

### FEATURES

- Fixed gain of 2000
- Access to internal nodes provides flexibility
- Low noise:  $1.5 \text{ nV}/\sqrt{\text{Hz}}$  input voltage noise
- High accuracy dc performance
  - Gain drift:  $10 \text{ ppm}/^\circ\text{C}$
  - Offset drift:  $1 \mu\text{V}/^\circ\text{C}$
  - Gain accuracy:  $0.2\%$
  - CMRR:  $130 \text{ dB min}$
- Excellent ac specifications
  - Bandwidth:  $3.5 \text{ MHz}$
  - Slew rate:  $40 \text{ V}/\mu\text{s}$
- Power supply range:  $\pm 4 \text{ V to } \pm 18 \text{ V}$
- 8-pin SOIC package
- ESD protection  $>5000 \text{ V (HBM)}$
- Temperature range for specified performance:  $-40^\circ\text{C to } +85^\circ\text{C}$
- Operational up to  $125^\circ\text{C}$

### APPLICATIONS

- Sensor interface
- Medical instrumentation
- Patient monitoring

### GENERAL DESCRIPTION

The **AD8428** is an ultralow noise instrumentation amplifier designed to accurately measure tiny, high speed signals. It delivers industry-leading gain accuracy, noise, and bandwidth.

All gain setting resistors for the **AD8428** are internal to the part and are precisely matched. Care is taken in both the chip pinout and layout. This results in excellent gain drift and quick settling to the final gain value after the part is powered on.

The high CMRR of the **AD8428** prevents unwanted signals from corrupting the signal of interest. The pinout of the **AD8428** is designed to avoid parasitic capacitance mismatches that can degrade CMRR at high frequencies.

### FUNCTIONAL BLOCK DIAGRAM

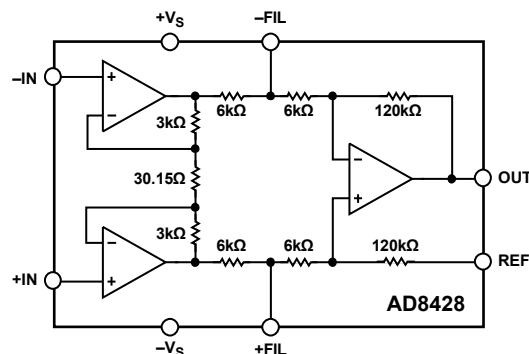


Figure 1.

Table 1. Instrumentation Amplifiers by Category<sup>1</sup>

General Purpose	Zero Drift	Military Grade	Low Power	Low Noise
AD8220	AD8231	AD620	AD627	AD8428
AD8221	AD8290	AD621	AD623	AD8429
AD8222	AD8293	AD524	AD8235	
AD8224	AD8553	AD526	AD8236	
AD8228	AD8556	AD624	AD8426	
AD8295	AD8557		AD8226	
			AD8227	

<sup>1</sup> See [www.analog.com](http://www.analog.com) for the latest instrumentation amplifiers.

The **AD8428** is one of the fastest instrumentation amplifiers available. The circuit architecture is designed for high bandwidth at high gain. The **AD8428** uses a current feedback topology for the initial preamplifier gain stage of 200, followed by a difference amplifier stage of 10. This architecture results in a  $3.5 \text{ MHz}$  bandwidth at a gain of 2000 for an equivalent gain bandwidth product of  $7 \text{ GHz}$ .

The **AD8428** pinout allows access to internal nodes between the first and second stages. This feature can be useful for modifying the frequency response between the two amplification stages, thereby preventing unwanted signals from contaminating the output results.

The performance of the **AD8428** is specified over the industrial temperature range of  $-40^\circ\text{C to } +85^\circ\text{C}$ . It is available in an 8-lead plastic SOIC package.

#### Rev. 0

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REVISION HISTORY

10/11—Revision 0: Initial Version

## SPECIFICATIONS

$V_S = \pm 15\text{ V}$ ,  $V_{REF} = 0\text{ V}$ ,  $T_A = 25^\circ\text{C}$ ,  $G = 2000$ ,  $R_L = 10\text{ k}\Omega$ , unless otherwise noted.

Table 2.

Parameter	Test Conditions/Comments	Min	Typ	Max	Unit
COMMON-MODE REJECTION RATIO (RTI)	$V_{CM} = \pm 10\text{ V}$				
CMRR, DC to 60 Hz		130			dB
CMRR at 50 kHz		110			dB
NOISE (RTI)	$V_{IN+}, V_{IN-} = 0\text{ V}$				
Voltage Noise	$f = 1\text{ kHz}$		1.3	1.5	nV/ $\sqrt{\text{Hz}}$
	$f = 0.1\text{ Hz to } 10\text{ Hz}$		40	50	nV p-p
Current Noise	$f = 1\text{ kHz}$		1.5		pA/ $\sqrt{\text{Hz}}$
	$f = 0.1\text{ Hz to } 10\text{ Hz}$		150		pA p-p
VOLTAGE OFFSET					
Input Offset, $V_{OSI}$				100	$\mu\text{V}$
Average TC	$T_A = -40^\circ\text{C to } +85^\circ\text{C}$			1	$\mu\text{V}/^\circ\text{C}$
Offset RTI vs. Supply (PSRR)		120			dB
INPUT CURRENT					
Input Bias Current				200	nA
Over Temperature	$T_A = -40^\circ\text{C to } +85^\circ\text{C}$		250		pA/ $^\circ\text{C}$
Input Offset Current				50	nA
Over Temperature	$T_A = -40^\circ\text{C to } +85^\circ\text{C}$		20		pA/ $^\circ\text{C}$
DYNAMIC RESPONSE					
–3 dB Small Signal Bandwidth			3.5		MHz
Settling Time to 0.01%	10 V step		0.75		$\mu\text{s}$
Settling Time to 0.001%	10 V step		1.4		$\mu\text{s}$
Slew Rate		40	50		V/ $\mu\text{s}$
GAIN					
First Stage Gain			200		V/V
Subtractor Stage Gain			10		V/V
Total Gain Error	$V_{OUT} = -10\text{ V to } +10\text{ V}$			0.2	%
Total Gain Nonlinearity	$V_{OUT} = -10\text{ V to } +10\text{ V}$			5	ppm
Total Gain vs. Temperature				10	ppm/ $^\circ\text{C}$
INPUT					
Impedance (Pin to Ground) <sup>1</sup>			1  2		G $\Omega$   pF
Input Operating Voltage Range	$V_S = \pm 4\text{ V to } \pm 18\text{ V}$	$-V_S + 2.5$		$+V_S - 2.5$	V
Over Temperature	$T_A = -40^\circ\text{C to } +85^\circ\text{C}$	$-V_S + 2.5$		$+V_S - 2.5$	V
OUTPUT					
Output Swing	$R_L = 2\text{ k}\Omega$	$-V_S + 1.7$		$+V_S - 1.2$	V
Over Temperature	$T_A = -40^\circ\text{C}$	$-V_S + 2.0$		$+V_S - 1.3$	V
	$T_A = +85^\circ\text{C}$	$-V_S + 1.6$		$+V_S - 1.1$	V
Output Swing	$R_L = 10\text{ k}\Omega$	$-V_S + 1.7$		$+V_S - 1.0$	V
Over Temperature	$T_A = -40^\circ\text{C}$	$-V_S + 1.8$		$+V_S - 1.2$	V
	$T_A = +85^\circ\text{C}$	$-V_S + 1.4$		$+V_S - 0.9$	V
Short-Circuit Current			30		mA
REFERENCE INPUT					
$R_{IN}$			132		k $\Omega$
$I_{IN}$	$V_{IN+}, V_{IN-} = 0\text{ V}$		6.5		$\mu\text{A}$
Voltage Range		$-V_S$		$+V_S$	V
Reference Gain to Output			1		V/V
Reference Gain Error			0.01		%

Parameter	Test Conditions/Comments	Min	Typ	Max	Unit
FILTER TERMINALS					
$R_{IN}$ <sup>2</sup>			6		k $\Omega$
Voltage Range		$-V_S$		$+V_S$	V
POWER SUPPLY					
Operating Range		$\pm 4$		$\pm 18$	V
Quiescent Current			6.5	6.8	mA
Over Temperature	$T_A = -40^{\circ}\text{C}$ to $+85^{\circ}\text{C}$			8	mA

<sup>1</sup> The differential and common-mode input impedances can be calculated from the pin impedance:  $Z_{DIFF} = 2(Z_{PIN})$ ;  $Z_{CM} = Z_{PIN}/2$ .

<sup>2</sup> To calculate the actual impedance, see Figure 1.

## ABSOLUTE MAXIMUM RATINGS

Table 3.

Parameter	Rating
Supply Voltage	$\pm 18\text{ V}$
Output Short-Circuit Current Duration	Indefinite
Maximum Voltage at $-\text{IN}$ , $+\text{IN}$ <sup>1</sup>	$\pm V_S$
Maximum Voltage at $-\text{FIL}$ , $+\text{FIL}$	$\pm V_S$
Differential Input Voltage <sup>1</sup>	$\pm 1\text{ V}$
Maximum Voltage at REF	$\pm V_S$
Storage Temperature Range	$-65^\circ\text{C}$ to $+150^\circ\text{C}$
Specified Temperature Range	$-40^\circ\text{C}$ to $+85^\circ\text{C}$
Maximum Junction Temperature	$140^\circ\text{C}$
ESD	
Human Body Model	5000 V
Charged Device Model	1250 V
Machine Model	400 V

<sup>1</sup> For voltages beyond these limits, use input protection resistors. See the Input Protection section for more information.

Stresses above those listed under Absolute Maximum Ratings may cause permanent damage to the device. This is a stress rating only; functional operation of the device at these or any other conditions above those indicated in the operational section of this specification is not implied. Exposure to absolute maximum rating conditions for extended periods may affect device reliability.

## THERMAL RESISTANCE

$\theta_{JA}$  is specified for the worst-case conditions, that is, a device soldered in a circuit board for surface-mount packages.

Table 4. Thermal Resistance

Package	$\theta_{JA}$	Unit
8-Lead SOIC_N	121	$^\circ\text{C}/\text{W}$

## ESD CAUTION



**ESD (electrostatic discharge) sensitive device.** Charged devices and circuit boards can discharge without detection. Although this product features patented or proprietary protection circuitry, damage may occur on devices subjected to high energy ESD. Therefore, proper ESD precautions should be taken to avoid performance degradation or loss of functionality.

PIN CONFIGURATION AND FUNCTION DESCRIPTIONS

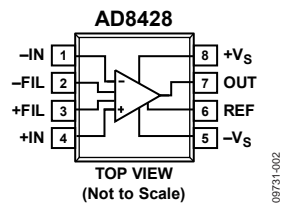


Figure 2. Pin Configuration

Table 5. Pin Function Descriptions

Pin No.	Mnemonic	Description
1	–IN	Negative Input Terminal.
2	–FIL	Negative Filter Terminal.
3	+FIL	Positive Filter Terminal.
4	+IN	Positive Input Terminal.
5	–V <sub>S</sub>	Negative Power Supply Terminal.
6	REF	Reference Voltage Terminal. Drive this terminal with a low impedance voltage source to level-shift the output.
7	OUT	Output Terminal.
8	+V <sub>S</sub>	Positive Power Supply Terminal.

## TYPICAL PERFORMANCE CHARACTERISTICS

$T_A = 25^\circ\text{C}$ ,  $V_S = \pm 15\text{ V}$ ,  $V_{\text{REF}} = 0\text{ V}$ ,  $R_L = 10\text{ k}\Omega$ , unless otherwise noted.

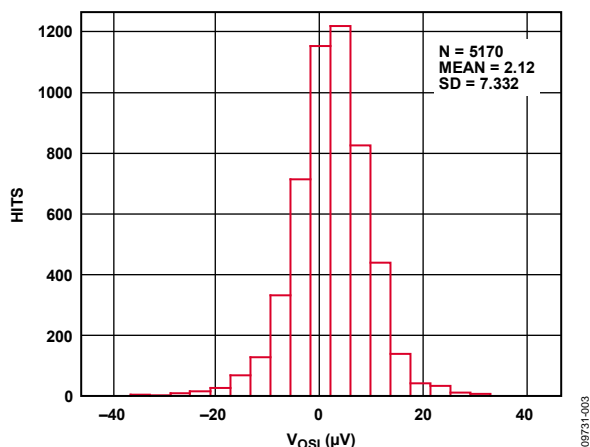


Figure 3. Typical Distribution of Input Offset Voltage,  $V_S = \pm 5\text{ V}$

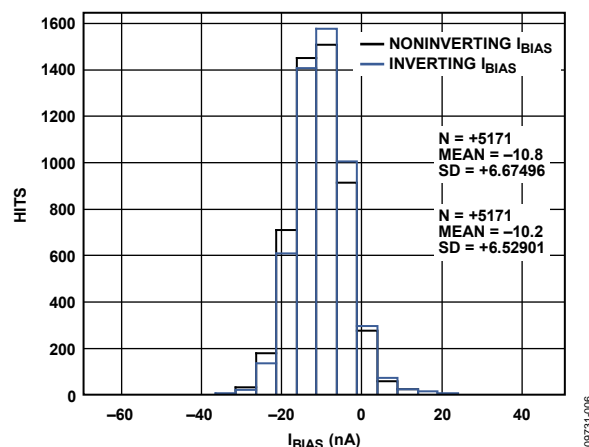


Figure 6. Typical Distribution of Input Bias Current

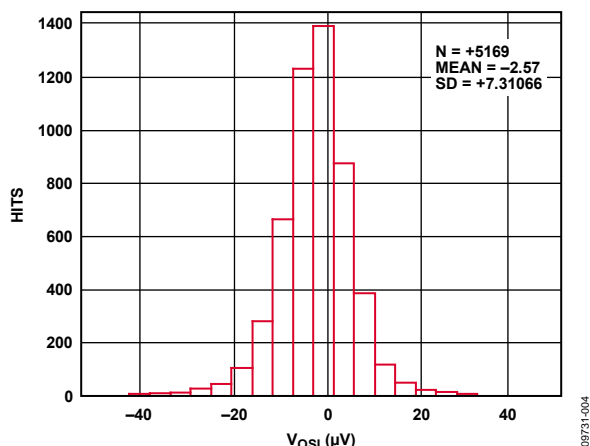


Figure 4. Typical Distribution of Input Offset Voltage,  $V_S = \pm 15\text{ V}$

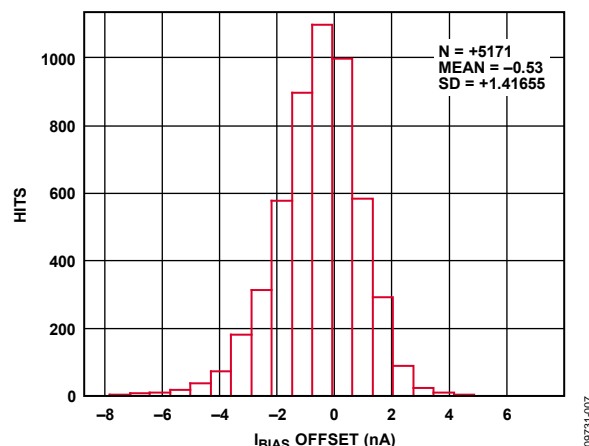


Figure 7. Typical Distribution of Input Bias Current Offset

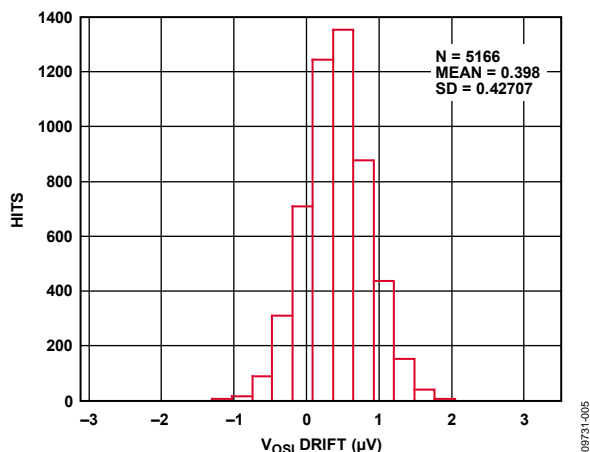


Figure 5. Typical Distribution of Input Offset Voltage Drift

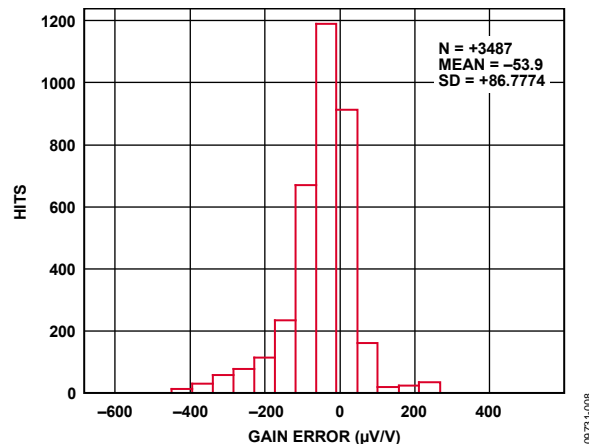


Figure 8. Typical Distribution of Gain Error, Gain = 2000,  $V_S = \pm 15\text{ V}$ ,  $R_L = 10\text{ k}\Omega$

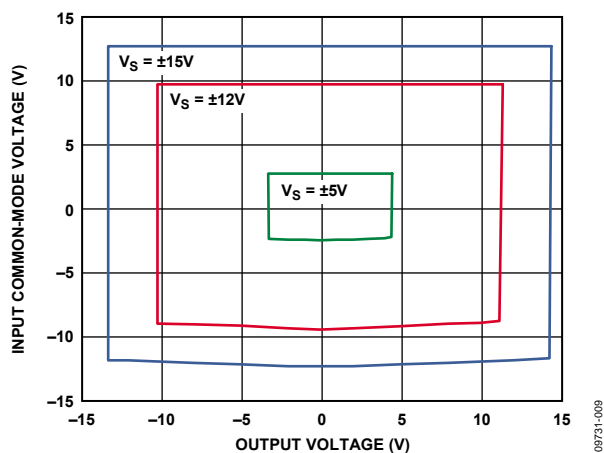


Figure 9. Input Common-Mode Voltage vs. Output Voltage,  
 $V_S = \pm 5\text{ V}$ ,  $V_S = \pm 12\text{ V}$ ,  $V_S = \pm 15\text{ V}$

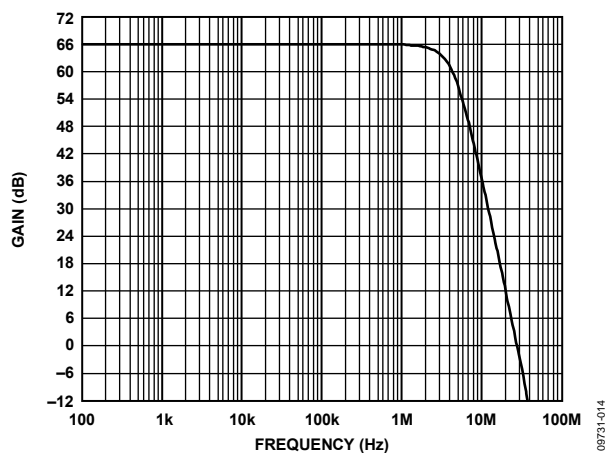


Figure 12. Gain vs. Frequency

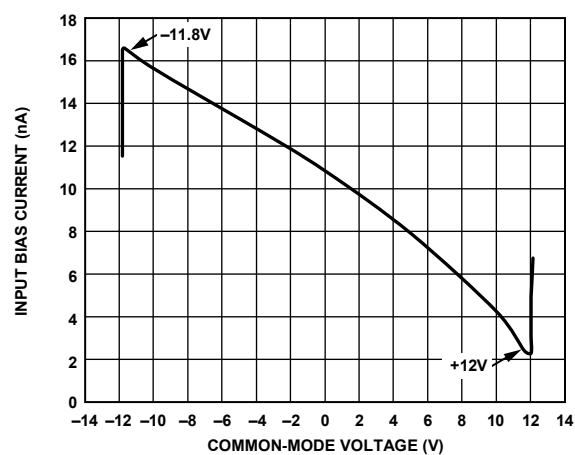


Figure 10. Input Bias Current vs. Common-Mode Voltage,  
 $V_S = \pm 15\text{ V}$

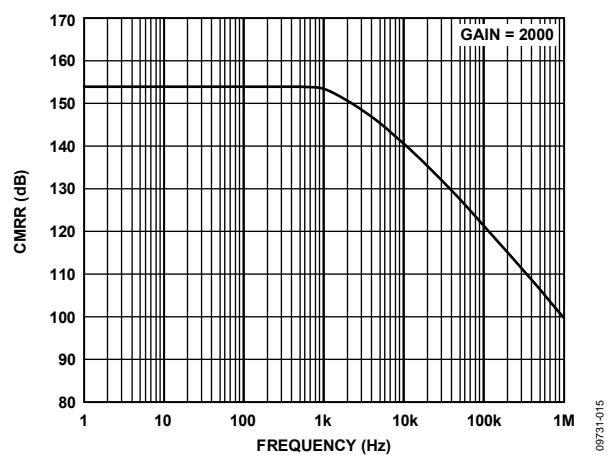


Figure 13. CMRR vs. Frequency

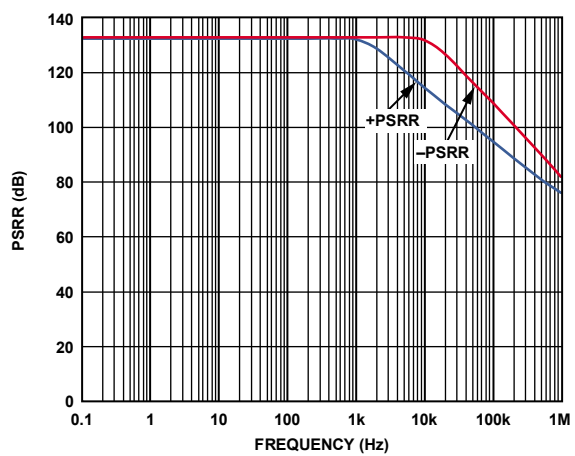


Figure 11. PSRR vs. Frequency

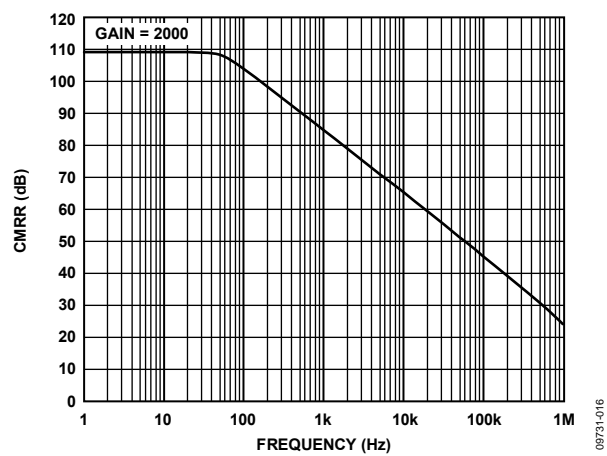


Figure 14. CMRR vs. Frequency, 1 kΩ Source Imbalance



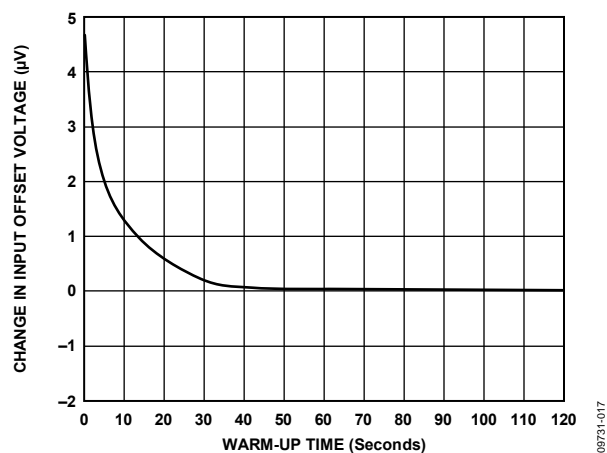
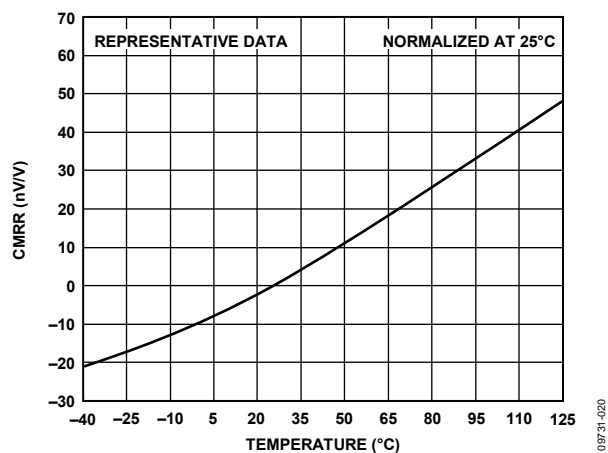
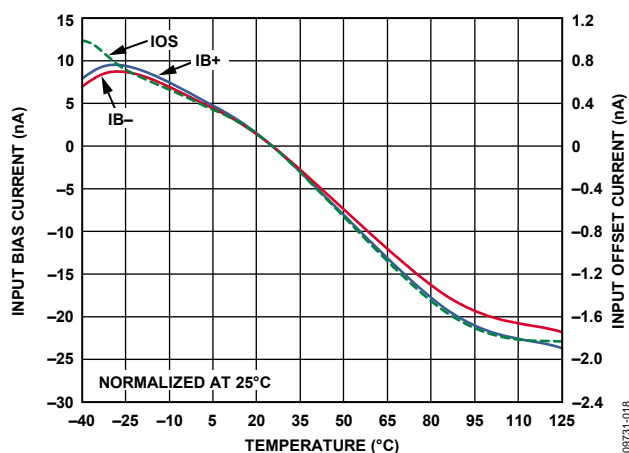
Figure 15. Change in Input Offset Voltage ( $V_{OS}$ ) vs. Warm-Up TimeFigure 18. CMRR vs. Temperature, Normalized at 25  $^{\circ}C$ 

Figure 16. Input Bias Current and Input Offset Current vs. Temperature

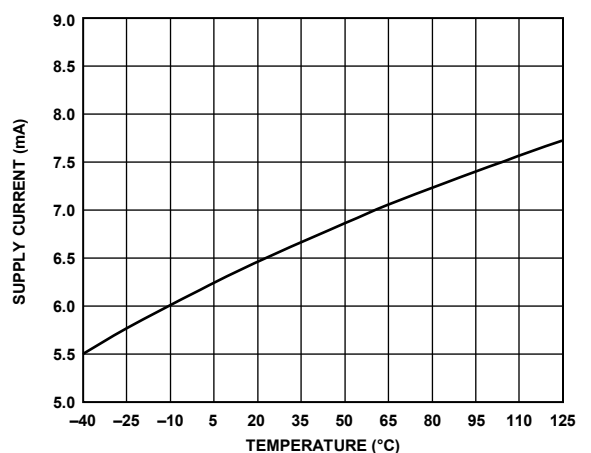


Figure 19. Supply Current vs. Temperature

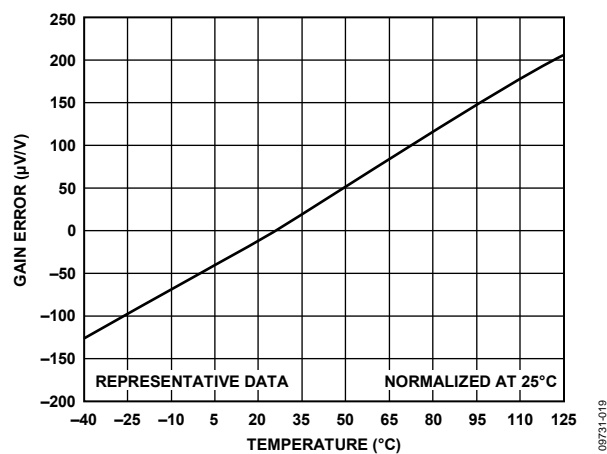
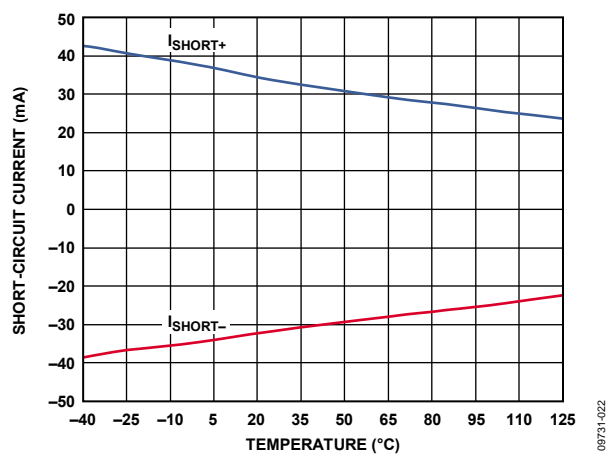
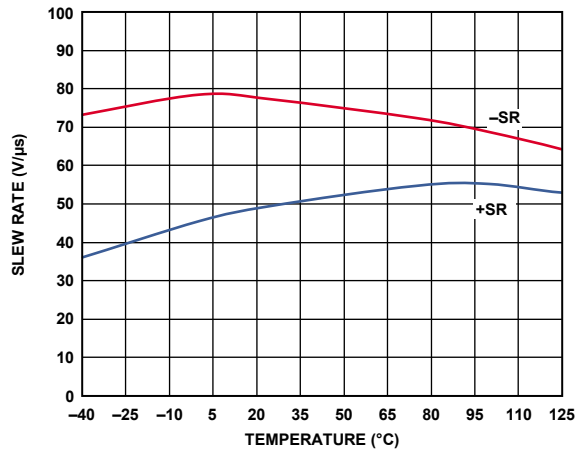
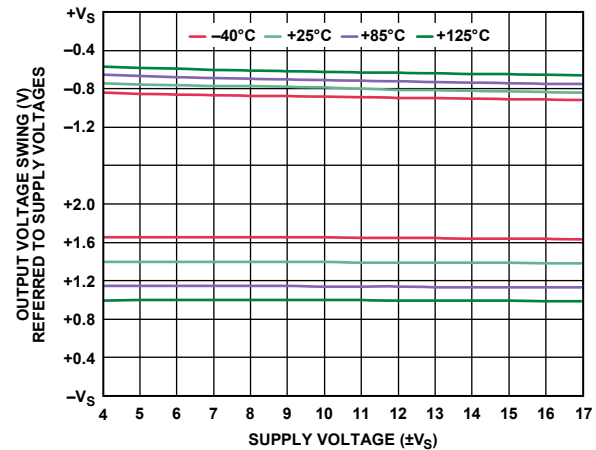
Figure 17. Gain Error vs. Temperature, Normalized at 25  $^{\circ}C$ 

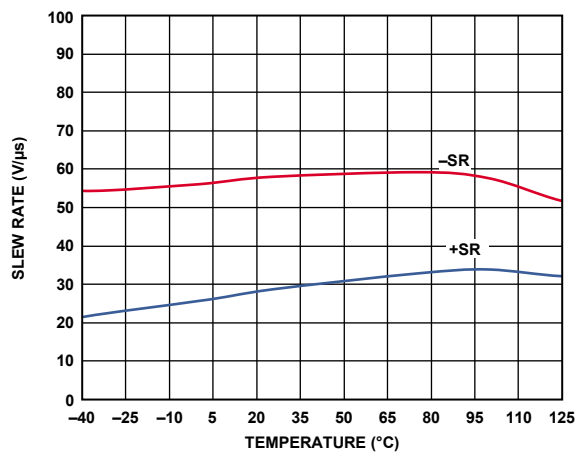
Figure 20. Short-Circuit Current vs. Temperature

Figure 21. Slew Rate vs. Temperature,  $V_S = \pm 15\text{ V}$ 

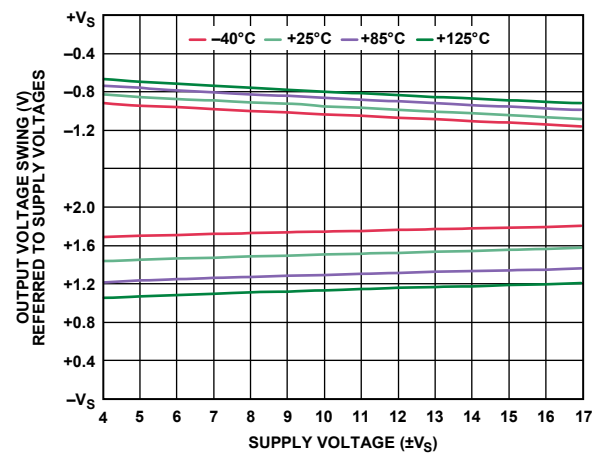
09731-023

Figure 24. Output Voltage Swing vs. Supply Voltage,  $R_L = 10\text{ k}\Omega$ 

09731-026

Figure 22. Slew Rate vs. Temperature,  $V_S = \pm 5\text{ V}$ 

09731-024

Figure 25. Output Voltage Swing vs. Supply Voltage,  $R_L = 2\text{ k}\Omega$ 

09731-027

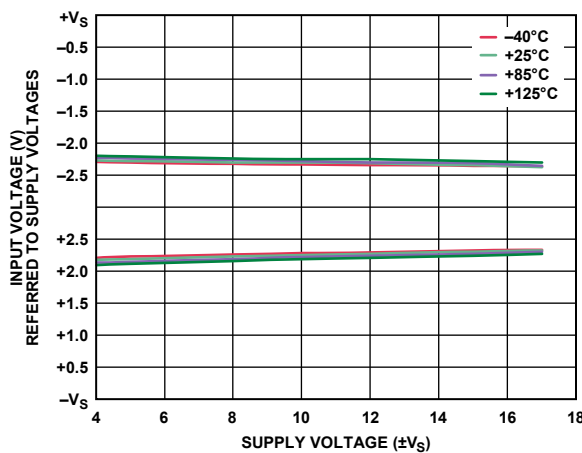
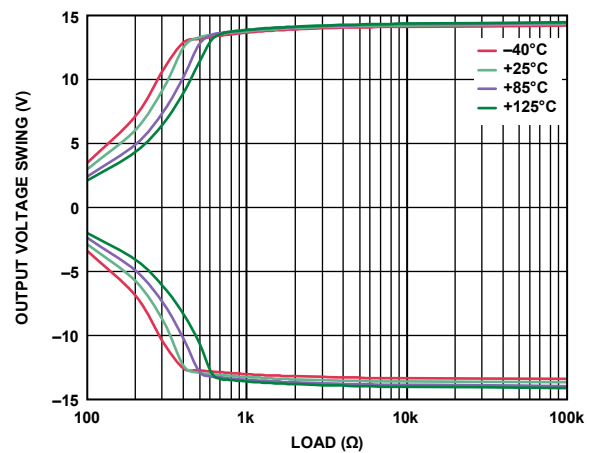


Figure 23. Input Voltage Limit vs. Supply Voltage

09731-025

Figure 26. Output Voltage Swing vs. Load Resistance,  $V_S = \pm 15\text{ V}$ 

09731-028

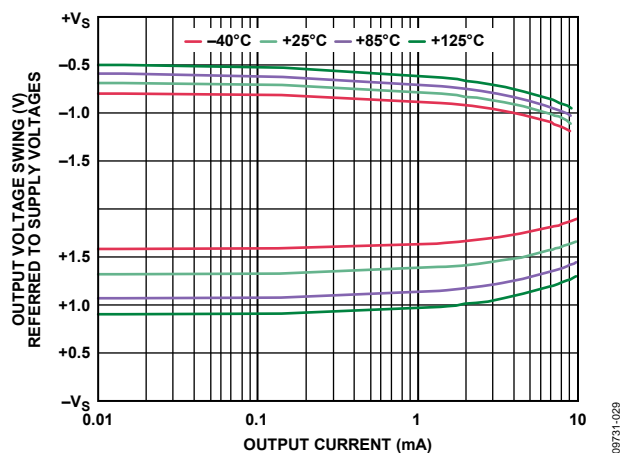
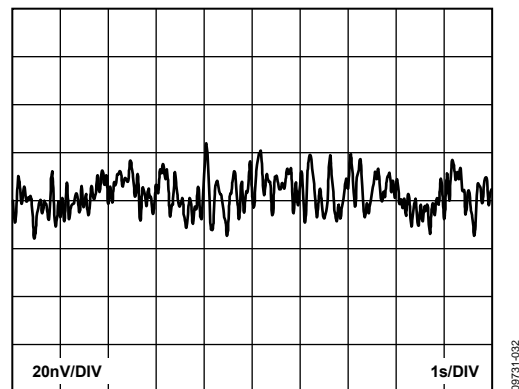
Figure 27. Output Voltage Swing vs. Output Current,  $V_S = \pm 15\text{ V}$ 

Figure 30. 0.1 Hz to 10 Hz RTI Voltage Noise

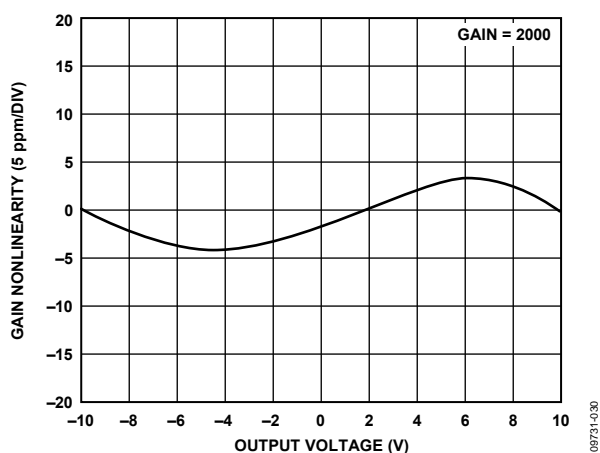
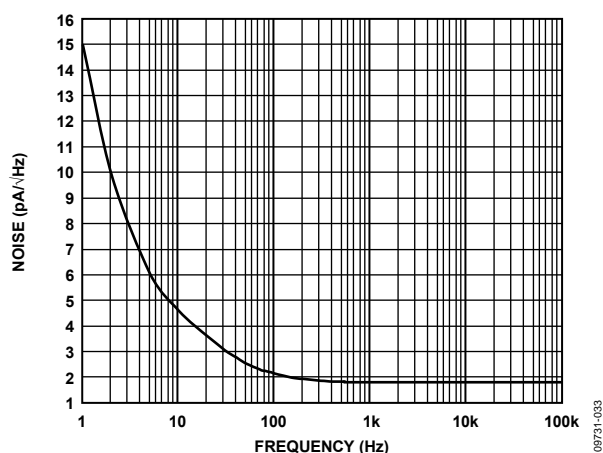
Figure 28. Gain Nonlinearity,  $R_L = 10\text{ k}\Omega$ 

Figure 31. Current Noise Spectral Density vs. Frequency

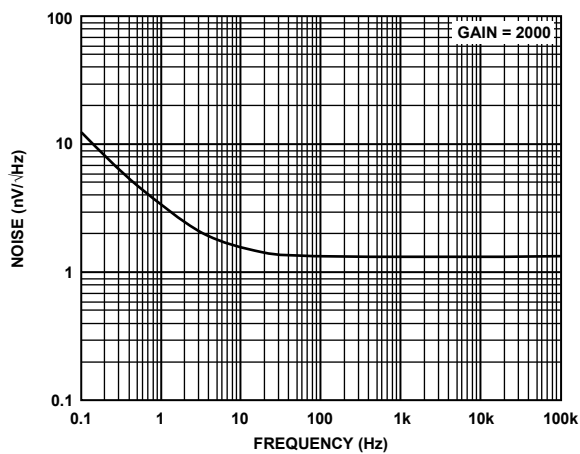


Figure 29. RTI Voltage Noise Spectral Density vs. Frequency

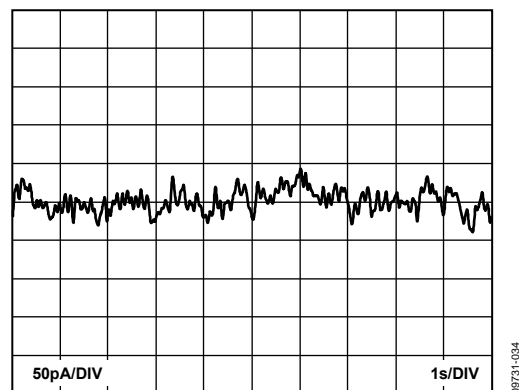


Figure 32. 0.1 Hz to 10 Hz Current Noise

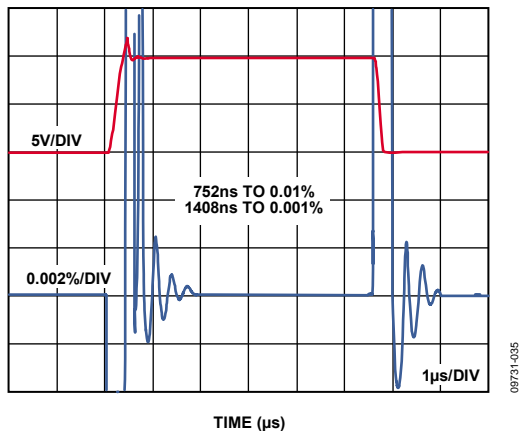


Figure 33. Large Signal Pulse Response and Settling Time, 10 V Step,  $V_S = \pm 15$  V

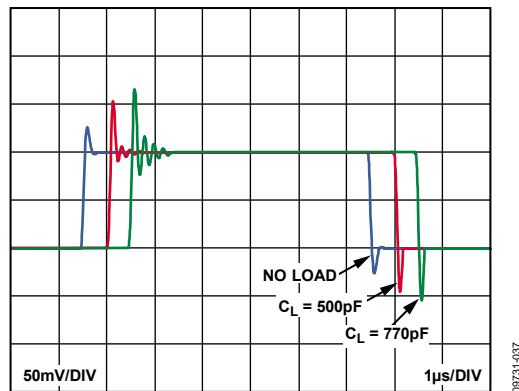


Figure 35. Small Signal Pulse Response with Various Capacitive Loads, No Resistive Load

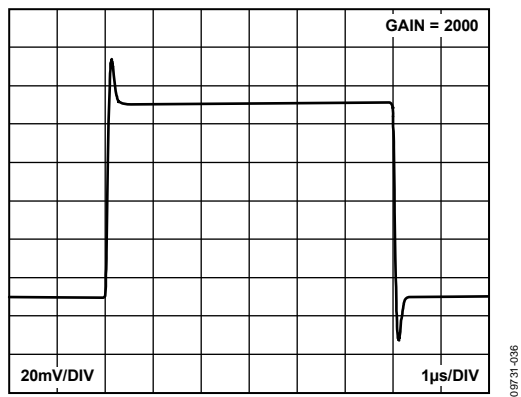


Figure 34. Small Signal Pulse Response,  $R_L = 10$  k $\Omega$ ,  $C_L = 100$  pF

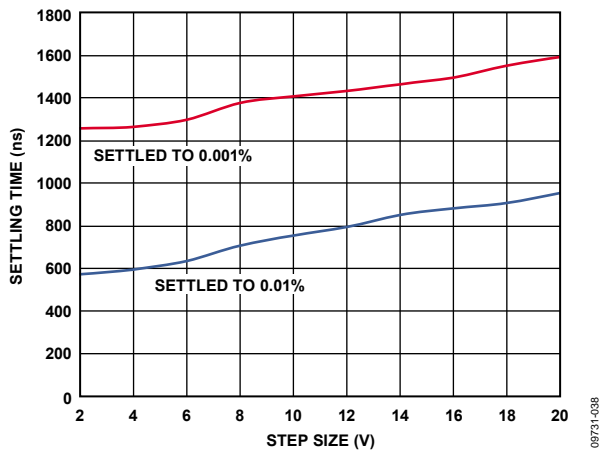


Figure 36. Settling Time vs. Step Size

## THEORY OF OPERATION

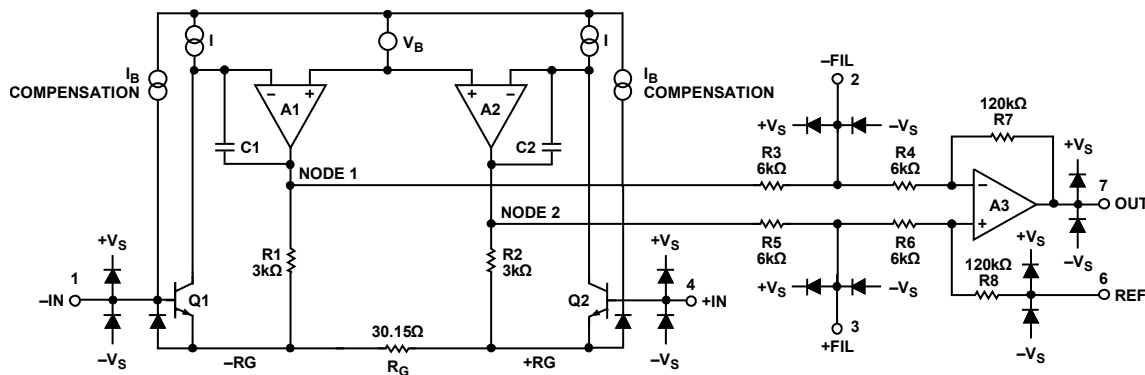


Figure 37. Simplified Schematic

### ARCHITECTURE

The **AD8428** is based on the classic 3-op-amp topology. This topology has two stages: a gain stage (preamplifier) to provide differential amplification by a factor of 200, followed by a difference amplifier stage to remove the common-mode voltage and provide additional amplification by a factor of 10. Figure 37 shows a simplified schematic of the **AD8428**.

The first stage works as follows. To keep its two inputs matched, Amplifier A1 must keep the collector of Q1 at a constant voltage. It does this by forcing  $-R_G$  to be a precise diode drop from  $-IN$ . Similarly, A2 forces  $+R_G$  to be a constant diode drop from  $+IN$ . Therefore, a replica of the differential input voltage is placed across the gain setting resistor,  $R_G$ . The current that flows across this resistance must also flow through the R1 and R2 resistors, creating a gained differential signal between the A2 and A1 outputs.

The second stage is a  $G = 10$  difference amplifier, composed of Amplifier A3 and Resistors R3 through R8. This stage removes the common-mode signal from the amplified differential signal.

The transfer function of the **AD8428** is

$$V_{OUT} = 2000 \times (V_{IN+} - V_{IN-}) + V_{REF}$$

### FILTER TERMINALS

The  $-FIL$  and  $+FIL$  terminals allow access between R3 and R4, and between R5 and R6, respectively. Adding a filter between these two terminals modifies the signal gain vs. frequency before it reaches the second amplifier stage.

### REFERENCE TERMINAL

The output voltage of the **AD8428** is developed with respect to the potential on the reference terminal. This is useful when the output signal must be offset to a precise midsupply level. For example, a voltage source can be tied to the REF pin to level-shift the output so that the **AD8428** can drive a single-supply ADC. The REF pin is protected with ESD diodes and should not exceed either  $+V_S$  or  $-V_S$ .

For best performance, the source impedance to the REF terminal should be kept well below  $1 \Omega$ . As shown in Figure 37, the reference terminal, REF, is at one end of a  $120 \text{ k}\Omega$  resistor. Additional impedance at the REF terminal adds to this  $120 \text{ k}\Omega$  resistor and results in amplification of the signal connected to the positive input. The amplification from the additional  $R_{REF}$  can be calculated as follows:

$$2 \times (120 \text{ k}\Omega + R_{REF}) / (240 \text{ k}\Omega + R_{REF})$$

Only the positive signal path is amplified; the negative path is unaffected. This uneven amplification degrades the CMRR of the amplifier.

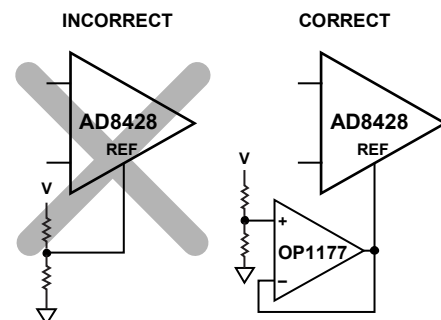


Figure 38. Driving the Reference Pin

## INPUT VOLTAGE RANGE

The 3-op-amp architecture of the AD8428 applies gain in the first stage before removing the common-mode voltage in the difference amplifier stage. Internal nodes between the first and second stages (Node 1 and Node 2 in Figure 37) experience a combination of an amplified differential signal, a common-mode signal, and a diode drop. This combined signal can be limited by the voltage supplies even when the individual input and output signals are not limited. Figure 9 shows the allowable input common-mode voltage ranges for various output voltages and supply voltages.

## LAYOUT

To ensure optimum performance of the AD8428 at the PCB level, care must be taken in the design of the board layout. The pins of the AD8428 are especially arranged to simplify board layout and to help minimize parasitic imbalance between the inputs.

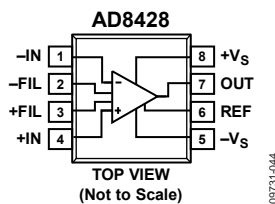


Figure 39. Pinout Diagram

## Common-Mode Rejection Ratio over Frequency

Poor layout can cause some of the common-mode signals to be converted to differential signals before reaching the in-amp. Such conversions occur when one input path has a frequency response that is different from the other. To maintain high CMRR over frequency, the input source impedance and capacitance of each path should be closely matched. Additional source resistance in the input paths (for example, for input protection) should be placed close to the in-amp inputs to minimize the interaction of the inputs with parasitic capacitance from the PCB traces.

Parasitic capacitance at the filter pins can also affect CMRR over frequency. If the board design has a component at the filter pins, the component should be chosen so that the parasitic capacitance is as small as possible.

## Power Supplies and Grounding

Use a stable dc voltage to power the instrumentation amplifier. Noise on the supply pins can adversely affect performance. See the PSRR performance curves in Figure 11 for more information.

Place a 0.1  $\mu\text{F}$  capacitor as close as possible to each supply pin. Because the length of the bypass capacitor leads is critical at high frequency, surface-mount capacitors are recommended. A parasitic inductance in the bypass ground trace works against the low impedance created by the bypass capacitor.

As shown in Figure 40, a 10  $\mu\text{F}$  capacitor can be used farther away from the device. For larger value capacitors, which are intended to be effective at lower frequencies, the current return path distance is less critical. In most cases, the 10  $\mu\text{F}$  capacitor can be shared by other precision integrated circuits.

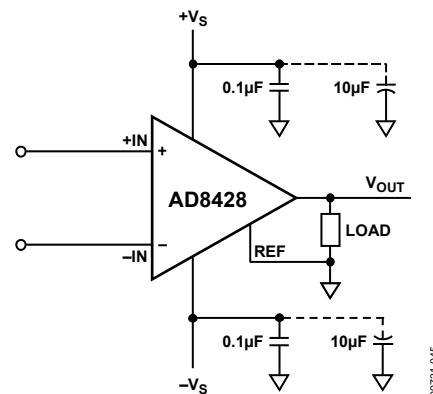


Figure 40. Supply Decoupling, REF, and Output Referred to Local Ground

A ground plane layer is helpful to reduce undesired parasitic inductances and to minimize voltage drops with changes in current. The area of the current path is directly proportional to the magnitude of parasitic inductances and, therefore, the impedance of the path at high frequency. Large changes in currents in an inductive decoupling path or ground return create unwanted effects due to the coupling of such changes into the amplifier inputs.

Because load currents flow from the supplies, the load should be connected at the same physical location as the bypass capacitor grounds.

## Reference Pin

The output voltage of the AD8428 is developed with respect to the potential on the reference terminal. Ensure that REF is tied to the appropriate local ground.

## INPUT BIAS CURRENT RETURN PATH

The input bias current of the AD8428 must have a return path to ground. When the source, such as a thermocouple, cannot provide a current return path, one should be created, as shown in Figure 41.

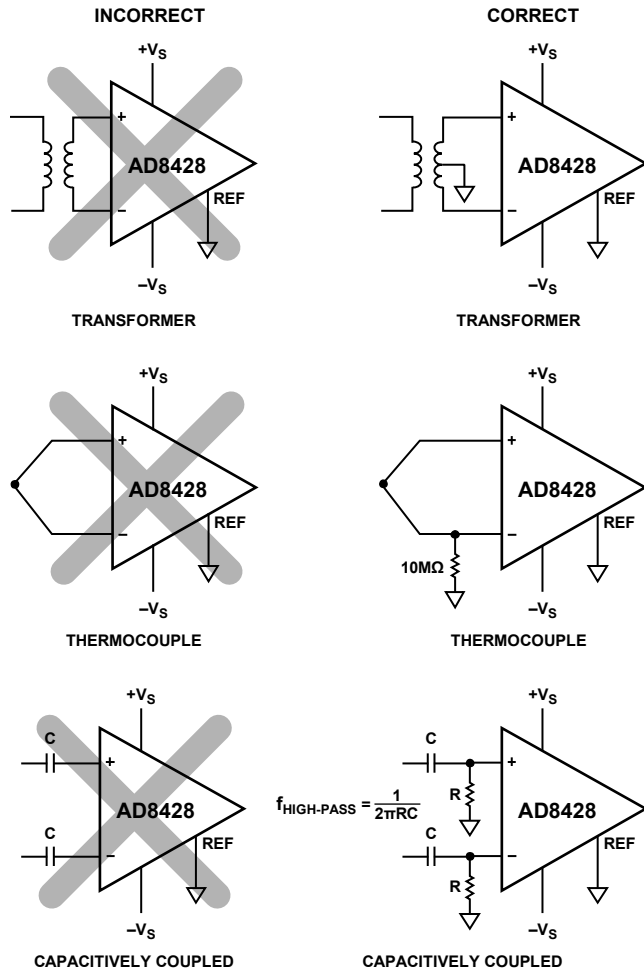


Figure 41. Creating an Input Bias Current Return Path

## INPUT PROTECTION

Do not allow the inputs of the AD8428 to exceed the ratings stated in the Absolute Maximum Ratings section. If these ratings cannot be adhered to, add protection circuitry in front of the AD8428 to limit the maximum current into the inputs (see the  $I_{MAX}$  section).

### $I_{MAX}$

The maximum current into the AD8428 inputs,  $I_{MAX}$ , depends on time and temperature. At room temperature, the device can withstand a current of 10 mA for at least one day. This time is cumulative over the life of the device.

## Input Voltages Beyond the Rails

If voltages beyond the rails are expected, use an external resistor in series with each input to limit current during overload conditions. The limiting resistor at each input can be computed using the following equation:

$$R_{PROTECT} \geq \frac{|V_{IN} - V_{SUPPLY}|}{I_{MAX}}$$

Noise sensitive applications may require a lower protection resistance. Low leakage diode clamps, such as the BAV199, can be used at the inputs to shunt current away from the AD8428 inputs and, therefore, allow smaller protection resistor values. To ensure that current flows primarily through the external protection diodes, place a small value resistor, such as a 33 Ω resistor, between the diodes and the AD8428.

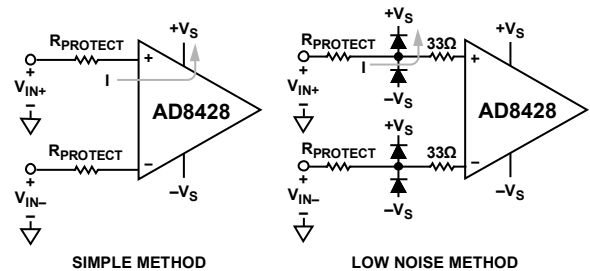


Figure 42. Protection for Voltages Beyond the Rails

## Large Differential Input Voltage at High Gain

If large differential voltages at high gain are expected, use an external resistor in series with each input to limit current during overload conditions. The limiting resistor at each input can be computed using the following equation:

$$R_{PROTECT} \geq \frac{1}{2} \times \left( \frac{|V_{DIFF}| - 1V}{I_{MAX}} - R_G \right)$$

Noise sensitive applications may require a lower protection resistance. Low leakage diode clamps, such as the BAV199, can be used across the AD8428 inputs to shunt current away from the inputs and, therefore, allow smaller protection resistor values.

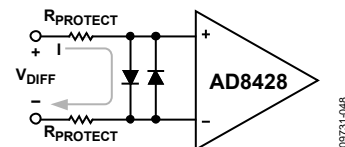


Figure 43. Protection for Large Differential Voltages

## RADIO FREQUENCY INTERFERENCE (RFI)

Because of its high gain and low noise properties, the AD8428 is a highly sensitive amplifier. Therefore, RF rectification can be a problem if the AD8428 is used in applications that have strong RF signal sources present. The problem is intensified if long leads or PCB traces are required to connect the amplifier to the signal source. The disturbance can appear as a dc offset voltage or a train of pulses.

High frequency signals can be filtered with a low-pass filter network at the input of the instrumentation amplifier, as shown in Figure 44.

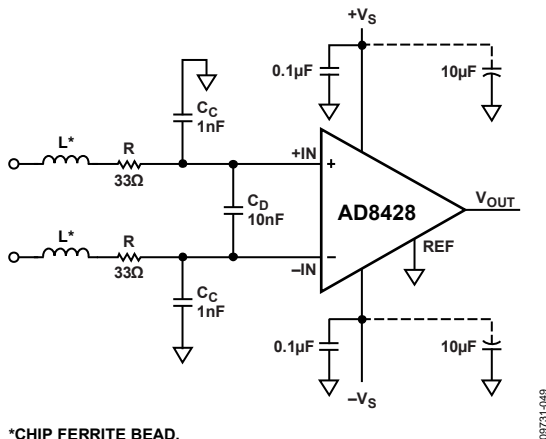


Figure 44. RFI Suppression

The filter limits both the differential and common-mode bandwidth, as shown in the following equations:

$$FilterFrequency_{DIFF} = \frac{1}{2\pi R(2C_D + C_C)}$$

$$FilterFrequency_{CM} = \frac{1}{2\pi RC_C}$$

where  $C_D \geq 10 C_C$ .

$C_D$  affects the differential signal, and  $C_C$  affects the common-mode signal. Choose values of  $R$  and  $C_C$  that minimize RFI. A mismatch between  $R \times C_C$  at the positive input and  $R \times C_C$  at the negative input degrades the CMRR of the AD8428. By using a value of  $C_D$  one order of magnitude larger than  $C_C$ , the effect of the mismatch is reduced, and performance is improved.

Resistors add noise; therefore, the choice of resistor and capacitor values depends on the desired trade-off between noise, input impedance at high frequencies, and RFI immunity. To achieve low noise and sufficient RFI filtering, the use of inductive ferrite beads is recommended (see Figure 44). Using inductive ferrite beads allows the value of the resistors to be reduced, which helps to minimize the noise at the input.

For best results, place the RFI filter network as close as possible to the amplifier. Layout is critical to ensure that RF signals are not picked up on the traces after the filter. If RF interference is too strong to be filtered, shielding is recommended.

Note that the resistors used for the RFI filter can be the same as those used for input protection (see the Input Protection section).

## CALCULATING THE NOISE OF THE INPUT STAGE

The total noise of the amplifier front end depends on much more than the specifications in this data sheet. The three main contributors to noise are as follows: the source resistance, the voltage noise of the instrumentation amplifier, and the current noise of the instrumentation amplifier.

In the following calculations, noise is referred to the input (RTI); that is, all sources of noise are calculated as if the source appeared at the amplifier input. To calculate the noise referred to the amplifier output (RTO), simply multiply the RTI noise by the gain of the instrumentation amplifier.

### Source Resistance Noise

Any sensor connected to the AD8428 has some output resistance. There may also be resistance placed in series with the inputs for protection from either overvoltage or radio frequency interference. This combined resistance is labeled  $R_1$  and  $R_2$  in Figure 45. Any resistor, no matter how well made, has an intrinsic level of noise. This noise is proportional to the square root of the resistor value. At room temperature, the value is approximately equal to  $4 \text{ nV}/\sqrt{\text{Hz}} \times \sqrt{(\text{resistor value in k}\Omega)}$ .

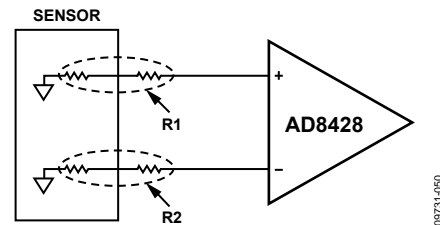


Figure 45. Source Resistance from Sensor and Protection Resistors

For example, assuming that the combined sensor and protection resistance is  $4 \text{ k}\Omega$  on the positive input and  $1 \text{ k}\Omega$  on the negative input, the total noise from the input resistance is

$$\sqrt{(4 \times \sqrt{4})^2 + (4 \times \sqrt{1})^2} = \sqrt{64 + 16} = 8.9 \text{ nV}/\sqrt{\text{Hz}}$$



**Voltage Noise of the Instrumentation Amplifier**

Unlike other instrumentation amplifiers in which an external resistor is used to set the gain, the voltage noise specification of the AD8428 already includes the input noise, output noise, and the  $R_G$  resistor noise.

**Current Noise of the Instrumentation Amplifier**

The contribution of current noise to the input stage in  $\text{nV}/\sqrt{\text{Hz}}$  is calculated by multiplying the source resistance in  $\text{k}\Omega$  by the specified current noise of the instrumentation amplifier in  $\text{pA}/\sqrt{\text{Hz}}$ .

For example, if the  $R_1$  source resistance in Figure 45 is  $4 \text{ k}\Omega$  and the  $R_2$  source resistance is  $1 \text{ k}\Omega$ , the total effect from the current noise is calculated as follows:

$$\sqrt{(4 \times 1.5)^2 + (1 \times 1.5)^2} = 6.2 \text{ nV}/\sqrt{\text{Hz}}$$

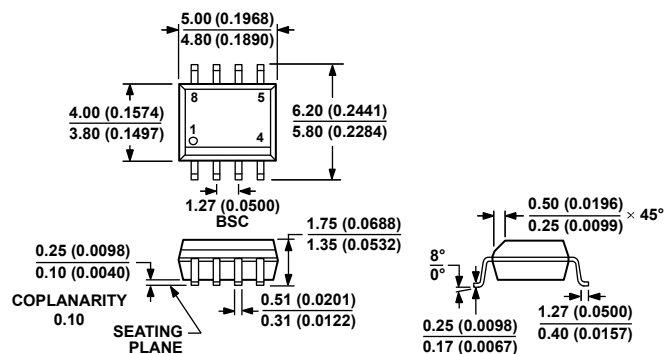
**Total Noise Density Calculation**

To determine the total noise of the in-amp, referred to input, combine the source resistance noise, voltage noise, and current noise contribution by the sum of squares method.

For example, if the  $R_1$  source resistance in Figure 45 is  $4 \text{ k}\Omega$  and the  $R_2$  source resistance is  $1 \text{ k}\Omega$ , the total noise, referred to input, is

$$\sqrt{8.9^2 + 1.5^2 + 6.2^2} = 11.0 \text{ nV}/\sqrt{\text{Hz}}$$

## OUTLINE DIMENSIONS



COMPLIANT TO JEDEC STANDARDS MS-012-AA  
CONTROLLING DIMENSIONS ARE IN MILLIMETERS; INCH DIMENSIONS  
(IN PARENTHESES) ARE ROUNDED-OFF MILLIMETER EQUIVALENTS FOR  
REFERENCE ONLY AND ARE NOT APPROPRIATE FOR USE IN DESIGN.

Figure 46. 8-Lead Standard Small Outline Package [SOIC\_N]  
Narrow Body  
(R-8)

Dimensions shown in millimeters and (inches)

012407-A

## ORDERING GUIDE

Model <sup>1</sup>	Temperature Range	Package Description	Package Option
AD8428ARZ	−40°C to +125°C	8-Lead SOIC_N	R-8
AD8428ARZ-RL	−40°C to +125°C	8-Lead SOIC_N, 13" Tape and Reel	R-8

<sup>1</sup> Z = RoHS Compliant Part.

NOTES

**NOTES**