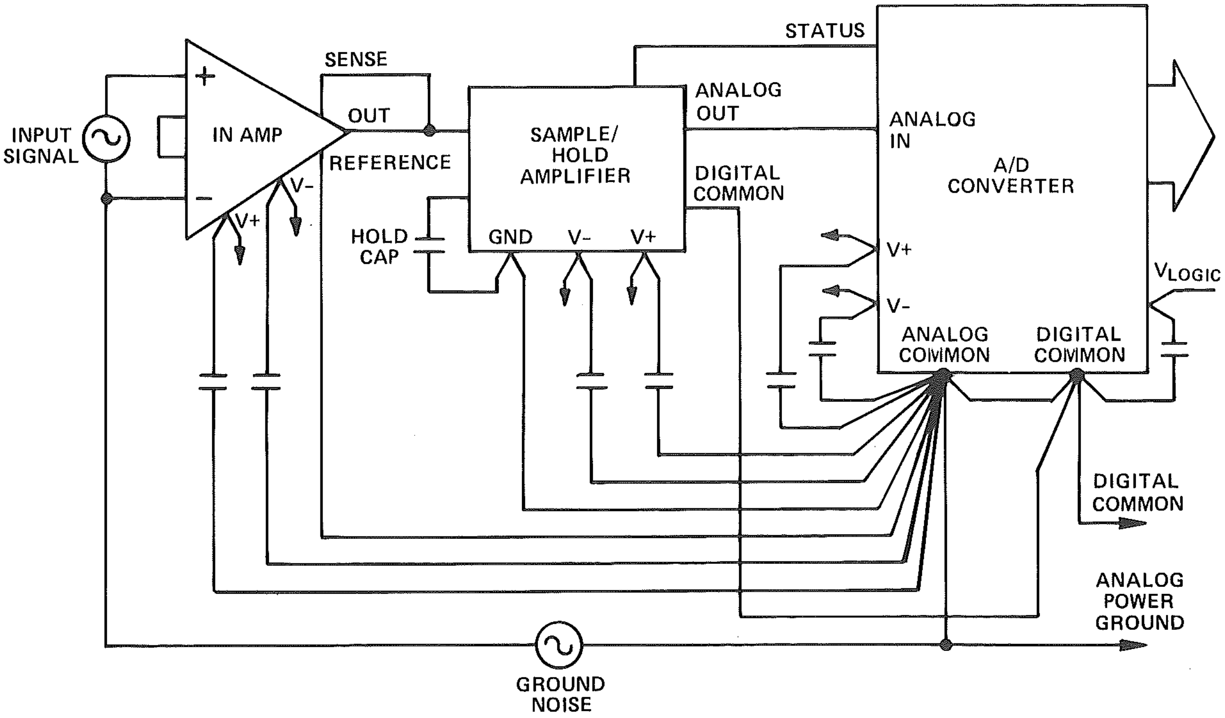
Some IAs are quite versatile and can be adapted to provide additional functions while they isolate common returns. For example, converting a bipolar input signal for use with a unipolar ADC. By referring the IA output to +5 volts, the ±5 volt amplifier input signal will appear as a 0 to +10 volt signal to the converter. This extra feature can be provided without compromising the ground noise rejection of the system.

INSTRUMENTATION AMPLIFIER USED FOR BIPOLAR OFFSET AND ISOLATION



In data acquisition systems containing instrumentation amplifiers, sample-and-hold amplifiers and A/D converters the grounding problems become complex. In such a case, the analog subsystem should be powered by a supply with a local common return which may be connected to the digital common but does not share any current carrying conductors. Ideally, there are not foreign currents which flow between the analog system and the digital system, except for those within the converter. If the two systems are joined only at the converter, the foreign currents share the shortest path, and their effects are minimized.

DATA ACQUISITION SYSTEM GROUNDING



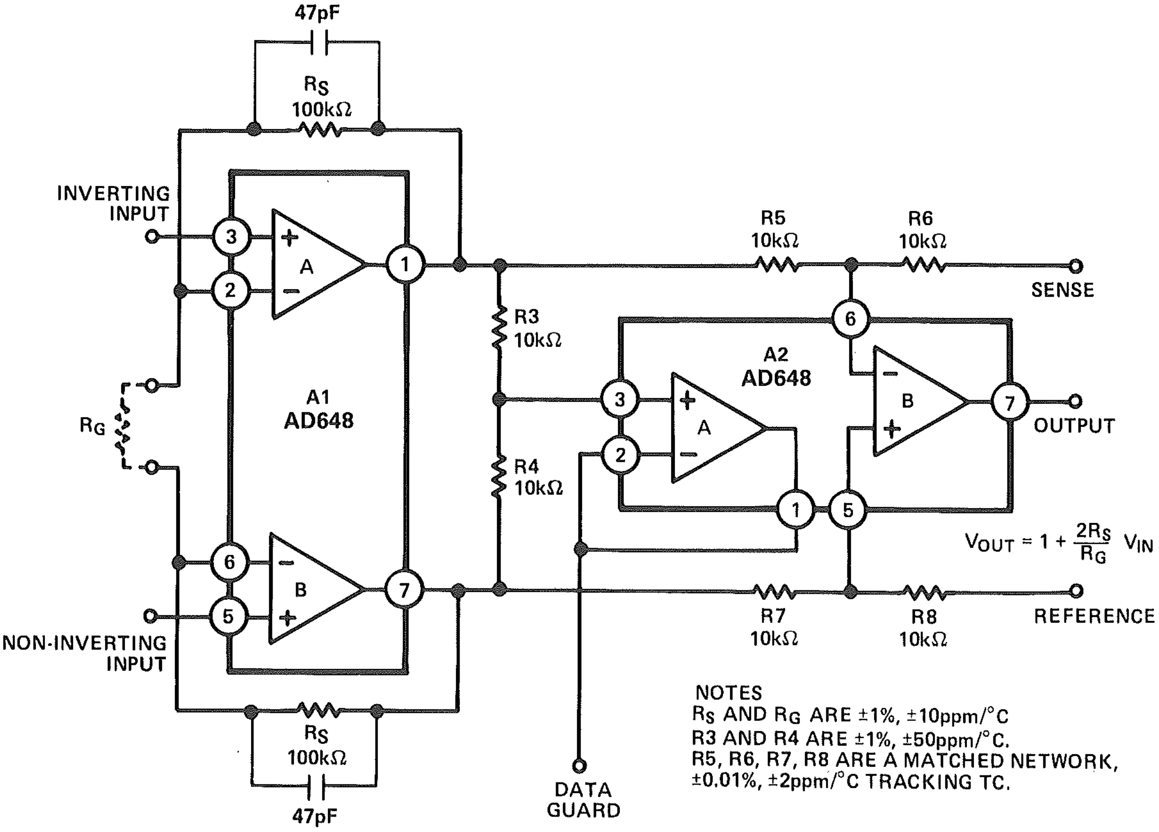
Discrete Instrumentation Amplifiers

Various forms of discrete instrumentation amplifiers can be constructed using IC operational amplifiers. The advantage of a discrete IA is flexibility in design, optimizing the IA for the most important aspect of a particular application, for instance, using FET input op amps for low bias currents or using fast, precision amplifiers for wideband applications. The disadvantage of a discrete design is that careful component selection is required.

The "classic" 3 op amp instrumentation amplifier can be implemented using 2 dual FET input amplifiers for applications requiring low bias currents and low power. In this application, the matching characteristics of the two amplifiers are critical to ensure high performance. The use of an AD648C (A_1) as the input amplifiers, guarantees a maximum input offset voltage of 300μV, input offset voltage drift of 3.0μV/°C and bias currents of 10pA. The AD712C may be substituted for greater speed, at the cost of increased supply current (200μA vs. 2.8mA per amplifier). A_2 serves less critical functions in the IA and, therefore can be an AD648A. Amplifier A of A_2 is an active data guard which increases ac CMR and minimizes extraneous signal pickup and leakage. Amplifier B of A_2 is the output amplifier of the IA.

The external resistor characteristics are very critical to achieve the precision available from the AD648C in this configuration, therefore, a great deal of care should be taken in their selection. CMRR will depend upon the matching of resistors R5, R6, R7 and R8. For example, a resistor mismatch of 0.1% results in a CMR of 60dB while a 0.01% will yield an 80dB CMR. The gain drift and CMR over temperature is directly effected by the TC match of the resistors used, therefore, it is recommended that a matched resistor network which tracks to ±2ppm/°C should be used.

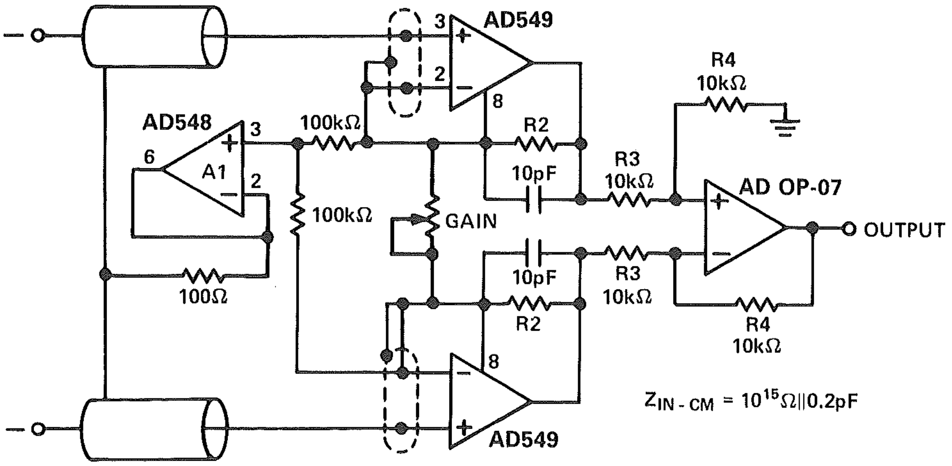
PRECISION FET INPUT INSTRUMENTATION AMPLIFIER



NOTES
RS AND RG ARE ±1%, ±10ppm/°C
R3 AND R4 ARE ±1%, ±50ppm/°C.
R5, R6, R7, R8 ARE A MATCHED NETWORK, ±0.01%, ±2ppm/°C TRACKING TC.

If there is a need for even lower bias current, a pair of AD549Cs (0.075pA) bias currents should be used as the input amplifiers. The AD549 has laser-trimmed offset voltage of 250µV, low drift (5.0µV/°C), low noise (4.0µV p-p, 0.1 to 10Hz) and low power consumption (600µA supply current).

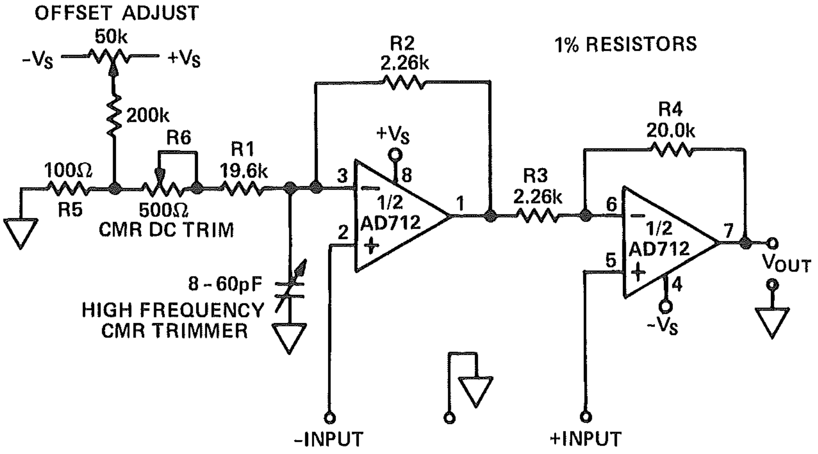
VERY HIGH IMPEDANCE INSTRUMENTATION AMPLIFIER



(ALL RESISTORS OF SAME NUMBER SHOULD BE MATCHED ±0.1%)
(BUFFER A1 BOOSTS COMMON MODE Z_{IN} BY DRIVING CABLE SHIELDS AT COMMON MODE VOLTAGE AND NEUTRALIZING CM CAPACITANCE)

A two op amp IA built with an AD712 can provide high accuracy signal conditioning with high frequency input signals. The circuit will have an offset voltage drift of 5V/°C, CMRR of 90dB over a range of dc to 1kHz and a bandwidth of 500kHz (−3dB) at 1V p-p output. The circuit can be configured for a gain range of 2 to 1000 with a typical nonlinearity of 0.01% at a gain of 10.

WIDE BANDWIDTH INSTRUMENTATION AMPLIFIER



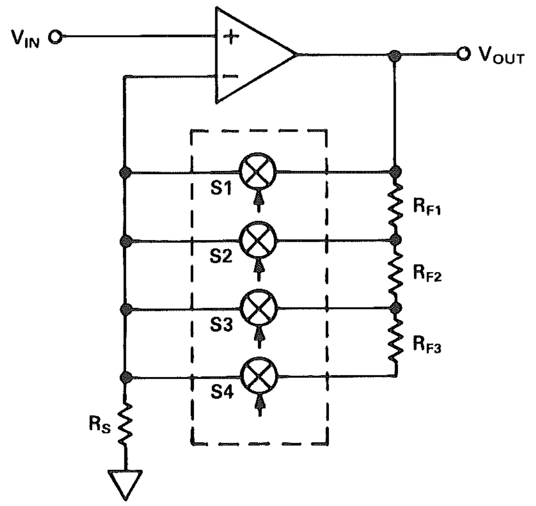
$$V_{OUT} = \left[\left(\frac{R_4}{R_3}+1\right)\left(+V_{IN}+V_{OS2}\right) - \left(-V_{IN}+V_{OS1}\right)\left(\frac{R_4}{R_3}\right)\left(\frac{R_2}{R_1}+1\right)\right]$$

INSTRUMENTATION AMPLIFIER WITH GAIN OF TEN

4. SINGLE-ENDED SPGA

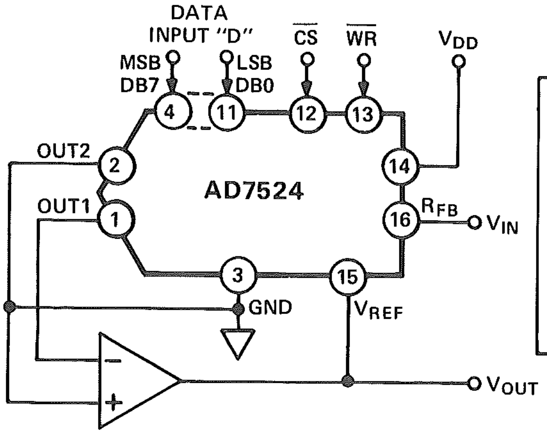
The previous section discussed software programmable gain configurations of instrumentation amplifiers. SPGA applications include gain ranging pre-amps and dynamic range extension for ADC systems. The advantages of an IA based system is its high common-mode rejection which reduces the effects of noise between grounding systems. Its disadvantage is that a minimum of two packaged devices is required and choosing precision, ratio matched resistors (or a network) may be necessary.

PGA USING OP AMP AND CMOS SWITCHES



In systems where ground management does not require high CMR the IA may be replaced by a less expensive operational amplifier. An op amp configured for noninverting gain has desirable high input impedance and can be made software programmable with a CMOS DAC in the feedback loop, or by using precision resistors and CMOS switches to select the gain setting resistance. An advantage to this approach (other than cost) is the op amp may be chosen for a particular attribute (FET input for high input impedance, wideband for fast response). Its drawback, like that of the IA based system, is the number of components required and the use of precision resistors or networks.

PGA USING OP AMP AND CMOS DAC



$V_{OUT} = \frac{-V_{IN}}{D}$

$A_V = \frac{-V_{OUT}}{V_{IN}} = -\frac{1}{D}$ WHERE: A_V = VOLTAGE GAIN

and where:

$D = \frac{DB7}{2^1} + \frac{DB6}{2^2} + \ldots \frac{DB0}{2^8}$

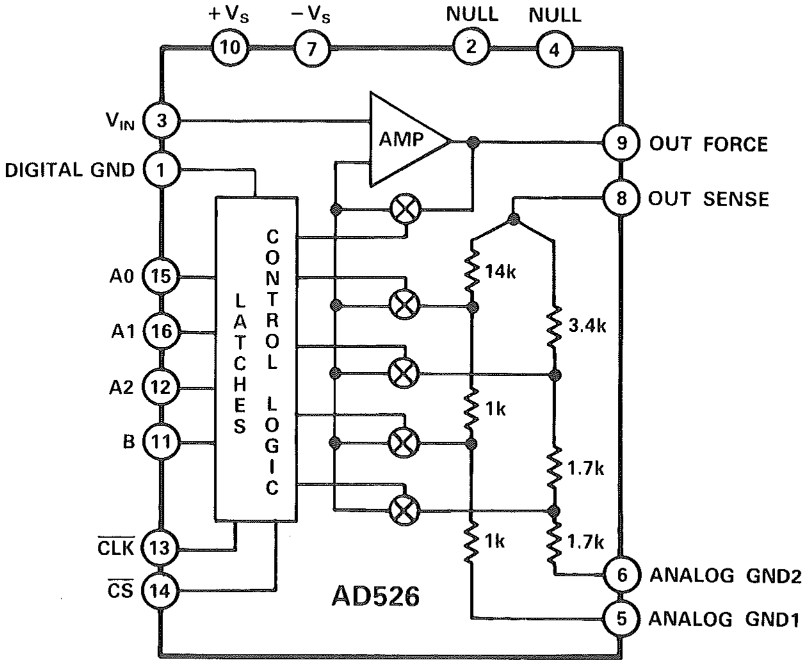
DB_N = 1 or 0

The AD526 is a complete single-ended SPGA in monolithic form. Its design includes a FET input op amp, precision laser-trimmed gain setting resistor ladders, JFET analog switches and microprocessor compatible logic. Its dc accuracies are consistent with the levels required for 12 bit systems, while its ac performance exceeds that of many IA based SPGAs.

AD526 FEATURES

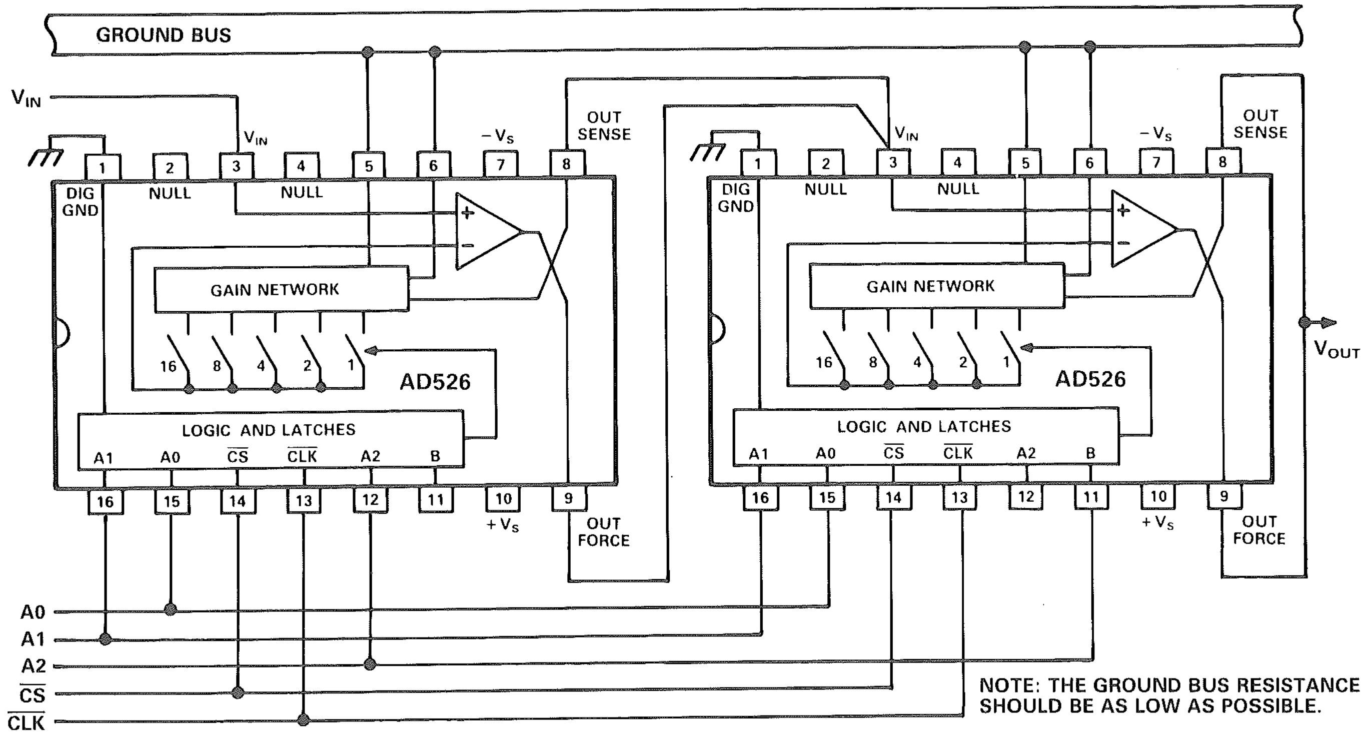
- **Digitally Programmable Binary Gains from 1 to 16**
- **Two-Chip Cascade Mode Achieves Binary Gain from 1 to 128 with No Additional Components**
- **Gain Error:**
  - **0.01% max, Gain = 1, 2, 4 (C Grade)**
  - **0.02% max, Gain = 8, 16 (C Grade)**
- **Fast Settling Time**
  - **Without Gain Change:**
    - **0.01% in 4.5µs (Gain = 16)**
  - **With Gain Change:**
    - **0.01% in 5.6µs (Gain = 16)**
- **Low Nonlinearity: ±0.035% FSR max (A Grade)**
- **Excellent dc Accuracy:**
  - **Offset Voltage: 0.7mV max (A Grade)**
  - **Max Offset Voltage Over Temperature**
    - **0.4mV (C Grade)**
- **TTL Compatible Digital Inputs**

AD526 BLOCK DIAGRAM



Binary gain settings (1, 2, 4, 8 and 16) make the AD526 ideal for use as dynamic range extension for analog-to-digital converter systems. A single AD526 can extend a 12-bit converters dynamic range to over 85dB. By cascading two AD526s, gains of 32, 64 and 128 can also be made available without using additional external components.

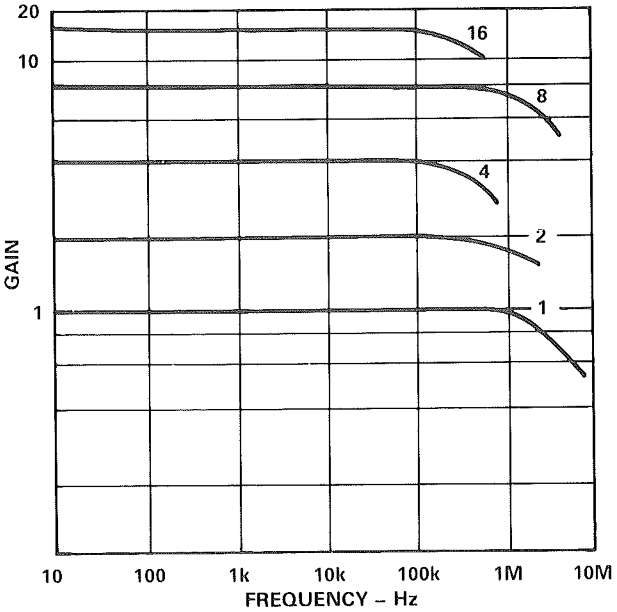
CASCADED OPERATION



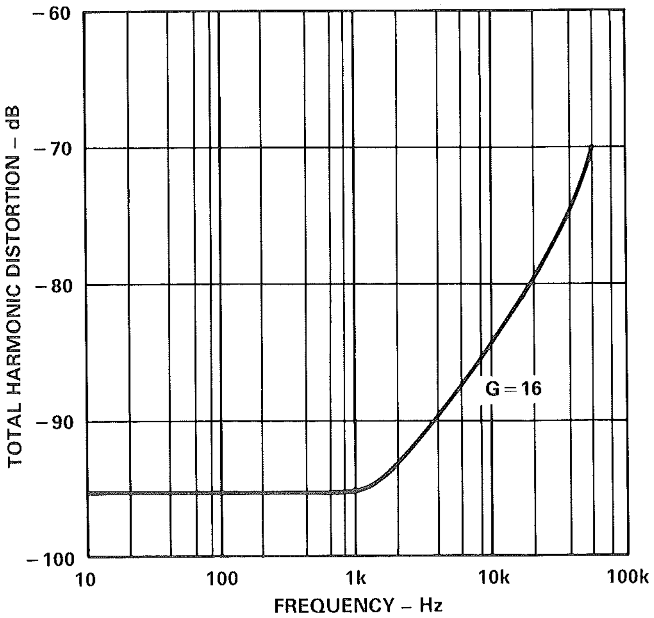
NOTE: THE GROUND BUS RESISTANCE SHOULD BE AS LOW AS POSSIBLE.

Small signal bandwidths of IA based SPGAs are typically below 1MHz at gains of one and fall lower as gain is increased. The AD526 maintains a small signal bandwidth greater than 1MHz for gains of 1, 2 and 4, while for gains of 8 and 16, the frequency compensation is automatically changed such that amplifier bandwidth is increased and SPGA settling time, slew rate, and bandwidth are improved over a constant compensation scheme. Settling time is typically 4µs regardless of gain, as opposed to 15µs for IA approach.

AD526 GAIN VS. FREQUENCY



AD526 TOTAL HARMONIC DISTORTION VS. FREQUENCY

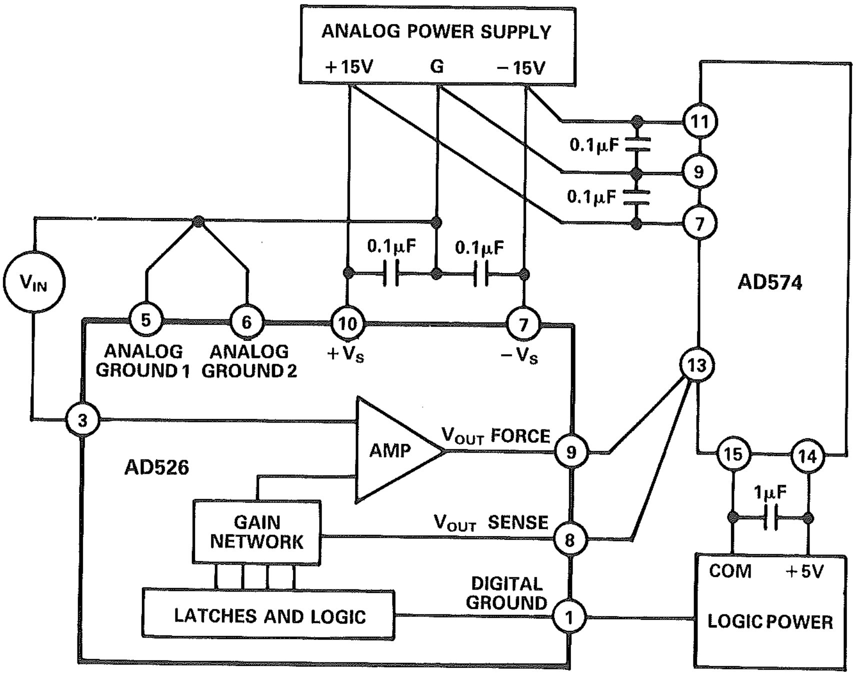


As previously mentioned, the AD526 dc accuracies are consistant with 12-bit systems (0.01% gain error, 0.005% nonlinearity error, and 0.5mV offset error; AD526B), therefore it is ideal for use in systems utilizing ADCs such as the AD574A.

In such a system, attention must be paid to signal and ground paths. Logic and signal grounds should be separate, and the signal source reference point must be connected locally to the AD526 ground pins (5 and 6) in order to maintain the gain accuracy of the device. Currents associated with other elements within the system should not be allowed to corrupt this ground connection.

The force and sense outputs of the AD526 are used in the same manner as those of an instrumentation amplifier. They should be connected together at the load to avoid signal drops when driving long distances and/or moderately low impedances. When driving very low impedances, a high current buffer stage can be added inside the loop.

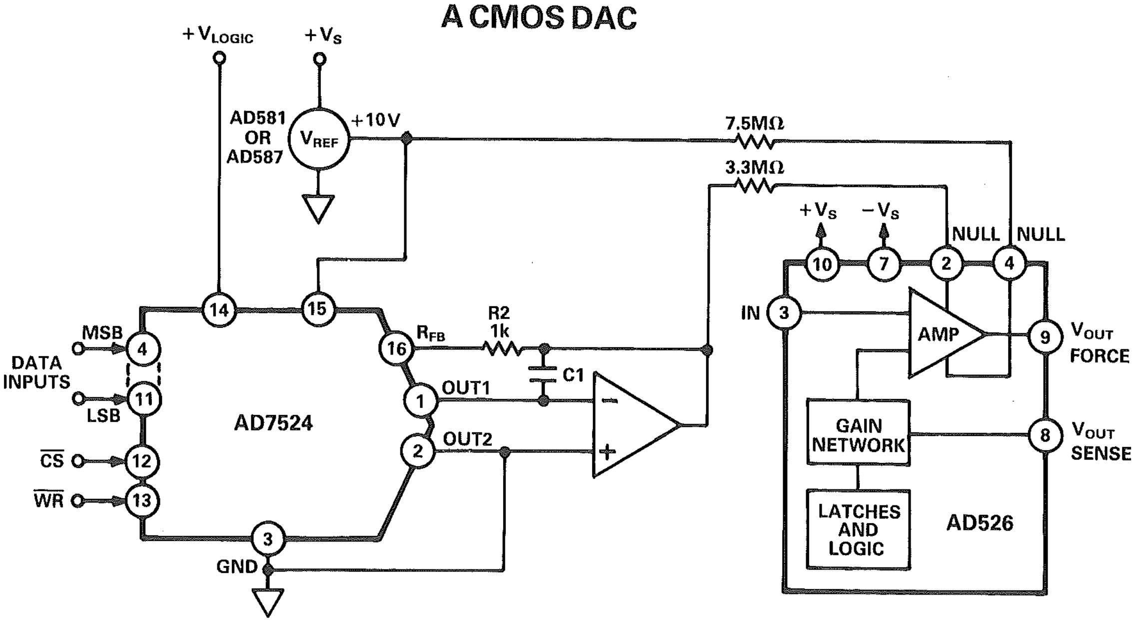
GROUNDING AND BYPASSING



The AD526 SPGA offers low, constant input offset voltage (250μV; AD526C) over all gain settings. To null this offset, a 20kΩ potentiometer should be used between the null pins (2 and 4) with the wiper tide to $-V_S$.

Microprocessor controlled offset nulling can be accomplished with an 8-bit CMOS DAC circuit instead of the trimpot configuration. Its advantage is that it can be implemented as part of an autocalibration scheme, with dipswitches, novram or RAM storing the 8-bit word after its value has been determined. Offset null sensitivity of this circuit is 80μV per LSB at a gain of 16 which guarantees dc accuracy to the 16-bit performance level.

AD526 OFFSET NULLING USING A CMOS DAC



5. ISOLATION AMPLIFIERS

Instrumentation amplifiers are restricted by the requirement that a return path for bias currents must be provided. Furthermore, large common-mode voltages can damage IA input circuitry. When the application involves galvanic or ohmic isolation of input and output circuitry, an Isolation Amplifier is required.

WHEN AN ISOLATION AMPLIFIER IS REQUIRED

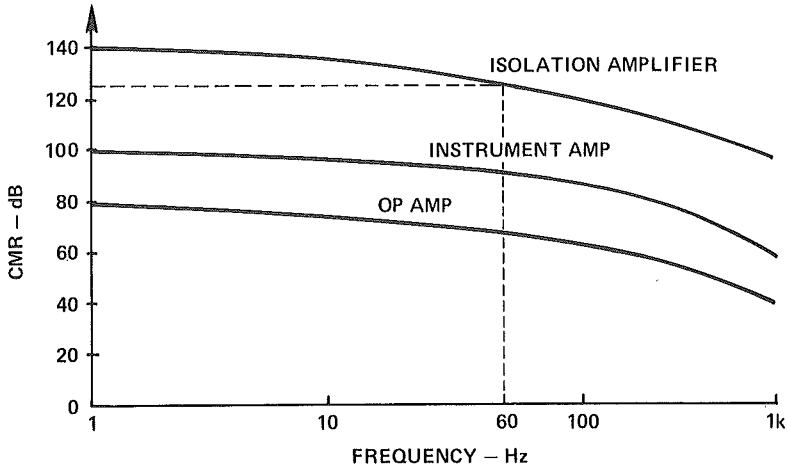
- Measure Low Level Signals in the Presence of High CMV.
- Eliminate Measurement Errors Caused by Disturbances on the Source Ground Network (from High Current Transients, etc.)
- Avoid Ground Loops and Their Attendant Pickup Problems. (No Need to Provide a Return Path for Bias Currents.)
- Protect Processing Circuitry from Damage from Large CMV Levels at Both Input and Output.
- Provide Patient-Safe Interface.

Ohmic and Galvanic Source Isolation

The isolation amplifier's floating input design provides complete decoupling between the source and amplifier output and, in most cases, power terminals. This offers other benefits beyond the high common-mode rejection.

Low capacitance and leakage from input to common gives very high CMR virtually independent of source imbalance.

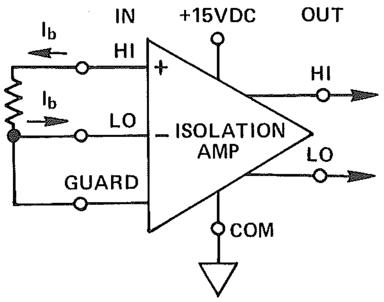
TYPICAL CMR VS FREQ WITH 1kΩ SOURCE IMBALANCE FOR VARIOUS TYPES OF DEVICES



No Bias Current

The front end circuitry of isolation amplifiers is fully floating and, therefore, no net bias current flows in the input leads. As shown below, the bias current of the LO input is supplied by the HI input. Thus, the isolator does not require connections to the source ground in order to establish the input bias current flow and, therefore, is not affected by disturbances on the source ground system.

ISOLATORS REQUIRE NO NET BIAS CURRENT



SELECTING AN ISOLATOR

Isolation amplifiers may be used to advantage in a limitless number of situations, but the vast majority of applications fall into one of three categories: Medical, Industrial (process control) and Instrumentation (data acquisition).

Medical

Medical amplifiers must first and foremost protect the patient from leakage currents and amplifier fault currents in excess of 10 microamps rms. It is just as important to be sure that the amplifier will not be damaged by 5kV defibrillator pulses; if an amplifier monitoring a patient's heartbeat should fail during defibrillation, the medical team may continue to defibrillate the patient in the belief that the heart has not restarted (while, in fact, the amplifier has failed). Continued defibrillation can kill the patient.

Industrial

Industrial amplifiers must provide accurate signal gain while rejecting common-mode noise and eliminating ground loops. Industrial malfunctions may cause power-line voltages to be imposed on low voltage signal lines. The industrial isolator should not be destroyed by such mistreatment, but more important, it must protect the expensive computer on the other end of the line from errant high voltage surges.

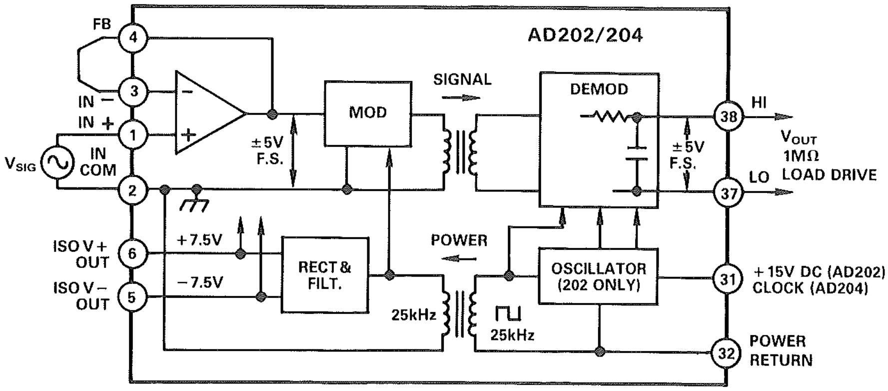
Instrumentation

Instrumentation (data acquisition) applications may not involve the extreme hazards of medical or industrial applications, but the precision required may demand the feature of an isolator. Twelve-bit systems require accuracies of ±0.01% and are therefore quite susceptible to ground loops or common-mode interference. Isolation amplifiers can eliminate ground loops and offer better common-mode rejection than conventional data amplifiers.

AD202/204 FEATURES

Low Cost
Small Size: 4 Channels/Inch
Low Power: 35mW (AD204)
High Accuracy: ±0.025% max Nonlinearity (K Grade)
High CMR: 130dB (Gain = 100 V/V)
Wide Bandwidth: 5kHz Full-Power (AD204)
High CMV Isolation: ±2000 V pk Continuous (K Grade) (Signal and Power)
Isolated Power Outputs
Uncommitted Input Amplifier

AD202/204 BLOCK DIAGRAM



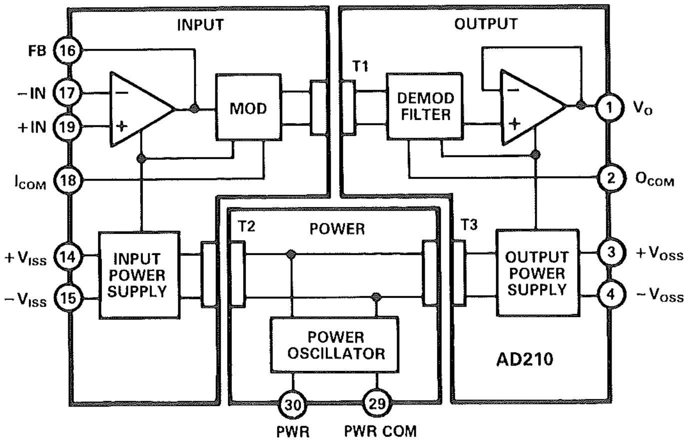
The AD202 and AD204 low cost, miniature isolation amplifiers use an amplitude modulation technique to permit transformer coupling of signals down to dc. Both models also contain an uncommitted input op amp and a power transformer which provides isolated power to the op amp, the modulator and any external load. The power transformer primary is driven by a 25kHz, 15V p-p square wave which is generated internally in the case of the AD202, or supplied externally for the AD204.

Within the signal swing limits of approximately ±5V, the output voltage of the isolator is equal to the output voltage of the op amp; that is, the isolation barrier has unity gain. The output signal is not internally buffered, so the user is free to interchange the output leads to get signal inversion. Additionally, in multichannel applications, the unbuffered outputs can be multiplexed with one buffer following the mux. This technique minimizes offset errors while reducing power consumption and cost. The output resistance of the isolator is typically 3kΩ for the AD204 (7kΩ for the AD202) and varies with signal level and temperature, so it should not be loaded. In many cases a high-impedance load will be present or a following circuit such as an output filter can serve as a buffer, so that a separate buffer function will not often be needed.

AD210 FEATURES

High CMV Isolation: 2500V rms Continuous
±3500V peak Continuous
Small Size: 1.00″ × 2.10″ × 0.350″
Three-Port Isolation: Input, Output, and Power
Low Nonlinearity: ±0.012% max
Wide Bandwidth: 20kHz Full-Power (−3dB)
Low Gain Drift: ±25ppm/°C max
High CMR: 120dB (G = 100V/V)
Isolated Power: ±15V @ ±5mA
Uncommitted Input Amplifier

AD210 BLOCK DIAGRAM



Referring to the AD210 block diagram above, a +15V supply is connected to the power port, and ±15V isolated power is supplied to both the input and output ports via a 50kHz carrier frequency. The uncommitted input amplifier can be used to supply gain or buffering of input signals to the AD210. The full wave modulator translates the signal to the carrier frequency for application to transformer T1. The synchronous demodulator in the output port reconstructs the input signal. A 20kHz, three-pole filter is employed to minimize output noise and ripple. Finally, an output buffer provides a low impedance output capable of driving a 2kΩ load.

6. TRANSDUCER INTERFACING

Ideally, a transducer should have a high level output, zero source impedance, low noise and be relatively linear, however they are not. Through precious portions of this section we have characterized instrumentation and isolation amplifiers, in this portion we will outline techniques for measuring physical phenomona. The emphasis will be on solutions to problems.

COMMON TRANSDUCERS SUMMARIZED

TEMPERATURE

TYPE	ELECTRICAL I/O CHARACTERISTICS	COMMENTS
Thermocouples	Low source impedance, typically 10Ω. Voltage-output devices. Output shift is 10's of microvolts/°C. Outputs typically in the millivolts at room temperature.	Low voltage output requires low-drift signal conditioning. Small size and wide temperature range are advantages. Requires reference to a known temperature. Nonlinear response.
Platinum and other RTD's	Resistance changes with temperature. Positive temperature coefficient. Typical impedance (0°C) 20Ω to 2kΩ. Typical sensitivities 0.1%/°C to 0.66%/°C, depending on material.	Highly repeatable. Good linearity over wide ranges. Requires bridge or other network for typical interface.
Thermistors	Resistance changes with temperature. Negative temperature coefficient. Typical impedances (25°C) of 50Ω to 1MΩ available. Sensitivity at 25°C is about 4%/°C. Linearized networks available with with 0.4%/°C sensitivity.	Highest sensitivity among common temperature transducers. Inherently nonlinear (exponential function) but accurate linearized networks available.
Semiconductor sensors	Voltage, current, or resistance functions. Voltage types (diodes) require excitation current. Current types (AD590) require excitation voltage. Resistive types (bulk silicon) may use either type of excitation.	Many devices are uncalibrated and require significant signal conditioning. AD590 is calibrated, linear, and requires minimal signal conditioning.

COMMON TRANSDUCERS SUMMARIZED

FORCE

TYPE	ELECTRICAL I/O CHARACTERISTICS	COMMENTS
Strain gages (metal)	Resistance shifts with applied strain. Almost always used in bridge configuration. Typical impedance levels of 120Ω and 350Ω. Typical change is 0.1% over the whole range.	Resistance change with strain small compared to initial value of device resistance. Requires high-quality low-level signal conditioning.
Strain-gage bridge, load cell	Voltage output with applied strain. Requires excitation potential or current to drive the bridge. Typical excitation is from 5 to 15 volts.	Small voltage outputs require low-drift signal conditioning with good common-mode rejection to achieve any degree degree of precision. Output is linear.
Semiconductor strain gages	Bridge types are assembled from individual gages and have a voltage output. Bridge requires excitation, typically 5V to 15V.	More output than metal strain gages, but with increased nonlinearity and sensitivity to temperature.
Piezoelectrics	True charge output device. Modeled as voltage source in series with capacitor. Physical input change produces corresponding charge change. AC and transient response only. Typical upper frequency limit is 20 to 50kHz. Typical output is 10^{-7} coulombs full-scale.	Requires low-bias-current charge amplifier configurations for signal conditioning. Responds to ac signals only.

PRESSURE

TYPE	ELECTRICAL I/O CHARACTERISTICS	COMMENTS
Rheostat/potentiometer	Resistance or ratio-of-resistance output. Requires voltage or current excitation. Typical impedance 500Ω to 5kΩ.	High-level easy-to-condition outputs are typical due to significant resistance or ratio
Strain gage	Resistance shift (single gage) or voltage output (strain-gage bridge). Requires excitation potential or current.	Small resistance change. Low-level signal requires good signal-conditioning amplifiers.

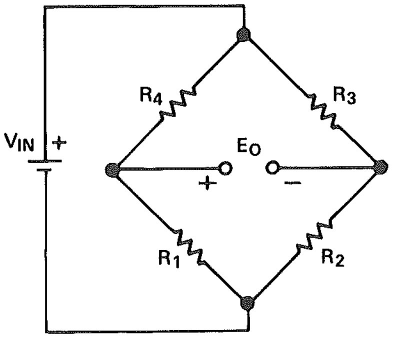
FLOW

TYPE	ELECTRICAL I/O CHARACTERISTICS	COMMENTS
Pressure-based	See PRESSURE transducers	Pressure types measure flow by measuring ΔP between static and flow-caused pressure, or pressure drop across a constriction. Differential pressure transducers are used to avoid common-mode pressure errors. Response is nonlinear.
Frequency-output types: paddle wheels, rotary types, vortex types	Digital output derived from frequency output are common. Optical or magnetic pickups provide non-invasive measurements. Photocell has 100Ω to 100MΩ on-to-off ratio. Magnetic employs switching or open-collector transistor.	Some types are directly logic-level compatible. Others require impedance and/or voltage amplification, level-shift, and buffering before signal is usable.

Bridge Circuits

The figure below shows the common Wheatstone bridge (actually developed by S.H. Christie in 1833). In its simplest form, a bridge consists of four two-terminal elements connected to form a quadrilateral, a source of excitation (voltage or current) connected along one of the diagonals, and a detector of voltage or current comprising the other diagonal. The detector measures the difference between the outputs of two potentiometric dividers connected across the excitation supply.

BASIC BRIDGE CIRCUIT – VOLTAGE EXCITATION AND VOLTAGE READOUT



$$E_O = \frac{R_1}{R_1+R_4} V_{IN} - \frac{R_2}{R_2+R_3} V_{IN}$$

$$= \frac{\frac{R_1}{R_4} - \frac{R_2}{R_3}}{\left(1+\frac{R_1}{R_4}\right)\left(1+\frac{R_2}{R_3}\right)} V_{IN}$$

AT BALANCE, $E_O = 0$ IF $\frac{R_1}{R_4} = \frac{R_2}{R_3}$

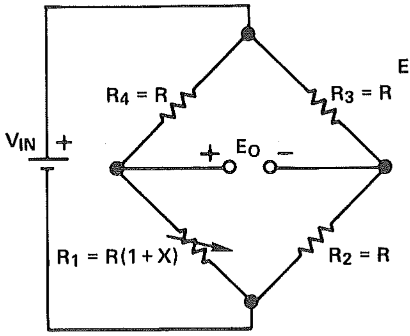
A bridge measures an electrical property of a circuit element indirectly, i.e., by comparison against a similar element. The two principle ways of operating a bridge are as a null detector and as a device that reads a difference directly in voltage or current.

When R1/R4 = R2/R3, the resistance bridge shown is at a *null*, irrespective of the mode of excitation (current or voltage, ac or dc), the magnitude of excitation, the mode of readout (current or voltage), or the impedance of the detector. Therefore if the ratio R2/R3 is fixed at *K*, a null is achieved when R1 = *K* R4. If R1 is unknown and R4 is an accurately determined variable resistance, the magnitude of R1 can be found by adjusting R4 until null is achieved. Conversely, in transducer-type measurements, R4 may be a fixed reference and a null occurs when the magnitude of the measurand is such that R1 is equal to *K* R4.

Null-type measurements are principally used in feedback systems, involving electromechanical and/or human elements. Such systems, as noted previously, seek to force the active element (strain gage, RTD, thermistor, mechanically coupled potentiometer) to balance the bridge by influencing the parameter being measured. Because the null is independent of the excitation, the null mode may also be used to discriminate between the two polarities of output, i.e., as a comparator. In such applications, the polarity of the off-null signal may be of greater significance than its magnitude (for example, if the level of a tank is below a preset value, a valve is caused to open to fill the tank).

For the majority of transducer applications employing bridges, the deviation of one or more resistors in a bridge from an initial value must be measured as an indication of the magnitude (or a change) of the measurand. The figure below shows a bridge with all resistances nominally equal; but one of them (R1) is variable by a factor, (1 + X), where X is a fractional deviation around zero, as a function of (say) strain. As the equation indicates, the relationship between the bridge output and X is not linear, but for small ranges of X it is sufficiently linear for many purposes. For example, if $V_{IN} = 10V$, and the maximum value of X is ±0.002, the output of the bridge will be linear to within 0.1% for a range of outputs from 0 to ±5mV, and to 1% for the range 0 to ±50mV (±0.02 range for X).

BRIDGE USED TO READ DEVIATION OF A SINGLE VARIABLE ELEMENT



$$E_O = \frac{R(1+X)}{R+R(1+X)} V_{IN} - \frac{1}{2} V_{IN}$$

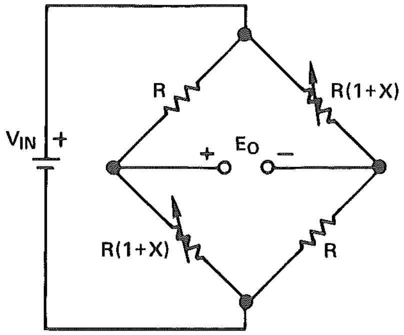
$$= \frac{2+2X-2-X}{2(2+X)} V_{IN}$$

$$= \frac{V_{IN}}{4} \cdot \frac{X}{1+\frac{X}{2}}$$

$$\cong \frac{V_{IN}}{4} X \text{ FOR } X<<1$$

The *sensitivity* of a bridge is the ratio of the maximum expected change in the output value to the excitation voltage. For the examples given in the previous paragraph, the sensitivities are ±500μV/V and ±5mV/V. The sensitivity can be doubled if two identical variable elements can be used, e.g., at positions R3 and R1, as shown below. An example of such a pair is two identically oriented strain gage resistances aligned in a single pattern. Note that the output of such a pair is doubled, but the nonlinearity remains the same.

BRIDGE WITH TWO VARIABLE ELEMENTS



$$E_O = \left[\frac{R(1+X)}{R+R(1+X)} - \frac{R}{R+R(1+X)}\right] V_{IN}$$

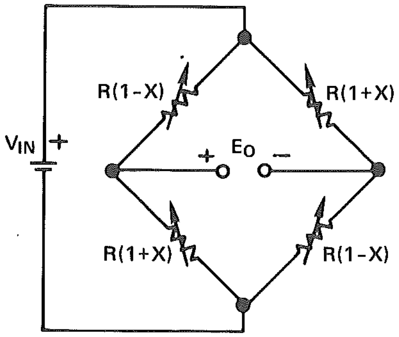
$$= \frac{X}{2+X}\, V_{IN}$$

$$= \frac{V_{IN}}{2}\,\frac{X}{1+\frac{X}{2}}$$

$$\cong \frac{V_{IN}}{2}\, X \text{ FOR } X<<1$$

In special cases, another doubling of the output can be achieved. The figure below shows a bridge consisting of four resistors, two of which increase and two of which decrease in the same ratio. Two identical two-element strain gages, attached to opposite faces of a thin carrier to measure its bending, could be electrically configured in this way. The output of such a bridge would be four times the output for a single-variable-element bridge; furthermore, the complementary nature of the resistance changes would result in a linear output.

ALL ELEMENTS VARIABLE



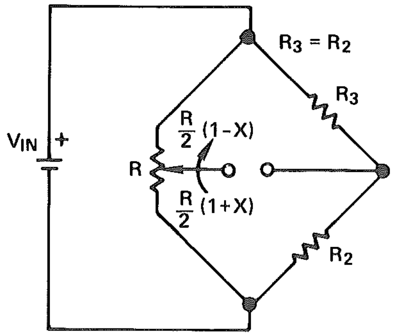
$$E_O = \left[\frac{R(1+X)}{2R} - \frac{R(1-X)}{2R}\right] V_{IN}$$

$$= \left[\frac{1+X-1+X}{2}\right] V_{IN}$$

$$= X\, V_{IN}$$

The next figure shows bridge employing a zero-centered potentiometer to constitute two adjacent arms; the position of the potentiometer is a measure of the physical phenomenon. Since it is a two-variable-element version, its output is twice that of the single-variable-element bridge, and being complimentary in nature, it is linear.

LINEAR POTENTIOMETER AS VARIABLE ARM



$$E_O = \left[\frac{\frac{R}{2}(1-X)}{R} - \frac{1}{2}\right] V_{IN}$$

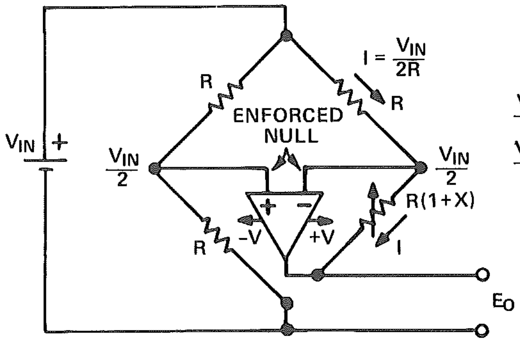
$$= \frac{R-XR-R}{2R}\, V_{IN}$$

$$= \frac{V_{IN}}{2}\, X$$

A distinction should be recognized between the linearity of the bridge equation and the linearity of the transducer response to the phenomenon being sensed. For example, if the active element is a potentiometer, a bridge used to implement the measurement would be adequately linear; yet the output could still be nonlinear due to the pot's nonlinearity.

Manufacturers of transducers employing bridges address the nonlinearity issue in a variety of ways, including keeping the resistive swings in the bridge small, shaping complementary nonlinear response into the active elements of the bridge, using resistive trims for first-order corrections, and a variety of proprietary magical techniques. A bridge can, of course, be linearized by making it less sensitive (e.g., by making the initial ratios, R4/R1 and R3/R2, large), but the tradeoff of sensitivity for linearity is painful.

ACTIVE BRIDGE



$$\frac{V_{IN}}{2} = E_O + IR(1+X)$$

$$\frac{V_{IN}}{2} = E_O + \frac{V_{IN}}{2}(1+X)$$

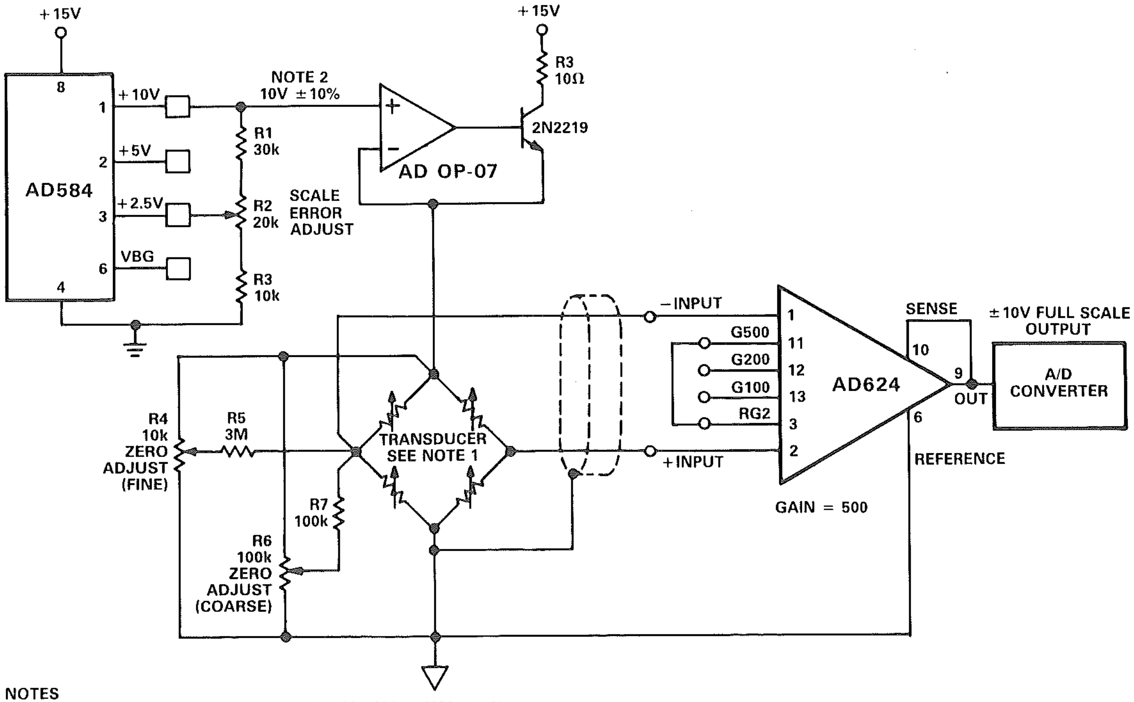
$$E_O = -\frac{V_{IN}}{2} X$$

The figure above shows an active bridge in which an op amp produces a null by adding a voltage in series with the variable arm. That voltage is equal in magnitude and opposite in polarity to the incremental voltage across Rx, and it is inherently linear with X. Since it is an op amp output, it can be used as a low-impedance output point for the bridge measurement. This active bridge has a gain of two over the standard one-active-element bridge, and the output is linear, even for large values of X.

Weigh Scale

The circuit below shows an example of how an AD624 can be used to condition the differential output voltage of a load cell. The 10% reference voltage adjustment range is required to accommodate the 10% transducer sensitivity tolerance. The high linearity and low noise of the AD624 make it ideal for use in applications of this type. The addition of an auto gain/auto tare cycle would enable the system to remove offsets, gain errors, and drifts making possible true 14-bit performance.

AD624 WEIGH SCALE APPLICATION

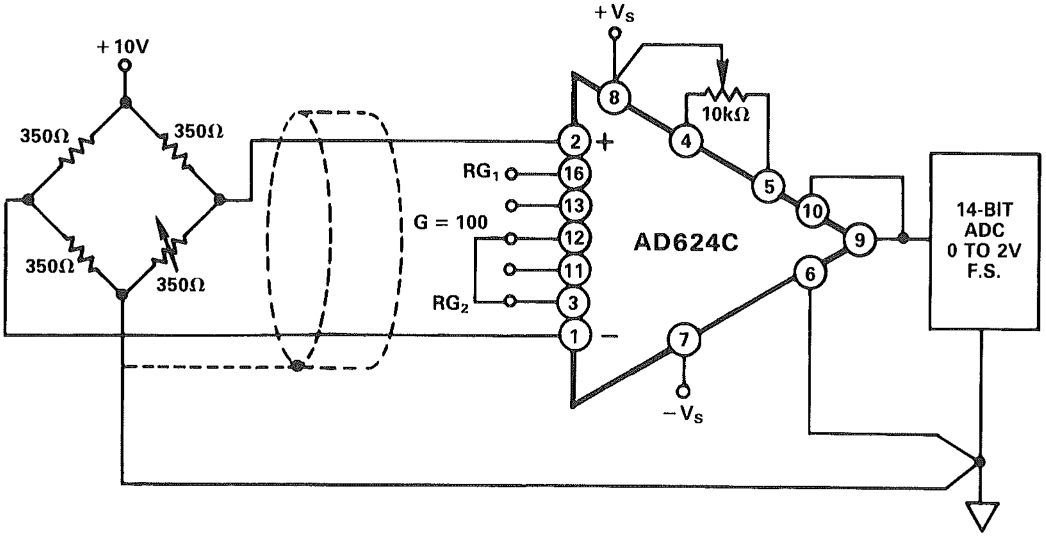


NOTES
1. LOAD CELL TEDEA MODEL 1010 10KG. OUTPUT 2mV/V ±10%.
2. R1, R2 and R3 SELECTED FOR AD584. OUTPUT 10V ±10%.

Error Budget Analysis

To illustrate how instrumentation amplifier specifications are applied, we will now examine a typical case where an AD624 is required to amplify the output of an unblanced transducer. The figure shows a differential transducer, unbalanced by ≈5Ω, supplying a 0 to 20mV signal to an AD624C. The output of the IA feeds a 14-bit A to D converter with a 0 to 2 volt input voltage range. The operating temperature range is −25°C to +85°C. Therefore the largest change in temperature, ΔT, within the operating range is from ambient to +85°C (85°C − 25°C = 60°C).

TYPICAL BRIDGE APPLICATION



In many applications, differential linearity and resolution are of prime importance. This would be so in cases where the absolute value of a variable is less important than changes in value. In these applications, only the irreducible errors (20ppm = 0.002%) are significant. Furthermore, if a system has an intelligent processor monitoring the A to D output, the addition of an auto-gain/auto-zero cycle will remove all reducible errors and may eliminate the requirement for initial calibration. This will also reduce errors to 0.002%.

ERROR BUDGET ANALYSIS OF AD624CD IN BRIDGE APPLICATION

Error Source	AD624C Specifications	Calculation	Effect on Absolute Accuracy at $T_A = 25°C$	Effect on Absolute Accuracy at $T_A = 85°C$	Effect on Resolution
Gain Error	±0.1%	±0.1% = 1000ppm	1000ppm	1000ppm	–
Gain Instability	10ppm/°C	(10ppm/°C)(60°C) = 600ppm	–	600ppm	–
Gain Nonlinearity	±0.001%	±0.001% = 10ppm	–	–	10ppm
Input Offset Voltage	±25µV, RTI	±25µV/20mV = ±1250ppm	1250ppm	1250ppm	–
Input Offset Voltage Drift	±0.25µV/°C	(±0.25µV/°C)(60°C) = 15µV 15µV/20mV = 750ppm	–	750ppm	–
Output Offset Voltage[1]	±2.0mV	±2.0mV/20mV = 1000ppm	1000ppm	1000ppm	–
Output Offset Voltage Drift[1]	±10µV/°C	(±10µV/°C)(60°C) = 600µV 600µV/20mV = 300ppm	–	300ppm	–
Bias Current – Source Imbalance Error	±15nA	(±15nA)(5Ω) = 0.075µV 0.075µV/20mV = 3.75ppm	3.75ppm	3.75ppm	–
Offset Current – Source Imbalance Error	±10nA	(±10nA)(5Ω) = 0.050µV 0.050µV/20mV = 2.5ppm	2.5ppm	2.5ppm	–
Offset Current – Source Resistance – Error	±10nA	(10nA)(175Ω) = 3.5µV 3.5µV/20mV = 87.5ppm	87.5ppm	87.5ppm	–
Offset Current – Source Resistance – Drift	±100pA/°C	(100pA/°C)(175Ω)(60°C) = 1µV 1µV/20mV = 50ppm	–	50ppm	–
Common Mode Rejection 5V dc	115dB	115dB = 1.8ppm × 5V = 9µV 9µV/20mV = 444ppm	450ppm	450ppm	–
Noise, RTI (0.1–10Hz)	0.22µV p-p	0.22µV p-p/20mV = 10ppm	–	–	10ppm
		Total Error	3793.75ppm	5493.75ppm	20ppm

[1] Output offset voltage and output offset voltage drift are given as RTI figures.

The preceding table lists all applicable error sources and their corresponding effects on accuracy. Initial errors are defined as those errors that can be reduced to a negligible amount by performing an initial calibration.

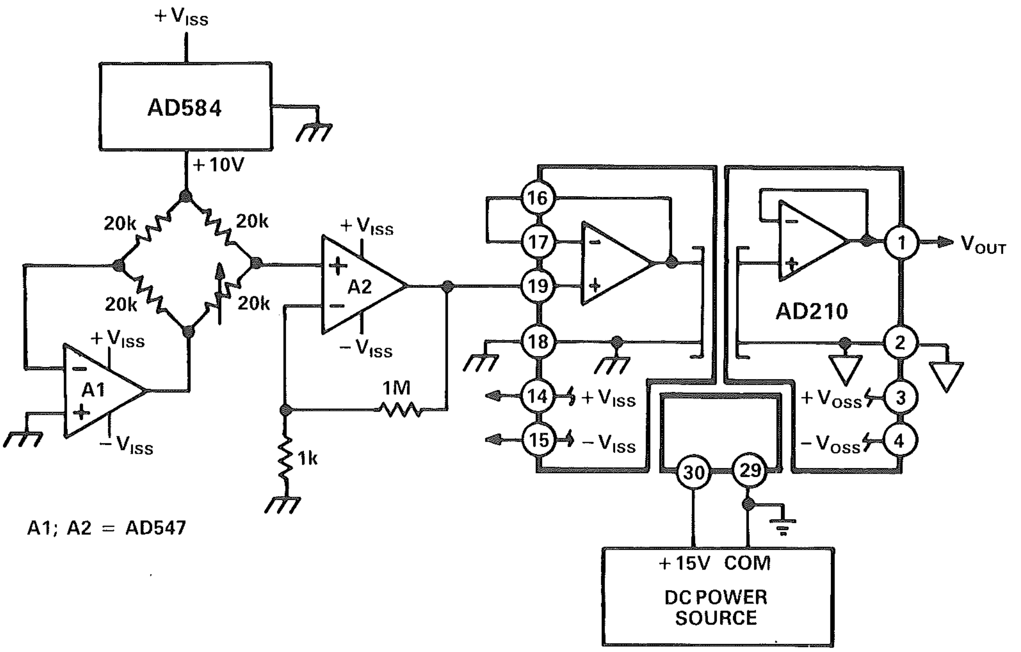
Reducible errors include these initial errors along with other errors that occur during normal operation that may be corrected by an adaptive or "intelligent" system. For example, changes in gain or offset may be measured during an auto-zero/auto-gain cycle by measuring two unknown voltages (a precision reference and ground, for example).

Irreducible errors are errors which can not be readily corrected either at initial calibration or in use.

Isolated Industrial Applications

The circuit below illustrates one possible configuration for isolated conditioning of a bridge circuit. The AD584 produces a +10V excitation voltage, while A1 inverts the voltage, producing negative excitation. A2 provides a gain of 1000V/V to amplify the low level bridge signal. Additional gain can be obtained by reconfiguring of the AD210's input amplifier. $\pm V_{ISS}$ provides the complete power for this circuit, eliminating the need for a separate isolated excitation source.

ISOLATED BRIDGE CIRCUIT



A1; A2 = AD547

Bridge Linearization

If one arm of a Wheatstone bridge varies from its nominal value by a factor, (1+2w), the voltage or current output of the bridge will be (with appropriate polarities and scale factors):

$$y = w/(1+w)$$

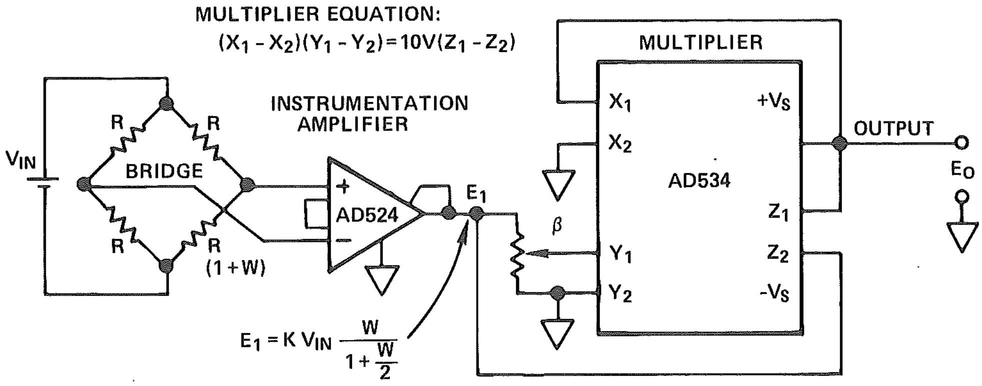
Linear response requires very small values of w (to make the denominator essentially independent of w) and, as a consequence, preamplification.

The circuit shown enables large-deviation bridges to be used without losing linearity or resorting to high attenuation. The circuit computes the inverse of the bridge function, i.e.,

$$w = y/(1-y)$$

Depending on which arm of the bridge varies, it may be necessary to reverse the polarity of the X connections. Any resistive, linearly responding transducer (one or more legs of the bridge proportional to the phenomenon being measured) may profit from the application of this circuit. Examples include position servos, linear thermistors, platinum-resistance-wire sensors, pressure transducers and strain gages.

BRIDGE LINEARIZATION USING ANALOG MULTIPLIER



MULTIPLIER EQUATION:

$(X_1 - X_2)(Y_1 - Y_2) = 10V(Z_1 - Z_2)$

$E_1 = K V_{IN} \frac{W}{1+\frac{W}{2}}$

$$(E_O)\left(K\beta V_{IN}\frac{W}{1+\frac{W}{2}}\right) = 10V\left(E_O - \frac{K V_{IN} W}{1+\frac{W}{2}}\right)$$

SOLVING FOR E_O,

$$E_O = \frac{K V_{IN} W}{1+\frac{W}{2} - \frac{K\beta V_{IN} W}{10V}}$$

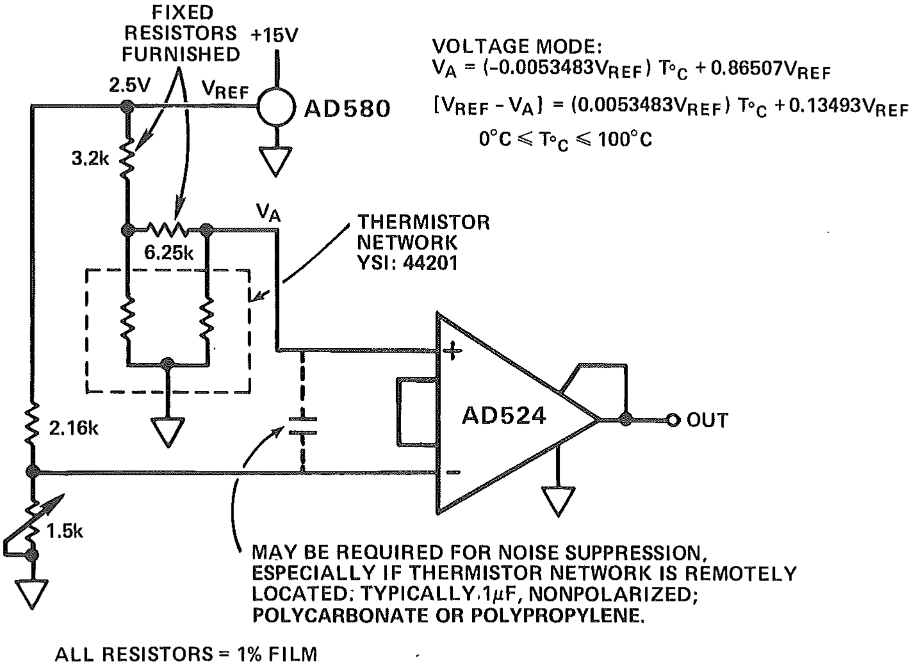
IF $K\beta V_{IN} = 5V$,

$E_O = K V_{IN} W$

Thermistor Interface

In the figure, a thermistor is used in the potentiometric mode. Both the sensor and the offset network are supplied by a 2.5V reference. The differential of the voltages is read out by an instrumentation amplifier, which may be connected for the desired gain and output configuration.

INSTRUMENTING LINEARIZED THERMISTORS VOLTAGE MODE



VOLTAGE MODE:

$V_A = (-0.0053483V_{REF})\,T°_C + 0.86507V_{REF}$

$[V_{REF} - V_A] = (0.0053483V_{REF})\,T°_C + 0.13493V_{REF}$

$0°C \leq T°_C \leq 100°C$

MAY BE REQUIRED FOR NOISE SUPPRESSION, ESPECIALLY IF THERMISTOR NETWORK IS REMOTELY LOCATED; TYPICALLY 1μF, NONPOLARIZED; POLYCARBONATE OR POLYPROPYLENE.

ALL RESISTORS = 1% FILM

Thermocouple Applications

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

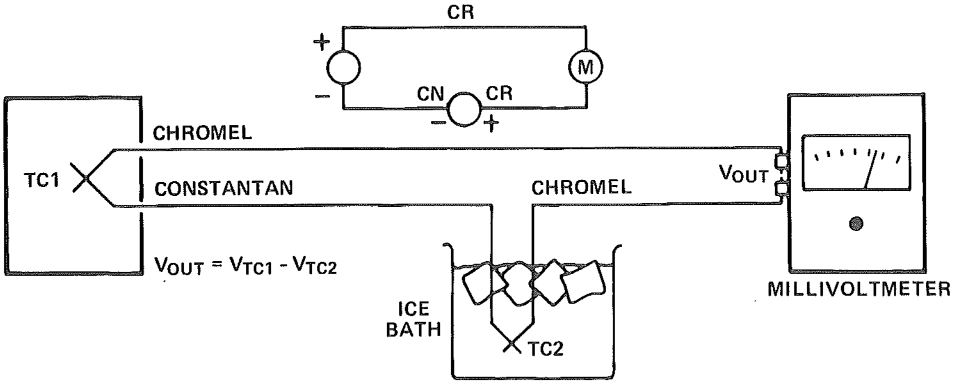
Because the number of free electrons in a piece of metal depends on both temperature and composition of the metal, two pieces of dissimilar metal in isothermal contact will exhibit a potential difference that is a repeatable function of temperature. In general, these voltages are small. The following table lists a number of standard thermocouples, their useful temperature range, and the voltage swing over that range; it can be seen that the average change of voltage with temperature ranges from 7 to about 75μV/°C.

SOME COMMON THERMOCOUPLES

Junction Materials	Typical Useful Temp Range (°C)	Voltage Swing Over Range (mV)	ANSI Designation
Platinum-6% Rhodium – Platinum-30% Rhodium	38 to 1800	13.6	B
Tungsten-5% Rhenium – Tungsten-26% Rhenium	0 to 2300	37.0	(C)
Chromel – Constantan	0 to 982	75.0	E
Iron – Constantan	-184 to 760	50.0	J
Chromel – Alumel	-184 to 1260	56.0	K
Platinum – Platinum-13% Rhodium	0 to 1593	18.7	R
Platinum – Platinum-10% Rhodium	0 to 1538	16.0	S
Copper – Constantan	-184 to 400	26.0	T

Since every pair of dissimilar metals in contact constitutes a thermocouple (including copper/solder, about 3μV/°C and Kovar/rhodium), and since a useful electrical circuit requires at least two contacts in series, measurements with thermocouples must be implemented in a manner which minimizes undesired contributions of incidental thermocouples and provides a suitable reference.

SIMPLE TEMPERATURE MEASURING CIRCUIT USING AN ICE BATH AT THE REFERENCE JUNCTION. THERMOCOUPLE MEASUREMENTS ARE INHERENTLY DIFFERENTIAL.



Because thermocouples are low-level devices, signal conditioning is not a trivial matter. The millivolt-level signals call for low-drift relatively expensive electronics if resolutions better than 1°C are required. Linearity in many types is poor, but the relationships are predictable and repeatable, so either analog or digital techniques can be used for linearizing downstream.

Providing a suitable temperature reference and minimizing the effects of unwanted thermocouples may prove challenging. Techniques include physical references (ice-point cells at +0.01°C, which are accurate and easy to construct but unwieldy to maintain); ambient-temperature reference junctions (acceptable so long as the ambient temperature range in the vicinity of the reference junction is smaller than the desired resolution of the temperature being measured); and electronic cold-junction compensators, which provide an artificial reference level and compensate for ambient temperature variations in the vicinity of the reference junction (this technique requires careful attention to both the electronics and the physical configuration at that location).

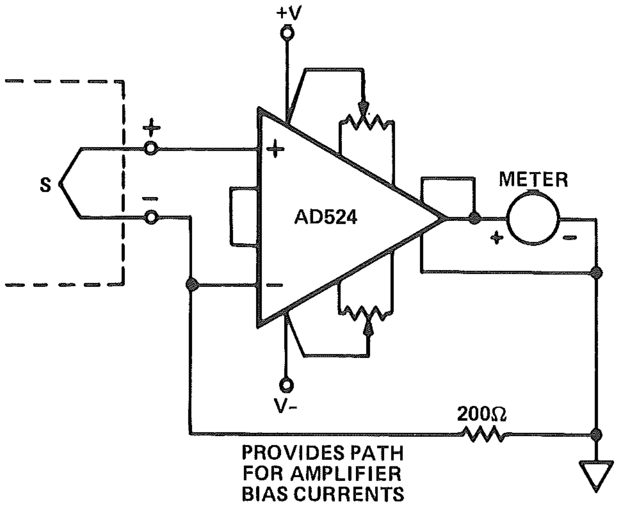
Ambient-Referenced Thermocouples

As we have noted, thermocouples require cold-junction compensation if they must resolve temperature changes with precision better than the ambient temperature range at the cold junction. However, for high-temperature measurements to within a few percent, the cold junction may often be profitably left at room ambient.

Suppose, for example, that a Type S thermocouple is used to measure temperatures of the order of 1500°C within a furnace, and the ambient temperature of the cold junction is 25°C ± 15°C, Since the sensitivity of the thermocouple is 12μV/°C at 1500°C, and a change from 10°C to 40°C at the cold junction produces a change of 180μV in the net output voltage, the equivalent ΔT at the active junction is 15°C for a full-scale change at the cold junction or 1% of 1500°C.

In the figure, an instrumentation amplifier is used to reject common-mode noise. If there is not conductive return path from the thermocouple, resistance may be used (as shown) to provide a path for the amplifiers bias currents.

THERMOCOUPLE PREAMPLIFIER USING AD524



Cold-Junction Compensation

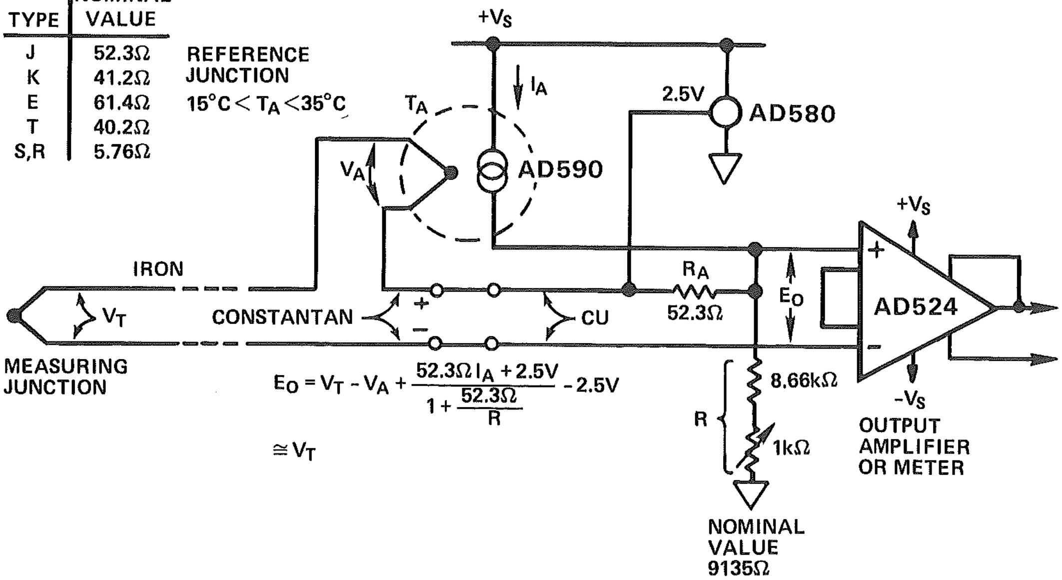
If ambient temperature variation of the cold junction can cause significant error in the output of the thermocouple pair, there are two alternatives: maintain the cold junction at constant temperature, by some such technique as an ice bath or a thermostatically controlled oven, or subtract a voltage that is equal to the voltage developed across the cold junction at any temperature in the expected ambient range.

The figure shows a simple application, in which the variation of the cold-junction voltage of a Type J thermocouple-iron(+)-constantan-is compensated for by a voltage developed in series by the temperature sensitive output current of an AD590 semiconductor temperature sensor.

The circuit is calibrated by adjusting R_T for proper output voltage with the measuring junction at a known reference temperature and the circuit near 25°C. If resistors with low tempcos are used, compensation accuracy will be to within ±0.5°C, for temperatures between +15°C and −35°C. Other thermocouple types may be accommodated with the standard resistance values shown in the table. For other ranges of ambient temperature, the equation in the figure may be solved for the optimum values of R_T and R_A. If an instrumentation amplifier is used, gain and offset specifications should be appropriate for the temperature being measured, the required precision, and the sensitivity of the thermocouple employed.

COLD-JUNCTION COMPENSATION

TYPE	R_A NOMINAL VALUE
J	52.3Ω
K	41.2Ω
E	61.4Ω
T	40.2Ω
S,R	5.76Ω



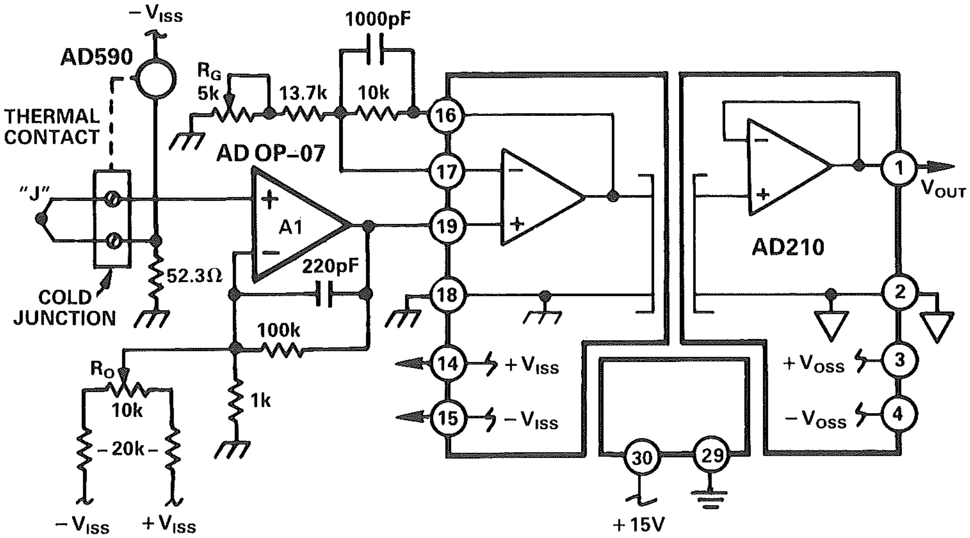
$$E_O = V_T - V_A + \frac{52.3\Omega\, I_A + 2.5V}{1 + \frac{52.3\Omega}{R}} - 2.5V$$

$$\cong V_T$$

Isolated Temperature Measurement and Cold-Junction Compensation

The circuit shown below provides amplification, isolation and cold junction compensation for a standard J type thermocouple. The AD590 temperature sensor accurately monitors the input terminal (cold junction). Ambient temperature changes from 0 to +40°C sensed by the AD590, are cancelled out at the cold junction. Total circuit gain equals 183; 100 and 1.83, from A1 and the AD210 respectively. Calibration is performed by replacing the thermocouple junction with plain thermocouple wire and a millivolt source set at 0.0000V (0°C) and adjusting R_O for V_{OUT} equal to 0.000V. Set the millivolt source to +0.02185V (400°C) and adjust RG for V_{OUT} equal to +4.000V. This application circuit will produce a nonlinearized output of about +10mV/°C for a 0 to +400°C range.

ISOLATED THERMOCOUPLE AMPLIFIER WITH COLD JUNCTION COMPENSATION



7. SPECIALIZED SIGNAL CONDITIONING CIRCUITS

A large portion of this section has been dedicated to instrumentation and isolation amplifiers and as the title suggests, they are special purpose amplifiers. However, we have found that although they are a specialized class of amplifiers, instrumentation and isolation amplifiers can be quite versatile. They can be configured to interface with transducers such as strain gages, load cells, thermocouples, RTDs and thermistors to name a few. The drawback to this is that additional external circuitry is quite often required (i.e., cold junction compensation for thermocouple, or excitation voltage/current for bridge circuits).

This segment will cover a number of highly specialized signal conditioning circuits which will be complete solutions in specific transducer interfacing applications.

MONOLITHIC THERMOCOUPLE SIGNAL CONDITIONER

AD594/AD595 FEATURES

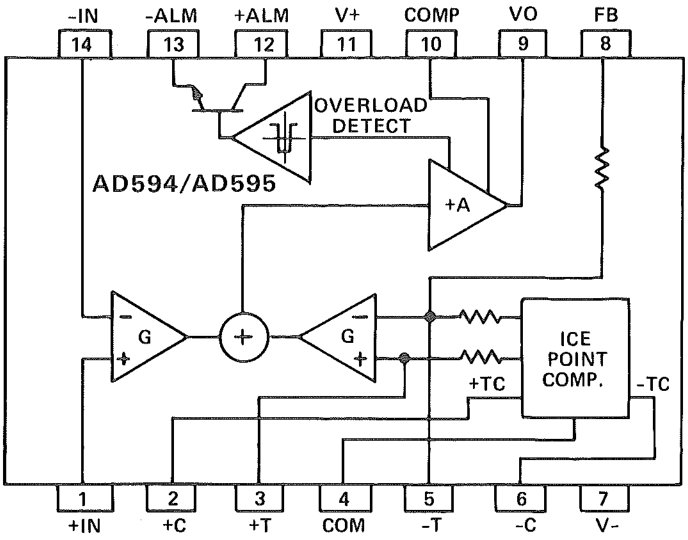
- **Pretrimmed for Type J (AD594) or Type K (AD595) Thermocouples**
- **Low Impedance Voltage Output 10mV/°C**
- **Internal Ice Point Compensation**
- **Wide Power Supply Range: +5V to ±15V**
- **Low Power: 1mW**
- **Thermocouple Failure Alarm**
- **Laser Wafer Trimmed to 1°C Calibration Accuracy**
- **Set Point Mode Operation**
- **Self Contained Centigrade Thermometer Operation**
- **High Impedance Differential Input**

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

Functional Description

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

AD594/AD595 BLOCK DIAGRAM



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

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

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

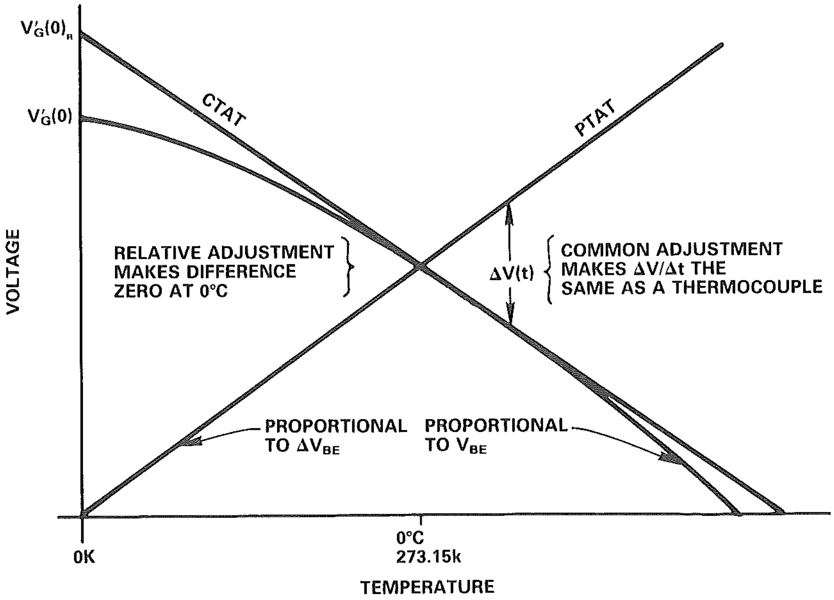
In order to operate properly the AD594/AD595 must have the thermocouple input within both the normal mode signal and common-mode operating range. A normally connected thermocouple within the common-mode operating range will meet these requirements. If one or both thermocouple input terminals are opened, however, an amplifier overload will result. The AD594/AD595 includes an input overload detector which switches on an alarm "transistor". This "transistor" is actually a current limited output buffer, but can be used, up to the limit as a switch transistor for either pull-up or pull-down operation of external alarms.

Cold Junction Reference Compensation

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

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

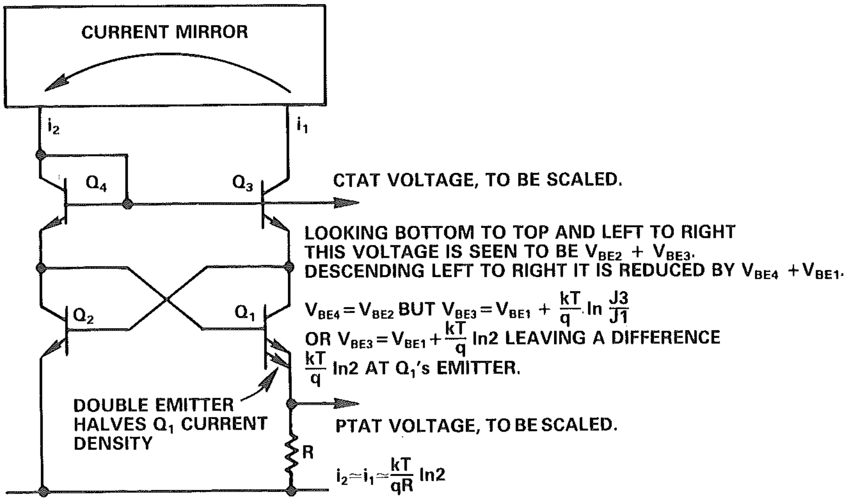
ICE POINT COMPENSATION FROM THE DIFFERENCE OF A PTAT AND A CTAT VOLTAGE



Two temperature sensitive voltages can be derived from the transistor base-emitter characteristics and can be scaled so that their difference approximates the output of an ice referenced thermocouple measuring the IC temperature. This difference is zero at zero Celsius and increases more-or-less linearly with temperature. These voltages are produced by four transistors in the AD594/AD595. A current mirror is used to force a pair of series connected transistors (Q2, Q4 in the figure below) to operate at the same current as another series connected pair (Q1, Q3). Three of these transistors are the same size and therefore operate at equal current densities. Consequently, they have the same base-emitter voltage. The fourth transistor is larger than the others so that at the same current it operates at lower current density. This implies that it has a lower base-emitter voltage. The base-emitter junctions of the four transistors connect in a loop which is completed by a resistor. Two of the voltages are connected to subtract from the others so that the net voltage across resistor is just the difference between the base-emitter voltages of the differently sized transistors.

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

A CROSS-CONNECTED TRANSISTOR QUAD PROVIDES CTAT VOLTAGE IN THE FORM OF $2V_{BE}$s AND PTAT VOLTAGE FROM THE DIFFERENCE OF V_{BE}s.



Applications

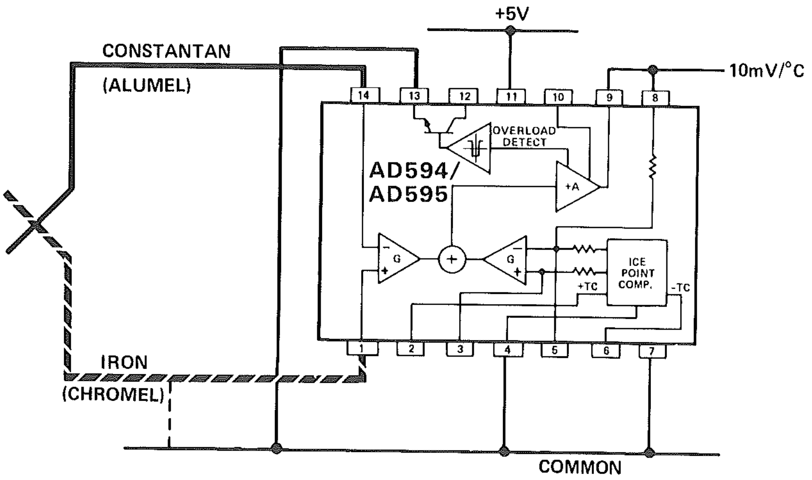
Single Supply Operation

The AD594/AD595 is completely self contained with the interconnections shown below and will provide a direct output from type J and K thermocouples measuring from 0 to +300°C. The measuring thermocouple wires connect to pins 1 and 14 either directly or through intervening connections. The connections at which the thermocouple wires terminate form the reference junction. This junction should be kept at the same temperature as the AD594/AD595 since this is the junction compensated by the ice point reference in the AD594/AD595. If the thermocouple is not directly connected to pins 1 and 14, the intervening connections must both be made of the same material.

In this single supply application the V− connection at pin 7 is strapped to power and signal common, pin 4. When the alarm is unused, pin 13 must connect to either pin 4 and pin 7 (common or V−). The positive 5 volt supply connects to pin 11. Any convenient supply voltage from +5 to +30 volt may be used, however, the lower the supply voltage the lower the power consumption. It is important to minimize power consumption so that self heating of the circuit can be neglected.

The output is taken from pin 9, with the precalibrated feedback network (pin 8) strapped to the output to provide a 10mV/°C nominal output scale.

BASIC CONNECTION, SINGLE SUPPLY OPERATION



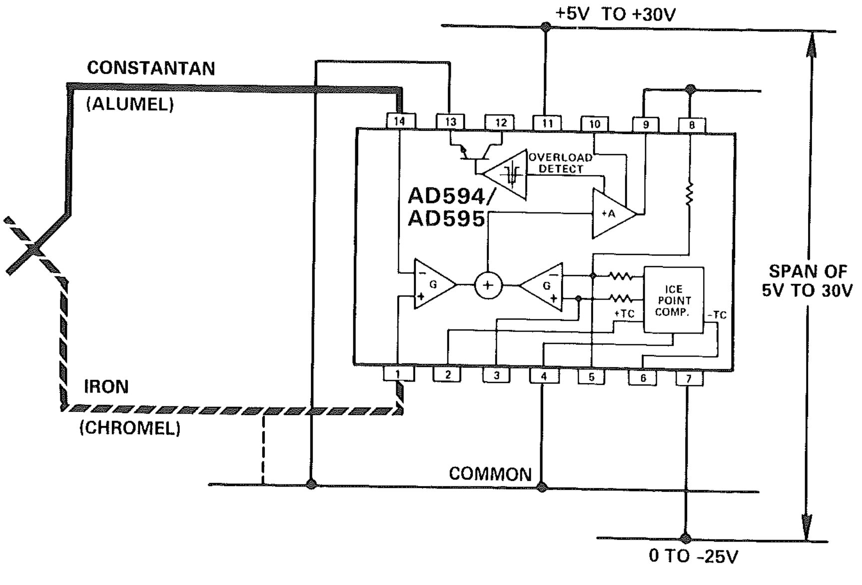
Dual Supply Operation

The AD594/AD595 requires dual supplies when operating with thermocouples at negative temperatures in order to allow the output voltage to assume negative values.

As shown in the figure below, the AD594/AD595 can operate on a total supply span between pins 11 and 7 from 5 to 30 volts. Also, at least 5 volts is required between pins 11 and 4 for proper operation of the circuit.

Since the output can swing to within 2 volt of the positive supply, a +15 volt supply will allow operation of the AD595 circuit at the maximum recommended type K thermocouple temperature of 1250°C. A negative

DUAL SUPPLY OPERATION



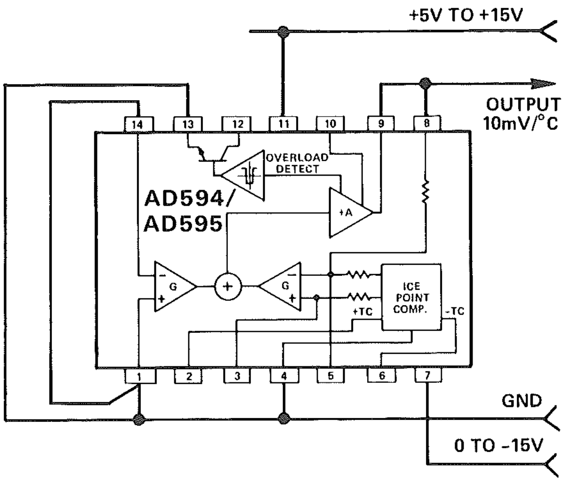
supply of −5 volts will allows the AD594/AD595 to function with J and K type thermocouples at their minimum recommended temperatures.

These flexible power supply requirements allow the AD594/AD595 to be included in most systems without the need for special power supplies or level translators.

Centigrade Thermometer

The AD594/AD595 contain a temperature reference which is internally offset to zero Celsius for use as cold junction compensation for the thermocouple. Without the thermocouple attached, this reference indicates the temperature of the IC and the self-contained fixed-gain amplifier can be made to scale up this signal for a 10mV/°C output. This arrangement as shown is a three-terminal (voltage output) temperature sensor referred to zero. Note that if negative temperature indications are desired, a negative supply should be connected to pin 7 of the device.

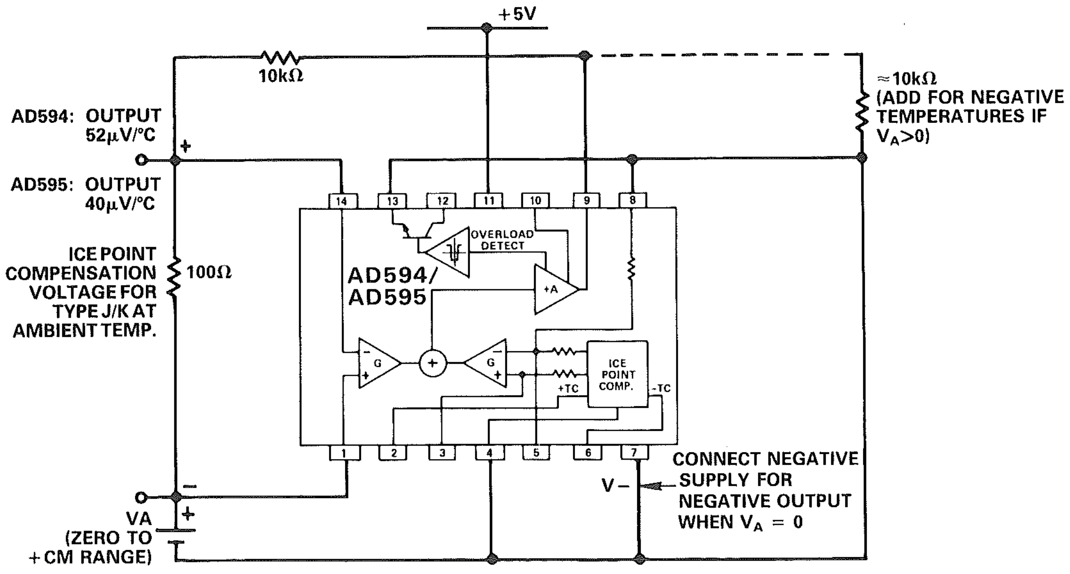
AD594/AD595 AS A STAND-ALONE CENTIGRADE THERMOMETER



Extracting CJC Voltage

Extracting the cold junction compensation voltage is possible using the circuit shown below. This voltage provides a low impedance drive signal for compensating one or more reference junctions that are at the same ambient temperature as the AD594/AD595.

ICE POINT COMPENSATION VOLTAGE



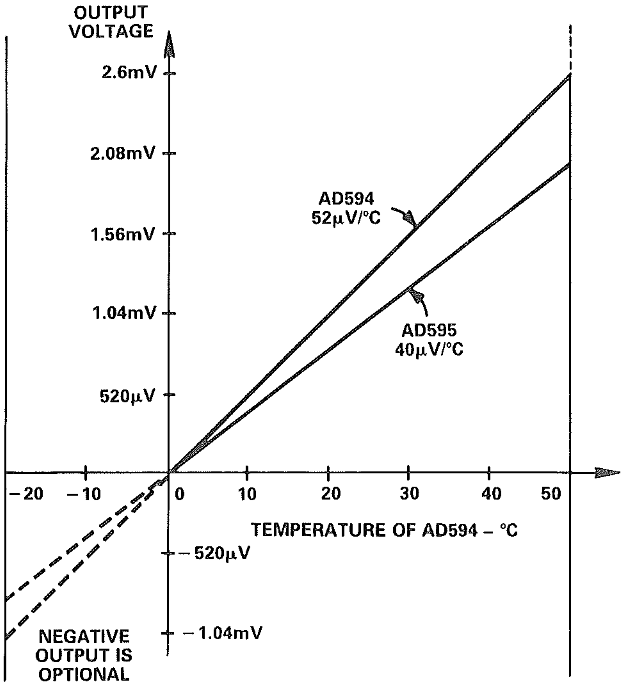
In practice, the feedback path at pin 8 is grounded, and the main circuit's control loop is completed by way of the thermocouple inputs (at pins 1 and 14). The exact values of the feedback resistors are not critical, since their only function is to frequency stabilize the loop by providing some attenuation in the feedback path.

Moreover, since the signal is derived across a differential input it may be either ground referenced or referred to an arbitrary voltage (Va) within the common-mode range of the amplifier.

A positive Va voltage will permit the output amplifier to be negative with respect to the compensation point and provide negative Celsius temperatures.

The figure below illustrates output voltages versus temperature with factory calibrated compensations of 52μV/°C and 40μV/°C from the AD594 and AD595 respectively.

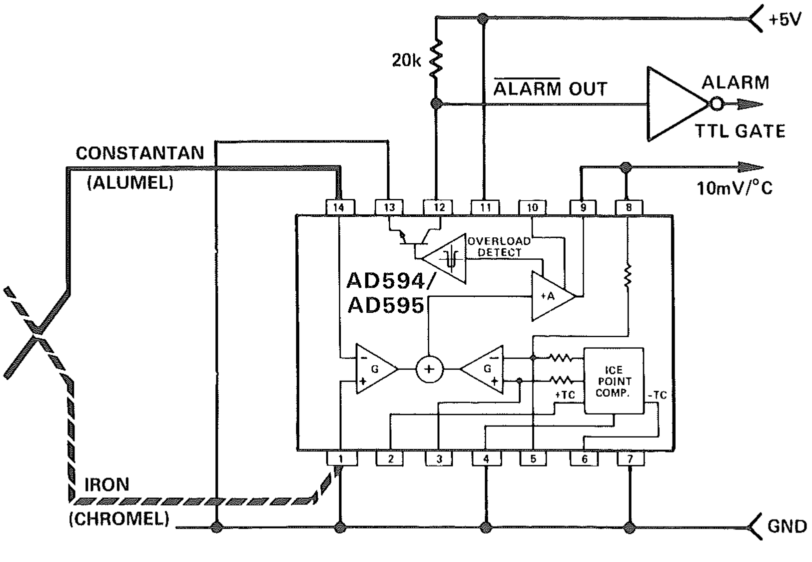
ICE POINT COMPENSATION VOLTAGE VS. TEMPERATURE



Alarm Circuit

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

USING THE ALARM TO DRIVE A TTL GATE ("GROUNDED" EMITTER CONFIGURATION)

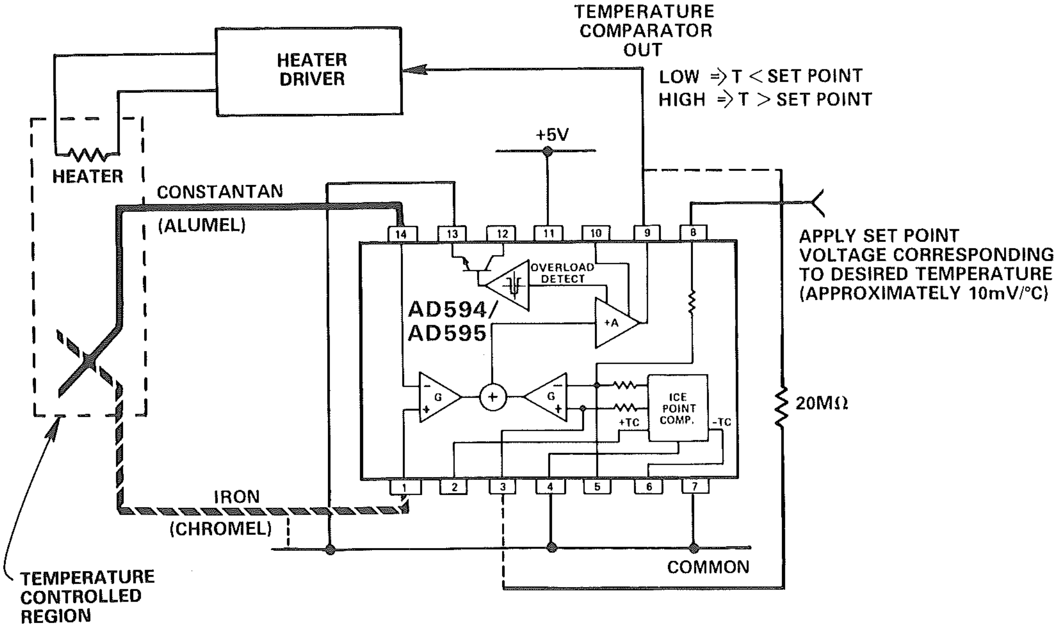


In this configuration the alarm transistor will be off in normal operation and the 20K pull up will cause the +ALM output on pin 12 to go high. If one or both of the thermocouple leads are interrupted, the +ALM pin will be driven low. As shown above, this signal is compatible with the input of a TTL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

The output alarm transistor is current limited to 20mA so a series limiting resistor is not required. The transistor, however, will operate in a high dissipation mode and the temperature of the circuit will rise well above ambient. The cold junction compensation will be affected whenever the alarm circuit is activated. The time required for the chip to return to ambient temperature will depend on the power dissipation of the alarm circuit, the nature of the thermal path to the environment and the alarm duration.

Set-Point Controller
The AD594/AD595 can readily be connected as a set-point controller as shown below.

SET-POINT CONTROLLER



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

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

AD693 FEATURES

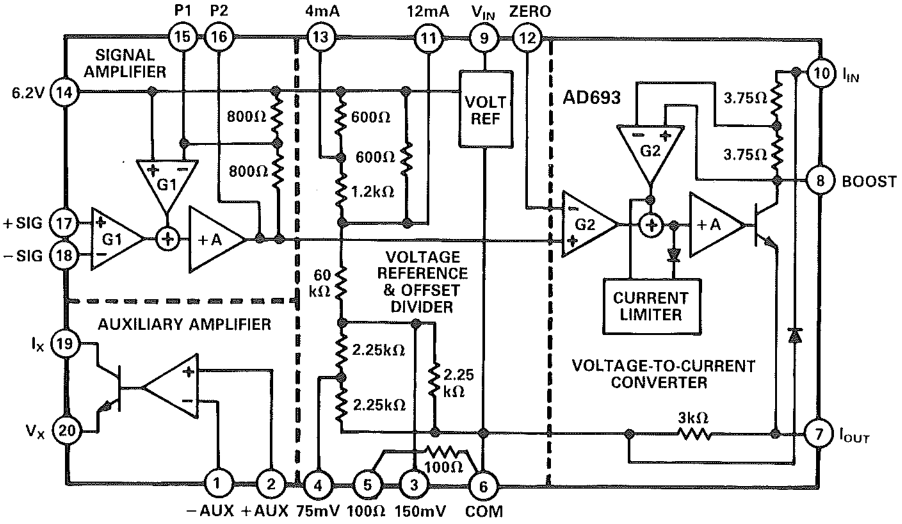
Instrumentation Amplifier Front End
Loop-Powered Operation
Precalibrated 30mV or 60mV Input Spans
Independently Adjustable Output Span and Zero
Precalibrated Output Spans: 4-20mA Unipolar
0-20mA Unipolar
12 ± 8mA Bipolar
Precalibrated 100Ω RTD Interface
6.2V Reference with Up to 3.5mA of Current Available
Uncommitted Auxiliary Amp for Extra Flexibility
Optional External Pass Transistor to Reduce Self-Heating Errors

The AD693 is a monolithic signal conditioning circuit which accepts low-level inputs from a variety of transducers to control a standard 4-20mA, two-wire current loop. An on-chip reference and auxillary amplifier are provided for transducer excitation. Alternatively, the device may be locally powered for three-wire applications when 0-20mA operation is desired.

Functional Description

The operation of the AD693 can be understood by dividing the circuit into four functional parts as shown below. First, an instrumentation amplifier front-end buffers and scales the low-level input signal. This amplifier drives the second section, a V/I converter, which provides the 4-to-20mA loop current. The third section, a voltage reference and resistance divider, provides application voltages for setting the various "live zero" currents. In addition to these three main sections, there is an on-chip auxillary amplifier which can be used for transducer excitation.

AD693 FUNCTIONAL BLOCK DIAGRAM

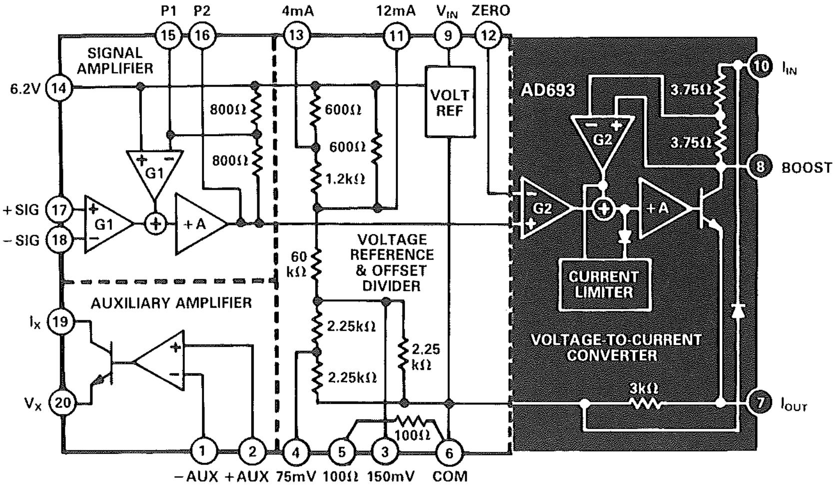


Voltage-to-Current (V/I) Converter

The output NPN transistor for the V/I section sinks loop current when driven by a high gain amplifier at its base. The input for this amplifier is derived from the difference between the input signal and the loop current sampling resistor between I_{IN} and Boost. The signal across this resistor is compared to the input of the left preamp and servos the loop current until both signals are equal. Accurate voltage-to-current transformation is thereby assured. The preamplifiers employ a special design which allows the active feedback amplifier to operate from the most positive point in the circuit, I_{IN}.

The V/I stage is designed to have a nominal transconductance of 0.2666 A/V. Thus, a 75mV signal applied to the inputs of the V/I (pin 16, noninverting; pin 12, inverting) results in a full-scale output current of 20mA.

AD693
VOLTAGE-TO-CURRENT CONVERTER

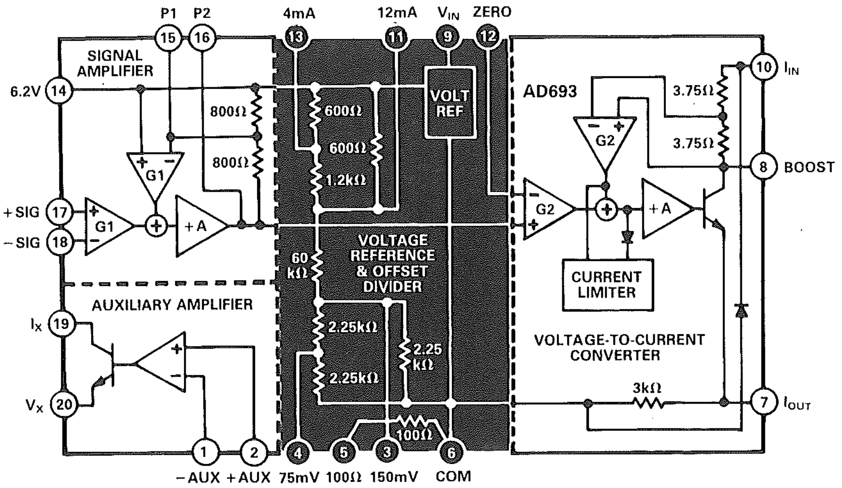


Voltage Reference and Divider
A stabilized bandgap voltage reference and laser-trimmed resistor divider provide for both transducer excitation as well as precalibrated offsets for the V/I converter. When not used for external excitation, the reference should be loaded by approximately 1mA (6.2kΩ to common).

The 4mA and 12mA taps on the resistor divider correspond to −15mV and −45mV, respectively, and result in a live zero of 4mA or 12mA of loop current when connected to the V/I converter's inverting input (pin 12). Arranging the zero offset in this way makes the zero signal output current independent of input span. When the input to the signal amp is zero, the noninverting input of the V/I is at 6.2V.

Since the standard offsets are laser trimmed at the factory, adjustment is seldom necessary except to accommodate the zero offset of the actual source.

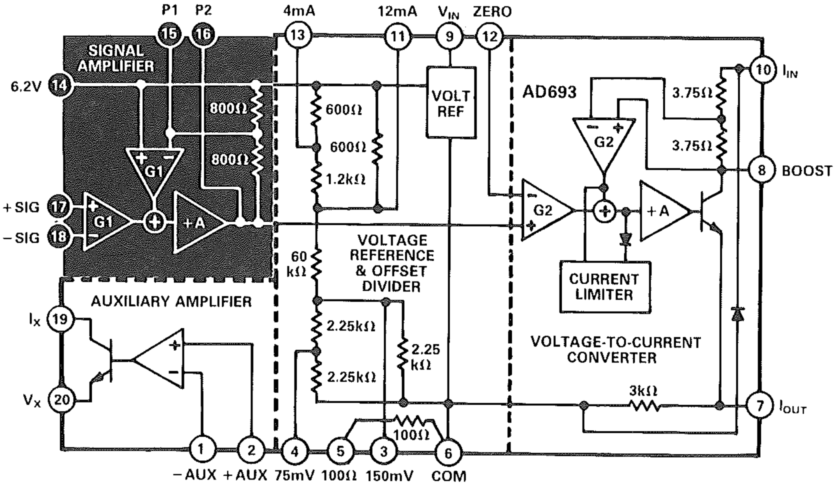
AD693
VOLTAGE REFERENCE & OFFSET DIVIDER



Signal Amplifier
The signal amplifier is an instrumentation amplifier used to buffer and scale the input to match the desired span. Inputs applied to the signal amplifier (at pins 17 and 18) are amplified and referred to the 6.2V reference output in much the same way as the level translation occurs in the V/I converter. Signals from the two preamplifiers are subtracted, the difference is amplified, and the result is fed back to the upper preamp to minimize the difference. Since the two preamps are identical, this minimum will occur when the voltage at the upper preamp just matches the differential input aplied to the signal amplifier at the left.

Since the signal which is applied to the V/I is attenuated across the two 800Ω resistors before driving the upper preamp, it will necessarily be an amplified version of the signal applied between pins 17 and 18. By changing this attenuation, you can control the span referred to the signal amplifier. To illustrate: a 75mV signal applied to the V/I results in a 20mA loop current. Nominally, 15mV is applied to offset the zero to 4mA leaving a 60mV range to correspond to the span. And, since the nominal attenuation of the resistors connected to pins 16, 15 and 14 is 2.00, a 30mV input signal will be doubled to result in 20mA of loop current.

AD693
SIGNAL AMPLIFIER



Shorting pins 15 and 16 results in unity gain and permits a 60mV input span. Other choices of span may be implemented with user supplied resistors to modify the attenuation.

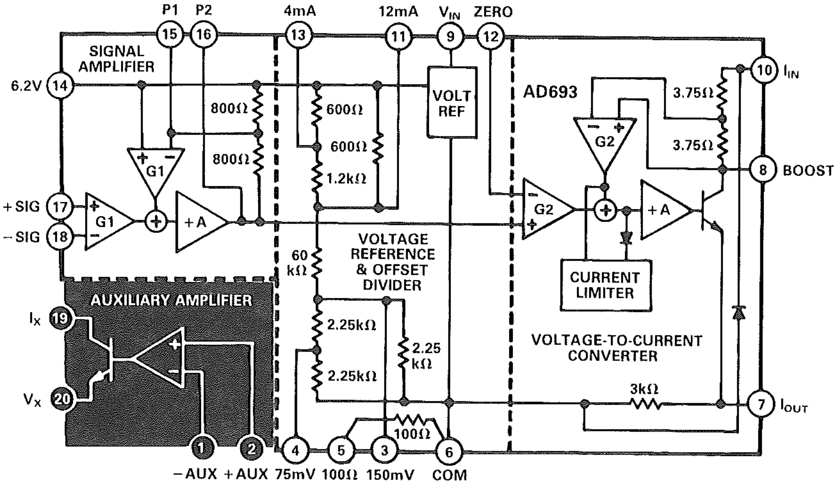
The signal amplifier is specially designed to accommodate a large common-mode range. Common-mode signals up to and beyond the 6.2V reference are easily handled as long as V_{IN} is sufficiently positive. The signal amplifier is biased with respect to V_{IN} and requires about 3.5 volts of headroom. The extended range will be useful when measuring sensors driven, for example, by the auxillary amplifier which may go above the 6.2V potential. In addition, the PNP input stage will continue to operate normally with common-mode voltages of several hundred millivolts, negative, with respect to common. This feature accommodates self-generating sensors, such as thermocouples, which may produce small negative normal-mode signals as well as common-mode noise on "grounded" signal sources.

Auxillary Amplifier

The auxillary amplifier is included in the AD693 as a signal conditioning aid. It can be used as an op amp in noninverting applications and has special provisions to provide a controlled current output. Designed with a differential input stage and an unbiased Class A output stage, the amplifier can be resistively loaded to common with the self-contained 100Ω resistor or with a user supplied resistor.

As a functional element, the auxillary amplifier can be used in dynamic bridges and arrangements such as RTD signal conditioning. It can be used to buffer, amplify and combine other signals with the main signal amplifier. The auxillary amplifier can also provide other voltages for excitation if the 6.2V of the reference is unsuitable. Configured as a simple follower, it can be driven from a user supplied voltage divider or the precalibrated outputs of the AD693 divider (pins 3 and 4) to provide a stiff voltage output at less than the 6.2V level, or by incorporating a voltage divider as feedback around the amplifier, one can gain-up the reference to levels higher than 6.2V. If large positive outputs are desired, I_X, the auxillary amplifier output current supply, should be strapped to either V_{IN} or Boost. Like the signal amplifier, the auxillary amplifier requires about 3.5V of headroom with respect to V_{IN} at its input and about 2V of difference between I_X and the voltage to which V_X is required to swing.

AD693
AUXILIARY AMPLIFIER



The output stage of the auxillary amplifier is actually a high gain Darlington transistor where I_X is the collector and V_X is the emitter. Thus, the auxillary amplifier can be used as a V/I converter when configured as a follower and resistively loaded. I_X functions as a high-impedance current source whose current is equal to the voltage at V_X divided by the load resistance. For example, using the onboard 100Ω resistor and the 75mV or 150mV application voltages, either a 750ΩA or a 1.5mA current source can be set up for transducer excitation.

The I_X terminal has a voltage compliance within 2V of V_X. If the auxillary amplifier is not used, then pin 2, the noninverting input, should be grounded.

Reverse Voltage Protection Feature

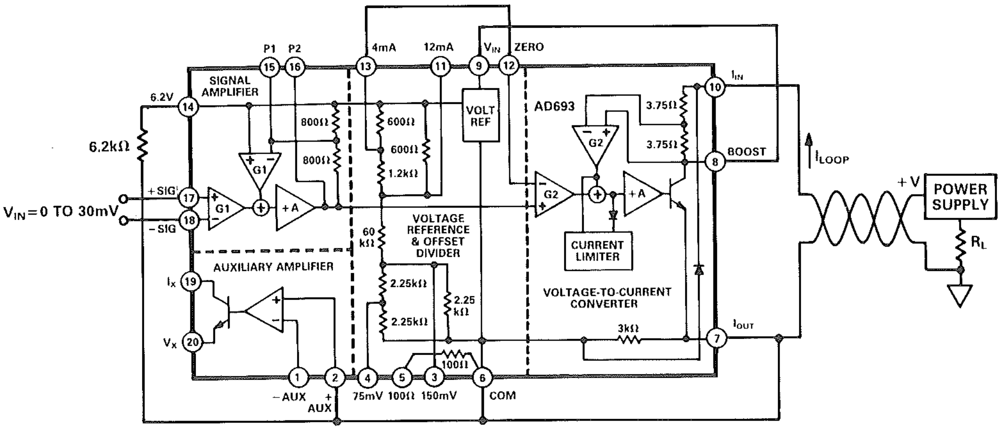
In the event of a reverse voltage being applied to the AD693 through a current-limited loop (limited to 200mA), an internal shunt diode protects the device from damage. This protection mode avoids the compliance voltage penalty which results from a series diode that must be added if reversal protection is required in high current loops.

Applications

Basic Operation

The figure below shows the minimal connections for basic operation: 0-30mV input span, 4-20mA output span in the two-wire, loop-powered mode. Loop power may be +12 to +36 volts.

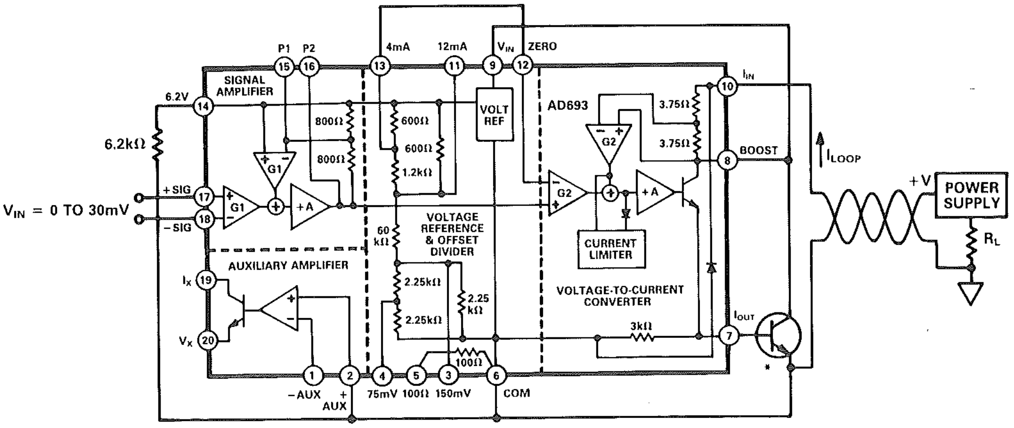
MINIMAL CONNECTION FOR 0–30mV UNIPOLAR INPUT, 4-20mA OUTPUT



Using an External Pass Transistor

The emitter of the NPN output section, I_{OUT}, of the AD693 is usually connected to common and the negative loop connection (pins 6 and 7). Provision has been made to reconnect I_{OUT} to the base of a user supplied NPN transistor as shown below. This permits the majority of the power dissipation to be moved off chip to enhance performance, improve reliability, and extend the operating temperature range. An internal hold-down resistor of about 3K is connected across the base emitter of the external transistor.

USING AN EXTERNAL PASS TRANSISTOR TO MINIMIZE SELF-HEATING ERRORS



*THE AD693 IS TESTED WITH A 2N3440
AT THE FACTORY. RECOMMENDED
TRANSISTOR TYPES ARE GIVEN
IN THE DATA SHEET TEXT.

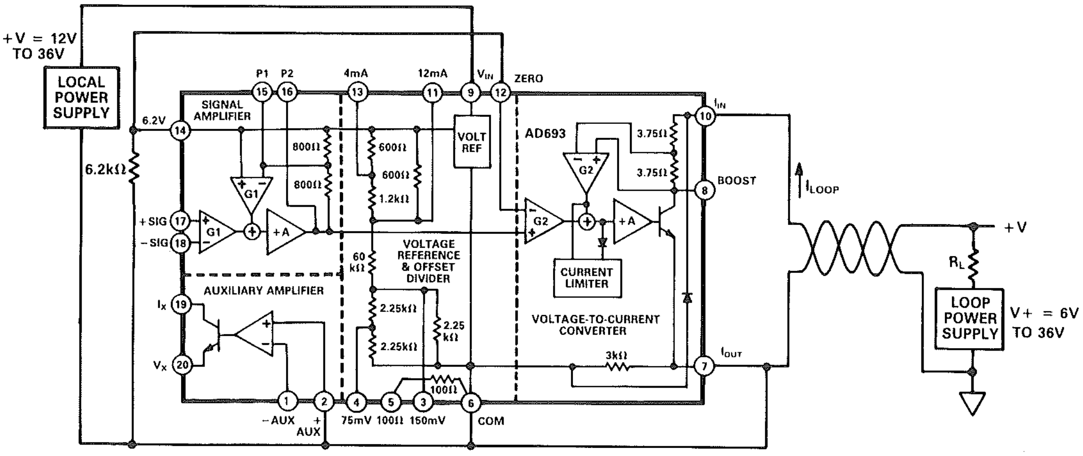
The external pass transistor selected should have a BV_{CEO} greater than the intended supply voltage with a sufficient power rating for continuous operation with 25mA current at the supply voltage. F_T should be in the 10MHz to 100MHz range and β should be greater than 10 at a 20mA emitter current. Some transistors that meet this criteria are the 2N1711 and 2N2219A. Heat sinking the external pass transistor is suggested.

The pass transistor option may also be emloyed for other applications as well. For example, I_{OUT} can be used to drive an LED connected to common, thus providing a local monitor of loop fault conditions without reducing the minimum compliance voltage.

Local-Powered Operation

The AD693 is designed for local-powered, three-wire systems as well as two-wire loops. All its usual ranges are available in three-wire operation, and in addition, the 0-to-20mA range can be used. The 0-20mA convention offers slightly more resolution and may simplify the loop receiver, two reasons why it is sometimes preferred.

LOCAL POWERED OPERATION WITH 0-20mA OUTPUT

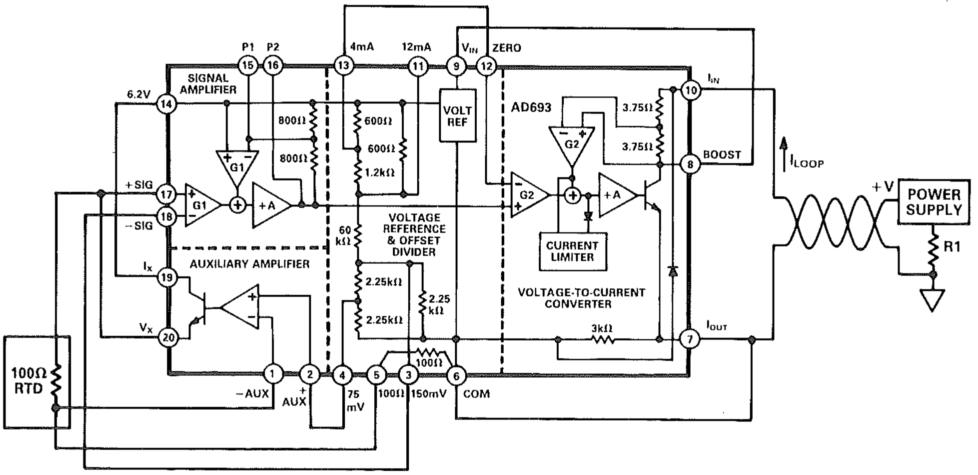


The arrangement, illustrated above, results in a 0-20mA transmitter where the precalibrated span is 37.5mV. Connecting P1 to P2 will double the span to 75mV. Sensor input and excitation is unchanged from the two-wire mode except for the 25% increase in span. Many sensors are ratiometric so that an increase in excitation can be used instead of a span adjustment.

In the local-powered mode, increases in excitation are made easier. Voltage compliance at the I_{IN} terminal is also improved; the loop voltage may be permitted to fall to 6 volts at the AD693, easing the tradeoff between loop voltage and loop resistance. Note that the load resistor, R1, should meter the current into pin 10, I_{IN}, so as not to confuse the loop curent with the local supply current.

Interfacing Platinum RTDs

0-TO-104°C DIRECT THREE-WIRE 100Ω RTD INTERFACE, 4-20mA OUTPUT



The AD693 has been specially configured to accept inputs from 100Ω Platinum RTDs (Resistance Temperature Detectors). Referring to the above circuit, the RTD and the temperature stable 100Ω resistor form a feedback network around the auxillary amplifier resulting in a noninverting gain of (1 + Rt/100Ω), where Rt is the temperature dependent resistance of the RTD. The noninverting input of the auxillary amplifier (pin 2) is then driven by the 75mV signal from the voltage divider (pin 4). When the RTD is at 0, its 100Ω resistance results in an amplifier gain of +2 causing V_X to be 150mV. The signal amplifier compares this voltage to the 150mV output (pin 3) so that zero differential signal results. As the temperature (and therefore, the resistance) of the RTD increases, V_X will likewise increase according to the gain relationship. The difference between this voltage and the zero degree value of 150mV drives the signal amp to modulate the loop current. The AD693 is precalibrated such that the full 4-20mA output span corresponds to 0 to 104°C range in the RTD. (This assumes the European Standard of $\alpha = 0.00385$) A total of 6 precalibrated ranges for three-wire (or two-wire) RTDs are available using only the pin strapping options as shown below.

PRECALIBRATED TEMPERATURE RANGE OPTIONS USING A EUROPEAN STANDARD 100Ω RTD AND THE AD693

Temperature Range	Pin Connections
0 to +104°C	12 to 13
0 to +211°C	12 to 13, and 15 to 16
+25°C to +130°C	12 to 14
+51°C to +266°C	12 to 14, and 15 to 16
−50°C to +51°C	12 to 11
−100°C to +104°C	12 to 11 and 15 to 16

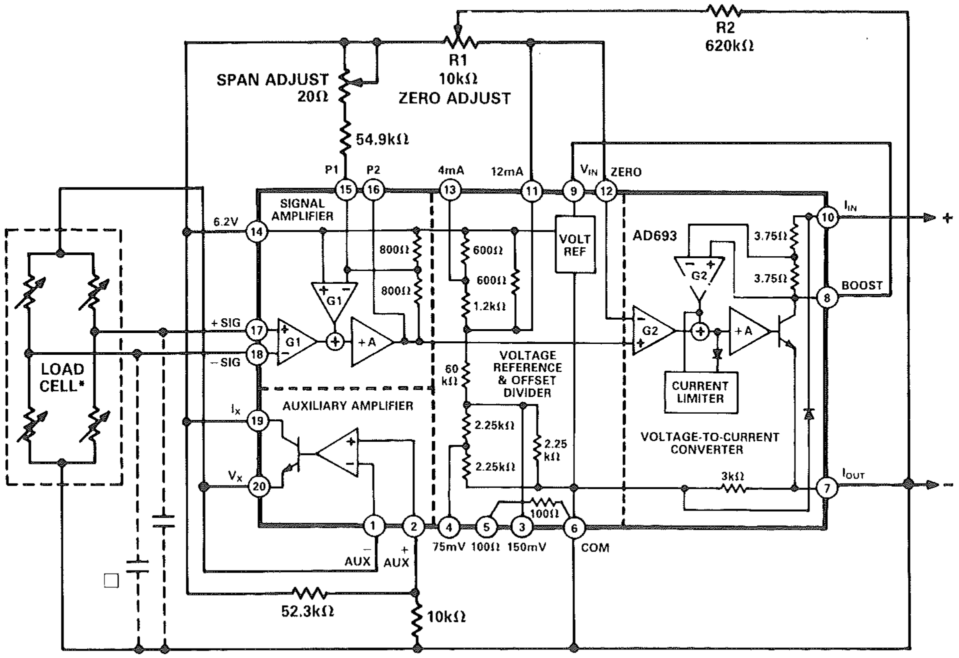
A variety of other temperature ranges can be realized by using different application voltages. For example, loading the voltage divider with a 1.5kΩ resistor from pin 3 to pin 6 (common) will approximately halve the original application voltages and allow for doubling of the range of resistance (and therefore, temperature) required to fill the two standard spans. Likewise, increasing the application voltages by adding resistance between pins 14 and 3 will decrease the temperature span.

The configuration for a three-wire RTD shown can accommodate two-wire sensors by simply joining pins 1 and 5 of the AD693.

Interfacing Load Cells

The availability of the on-chip voltage reference, auxillary amplifier and 3mA of excitation current make it easy to adapt the AD693 to a variety of load cells and strain gages.

UTILIZING THE AUXILIARY AMPLIFIER TO DRIVE A LOAD CELL, 12mA ±8mA OUTPUT



*350Ω, 2mV/V
E.G., SENSOTEC MODEL 41
AL DESIGN MODEL ALD-UTC

☐ OPTIONAL INPUT FILTER
(SEE TEXT)

The circuit above illustrates a generalized approach in which the full flexibility of the AD693 is required to interface to a low resistance bridge. For a high impedance transducer the bridge can be directly powered from the 6.2V reference.

Component values in this example have been selected to match the popular standard of 2mV/V sensitivity and 350Ω bridge resistance. Load cells are generally made for either tension and compression, or compression only; use of the 12mA zero tap allows for operation in the tension and compression mode. An optional zero adjustment is provided with values selected for ±2% FS adjustment range.

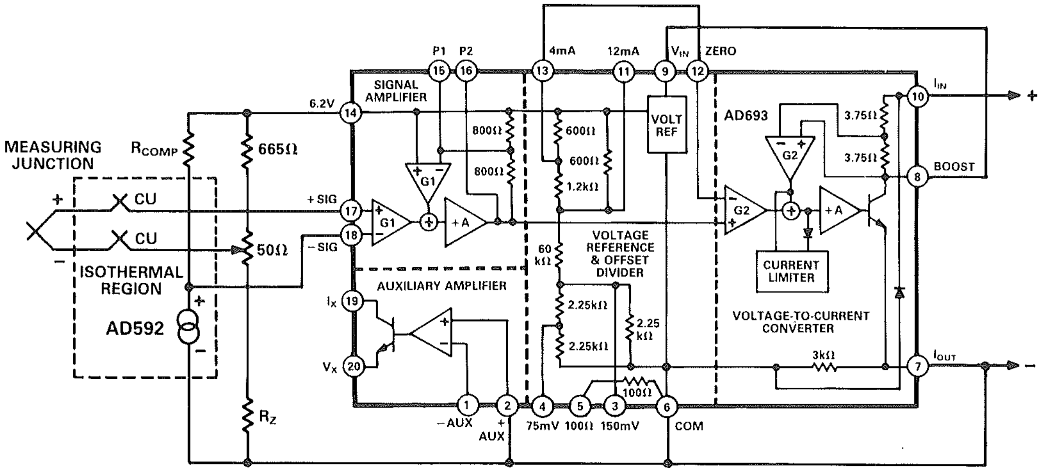
Because of the low resistance of most foil bridges, the excitation voltage must be low so as not to exceed the available 4mA zero current. About 1V is derived from the 6.2V reference and an external voltage divider; the auxillary amp is then used as a follower to make a stiff drive for the bridge. Similar applications with higher resistance sensors can use proportionally higher voltage.

Finally, to accommodate the 2mV/V sensitivity of the bridge, the full scale span of the signal amplifier must be reduced. Using the load cell in both tension and compression with 1V of excitation, therefore, dictates that the span be adjusted to 4mV. By substituting in the expression, $R_{S1} = 400\Omega/[(30mV/Span) - 1]$, the nominal resistance required to achieve this span is found to be 61.54Ω. Calculate the minimum resistance required by subtracting 10% from 61.54Ω to allow for the internal resistor tolerance of the AD693, leaving 55.38Ω. The standard value of 54.9Ω is used with a 20Ω potentiometer for full-scale adjustment.

Thermocouple Interface

The AD693 can be used with several types of thermocouple inputs to provide a 4-20mA current loop output corresponding to a variety of measurement temperature ranges. Cold junction compensation (CJC) can be implemented using the AD590 or AD592 and a few resistors as shown below.

THERMOCOUPLE INPUTS WITH COLD JUNCTION COMPENSATION



From the table below, simply choose the type of thermocouple and the appropriate average reference junction temperature to select values for R_{COMP} and R_Z. The CJC voltage is developed across R_{COMP} as a result of the AD592 1μA/K output and is added to the thermocouple loop voltage. The 50Ω potentiometer is biased by R_Z to provide the correct zero adjustment range appropriate for the divider and also translates the Kelvin scale of the AD592 to °Celsius. To calibrate the circuit, put the thermocouple in an ice bath (or use a thermocouple simulator set to 0) and adjust the potentiometer for a 4mA loop current.

THERMOCOUPLE APPLICATION – COLD JUNCTION COMPENSATION TABLE

POLARITY	MATERIAL	TYPE	AMBIENT TEMP	RCOMP	RZ	30mV TEMP RANGE	60mV TEMP RANGE
+	IRON	J	25°	51.7Ω	301K	546°C	1035°C
–	CONSTANTAN		75°	53.6Ω	294K		
+	NICKEL-CHROME		25°	40.2Ω	392K	721°C	—
–	NICKEL-ALUMINUM	K	75°	42.2Ω	374K		
+	NICKEL-CHROME	E	25°	60.4Ω	261K	413°C	787°C
–	COPPER-NICKEL		75°	64.9Ω	243K		
+	COPPER	T	25°	40.2Ω	392K	USE WITH GAIN>2	
–	COPPER-NICKEL		75°	45.3Ω	340K		

The span of the circuit in °C is determined by matching the signal amplifier input voltage range to its temperature equivalent via a set of thermocouple tables referenced to 0°C. For example, the output of a properly referenced type J thermocouple is 60mV when the hot junction is at 1035°C. The table lists the maximum measurement temperature for several thermocouple types using the preadjusted 30mV and 60mV input ranges.

More convenient temperature ranges can be selected by determining the full-scale input voltages via standard thermocouple tables and adjusting the AD693 span. For example, suppose only a 300°C span is to be measured with a type K thermocouple. From a standard table, the thermocouple to a 16mA span at the output, a gain of 5 (or more precisely 60mV/12.207mV = 4.915) will be needed. To achieve the 12.207mV span a resistance of 270Ω is required from P1 to 6.2V. Adding a 50Ω potentiometer will allow ample adjustment range.

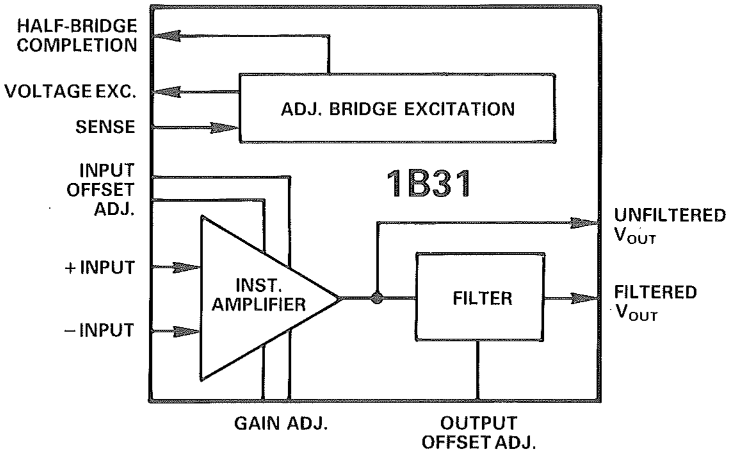
With the connection illustrated, the AD693 will give a full-scale indication with an open thermocouple.

COMPLETE STRAIN GAGE AND BRIDGE TRANSDUCER SIGNAL CONDITIONERS

1B31 FEATURES

Low Cost
Complete Signal-Conditioning Solution
Small Package: 28-Pin Double DIP
Internal Half-Bridge Completion Resistors
Remote Sensing
High Accuracy
  Low Drift: ±0.25μV/°C
  Low Noise: 0.3μV p-p
  Low Nonlinearity: ±0.005% max
  High CMR: 140dB min (60Hz, G = 1000V/V)
Programmable Bridge Excitation: +4V to +15V
Adjustable Low Pass Filter: f_C = 10Hz to 20kHz

1B31 BLOCK DIAGRAM



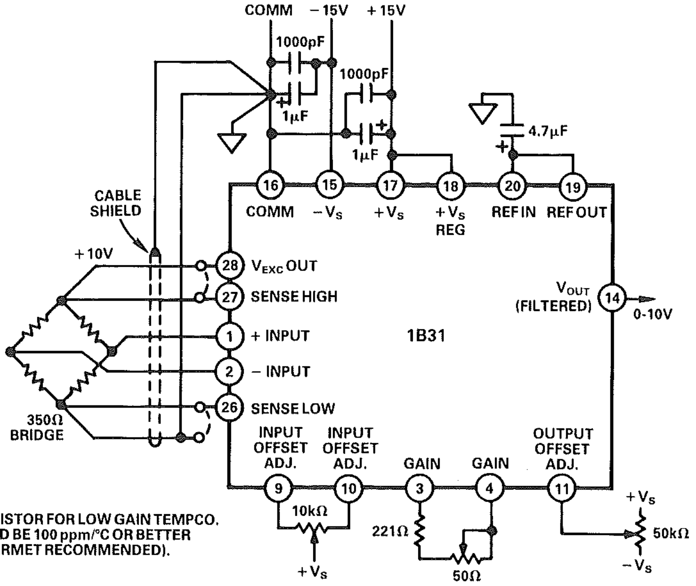
Model 1B31 is based on a two-stage amplifier design and an adjustable voltage regulator section, as shown above. The front end is a low noise, low drift, instrumentation amplifier (IA) that is optimized to amplify low level transducer signals (from 2mV full scale) riding an high common-mode voltages (±9.5V). The gain of the IA is programmed by a single resistor (1V/V to 2500V/V) and the input offset nulled by an external potentiometer across the offset adjust pins 9 and 10. The inverted signal ($V_{-INPUT} - V_{+INPUT}$) is brought out to pin 8 for applications such as vibration and torque testing where the unfiltered output is required.

The signal is also fed to an inverting Butterworth filter with a fixed gain of −2V/V. This two-pole filter is preset with a 1kHz corner frequency which can be adjusted downwards to 10Hz by using two external capacitors or upwards to 20kHz by three resistors. This stage also provides a convenient means of adjusting output offset voltage (±10V) by connecting a 50kΩ potentiometer to pin 11.

The bridge excitation section is an adjustable output, regulated supply with an internally provided reference voltage (+6.8V). It is configured as a gain stage with the output preset at +10V. The excitation voltage is adjusted by connecting resistors between pins 21 and 26, and between pins 19 and 20. Sense lines are provided to compensate for lead-wire resistance by effectively bringing the leads into the feedback loop.

For half-bridge applications, two tracking thin film resistors (20kΩ, ±5ppm/°C max) are connected from V_{EXC} OUT (pin 28) to SENSE LOW (pin 26).

1B31 BASIC CONFIGURATION



NOTES:
- USE 5ppm/°C GAIN RESISTOR FOR LOW GAIN TEMPCO.
- ALL TRIM POTS SHOULD BE 100 ppm/°C OR BETTER (TYPE 79PR 15 TURN CERMET RECOMMENDED).

Basic Operation

The figure above shows the 1B31 configured for basic operation. The differential gain, G, is determined by the equation:

$$G = 1 + (80k\Omega/R_G)$$

where R_G is connected between the GAIN terminals (pins 3 and 4) of the 1B31. For best performance, a low temperature coefficient (5ppm/°C) R_G is recommended. For fine span adjustment, a 50Ω potentiometer may be connected in series with R_G.

Input offset can be nulled with a 10kΩ potentiometer between pins 9 and 10 and tapped to $+V_S$ (if not used leave pins open). The output can be offset over the ± 10V range to compensate for dead load of source imbalance by using a 50kΩ potentiometer connected between the supply rails and tapped to pin 11 (pin 11 should be grounded if not used).

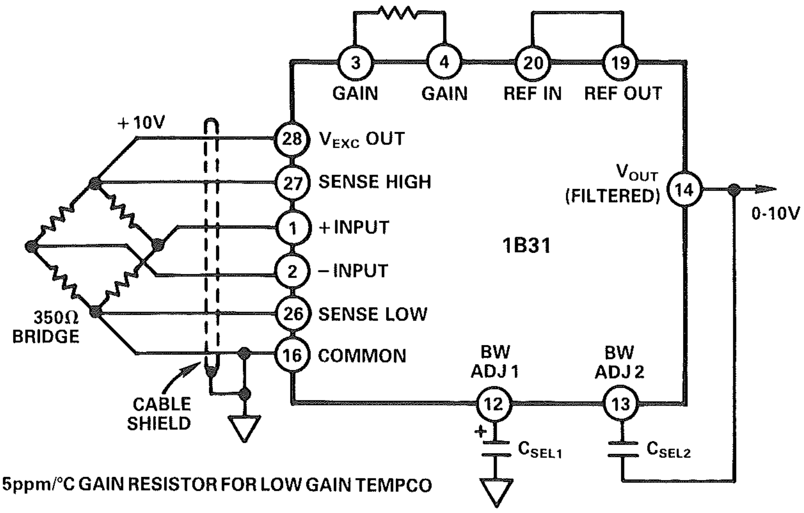
Filter Cutoff Frequency Programming

The low pass filter cutoff frequency is internally set at 1kHz. It may be decreased from 1kHz by the addition of two external capacitors connected as shown below (from pin 12 to common and between pins 13 and 14). The values of capacitors required from a desired cutoff frequency, f_C, below 1kHz are obtained by the equations:

$$C_{SEL1} = 0.015\mu F[(1kHz/f_C) - 1]$$

$$C_{SEL2} = 0.0022\mu F[(1kHz/f_C) - 1]$$

NARROW BANDWIDTH APPLICATION



NOTE
- USE 5ppm/°C GAIN RESISTOR FOR LOW GAIN TEMPCO

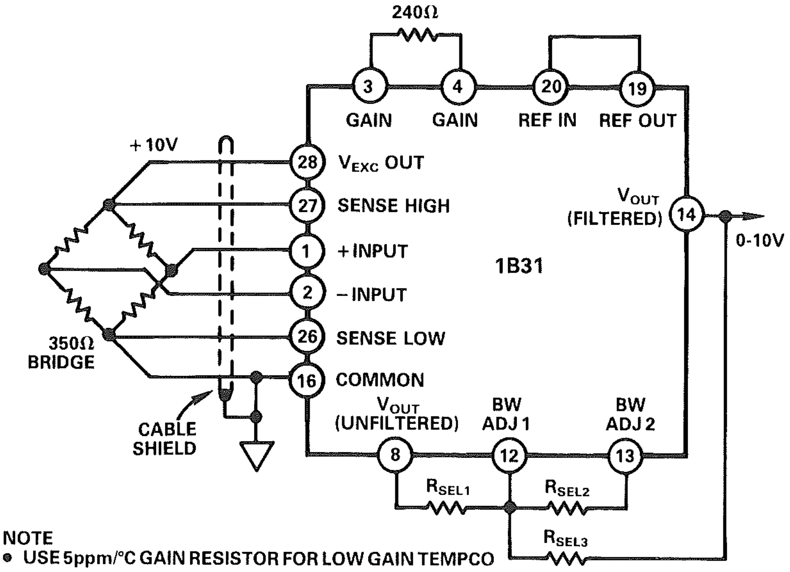
The cutoff frequency may also be increased from 1kHz to 20kHz by the addition of three external resistors, connected as shown in the circuit below. The equations for determining the resistor values are:

$R_{SEL1} = 20k\Omega/[(f_C/1kHz) - 1]$

$R_{SEL2} = 16k\Omega/[(f_C/1kHz) - 1]$

$R_{SEL3} = 40k\Omega/[(f_C/1kHz) - 1]$

WIDE BANDWIDTH APPLICATION



NOTE
- USE 5ppm/°C GAIN RESISTOR FOR LOW GAIN TEMPCO

The table below gives the nearest resistor and capacitor values for several common filter cutoff frequencies.

FILTER CUTOFF FREQUENCY VS R_{SEL} AND C_{SEL}

f_C(Hz)	C_{SEL1} (μF)	C_{SEL2} (μF)	
10	1.5	0.2	
50	0.27	0.039	
100	0.15	0.02	
200	0.056	0.0082	
500	0.015	0.0022	
	R_{SEL1} (kΩ)	**R_{SEL2} (kΩ)**	**(R_{SEL3} (kΩ)**
2000	20	16.2	40.2
5000	4.99	4.12	10.0
10000	2.21	1.78	4.42
20000	1.05	0.866	2.21

Voltage Excitation Programming

The excitation voltage is preset to +10V. To increase V_{EXC} up to +15V a resistor must be connected between EXC ADJ and SENSE LOW (pins 21 and 26) as shown below.

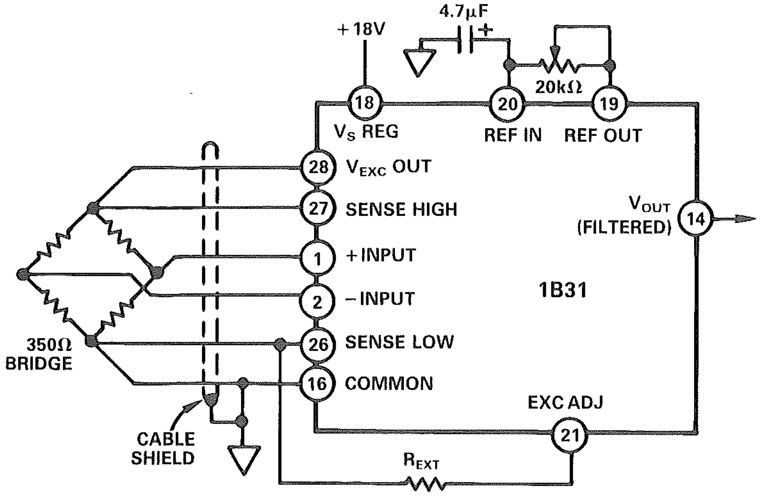
The $+V_S$ (REG) input (pin 18) must be raised to +18V to satisfy the +3V min input-output voltage differential of the regulator. For a desired V_{EXC} the resistor value, R_{EXT}, is determined by the equations:

$R_T = (10k\Omega \times V_{REF\ OUT})/(V_{EXC} - V_{REF\ OUT})$; $V_{REF\ OUT} = 6.8V$

$R_{EXT} = (20k\Omega \times R_T)/(20k\Omega - R_T)$

The +10V to +15V range can be covered by a 20kΩ potentiometer between pins 19 and 20.

CONSTANT VOLTAGE EXCITATON: +10V TO +15V RANGE

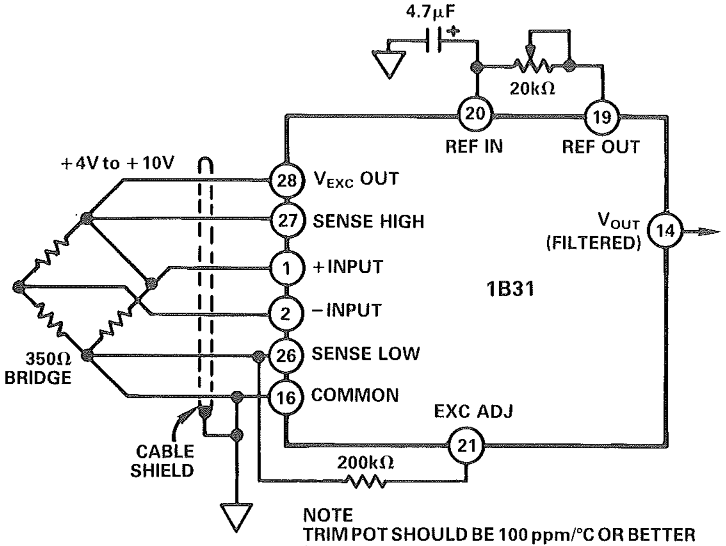


NOTE
TRIM POT SHOULD BE 100 ppm/°C OR BETTER
(TYPE 79PR 15 TURN CERMET RECOMMENDED).

Similarly, to decrease V_{EXC} down to +4V, a 200kΩ resistor and a 20kΩ potentiometer should be connected as shown in the diagram below.

A 4.7μF tantalum capacitor from REF IN (pin 20) to COMMON (pin 16) is recommended in all cases to lower the voltage noise at the reference input.

CONSTANT VOLTAGE EXCITATION: +4V TO +10V RANGE

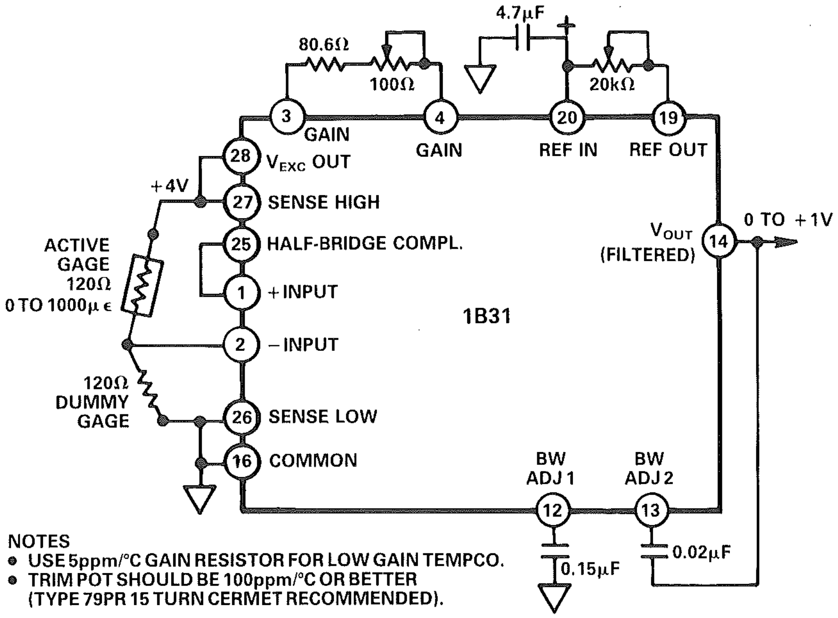


NOTE
TRIM POT SHOULD BE 100 ppm/°C OR BETTER
(TYPE 79PR 15 TURN CERMET RECOMMENDED).

Using the Internal Half-Bridge

The figure below shows a 1B31 in a strain measurement system. A single active gage (120Ω, Gage Factor = 2) is used in a bridge configuration to detect fractional changes in a gage resistance caused by strain. An equivalent resistance dummy gage mounted adjacent to the active gage provides temperature compensation. The rest of the bridge is completed by the 1B31 internal half-bridge network which consists of two 20kΩ, 1% thin-film resistors tracking to within 5ppm/°C. Bridge excitation is set at +4V to avoid self-heating errors from the strain gage. System calibration produces a +1V output for an input of 1000 microstrains. The filter cutoff frequency is set at approximately 100Hz.

STRAIN GAGE APPLICATION USING INTERNAL HALF-BRIDGE

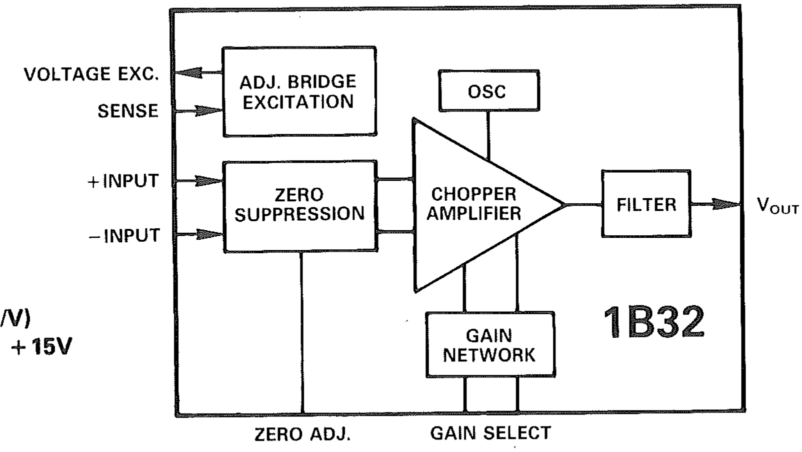


NOTES
- USE 5ppm/°C GAIN RESISTOR FOR LOW GAIN TEMPCO.
- TRIM POT SHOULD BE 100ppm/°C OR BETTER (TYPE 79PR 15 TURN CERMET RECOMMENDED).

1B32 FEATURES

Low Cost
Complete Signal-Conditioning Solution
Small Package: 28-Pin Double DIP
Internal Thin-Film Gain Network
High Accuracy
  Low Input Offset Tempco: ±0.07µV/°C
  Low Gain Tempco: ±2ppm/°C
  Low Nonlinearity: ±0.005% max
  High CMR: 140dB min (60Hz, G=1000V/V)
Programmable Bridge Excitation: +4V to +15V
Remote Sensing
Low Pass Filter (f_C=4Hz)

1B32 BLOCK DIAGRAM

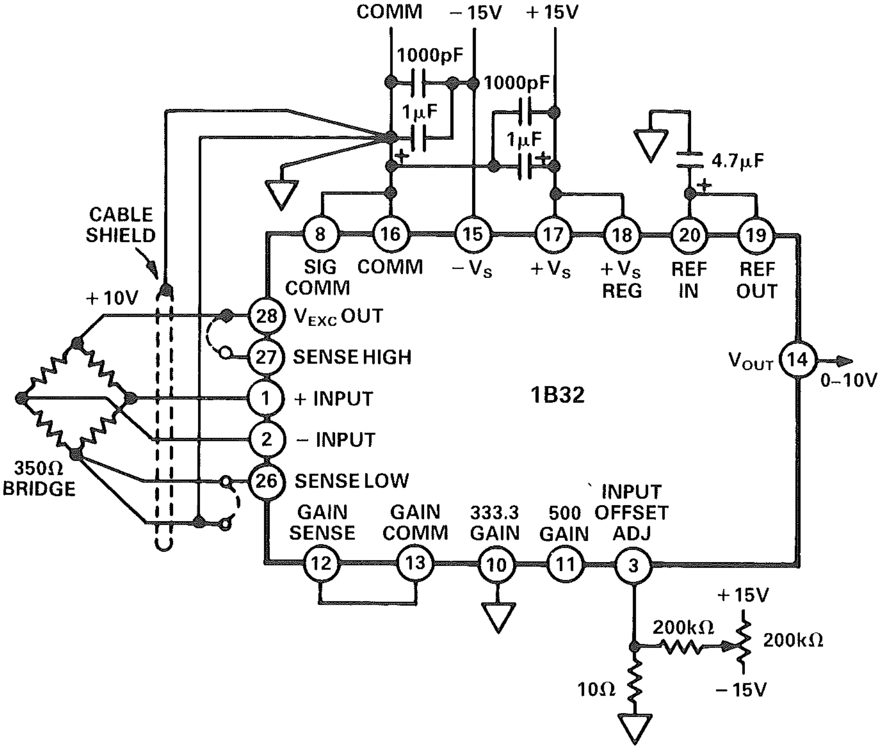


The 1B32 is based on a switched capacitor, chopper stabilized amplifier followed by an active filter and an adjustable voltage regulator section for excitation. The ultralow drift chopper samples the difference between the +INPUT and −INPUT at 190Hz. The signal is modulated, amplified and then demodulated. This stage introduces a pole with a 20dB/decade rolloff from 4Hz. The high-level signal is then filtered by a two-pole active filter with a 4Hz cutoff frequency to give a ±10V output. The clock signal for the chopper is generated by an on-board oscillator.

The gain of the 1B32 can be pin-strapped by an internal resistor network. Standard gains of 333.3 and 500 can be achieved by this method with gain tempco of ±6ppm/°C max. Finally, the offset adjust of the amplifier is input referred, and requires a voltage input similar to the differential input voltage to implement wide range suppression.

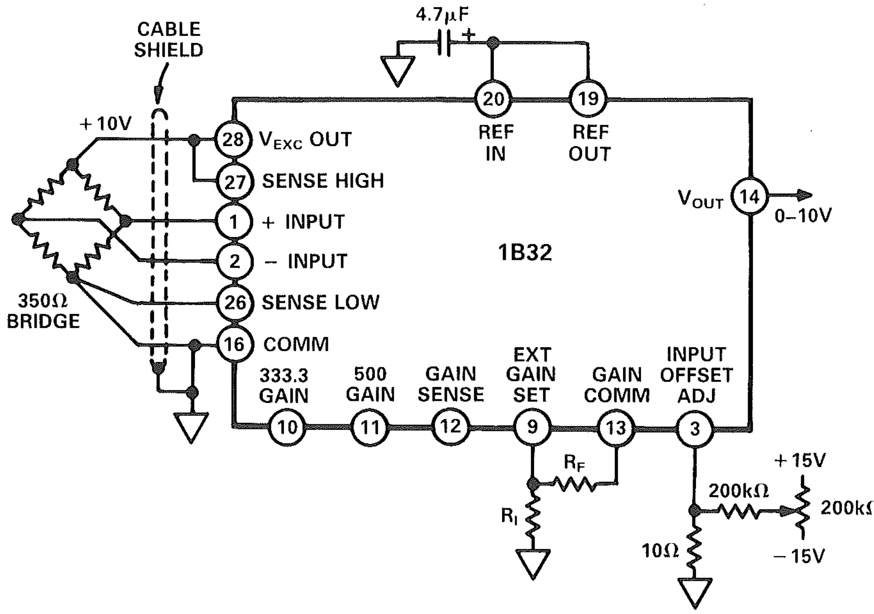
The adjustable bridge excitation section is essentially identical to that of the 1B31.

INTERNAL GAIN STRAPPING



The figure above shows a typical circuit configuration for the 1B32. As previously stated, the differential gain of the 1B32 can be either pin-strapped or programmed externally with two resistors. The internal thin-film network provides gain of 500 and 333.3 for standard load cell sensitivities of 2mV/V and 3mV/V. As shown above, this is achieved by connecting GAIN SENSE (pin 12) to GAIN COMM (pin 13) and grounding either pin 10 or pin 12. The gain tempco using the internal network is an excellent ±2ppm/°C typ (±6ppm/°C max).

EXTERNAL GAIN STRAPPING



To program gain externally, two resistors are connected as shown in the figure above. The gain equation is:

$$G = 1 + R_F/R_I$$

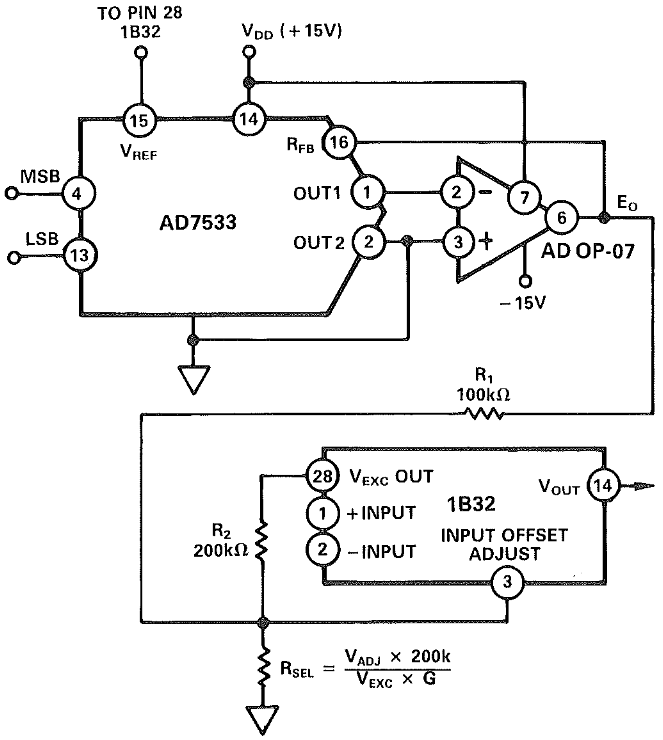
The gain strapping pins (10 and 11) and GAIN SENSE (pin 12) are left unconnected, effectively floating the internal network.

The input-referred offset adjust has the same sensitivity as the inputs of the 1B32. The voltage level at INPUT OFFSET ADJ (pin 3) is gained by the same factor as the input signal to provide a ±10V output adjust. The figures above show an external network and potentiometer set up for a ±7.5mV span at the input, which will give a ±2.5V output adjust capability when gain is set at 333.3 (7.5mV × 333.3 = 2.5V). Wider ranges can be chosen with the appropriate resistor and potentiometer values. If offset adjustment is not required, pin 3 must be grounded.

Voltage excitation programming for the 1B32 is done in the same manner as that of the 1B31.

Digital Output Offset Adjustment
A 10-bit multiplying DAC such as the AD7533 can be used to control the output offset of the 1B32 as shown below. The DAC is configured for unipolar operation with an AD OP-07 generating a voltage output. This 0-10V output is attenuated by R1 and R_{SEL} and superimposed on another fixed voltage derived from V_{EXC}. Thus the voltage at pin 3 (INPUT OFFSET ADJUST) is insensitive to the tempco of the excitation voltage since it is also used as the reference of the DAC. For best performance R1 and R2 should track to ±5ppm/°C. As an example, a ±5V output adjustment can be obtained by using $R_{SEL} = 200\Omega$ for $G = 500$ and $V_{EXC} = 10V$.

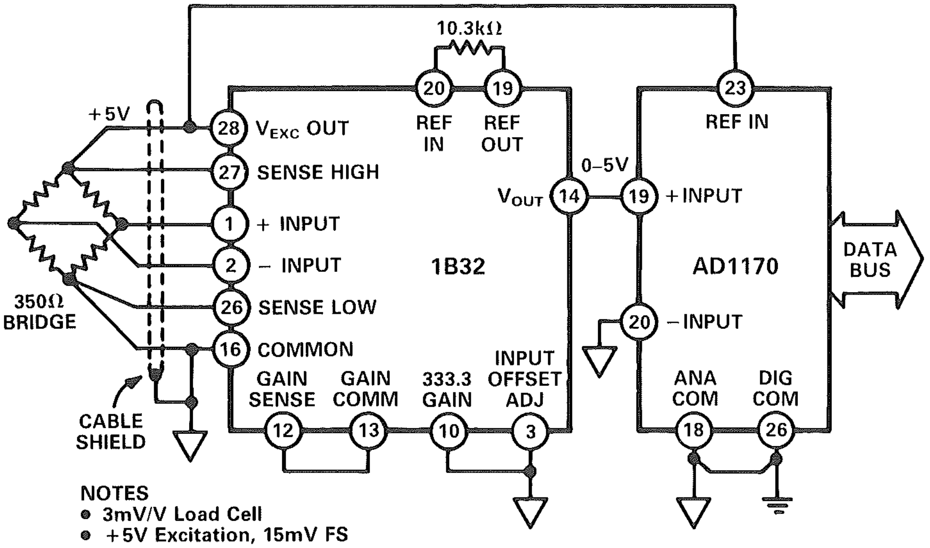
OUTPUT OFFSET ADJUST USING A 10-BIT DAC



Pressure Transducer Data Acquisition System

A two device solution for microcomputer based data acquisition using a 1B32 and an AD1170 18-bit A/D converter is shown below. A 3mV/V pressure transducer (e.g. Dynisco 800 series) is interfaced to a 1B32 configured with a gain of 333.3, to provide a 0 to 5V output. The regulated excitation is +5V, and is used as the reference input for the AD1170 to produce ratiometric operation. This configuration yields very high CMR enhanced by the 1B32 low pass filter and the integrating conversion scheme of the AD1170.

AUTO-CALIBRATING DATA ACQUISITION USING 1B32 AND AD1170



NOTES
- 3mV/V Load Cell
- +5V Excitation, 15mV FS
- Gain = 333.3

In addition, fixed offsets caused by bridge imbalance can be nulled out by the AD1170 with a power-up initialization command from the microcomputer. The full-scale output of the 1B32 and transducer can be normalized to the AD1170 full scale through the electronic calibration command ECAL. Both the offset and full scale correction data will then be stored nonvolatile memory to eliminate the need for the trim process after each power-up. The AD1170 eliminates a potentiometer or software overhead which might otherwise be needed for these functions.