Chapter VI

IN-AMP AND DIFF AMP APPLICATIONS CIRCUITS

Composite In-Amp Circuit Has Excellent High Frequency CMR

The primary benefit of an in-amp circuit is that it provides common-mode rejection. While the AD8221 and AD8225 both have an extended CMR frequency range, most in-amps fail to provide decent CMR at frequencies above the audio range.

The circuit in Figure 6-1 is a composite instrumentation amplifier with a high common-mode rejection ratio. It features an extended frequency range over which the instrumentation amplifier has good common-mode rejection (Figure 6-2). The circuit consists of three instrumentation amplifiers. Two of these, U1 and U2, are correlated to one another and connected in antiphase. It is not necessary to match these devices because they are correlated by design. Their outputs, OUT1 and OUT2, drive a third instrumentation amplifier that rejects common-mode signals and amplifiers' differential signals. The overall gain of the system can be determined by adding external resistors. Without any external resistors, the system gain is 2 (Figure 6-3). The performance of the circuit with a gain of 100 is shown in Figure 6-4.

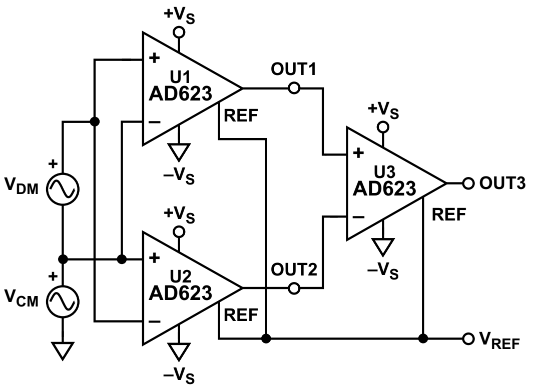


Figure 6-1. A Composite Instrumentation Amplifier

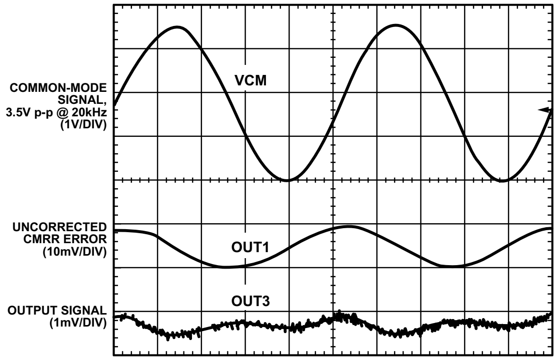


Figure 6-2. CMR of the Circuit in Figure 6-1 at 20 kHz

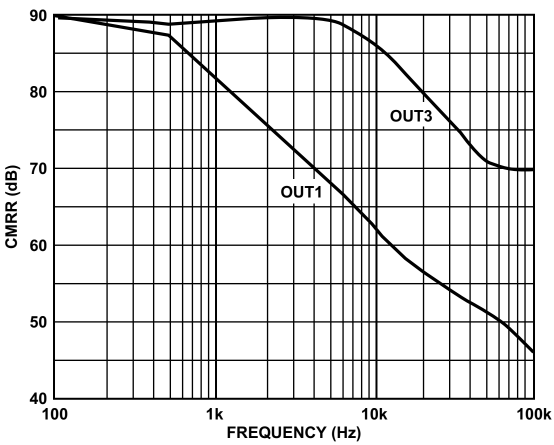


Figure 6-3. CMRR vs. Frequency at a Gain of 2

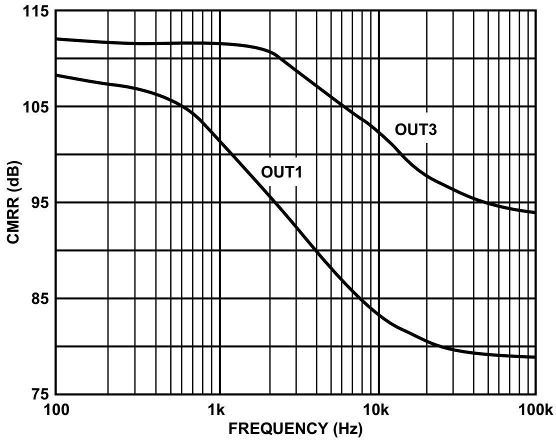


Figure 6-4. CMR of the System at a Gain of 100

Since U1 and U2 are correlated, their common-mode errors are the same. Therefore, these errors appear as common-mode input signal to U3, which rejects them. In fact, if it is necessary, OUT1 and OUT2 can directly drive an analog-to-digital converter (ADC). The differential-input stage of the ADC will reject the common-mode signal, as seen in Figure 6-5.

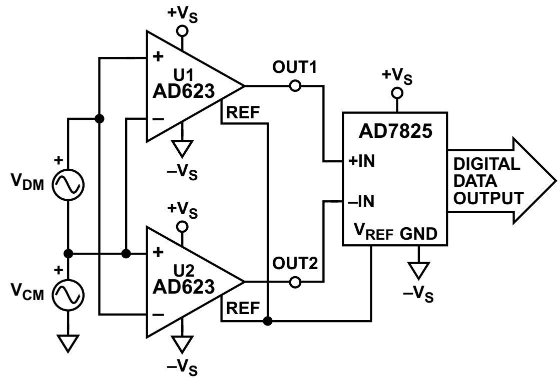


Figure 6-5. The OUT1 and OUT2 signals of the first stage can directly drive an analog-to-digital converter, allowing the ADC to reject the common-mode signal.

STRAIN GAGE MEASUREMENT USING AN AC EXCITATION

Strain gage measurements are often plagued by offset drift, 1/f noise, and line noise. One solution is to use an ac signal to excite the bridge, as shown in Figure 6-6. The AD8221 gains the signal and an **AD630AR** synchronously demodulates the waveform. What results is a dc output proportional to the strain on the bridge. The output signal is devoid of all dc errors associated with the in-amp and the detector, including offset and offset drift.

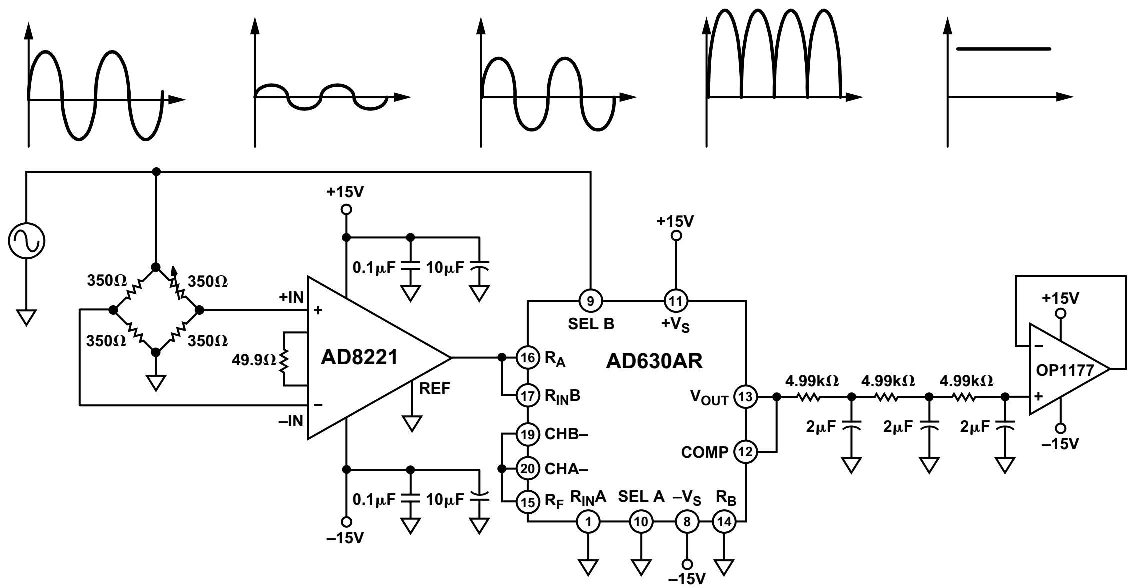


Figure 6-6. Using an AC Signal to Excite the Bridge

In Figure 6-6, a 400 Hz signal excites the bridge. The signal at the AD8221's input is an ac voltage. Similarly, the signal at the input of the AD630 is ac; the signal is dc at the end of the low-pass filter following the AD630.

The 400 Hz ac signal is rectified and then averaged; dc errors are converted in an ac signal and removed by the AD630. Ultimately, a precision dc signal is obtained.

The AD8221 is well suited for this application because its high CMRR over frequency ensures that the signal of interest, which appears as a small difference voltage riding on a large sinusoidal common-mode voltage, is gained and the common-mode signal is rejected. In typical instrumentation amplifiers, CMRR falls off at about 200 Hz. In contrast, the AD8221 continues to reject common-mode signals beyond 10 kHz.

If an ac source is not available, a commutating voltage may be constructed using switches. The AD8221's high CMRR over frequency rejects high frequency harmonics from a commutating voltage source.

APPLICATIONS OF THE AD628 PRECISION GAIN BLOCK

The AD628 can be operated as either a differential/scaling amplifier or as a pin-strapped precision gain block. Specifically designed for use ahead of an analog-to-digital converter, the AD628 is extremely useful as an input scaling and buffering amplifier. As a differential amplifier, it can extract small differential voltages riding on large common-mode voltages up to ±120 V. As a prepackaged precision gain block, the pins of the AD628 can be strapped to provide a wide range of precision gains, allowing for high accuracy data acquisition with very little gain or offset drift.

The AD628 uses an absolute minimum of external components. Its tiny MSOP provides these functions in the smallest size package available on the market. Besides high gain accuracy and low drift, the AD628 provides a very high common-mode rejection, typically more than 90 dB at 1 kHz while still maintaining a 60 dB CMRR at 100 kHz.

The AD628 includes a V_{REF} pin to allow a dc (midscale) offset for driving single-supply ADCs. In this case, the V_{REF} pin may simply be tied to the ADC's reference pin, which also allows easy ratiometric operation.

Why Use a Gain Block IC?

Real-world measurement requires extracting weak signals from noisy sources. Even when a differential measurement is made, high common-mode voltages are often present. The usual solution is to use an op amp or, better still, an in-amp, and then perform some type of low-pass filtering to reduce the background noise level.

The problem with this traditional approach is that a discrete op amp circuit will have poor common-mode rejection and its input voltage range will always be less than the power supply voltage. When used with a differential signal source, an in-amp circuit using a monolithic IC will improve common-mode rejection. However, signal sources greater than the power supply voltage, or signals riding on high common-mode voltages, cannot handle

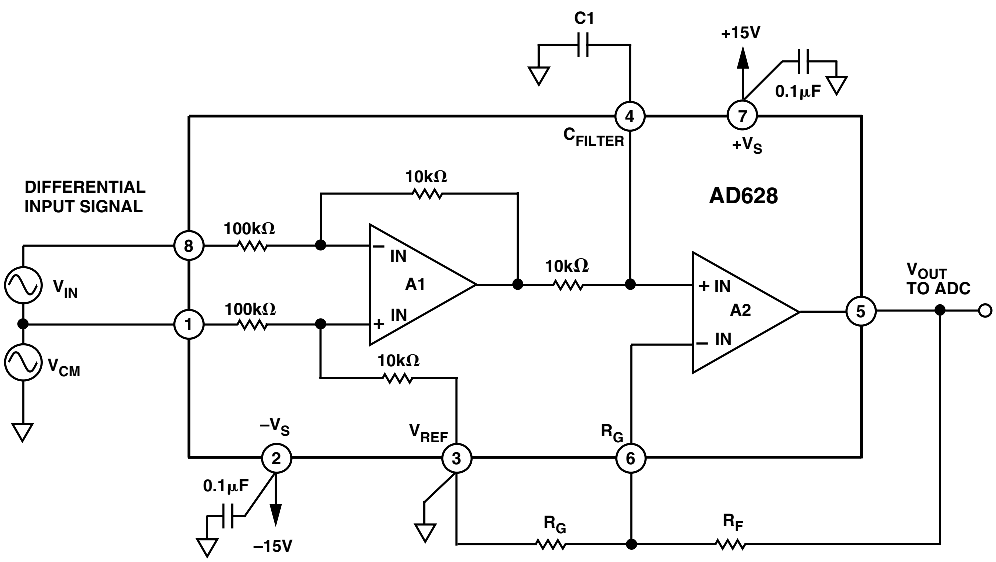


Figure 6-7. Basic Differential Input Connection with Single-Pole LP Filter

standard in-amps. In addition, in-amps using a single external gain resistor suffer from gain drift. Finally, low-pass filtering usually requires the addition of a separate op amp, along with several external components. This drains valuable board space.

The **AD628** eliminates these common problems by functioning as a scaling amplifier between the sensor, the shunt resistor, or another point of data acquisition, as well as the ADC. Its 120 V max input range permits the direct measurement of large signals or small signals riding on large common-mode voltages.

Standard Differential Input ADC Buffer Circuit with Single-Pole LP Filter

Figure 6-7 shows the AD628 connected to accept a differential input signal riding on a very high common-mode voltage. The AD628 gain block has two internal amplifiers: A1 and A2. Pin 3 is grounded, thus operating amplifier A1 at a gain of 0.1. The 100 kΩ input resistors and other aspects of its design allow the AD628 to process small input signals riding on common-mode voltages up to ±120 V.

The output of A1 connects to the plus input of amplifier A2 through a 10 kΩ resistor. Pin 4 allows connecting an external capacitor to this point, providing single-pole low-pass filtering.

Changing the Output Scale Factor

Figure 6-7 reveals that the output scale factor of the AD628 may be set by changing the gain of amplifier A2. This uncommitted op amp may be operated at any convenient gain higher than unity. When configured, the AD628 may be set to provide circuit gains between 0.1 and 1,000.

Since the gain of A1 is 0.1, the combined gain of A1 and A2 equals

$$\frac{V_{OUT}}{V_{IN}} = G = 0.1\left(1+\left(R_F/R_G\right)\right)$$

Therefore

$$(10G-1)=\frac{R_F}{R_G}$$

For ADC buffering applications, the gain of A2 should be chosen so that the voltage driving the ADC is close to its full-scale input range. The use of external resistors, R_F and R_G to set the output scale factor (i.e., gain of A2) will degrade gain accuracy and drift essentially to the resistors themselves.

A separate V_{REF} pin is available for offsetting the AD628 output signal, so it is centered in the middle of the ADC's input range. Although Figure 6-7 indicates ±15 V, the circuit may be operated from ±2.25 V to ±18 V dual supplies. This V_{REF} pin may also be used to allow single-supply operation; V_{REF} may simply be biased at $V_S/2$.

Using an External Resistor to Operate the AD628 at Gains Below 0.1

The AD628 gain block may be modified to provide any desired gain from 0.01 to 0.1, as shown in Figure 6-8.

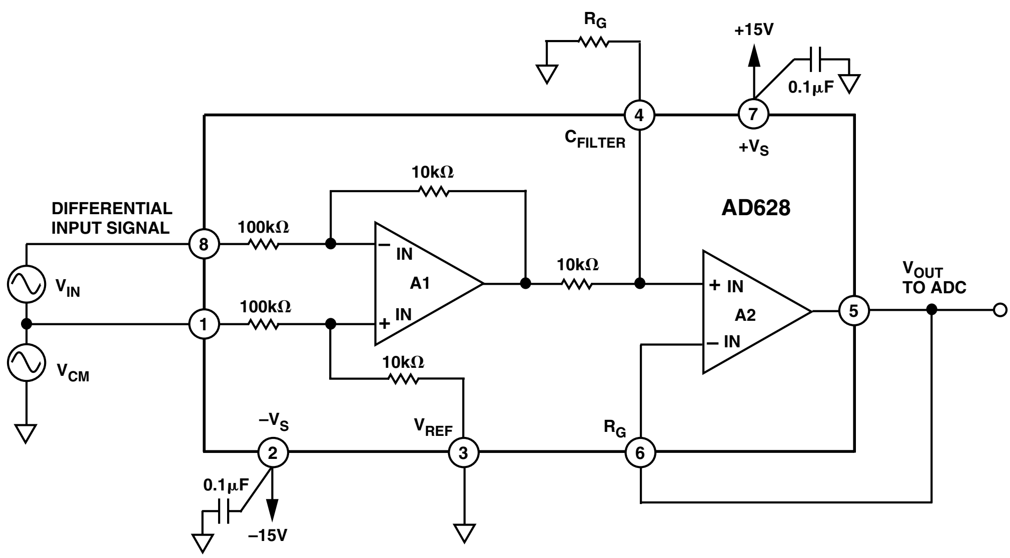


Figure 6-8. AD628 Connection for Gains Less Than 0.1

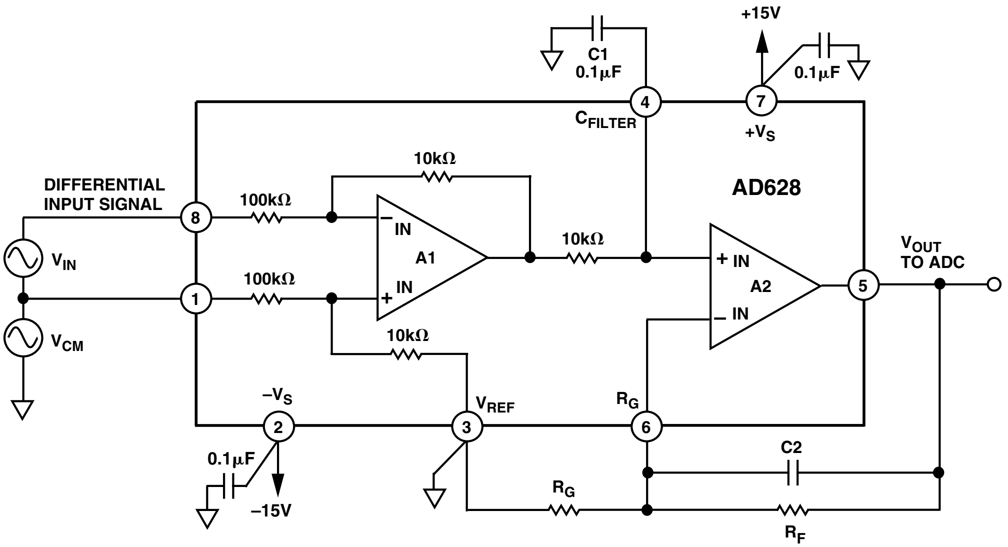


Figure 6-9. Differential Input Circuit with Two-Pole Low-Pass Filtering

This connection is the same as the basic wide input range circuit of Figure 6-7, but with Pins 5 and 6 strapped, and with an external resistor, R_G, connection between Pin 4 and ground. The pin strapping operates amplifier A2 at unity gain. Acting with the on-chip 10 kΩ resistor at the output of A1, R_{GAIN} forms a voltage divider that attenuates the signal between the output of A1 and the input of A2. The gain for this connection equals 0.1 V_{IN} ((10 kΩ + R_G)/R_G).

Differential Input Circuit with Two-Pole Low-Pass Filtering

The circuit in Figure 6-9 is a modification of the basic ADC interface circuit. Here, two-pole low-pass filtering is added for the price of one additional capacitor (C2).

As before, the first pole of the low-pass filter is set by the internal 10 kΩ resistor at the output of A1 and the external capacitor C1. The second pole is created by an external RC time constant in the feedback path of A2, consisting of capacitor C2 across resistor R_F. Note that this second pole provides a more rapid roll-off of frequencies above its RC *corner* frequency (1/(2πRC)) than a single-pole LP filter. However, as the input frequency is increased, the gain of amplifier A2 eventually drops to unity and will not be further reduced. So, amplifier A2 will have a voltage gain set by the ratio of R_F/R_G at frequencies below its –3 dB corner and have unity gain at higher frequencies.

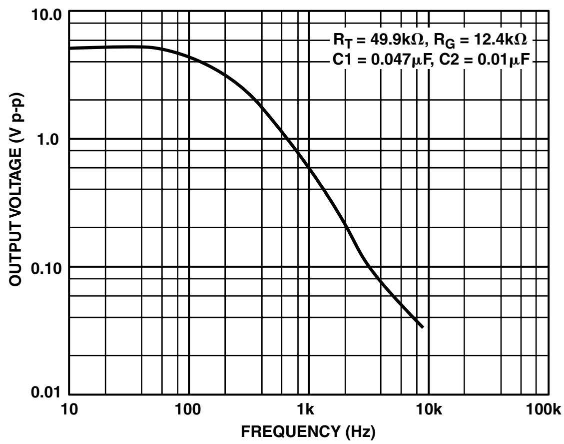


Figure 6-10. Frequency Response of the Two-Pole LP Filter

Figure 6-10 shows the filter's output vs. frequency using components chosen to provide a 200 Hz –3 dB corner frequency. There is a sharp roll-off between the corner frequency and approximately 10× the corner frequency. Above this point, the second pole starts to become less effective and the rate of attenuation is close to that of a single-pole response.

Table 6-1.

Two-Pole LP Filter Input Range: 10 V p-p F.S. for a 5 V p-p Output R_F = 49.9 kΩ, R_G = 12.4 kΩ				
	–3 dB Corner Frequency			
	200 Hz	**1 kHz**	**5 kHz**	**10 kHz**
Capacitor C2	0.01 μF	0.002 μF	390 pF	220 pF
Capacitor C1	0.047 μF	0.01 μF	0.002 μF	0.001 μF

Table 6-2.

Two-Pole LP Filter Input Range: 20 V p-p F.S. for a 5 V p-p Output R_F = 24.3 kΩ, R_G = 16.2 kΩ				
	–3 dB Corner Frequency			
	200 Hz	**1 kHz**	**5 kHz**	**10 kHz**
Capacitor C2	0.02 μF	0.0039 μF	820 pF	390 pF
Capacitor C1	0.047 μF	0.01 μF	0.002 μF	0.001 μF

Tables 6-1 and 6-2 provide typical filter component values for various –3 dB corner frequencies and two different full-scale input ranges. Values have been rounded off to match standard resistor and capacitor values. Capacitors C1 and C2 need to be high Q, low drift devices; low grade disc ceramics should be avoided. High quality NPO ceramic, Mylar, or polyester film capacitors are recommended for the lowest drift and best settling time.

Using the AD628 to Create Precision Gain Blocks

Real-world data acquisition systems require amplifying weak signals enough to apply them to an ADC. Unfortunately, when configured as gain blocks, most common amplifiers have both gain errors and offset drift.

In op amp circuits, the usual two resistor gain setting arrangement has accuracy and drift limitations. Using standard 1% resistors, amplifier gain can be off by 2%. The gain will also vary with temperature because each resistor will drift differently. Monolithic resistor networks can be used for precise gain setting, but these components increase cost, complexity, and board space.

The gain block circuits of Figures 6-11 to 6-15 overcome all of these performance limitations, are very inexpensive, and offer a single MSOP solution. The AD628 provides this complete function using the smallest IC package available. Since all resistors are internal to the AD628 gain block, both accuracy and drift are excellent.

All of these pin-strapped circuits (using no external components) have a gain accuracy better than 0.2%, with a gain TC better than 50 ppm/°C.

Operating the AD628 as a +10 or –10 Precision Gain Block

Figure 6-11 shows an AD628 precision gain block IC connected to provide a voltage gain of +10. The gain block may be configured to provide different gains by strapping or grounding the appropriate pin. The gain block itself consists of two internal amplifiers: a gain of 0.1 difference amplifier (A1) followed by an uncommitted buffer amplifier (A2).

The input signal is applied between the V_{REF} pin (Pin 3) and ground. With the input tied to Pin 3, the voltage at the positive input of A1 equals V_{IN} (100 kΩ/110 kΩ) which is V_{IN} (10/11). With Pin 6 grounded, the minus

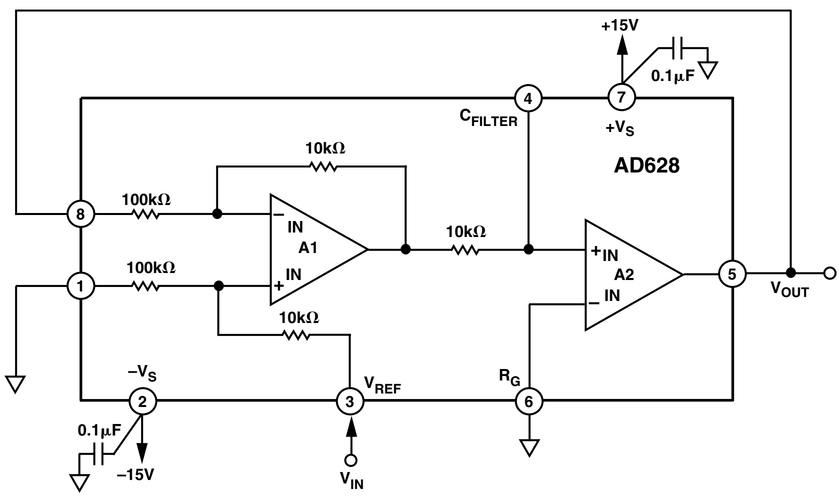


Figure 6-11. Circuit with a Gain of +10 Using No External Components

input of A2 equals 0 V. Therefore, the positive input of A2 will be forced by feedback from the output of A2 to be 0 V as well. The output of A1 then must also be at 0 V. Since the negative input of A1 must be equal to the positive input of A1, both will equal V_{IN} (10/11).

This means that the output voltage of A2 (V_{OUT}) will equal

$$V_{OUT} = V_{IN}(10/11)(1+100k/10k) =$$
$$V_{IN}(10/11)\,11 = 10\ V_{IN}$$

The companion circuit in Figure 6-12 provides a gain of –10. This time the input is applied between the negative input of A2 (Pin 6) and ground. Operation is exactly the same, but now the input signal is inverted 180° by A2. With Pin 3 grounded, the positive input of A1 is at 0 V, so feedback will force the negative input of A1 to zero as well. Since A1 operates at a gain of 1/10 (0.1), the output of A2 that is needed to force the negative input of A1 to zero is *minus* 10 V_{IN}.

The two connections will have different input impedances. When driving Pin 3 (Figure 6-11), the input impedance to ground is 110 kΩ, while it is approximately 50 GΩ when driving Pin 6 (Figure 6-12). The –3 dB bandwidth for both circuits is approximately 110 kHz for 10 mV and 95 kHz for 100 mV input signals.

Operating the AD628 at a Precision Gain of +11

The gain of +11 circuit (Figure 6-13) is almost identical to the gain of +10 connection, except that Pin 1 is strapped to Pin 3, rather than being grounded. This

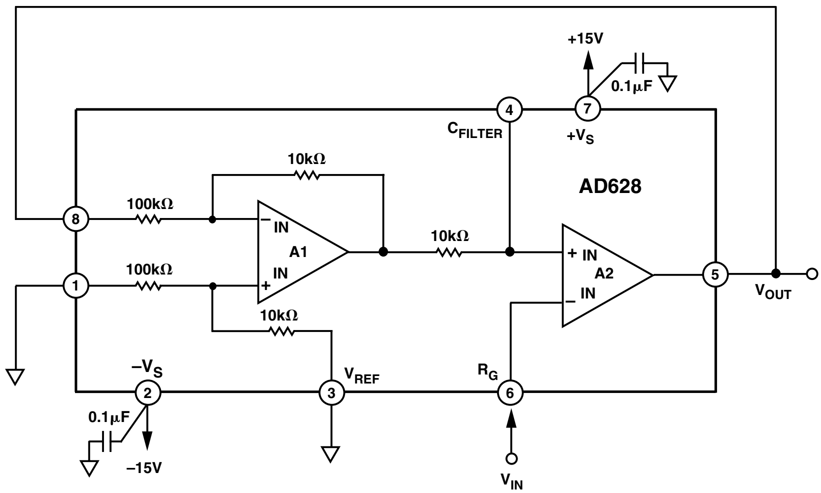


Figure 6-12. Companion Circuit Providing a Gain of –10

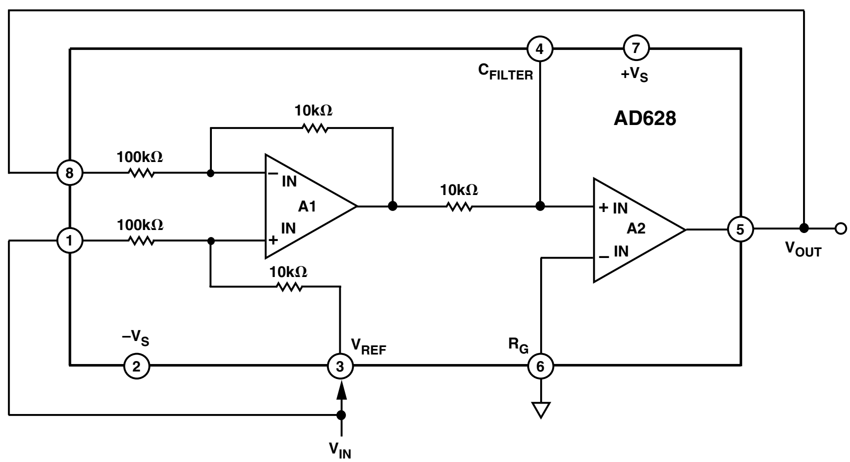


Figure 6-13. A Gain of +11 Circuit

connects the two internal resistors (100 kΩ and 10 kΩ) that are tied in parallel to the plus input of A1. So, this now removes the 10 kΩ/110 kΩ voltage divider between V_{IN} and the positive input of A1. Thus modified, V_{IN} drives the positive input through approximately a 9 kΩ resistor. Note that this series resistance is negligible compared to the very high input impedance of amplifier A1. The gain from Pin 8 to the output of A1 is 0.1. Therefore, feedback will force the output of A2 to equal 10 V_{IN}. The –3 dB bandwidth of this circuit is approximately 105 kHz for 10 mV and 95 kHz for 100 mV input signals.

Operating the AD628 at a Precision Gain of +1

Figure 6-14 shows the AD628 connected to provide a precision gain of +1. As before, this connection uses the gain block's internal resistor networks for high gain accuracy and stability.

The input signal is applied between the V_{REF} pin and ground. Because Pins 1 and 8 are grounded, the input signal runs through a 100 kΩ/110 kΩ input attenuator to the plus input of A1. The voltage equals V_{IN} (10/11) = 0.909 V_{IN}. The gain from this point to the output of A1 will equal 1 + (10 kΩ/100 kΩ) = 1.10. Therefore, the voltage at the output of A1 will equal V_{IN} (1.10) (0.909) = 1.00. Amplifier A2 is operated as a unity gain buffer (as Pins 5 and 6 tied together), providing an overall circuit gain of +1.

Increased BW Gain Block of –9.91 Using Feedforward

The circuit of Figure 6-12 can be modified slightly by applying a small amount of positive feedback to increase its bandwidth, as shown in Figure 6-15. The output of amplifier A1 feeds back its positive input by connecting Pin 4 and Pin 1 together. Now, Gain = –(10 – 1/11) = –9.91.

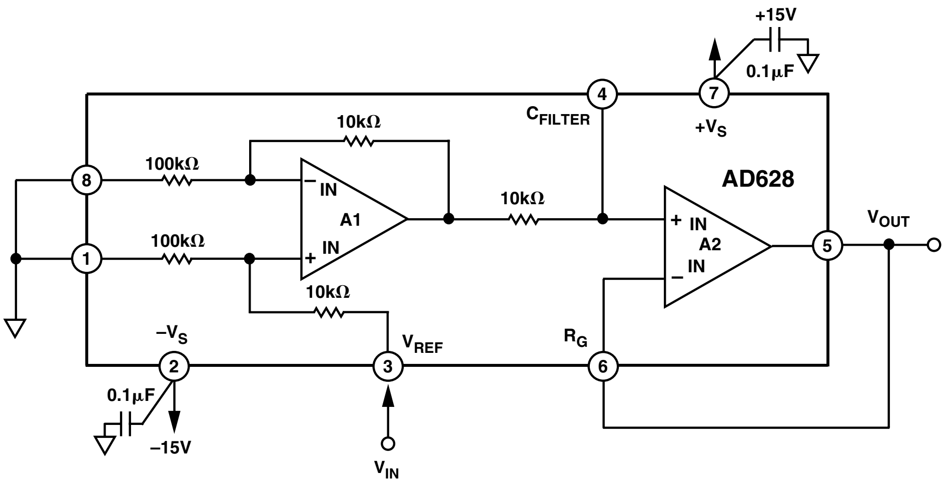


Figure 6-14. AD628 Precision Gain of +1

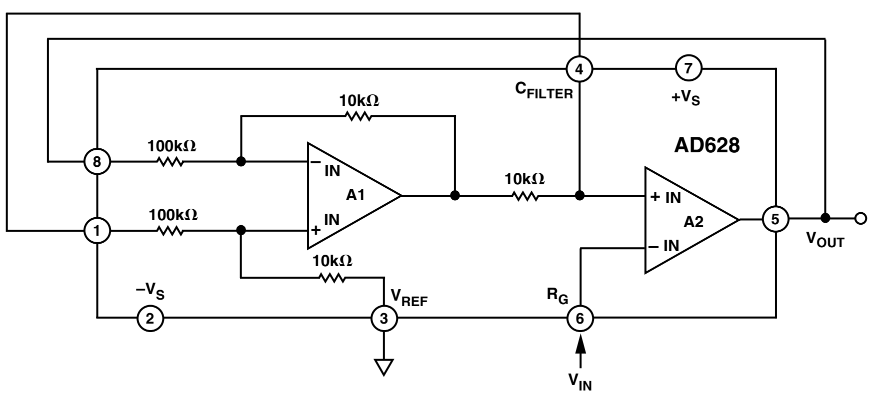


Figure 6-15. Precision –10 Gain Block with Feedforward

The resulting circuit is still stable because of the large amount of negative feedback applied around the entire circuit (from the output of A2 back to the negative input of A1). This connection actually results in a small signal –3 dB bandwidth of approximately 140 kHz. This is a 27% increase in bandwidth over the unmodified circuit in Figure 6-9. However, gain accuracy is reduced to ±2%.

CURRENT TRANSMITTER REJECTS GROUND NOISE

Many systems use current flow to control remote instrumentations. The advantage of such a system is its ability to operate with two remotely connected power supplies, even if their grounds are not the same. In such cases, it is necessary for the output to be linear with respect to the input signal, and any interference between the grounds must be rejected. Figure 6-16 shows such a circuit.

For this circuit

$$I_{OUT} = \frac{\left(V_{IN}/10\right)}{1\text{k}\Omega}$$

$$I_{OUT} = \frac{V_{IN}(\text{V})}{1\text{k}\Omega}$$

The **AD629**, a difference amplifier with very high common-mode range, is driven by an input signal Pin 3. Its transfer function is

$$V_{OUT} = V_{IN}$$

Where:

V_{OUT} is measured between Pin 6 and its reference (Pin 1 and Pin 5), and the input V_{IN} is measured between Pin 3 and Pin 2. The common-mode signal, VCM, will be rejected.

In order to reduce the voltage at Pin 6, an inverter with a gain of 9 is connected between Pin 6 and its reference. The inverter sets the gain of the transmitter such that for a 10 V input, the voltage at Pin 6 only changes by 1 V; yet, the difference between Pin 6 and its reference is 10 V.

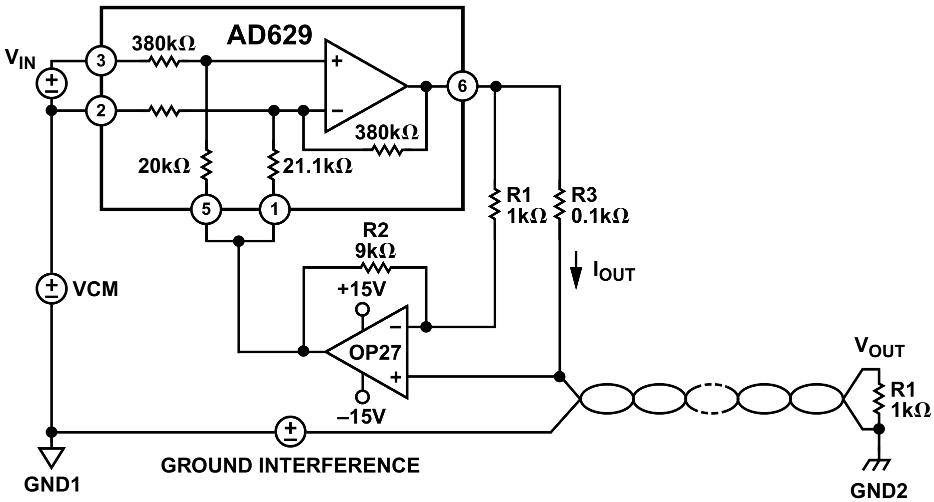


Figure 6-16. Current Transmitter

Since the gain between the noninverting terminal of the OP27 and the output of the AD629 is 1, no modulation of the output current will take place as a function of the output voltage V_{OUT}. The scaling resistor R3 is 100 Ω to make 1 mA/V of input signal.

OP27 was chosen because, at a noise gain of 10, its bandwidth does not compromise the transmitter. Figure 6-17 is the transfer function of the output voltage V_{OUT} vs. the input voltage V_{IN}. Figure 6-18 is a demonstration of how well the transmitter rejects ground noise.

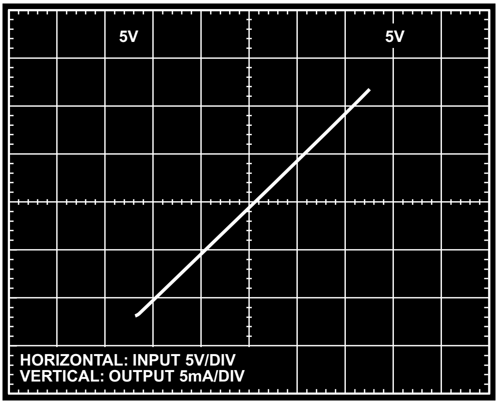


Figure 6-17. Transfer Function

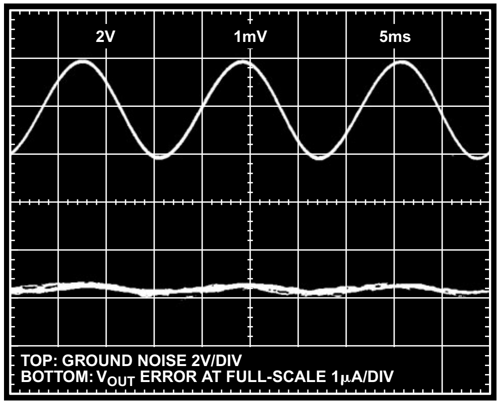


Figure 6-18. Interference Rejection

HIGH LEVEL ADC INTERFACE

The circuit of Figure 6-19 provides an interface between large level analog inputs as high as ±10 V operating on dual supplies and a low level, differential input ADC, operating on a single supply.

As shown, two AD628 difference amplifiers are connected in antiphase. The differential output, V_1–V_2, is an attenuated version of the input signal

$$V_1 - V_2 = \frac{(V_A - V_B)}{5}$$

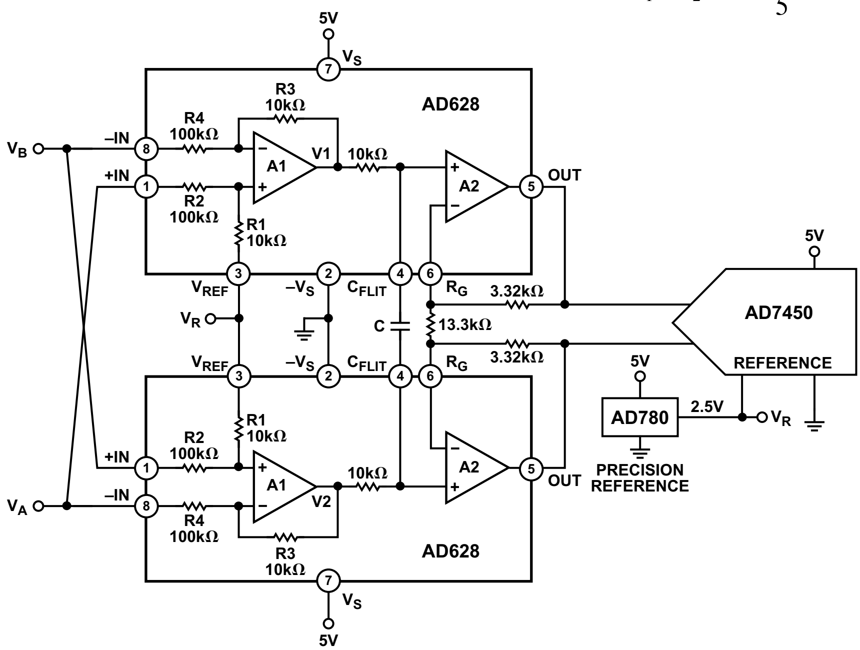


Figure 6-19. This ADC Interface Circuit Attenuates and Level Shifts a ±10 V Differential Signal While Operating from a Single 5 V Supply

The difference amplifiers reject the common-mode voltage on inputs V_A and V_B. The reference voltage, V_R, which the **AD780** develops and the ADC and the amplifier share, sets the output common-mode voltage. A single capacitor, C, placed across the C_{FILT} pins, low-pass filters the difference signal, $V_1 - V_2$. The –3 dB pole frequency is $f_P = 1/(40{,}000 \times \pi \times C)$. The difference signal is amplified by 1.5. Thus, the total gain of this circuit is 3/10.

Figure 6-20 shows a 10 V input signal (top), the signals at the output of each AD628 (middle), and the differential output (bottom). The benefits of this configuration go beyond simply interfacing with the ADC. The circuit improves specifications such as common-mode rejection ratio, offset voltage, drift, and noise by a factor of $\sqrt{2}$ because the errors of each AD628 are not correlated.

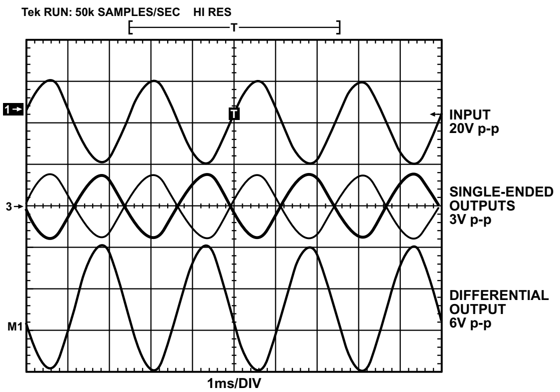


Figure 6-20. The Waveforms Show a 10 V Input Signal (top), the Signals at the Output of Each AD628 (middle), and the Differential Output (bottom)

The output demonstrates an 85 dB SNR (Figure 6-21). The two AD628s interface with an AD7450 12-bit, differential-input ADC. The AD7450 easily rejects residual common-mode signals at the output of the difference amplifiers. Figure 6-22 shows the common-mode error at the output of the AD628.

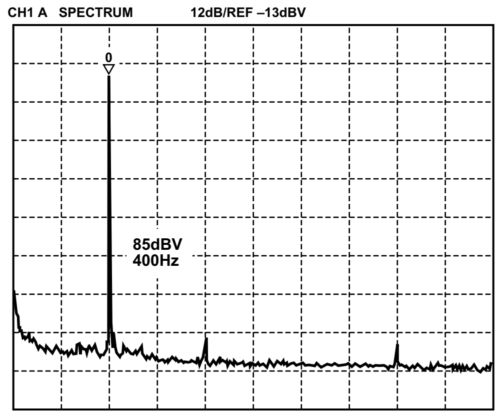


Figure 6-21. The Circuit in Figure 6-19 has an 85 dBV SNR

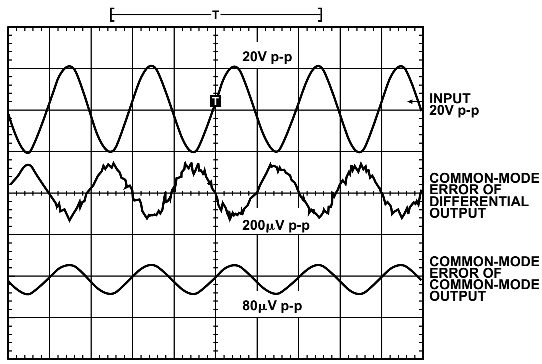


Figure 6-22. The Common-Mode Input (top) Measures 20 V p-p. The common-mode error of the differential output (middle) is 20 µV p-p. The error of the common-mode output (bottom) is 80 µV p-p.

The topmost waveform is a 10V, common-mode input signal. The middle waveform, measuring 150 µV, is the common-mode error measured differentially from the output of the two AD628s. The bottom waveform, measuring 80 µV, is the common-mode error that results.

A HIGH SPEED NONINVERTING SUMMING AMPLIFIER

The schematic in Figure 6-23 is that of a common summing amplifier with multiple inputs and one single-ended output. It is a variation of an inverting amplifier. Point X is a virtual ground and referred to as a summing junction. The transfer function for this circuit is

$$V_O = -\left[\left(RF/R1\right)V1 + \left(RF/R2\right)V2 + \left(RF/R3\right)V3\right]$$

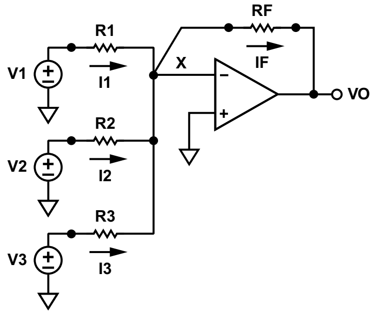


Figure 6-23. A Traditional Summing Amplifier

This indicates that the output is a weighted sum of the inputs with the weights being determined by the resistance ratio. If all resistances are equal, the circuit yields the inverted sum of its inputs.

$$V_O = -(V1 + V2 + V3)$$

Note that if we want the result V_O = (V1 + V2 + V3), we need an additional inverter with Gain = –1. Furthermore, this circuit has many disadvantages, such as low input impedance, different input impedance for positive and negative inputs, low bandwidth and highly matched resistors are needed.

Figure 6-24 is the schematic of a high speed summing amplifier, which can sum up as many as four input voltages without the need for an inverter to change the sign of the output. This could prove very useful in audio and video applications. The circuit contains three low cost high speed instrumentation amplifiers. The first two interface with input signals and their total sum is taken at the third amplifier's output with respect to ground. The inputs are very high impedance and the signal that appears at the network output is noninverting.

Figure 6-25 is the performance photo at 1 MHz. The top trace is the input signal for all four inputs. The middle trace is the sum of inputs V1 and V2. The bottom trace is the output of the system, which is the total sum of all four inputs.

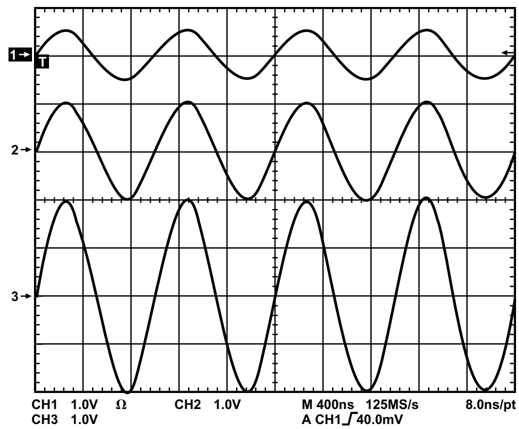


Figure 6-25. Performance Photo of the Circuit in Figure 6-24

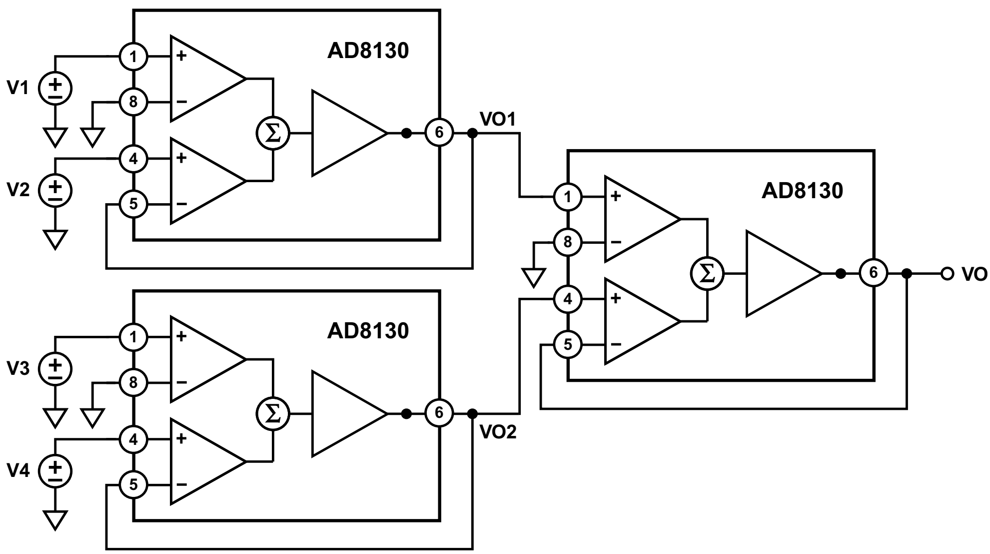


Figure 6-24. A Summing Circuit with High Input Impedance

Figure 6-26 demonstrates the high bandwidth of the system in Figure 6-24. As we can see, the –3 dB point is about 220 MHz.

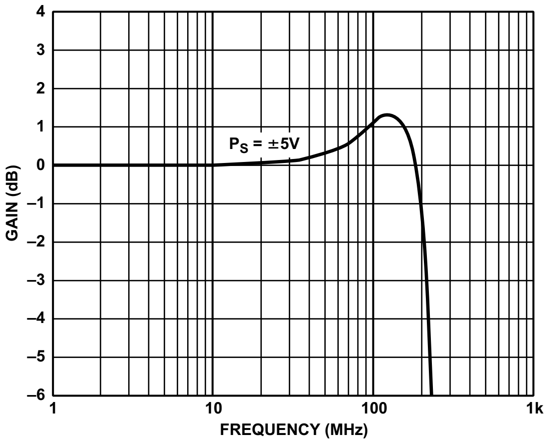


Figure 6-26. Frequency Response of Summing Circuit in Figure 6-24

HIGH VOLTAGE MONITOR

A high accuracy, high voltage monitor is shown in Figure 6-27.

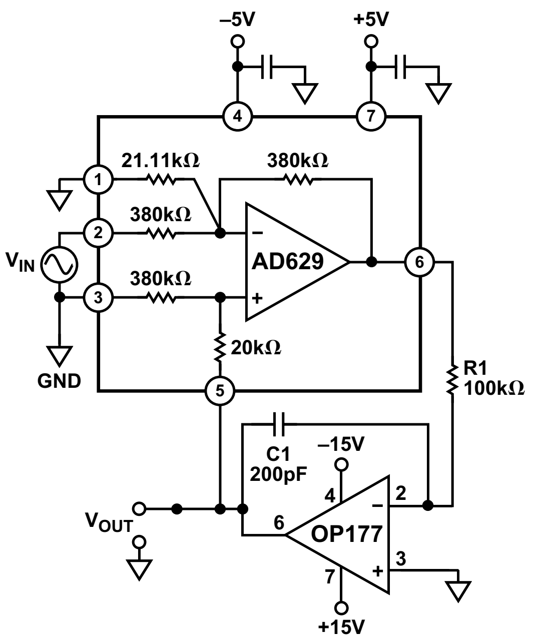


Figure 6-27. High Voltage Monitor

An integrator (OP177) supplies negative feedback around a difference amplifier (AD629), forcing its output to stay at 0 V. The voltage divider on the inverting input sets the common-mode voltage of the difference amplifier to $V_{IN}/20$. V_{OUT}, the integrator output and the measurement output, sources the required current to maintain the common-mode voltage. R1 and C1 compensate the system to a bandwidth of 200 kHz.

The transfer function is $V_{OUT} = V_{IN}/19$. For example, a 400 V p-p input signal will produce a 21 V p-p output.

Figure 6-28 shows that the measured system nonlinearity is less than 20 ppm over the entire 400 V p-p input range. System noise is about 550 nV/$\sqrt{\text{Hz}}$ referred to the input, or around 2 mV peak noise voltage (10 ppm of full scale) over a 300 kHz bandwidth.

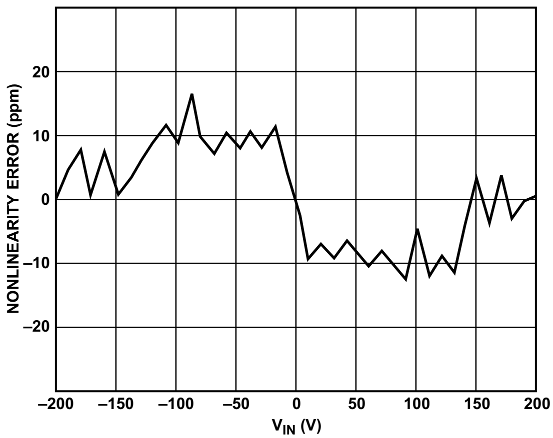


Figure 6-28. Nonlinearity vs. V_{IN}

HIGH COMMON-MODE REJECTION SINGLE-SUPPLY CIRCUIT

The circuit of Figure 6-29 can extract tiny signals riding on very large common-mode voltages and its single supply. Also, unless the converter is driven differentially, the noise on the analog-to-digital converter (ADC) reference pin is indistinguishable from a real signal.

The circuit of Figure 6-29 solves both of these problems. It provides a gain of 2, along with differential inputs and a differential output. The ADC reference sets the output common-mode level. The amplifier is constructed with two subtractors, each compliant to high common-mode voltage. These subtractors are set up so that the positive input of one connects to the negative input of the other, and vice versa. Their reference pins are tied together and connected to the ADC's reference pin.

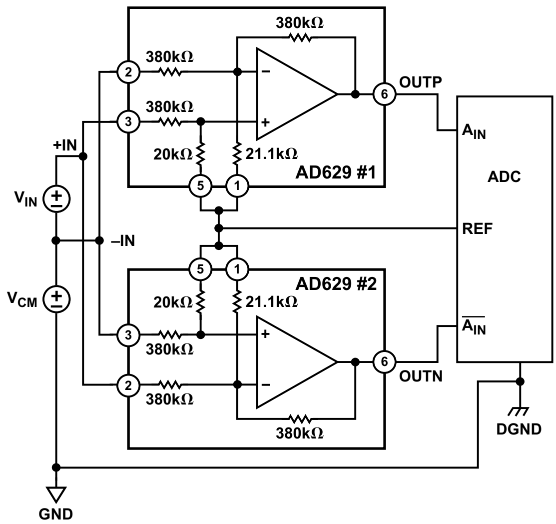


Figure 6-29. A High CMV Single-Supply Circuit

As the input signal increases, one output, OUTP, increases, while the other output, OUTN, decreases. Both outputs remain centered with respect to the common-mode level set by the ADC's reference.

Figure 6-30 illustrates the circuit's performance with a single 5 V power supply. At the top is a 1 kHz, 3 V p-p input signal. At the bottom are the two outputs in antiphase to produce a 3 V p-p signal centered around the 2.5 V reference.

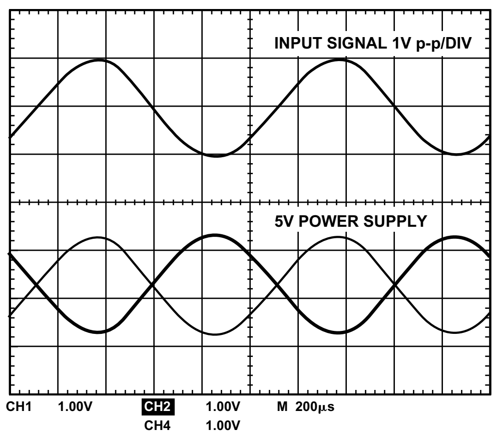


Figure 6-30. Top Trace is the Input Signal; Bottom Trace Is Antiphase Output, 40 V p-p on +2.5 dc

Figure 6-31 demonstrates the system's ability to reject a 1 kHz, 60 V p-p common-mode signal. The upper waveform shows the common-mode input, while the lower waveform shows the output.

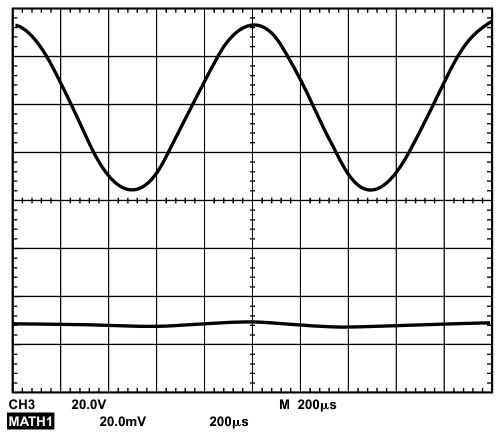


Figure 6-31. With a 5 V Supply and a 1 kHz, 60 V p-p Common-Mode Signal (Upper Trace), the Circuit's Output (Lower Trace) Illustrates the High Common-Mode Rejection

Bigger power supplies, such as ±15 V, can be used for larger common-mode signals. Figure 6-32 shows that the system can reject a 400 V p-p common-mode signal (upper waveform), with the residual error of less than 100 mV p-p shown in the lower waveform.

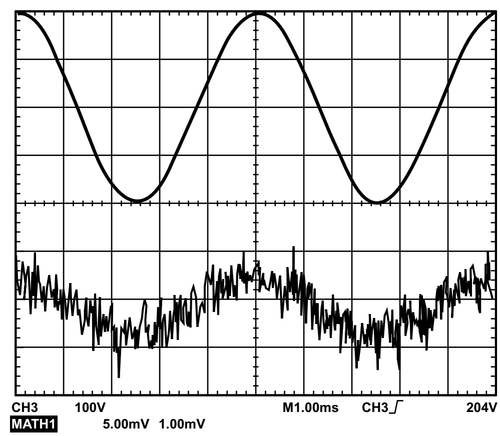


Figure 6-32. Using a ±15 V Supply, the Circuit Reduces a 400 V p-p Common-Mode Signal (Upper Trace) to Under 10 mV p-p (Lower Trace)

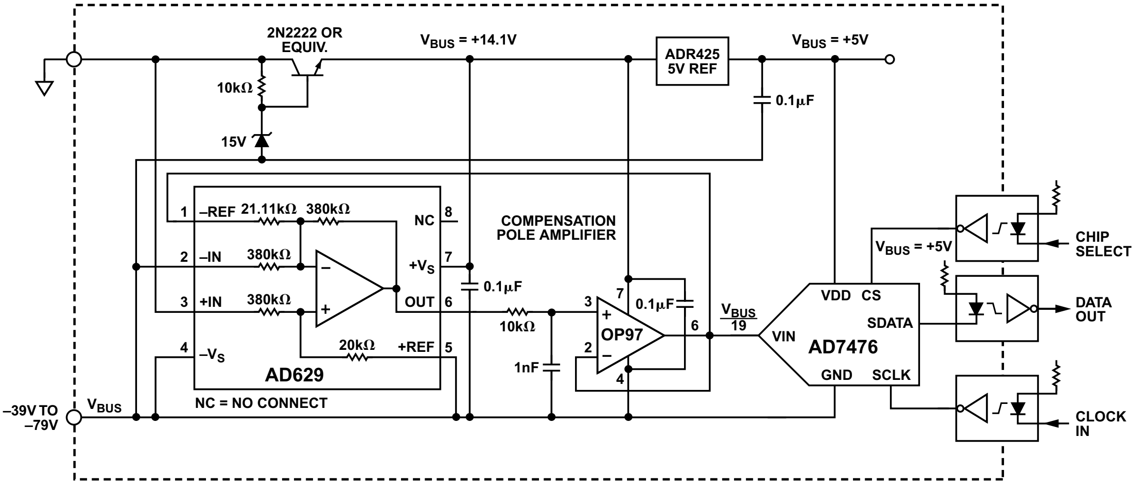


Figure 6-33. Precision Remote Voltage Measurement of –48 V Power Distribution Bus

PRECISION 48 V BUS MONITOR

Telephone equipment power supplies normally consist of a 48 V dc power source and an array of batteries. The batteries provide backup power during ac power line outages and help regulate the 48 V dc supply voltage.

Although nominally –48 V, the dc voltage on the telephone lines can vary anywhere from –40V to –80V and is subject to surges and fluctuations. Supply regulation at the source has little effect on remote voltage levels and equipment failures resulting from surges, brownouts, or other line faults, may not always be detected.

Capturing power supply information from remote communications equipment requires precise measurement of the voltages, sometimes under outdoor temperature conditions. High common-mode voltage difference amplifiers have been used to monitor current. However, these versatile components can also be used as voltage dividers, enabling remote monitoring of voltage levels as well.

Figure 6-33 shows a precision monitor using just two integrated circuits, which derives its power from the –48 V supply. A low cost transistor and Zener diode combination provide 15 V supply voltage for the amplifiers.

The AD629 IC is a self-contained high common-mode voltage difference amplifier. Connected as shown, it reduces the differential input voltage by approximately 19 V, thus acting as a precision voltage divider. An additional amplifier is required for loop stability.

The output from the OP-07 drives an AD7476 ADC.

The circuit features several advantages over alternative solutions. The AD629's laser trimmed divider resistors exhibit essentially perfect matching and tracking over temperature. Linearity errors from –40 V to –80 V are nearly immeasurable. Figures 6-34 and 6-35 are linearity and temperature drift curves for this circuit.

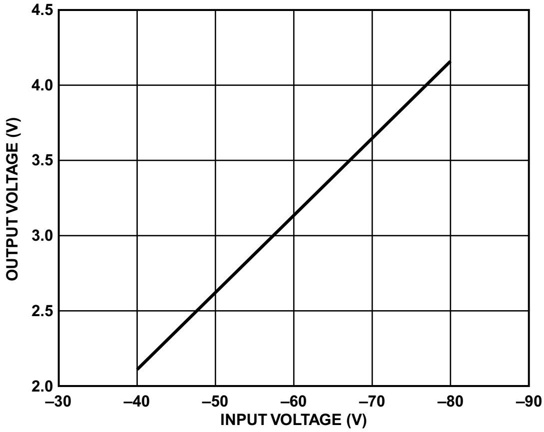


Figure 6-34. Output vs. Input Linearity for the Circuit of the 48 V Bus Monitor

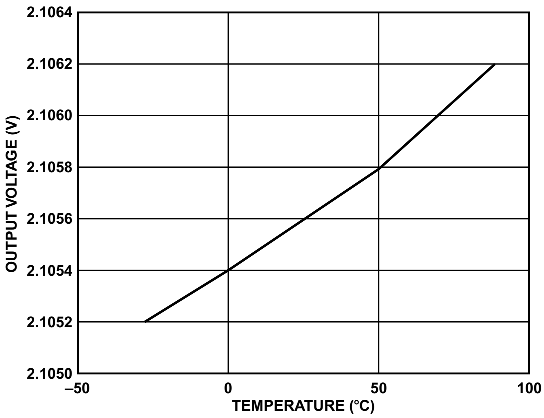


Figure 6-35. Temperature Drift of the 48 V Bus Monitor

HIGH-SIDE CURRENT SENSE WITH A LOW-SIDE SWITCH

A typical application for the AD8205 is high-side measurement of a current through a solenoid for PWM control of the solenoid opening. Typical applications include hydraulic transmission control and diesel injection control.

Two typical circuit configurations are used for this type of application.

In this case, the PWM control switch is ground referenced. An inductive load (solenoid) is tied to a power supply. A resistive shunt is placed between the switch and the load (see Figure 6-36). An advantage of placing the shunt on the high side is that the entire current, including the recirculation current, can be measured since the shunt remains in the loop when the switch is off. In addition, diagnostics can be enhanced because shorts to ground can be detected with the shunt on the high side.

In this circuit configuration, when the switch is closed, the common-mode voltage moves down to near the negative rail. When the switch is opened, the voltage reversal across the inductive load causes the common-mode voltage to be held one diode drop above the battery by the clamp diode.

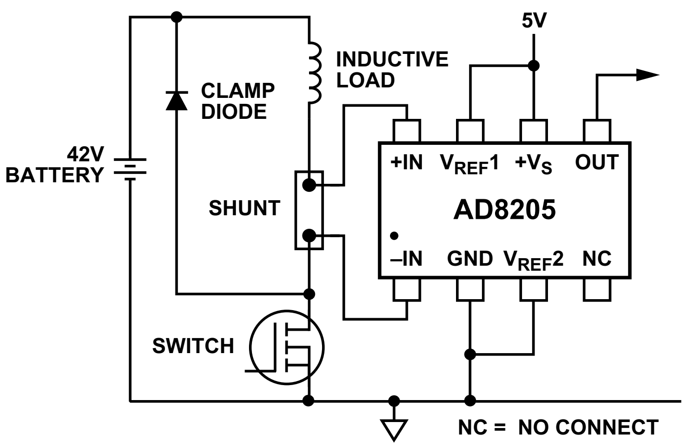


Figure 6-36. Low-Side Switch

HIGH-SIDE CURRENT SENSE WITH A HIGH-SIDE SWITCH

This configuration minimizes the possibility of unexpected solenoid activation and excessive corrosion (see Figure 6-37). In this case, both the switch and the shunt are on the high side. When the switch is off, this removes the battery from the load, which prevents damage from potential shorts to ground, while still allowing the recirculating current to be measured and providing for diagnostics. Removing the power supply from the load for the majority of the time minimizes the corrosive effects that could be caused by the differential voltage between the load and ground.

When using a high-side switch, the battery voltage is connected to the load when the switch is closed, causing the common-mode voltage to increase to the battery voltage. In this case, when the switch is opened, the voltage reversal across the inductive load causes the common-mode voltage to be held one diode drop below ground by the clamp diode.

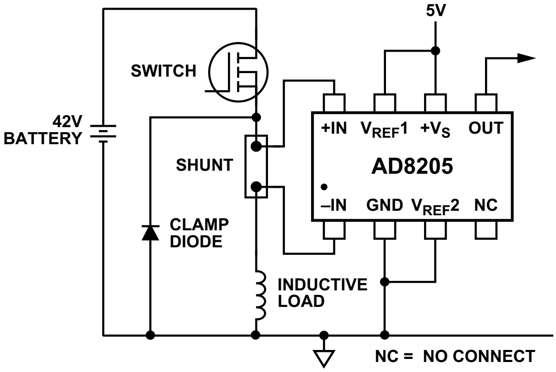


Figure 6-37. High-Side Switch

Another typical application for the AD8205 is as part of the control loop in H-bridge motor control. In this case, the AD8205 is placed in the middle of the H-bridge (see Figure 6-38) so that it can accurately measure current in both directions by using the shunt available at the motor. This is a better solution than a ground referenced op amp because ground is not typically a stable reference voltage in this type of application. This instability in the ground reference causes the measurements that could be made with a simple ground referenced op amp to be inaccurate.

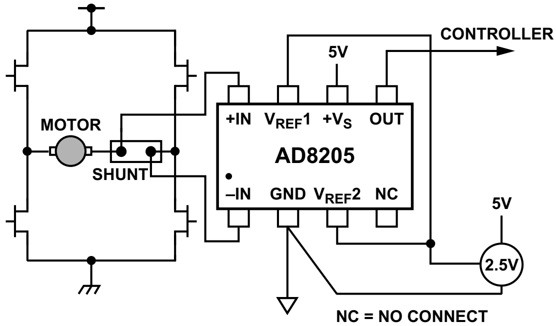


Figure 6-38. Motor Control Application

The AD8205 measures current in both directions as the H-bridge switches and the motor changes direction. The output of the AD8205 is configured in an external reference bidirectional mode.

BRIDGE APPLICATIONS

Instrumentation amplifiers are widely used for buffering and amplifying the small voltage output from transducers that make use of the classic four resistor Wheatstone bridge.

A Classic Bridge Circuit

Figure 6-39 shows the AD627 configured to amplify the signal from a classic resistive bridge. This circuit will work in either dual- or single-supply mode. Typically, the bridge will be excited by the same voltage used to power the in-amp. Connecting the bottom of the bridge to the negative supply of the in-amp (usually either 0, –5 V, –12 V, or –15 V) sets up an input common-mode voltage that is optimally located midway between the supply voltages. It is also appropriate to set the voltage on the

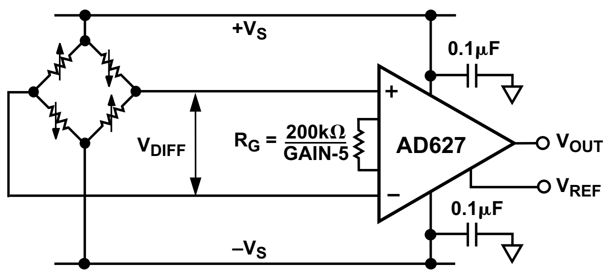


Figure 6-39. A Classic Bridge Circuit for Low Power Applications

REF pin to midway between the supplies, especially if the input signal will be bipolar. However, the voltage on the REF pin can be varied to suit the application. A good example of this is when the REF pin is tied to the V_{REF} pin of an analog-to-digital converter (ADC) whose input range is ($V_{REF} \pm V_{IN}$). With an available output swing on the AD627 of ($-V_S$ + 100 mV) to ($+V_S$ – 150 mV), the maximum programmable gain is simply this output range divided by the input range.

A Single-Supply Data Acquisition System

The bridge circuit of Figure 6-40 is excited by a +5 V supply. The full-scale output voltage from the bridge (±10 mV), therefore, has a common-mode level of 2.5 V. The AD623 removes the common-mode voltage component and amplifies the input signal by a factor of 100 (R_{GAIN} = 1.02 kΩ). This results in an output signal of ±1 V.

In order to prevent this signal from running into the AD623's ground rail, the voltage on the REF pin has to be raised to at least 1 V. In this example, the 2 V reference voltage from the AD7776 ADC is used to bias the AD623's output voltage to 2 V ± 1 V. This corresponds to the input range of the ADC.

A Low Dropout Bipolar Bridge Driver

The AD822 can be used for driving a 350 Ω Wheatstone bridge. Figure 6-41 shows one-half of the AD822 being used to buffer the **AD589**, a 1.235 V low power reference. The output of +4.5 V can be used to drive an A/D converter front end. The other half of the AD822 is configured as a unity-gain inverter and generates the other bridge input of –4.5 V.

Resistors R1 and R2 provide a constant current for bridge excitation. The AD620 low power instrumentation amplifier is used to condition the differential output voltage of the bridge. The gain of the AD620 is programmed using an external resistor, R_G, and determined by

$$G = \frac{49.4\,\text{k}\Omega}{R_G} + 1$$

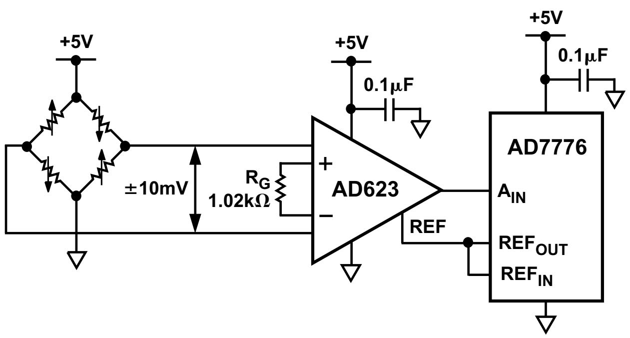


Figure 6-40. A Single-Supply Data Acquisition System

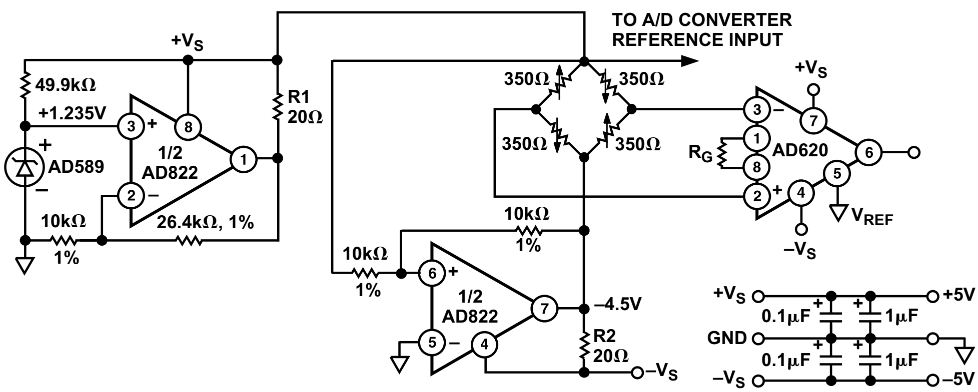


Figure 6-41. Low Dropout Bipolar Bridge Driver

TRANSDUCER INTERFACE APPLICATIONS

Instrumentation amplifiers have long been used as preamplifiers in transducer applications. High quality transducers typically provide a highly linear output, but at a very low level, and a characteristically high output impedance. This requires the use of a high gain buffer/preamplifier that will not contribute any discernible noise of its own to that of the signal. Furthermore, the high output impedance of the typical transducer may require that the in-amp have a low input bias current.

Table 6-3 gives typical characteristics for some common transducer types.

Since most transducers are slow, bandwidth requirements of the in-amp are modest: a 1 MHz small signal bandwidth at unity gain is adequate for most applications.

MEDICAL EKG APPLICATIONS

An EKG is a challenging real-world application, as a small 5 mV signal must be extracted in the presence of much larger 60 Hz noise and large dc common-mode offset variations. Figure 6-42 shows a block diagram of a typical EKG monitor circuit. The value of capacitor C_X is chosen to maintain stability of the right leg drive loop.

Three outputs from the patient are shown here, although several more may be used. The output buffer amplifiers should be low noise, low input bias current FET op amps, since the patient sensors are typically very high impedance and signal levels may be quite low. A three resistor summing network is used to establish a common sense point to drive the force amplifier. The output from the force amplifier servos current through the patient until the net sum output from the three buffer amplifiers is zero.

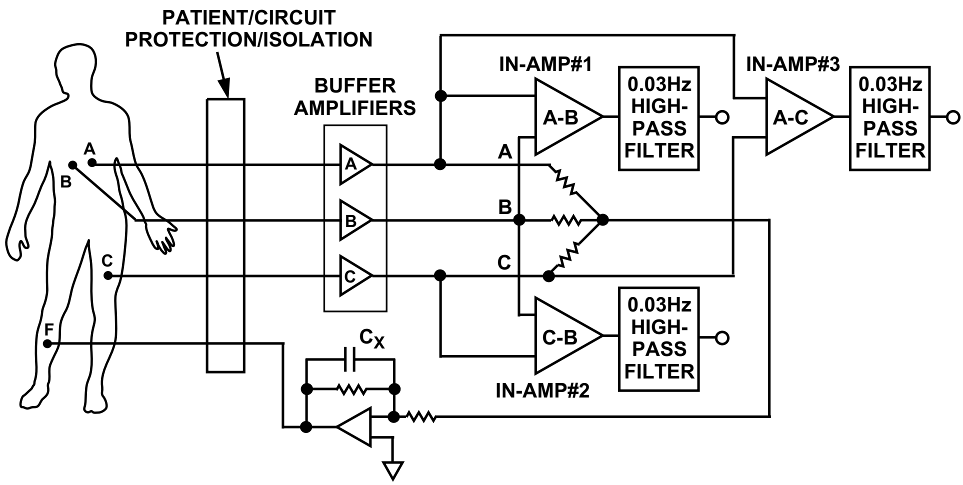


Figure 6-42. A Medical EKG Monitor Circuit

Table 6-3. Typical Transducer Characteristics

Transducer Type	Type of Output	Output Z	Recommended ADI In-Amp/Diff Amp
Thermistor	Resistance Changes with Temperature (–TC) 4%/°C @ +25°C High Nonlinear Output Single-Supply	50 Ω to 1 MΩ @ +25°C	AD620, AD621, AD623, AD627, AD629, AD8221, AD8225
Thermocouple	Low Source Z 10 μV/°C to 100 μV/°C mV Output Level @ +25°C Single-Supply	20 Ω to 20 kΩ (10 Ω typ)	AD620, AD621, AD623, AD627, AD8221
Resistance Temperature Detector (RTD) (In Bridge Circuit)	Low Source Z with Temperature (+TC) 0.1%/°C to 0.66%/°C Single- or Dual-Supply	20 Ω to 20 kΩ @ 0°C	AD620, AD621, AD623, AD627, AD8221, AD8225
Level Sensors Thermal Types Float Types	Thermistor Output (Low) Variable Resistance Output of mV to Several Volts Single-Supply	500 Ω to 2 kΩ 100 Ω to 2 kΩ	AD626, AD628, AD629, AD8225
Load Cell (Strain Gage Bridge) (Weight Measurement)	Variable Resistance 2 mV/V of Excitation 0.1% Typical Full-Scale Change Single- or Dual-Supply	120 Ω to 1 kΩ	AD620, AD621, AD8221, AD8225
Current Sense (Shunt)	Low Value Resistor Output High Common-Mode Voltage	A Few Ohms (or less)	AD626, AD628, AD629, AD8202, AD8205
EKG Monitors (Single-Supply Bridge Configuration)	Low Level Differential Output Voltage 5 mV Output Typical Single- or Dual-Supply	500 kΩ	AD620, AD621, AD623, AD627, AD8221, AD8225
Photodiode Sensor	Current Increases with Light Intensity 1 pA to 1 μA I_{OUTPUT} Single-Supply	10^9 Ω	AD620, AD621, AD622, AD623, AD627, AD8221
Hall Effect Magnetic	5 mV/kG to 120 mV/kG	1 Ω to 1 kΩ	AD620, AD621, AD622, AD623, AD627, AD8221

Three in-amps are used to provide three separate outputs for monitoring the patient's condition. Suitable ADI products include AD8221, AD627, and AD623 in-amps and **AD820**, **AD822** (dual), and **AD824** (quad) op amps for use as the buffer. Each in-amp is followed by a high-pass filter that removes the dc component from the signal. It is common practice to omit one of the in-amps and determine the third output by software (or hardware) calculation.

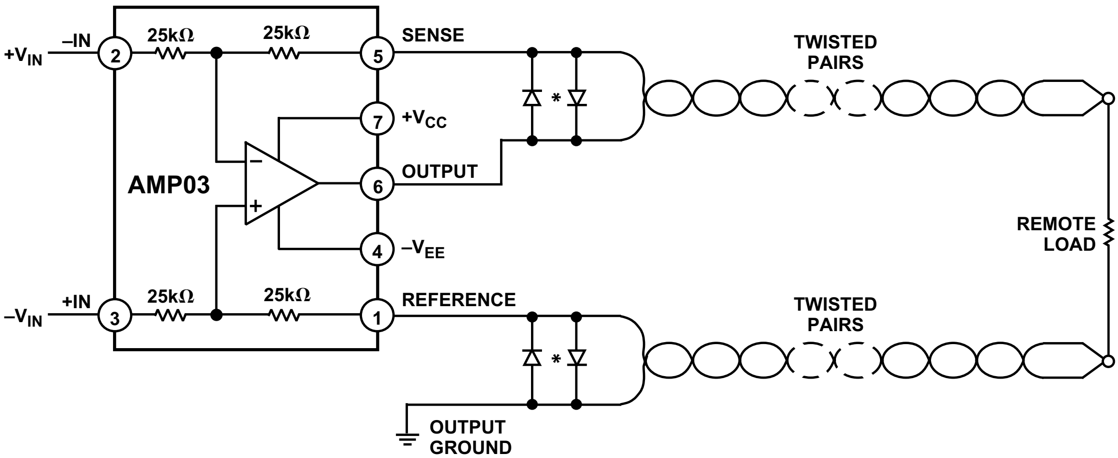
Proper safeguards, such as isolation, must be added to this circuit to protect the patient from possible harm.

REMOTE LOAD-SENSING TECHNIQUE

The circuit of Figure 6-43 is a unity gain instrumentation amplifier that uses its sense and reference pins to minimize any errors due to parasitic voltage drops within the circuit. If heavy output currents are expected, and there is a need to sense a load that is some distance away from the circuit, voltage drops due to trace or wire resistance can cause errors. These voltage drops are particularly troublesome with low resistance loads, such as 50 Ω.

The sense terminal completes the feedback path for the instrumentation amplifier output stage and is normally connected directly to the in-amp output. Similarly, the reference terminal sets the reference voltage about which the in-amp's output will swing. This connection puts the IR drops inside the feedback loop of the in-amp and virtually eliminates any IR errors.

This circuit will provide a 3 dB bandwidth better than 3 MHz. Note that any net capacitance between the twisted pairs is isolated from the in-amp's output by 25 kΩ resistors, but any net capacitance between the twisted pairs and ground needs to be minimized to maintain stability. So, *unshielded* twisted-pair cable is recommended for this circuit. For low speed applications that require driving long lengths of *shielded* cable, the **AMP01** should be substituted for the AMP03 device. The AMP01 can drive capacitance loads up to 1 µF, while the AMP03 is limited to driving a few hundred pF.



*1N4148 DIODES ARE OPTIONAL. DIODES LIMIT THE OUTPUT VOLTAGE EXCURSION IF SENSE AND/OR REFERENCE LINES BECOME DISCONNECTED FROM THE LOAD.

Figure 6-43. A Remote Load Sensing Connection

A PRECISION VOLTAGE-TO-CURRENT CONVERTER

Figure 6-44 is a precision voltage-to-current converter whose scale factor is easily programmed for exact decade ratios using standard 1% metal film resistor values. The **AD620** operates with full accuracy on standard 5 V power supply voltages. Note that although the quiescent current of the AD620 is only 900 µA, the addition of the **AD705** will add an additional 380 µA current consumption.

A CURRENT SENSOR INTERFACE

Figure 6-45 shows a novel circuit for sensing low-level currents. It makes use of the large common-mode range of the AD626. The current being measured is sensed across resistor R_S. The value of R_S should be less than 1 kΩ and should be selected so that the average differential voltage across this resistor is typically 100 mV.

To produce a full-scale output of +4 V, a gain of 40 is used, adjustable by +20% to absorb the tolerance in the sense resistor. Note that there is sufficient headroom to allow at least a 10% overrange (to +4.4 V).

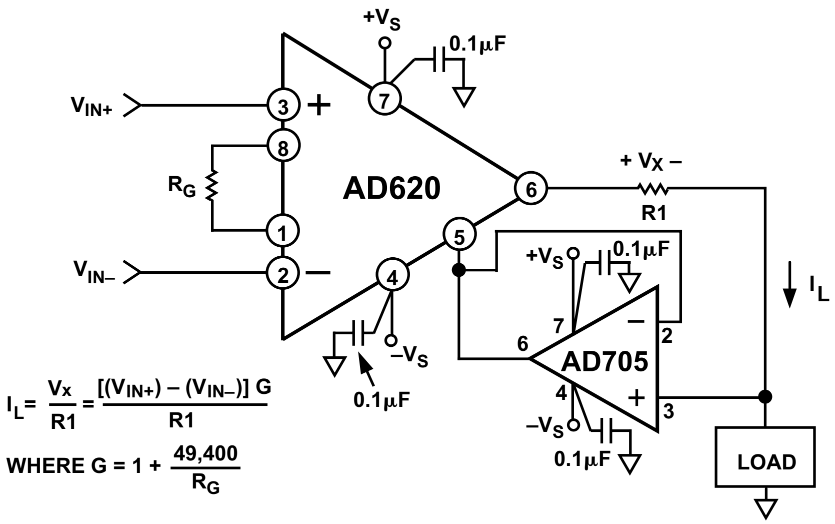


Figure 6-44. A Precision Voltage-to-Current Converter that Operates on ±5 V Supplies

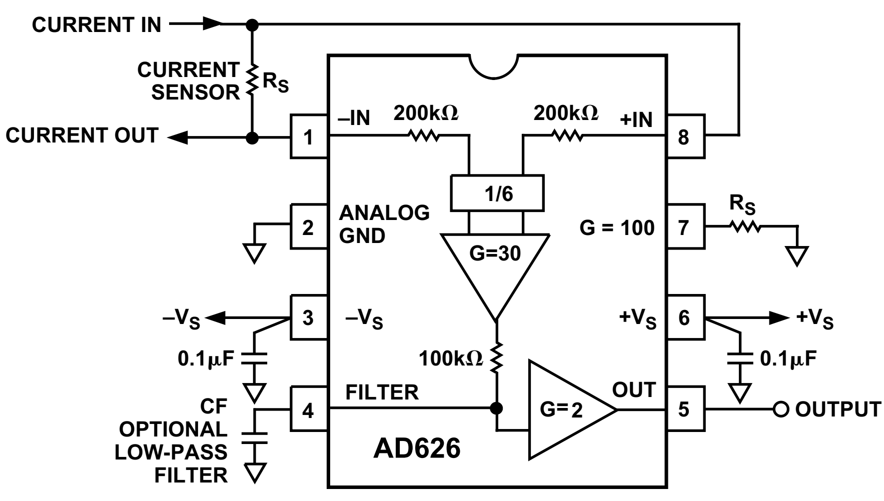


Figure 6-45. Current Sensor Interface

OUTPUT BUFFERING LOW POWER IN-AMPS

The AD627 low power in-amp is designed to drive load impedances of 20 kΩ or higher, but can deliver up to 20 mA to heavier loads with low output voltage swings. If more than 20 mA of output current is required, the AD627's output should be buffered with a precision low power op amp, such as the AD820, as shown in Figure 6-46. This op amp can swing from 0 V to 4 V on its output while driving a load as small as 600 Ω. The addition of the AD820 isolates the in-amp from the load, thus greatly reducing any thermal effects.

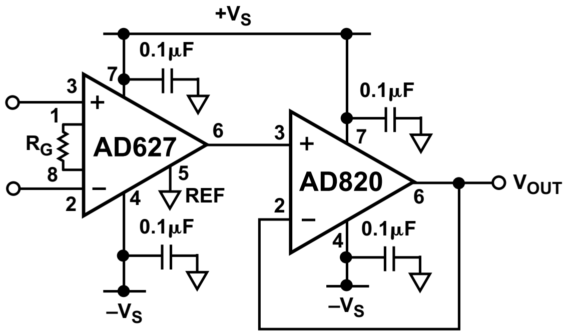


Figure 6-46. Output Buffer for Low Power In-Amps

A 4 mA TO 20 mA SINGLE-SUPPLY RECEIVER

Figure 6-47 shows how a signal from a 4 mA to 20 mA transducer can be interfaced to the **ADuC812**, a 12-bit ADC with an embedded microcontroller. The signal from a 4 mA to 20 mA transducer is single-ended. This initially suggests the need for a simple shunt resistor to convert the current to a voltage at the high impedance analog input of the converter. However, any line resistance in the return path (to the transducer) will add a current-dependent offset error. So, the current must be sensed differentially. In this example, a 24.9 Ω shunt resistor generates a maximum differential input voltage to the AD627 of between 100 mV (for 4 mA in) and 500 mV (for 20 mA in). With no gain resistor present, the AD627 amplifies the 500 mV input voltage by a factor of 5 to 2.5 V, the full-scale input voltage of the ADC. The zero current of 4 mA corresponds to a code of 819 and the LSB size is 4.9 mV.

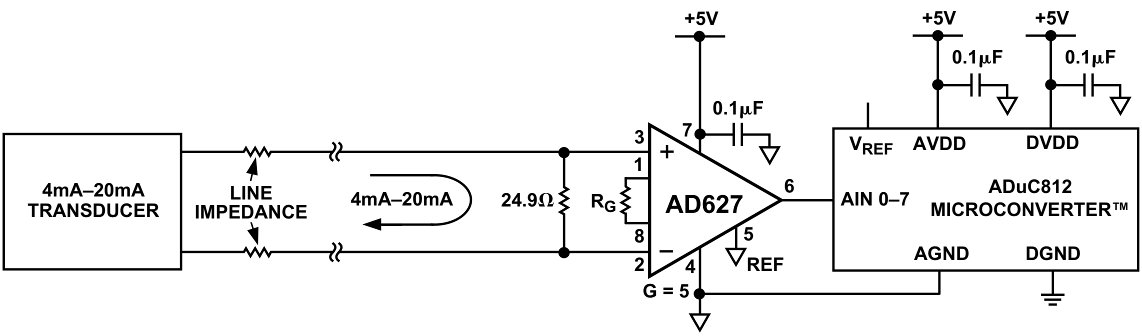


Figure 6-47. A 4 mA to 20 mA Receiver Circuit

A SINGLE-SUPPLY THERMOCOUPLE AMPLIFIER

Because the common-mode input range of the AD627 extends 0.1 V below ground, it is possible to measure small differential signals with little or no common-mode component. Figure 6-48 shows a thermocouple application where one side of the J-type thermocouple is grounded. Over a temperature range from –200°C to +200°C, the J-type thermocouple delivers a voltage ranging from –7.890 mV to +10.777 mV.

A programmed gain on the AD627 of 100 (R_G = 2.1 kΩ) and a voltage on the AD627 REF pin of 2 V results in the AD627's output voltage ranging from 1.110 V to 3.077 V relative to ground.

SPECIALTY PRODUCTS

Analog Devices sells a number of specialty products, many of which were designed for the audio market that are useful for some in-amp applications. Table 6-4 lists some of these products.

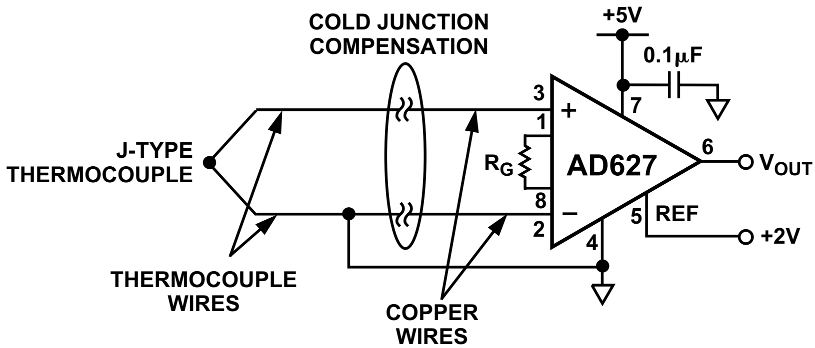


Figure 6-48. A Thermocouple Amplifier Using a Low Power, Single-Supply In-Amp

Table 6-4. Specialty Products Available from Analog Devices

Model Number	Description	BW	CMR (DC)	Supply	Features
SSM2141	Diff Line Receiver	3 MHz	100 dB	±18 V	High CMR, Audio Subtractor
SSM2143	Diff Line Receiver	7 MHz (G = 0.5)	90 dB	±6 V to ±18 V	Low Distortion, Audio Subtractor
SSM2019	Audio Preamp	2 MHz (G = 1)	74 dB	±5 V to ±18 V	Low Noise, Low Distortion, Audio IA