

## Low Power Instrumentation Amplifier

The HT8620 is a low cost, high accuracy instrumentation amplifier that requires only one external resistor to set gains of and DIP packaging that is smaller than discrete designs and offers lower power (only 1.3 mA max supply current), making The HT8620, with its high accuracy of 40 ppm maximum nonlinearity, low offset voltage of 50  $\mu\text{V}$  max, and offset drift of 0.6  $\mu\text{V}/^\circ\text{C}$  max, is ideal for use in precision data acquisition systems, such as weigh scales and transducer interfaces. Furthermore, the low noise, low input bias current, and low power of the HT8620 make it well suited for medical applications, such as ECG and noninvasive blood pressure monitors.

The low input bias current of 1.0 nA max is made possible with the use of Superbeta processing in the input stage. The HT8620 works well as a preamplifier due to its low input voltage noise of 9 nV/ $\sqrt{\text{Hz}}$  at 1 kHz, 0.28  $\mu\text{V}$  p-p in the 0.1 Hz to 10 Hz band, and 0.1 pA/ $\sqrt{\text{Hz}}$  input current noise. Also, the HT8620 is well suited for multiplexed applications with its settling time of 15  $\mu\text{s}$  to 0.01%, and its cost is

### FEATURES

#### Easy to use

Gain set with one external resistor

(Gain range 1 to 10,000)

Wide power supply range ( $\pm 2.3\text{ V}$  to  $\pm 18\text{ V}$ )

Higher performance than 3 op amp IA designs

Available in 8-lead DIP and SOIC packaging

Low power, 1.3 mA max supply current

#### Excellent dc performance (B grade)

50  $\mu\text{V}$  max, input offset voltage

0.6  $\mu\text{V}/^\circ\text{C}$  max, input offset drift

1.0 nA max, input bias current

100dB min common-mode rejection ratio (G)

#### Low noise

9 nV/ $\sqrt{\text{Hz}}$  @ 1 kHz, input voltage noise

0.28  $\mu\text{V}$  p-p noise (0.1 Hz to 10 Hz)

#### Excellent ac specifications

120 kHz bandwidth (G = 100)

15  $\mu\text{s}$  settling time to 0.01%

### APPLICATIONS

Weigh scales

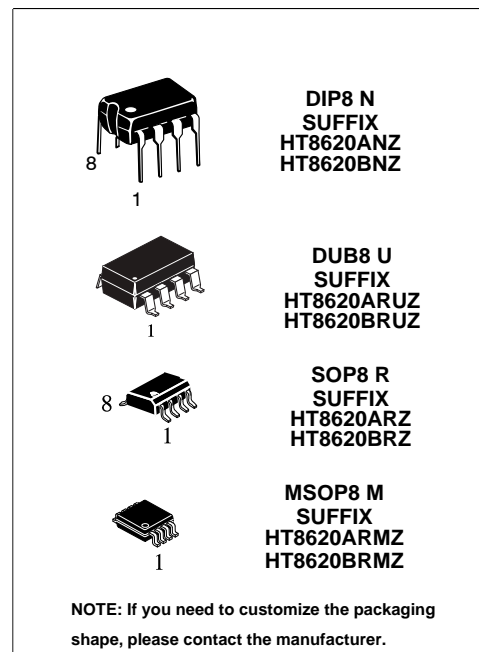
ECG and medical instrumentation

Transducer interface

Data acquisition systems

Industrial process controls

Battery-powered and portable equipment



### CONNECTION DIAGRAM

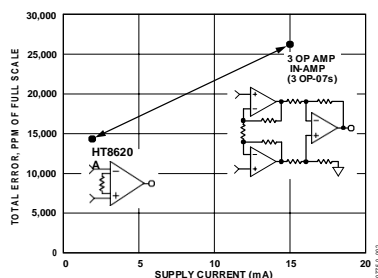
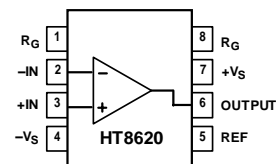


Figure 2. Three Op Amp IA Designs vs. HT8620

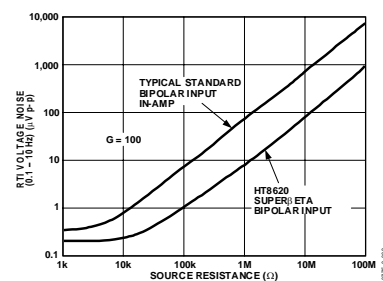


Figure 3. Total Voltage Noise vs. Source Resistance

## SPECIFICATIONS

Typical @ 25 °C,  $V_S = \pm 15$  V, and  $R_L = 2$  k $\Omega$ , unless otherwise noted.

Table 1.

Parameter	Conditions	HT8620A			HT8620B			HT8620S			Unit
		Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	
<b>GAIN</b>	$G = 1 + (49.4 \text{ k}\Omega / R_G)$										
Gain Range		1		10,000	1		10,000	1		10,000	
Gain Error <sup>2</sup>	$V_{OUT} = \pm 10$ V										
G = 1			0.03	0.10		0.01	0.02		0.03	0.10	%
G = 10			0.15	0.30		0.10	0.15		0.15	0.30	%
G = 100			0.15	0.30		0.10	0.15		0.15	0.30	%
G = 1000			0.40	0.70		0.35	0.50		0.40	0.70	%
Nonlinearity	$V_{OUT} = -10$ V to $+10$ V										
G = 1–1000	$R_L = 10$ k $\Omega$		10	40		10	40		10	40	ppm
G = 1–100	$R_L = 2$ k $\Omega$		10	95		10	95		10	95	ppm
Gain vs. Temperature	G = 1			10			10			10	ppm/°C
	Gain $> 1^2$			-50			-50			-50	ppm/°C
<b>VOLTAGE OFFSET</b>	(Total RTI Error = $V_{OSI} + V_{OSO}/G$ )										
Input Offset, $V_{OSI}$	$V_S = \pm 5$ V to $\pm 15$ V		30	125		15	50		30	125	$\mu$ V
Overtemperature	$V_S = \pm 5$ V to $\pm 15$ V			185			85			225	$\mu$ V
Average TC	$V_S = \pm 5$ V to $\pm 15$ V		0.3	1.0		0.1	0.6		0.3	1.0	$\mu$ V/°C
Output Offset, $V_{OSO}$	$V_S = \pm 15$ V		400	1000		200	500		400	1000	$\mu$ V
Overtemperature	$V_S = \pm 5$ V to $\pm 15$ V			1500			750			1500	$\mu$ V
Average TC	$V_S = \pm 5$ V to $\pm 15$ V		5.0	15		2.5	7.0		5.0	15	$\mu$ V/°C
Offset Referred to the Input vs. Supply (PSR)	$V_S = \pm 2.3$ V to $\pm 18$ V										
G = 1		80	100		80	100		80	100		dB
G = 10		95	120		100	120		95	120		dB
G = 100		110	140		120	140		110	140		dB
G = 1000		110	140		120	140		110	140		dB
<b>INPUT CURRENT</b>											
Input Bias Current			0.5	2.0		0.5	1.0		0.5	2	nA
Overtemperature				2.5			1.5			4	nA
Average TC			3.0			3.0			8.0		pA/°C
Input Offset Current			0.3	1.0		0.3	0.5		0.3	1.0	nA
Overtemperature				1.5			0.75			2.0	nA
Average TC			1.5			1.5			8.0		pA/°C
<b>INPUT</b>											
Input Impedance											
Differential			10  2			10  2			10  2		G $\Omega$ _pF
Common-Mode			10  2			10  2			10  2		G $\Omega$ _pF
Input Voltage Range <sup>3</sup>	$V_S = \pm 2.3$ V to $\pm 5$ V	$-V_S + 1.9$		$+V_S - 1.2$	$-V_S + 1.9$		$+V_S - 1.2$	$-V_S + 1.9$		$+V_S - 1.2$	V
Overtemperature	$V_S = \pm 5$ V to $\pm 18$ V	$-V_S + 2.1$		$+V_S - 1.3$	$-V_S + 2.1$		$+V_S - 1.3$	$-V_S + 2.1$		$+V_S - 1.3$	V
		$-V_S + 1.9$		$+V_S - 1.4$	$-V_S + 1.9$		$+V_S - 1.4$	$-V_S + 1.9$		$+V_S - 1.4$	V
Overtemperature		$-V_S + 2.1$		$+V_S - 1.4$	$-V_S + 2.1$		$+V_S + 2.1$	$-V_S + 2.3$		$+V_S - 1.4$	V

Parameter	Conditions	HT8620A			HT8620B			HT8620S			Unit	
		Min	Typ	Max	Min	Typ	Max	Min	Typ	Max		
Common-Mode Rejection												
Ratio DC to 60 Hz with 1 kΩ Source Imbalance	$V_{CM} = 0\text{ V to } \pm 10\text{ V}$											
G = 1		73	90		80	90		73	90		dB	
G = 10		93	110		100	110		93	110		dB	
G = 100		110	130		120	130		110	130		dB	
G = 1000		110	130		120	130		110	130		dB	
OUTPUT												
Output Swing	$R_L = 10\text{ k}\Omega$ $V_S = \pm 2.3\text{ V}$ to $\pm 5\text{ V}$  $V_S = \pm 5\text{ V}$ to $\pm 18\text{ V}$											
		$-V_S + 1.1$		$+V_S - 1.2$	$-V_S + 1.1$		$+V_S - 1.2$	$-V_S + 1.1$		$+V_S - 1.2$	V	
Overttemperature		$-V_S + 1.4$		$+V_S - 1.3$	$-V_S + 1.4$		$+V_S - 1.3$	$-V_S + 1.6$		$+V_S - 1.3$	V	
		$-V_S + 1.2$		$+V_S - 1.4$	$-V_S + 1.2$		$+V_S - 1.4$	$-V_S + 1.2$		$+V_S - 1.4$	V	
Overttemperature		$-V_S + 1.6$		$+V_S - 1.5$	$-V_S + 1.6$		$+V_S - 1.5$	$-V_S + 2.3$		$+V_S - 1.5$	V	
Short Circuit Current		$\pm 18$			$\pm 18$			$\pm 18$			mA	
DYNAMIC RESPONSE												
Small Signal -3 dB Bandwidth	10 V Step											
G = 1		1000			1000			1000			kHz	
G = 10		800			800			800			kHz	
G = 100		120			120			120			kHz	
G = 1000		12			12			12			kHz	
Slew Rate		0.75	1.2		0.75	1.2		0.75	1.2		V/μs	
Settling Time to 0.01%												
G = 1-100		15			15			15			μs	
G = 1000	150			150			150			μs		
NOISE												
Voltage Noise, 1 kHz	$Total\ RTI\ Noise = \sqrt{(e_{ni}^2) + (e_{no}/G)^2}$											
Input, Voltage Noise, $e_{ni}$		9	13		9	13		9	13		nV/√Hz	
Output, Voltage Noise, $e_{no}$		72	100		72	100		72	100		nV/√Hz	
RTI, 0.1 Hz to 10 Hz												
G = 1		3.0			3.0	6.0		3.0	6.0		μV p-p	
G = 10		0.55			0.55	0.8		0.55	0.8		μV p-p	
G = 100-1000		0.28			0.28	0.4		0.28	0.4		μV p-p	
Current Noise		f = 1 kHz	100			100			100			fA/√Hz
0.1 Hz to 10 Hz			10			10			10			pA p-p
REFERENCE INPUT												
$R_{IN}$	$V_{IN+}, V_{REF} = 0$	20			20			20			kΩ	
$I_{IN}$		50	60		50	60		50	60		μA	
Voltage Range		$-V_S + 1.6$		$+V_S - 1.6$	$-V_S + 1.6$		$+V_S - 1.6$	$-V_S + 1.6$		$+V_S - 1.6$	V	
Gain to Output		$1 \pm 0.0001$			$1 \pm 0.0001$			$1 \pm 0.0001$				
POWER SUPPLY												
Operating Range <sup>4</sup>	$V_S = \pm 2.3\text{ V}$ to $\pm 18\text{ V}$	$\pm 2.3$		$\pm 18$	$\pm 2.3$		$\pm 18$	$\pm 2.3$		$\pm 18$	V	
Quiescent Current		0.9	1.3		0.9	1.3		0.9	1.3		mA	
Overttemperature		1.1	1.6		1.1	1.6		1.1	1.6		mA	
TEMPERATURE RANGE												
For Specified Performance		$-40\text{ to }+125$			$-40\text{ to }+125$			$-55\text{ to }+125$			°C	

<sup>1</sup> See Analog Devices military data sheet for 883B tested specifications.

<sup>2</sup> Does not include effects of external resistor  $R_G$ .

<sup>3</sup> One input grounded.  $G = 1$ .

<sup>4</sup> This is defined as the same supply range that is used to specify PSR.

## TYPICAL PERFORMANCE CHARACTERISTICS

(@ 25 °C,  $V_S = \pm 15$  V,  $R_L = 2$  k $\Omega$ , unless otherwise noted.)

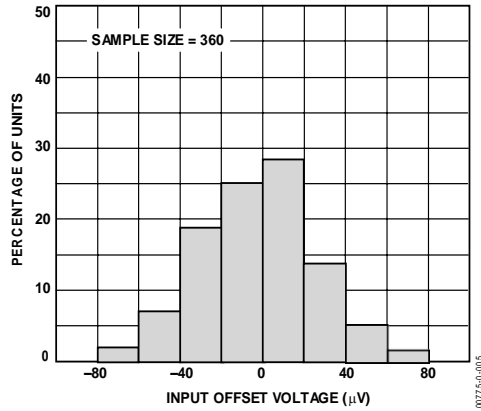


Figure 5. Typical Distribution of Input Offset Voltage

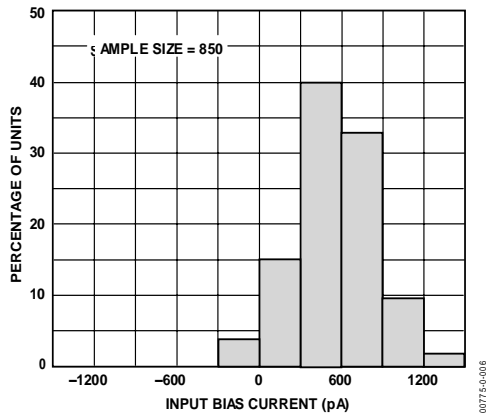


Figure 6. Typical Distribution of Input Bias Current

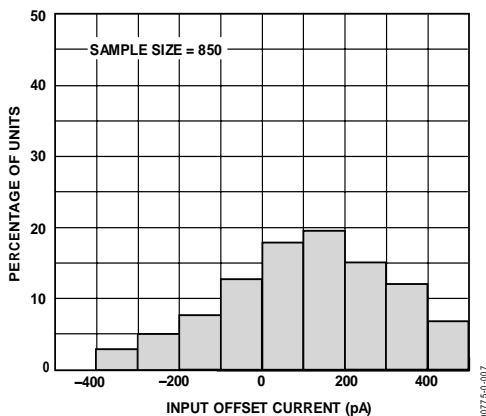


Figure 7. Typical Distribution of Input Offset Current

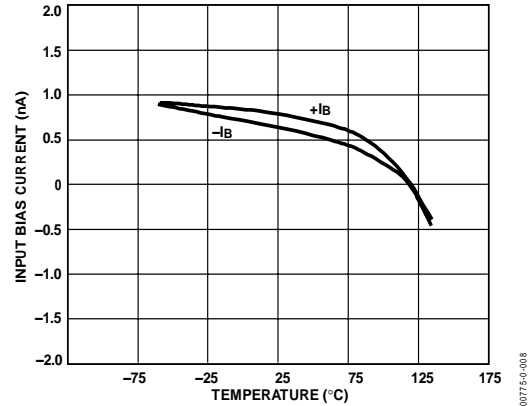


Figure 8. Input Bias Current vs. Temperature

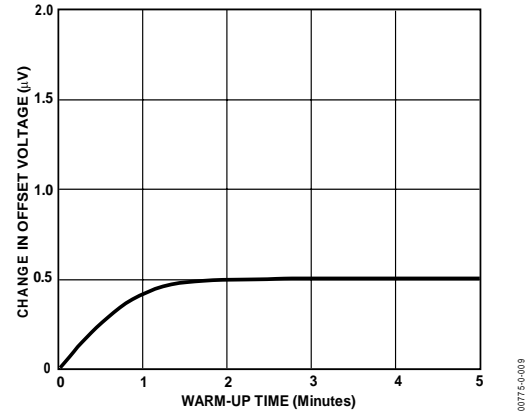


Figure 9. Change in Input Offset Voltage vs. Warm-Up Time

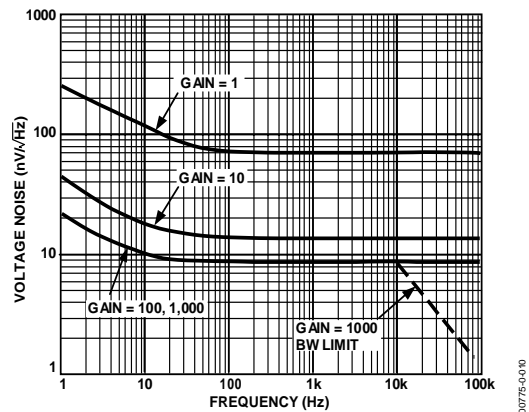


Figure 10. Voltage Noise Spectral Density vs. Frequency ( $G = 1-1000$ )

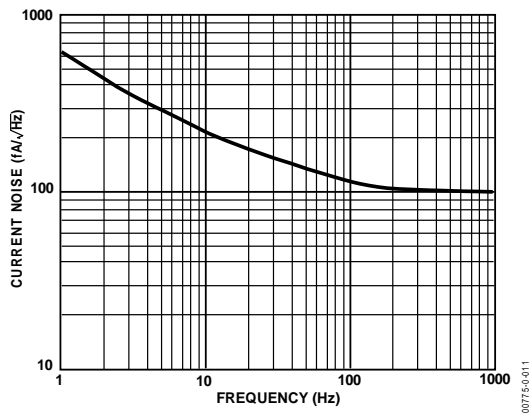


Figure 11. Current Noise Spectral Density vs. Frequency

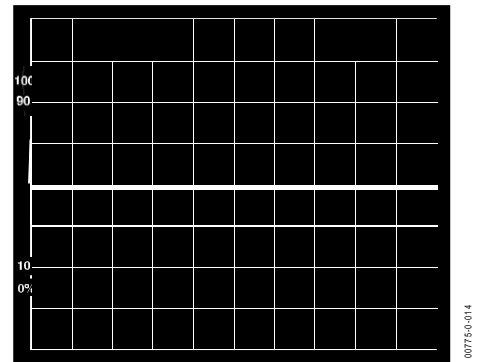


Figure 14. 0.1 Hz to 10 Hz Current Noise, 5 pA/Div

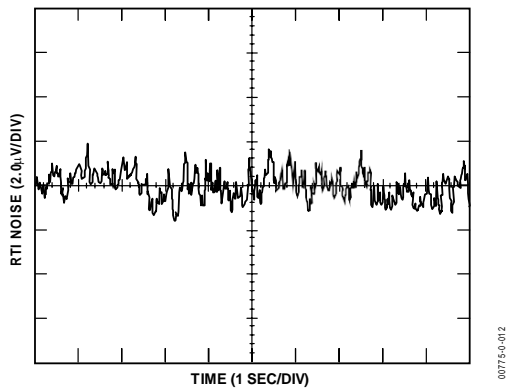


Figure 12. 0.1 Hz to 10 Hz RTI Voltage Noise (G = 1)

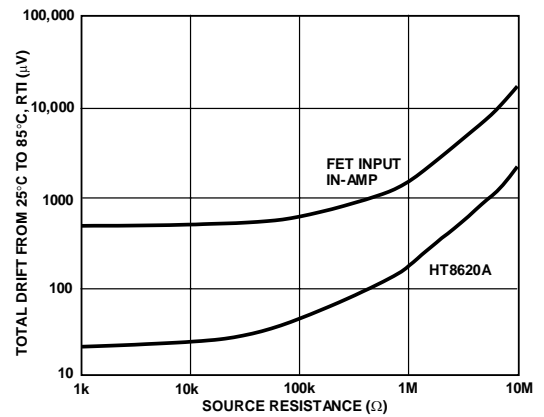


Figure 15. Total Drift vs. Source Resistance

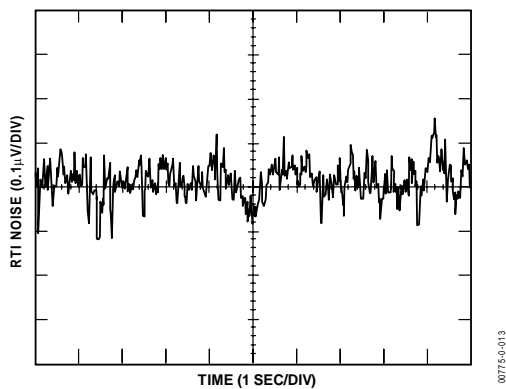


Figure 13. 0.1 Hz to 10 Hz RTI Voltage Noise (G = 1000)

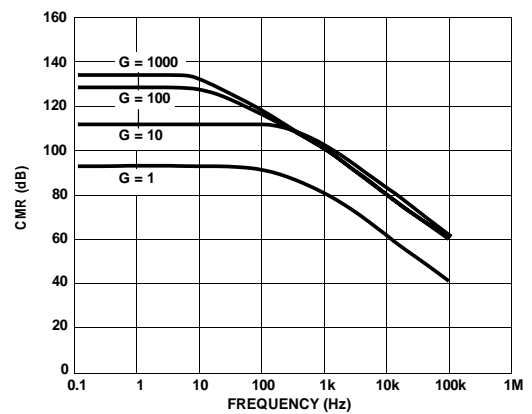


Figure 16. Typical CMR vs. Frequency, RTI, Zero to 1 kΩ Source Imbalance

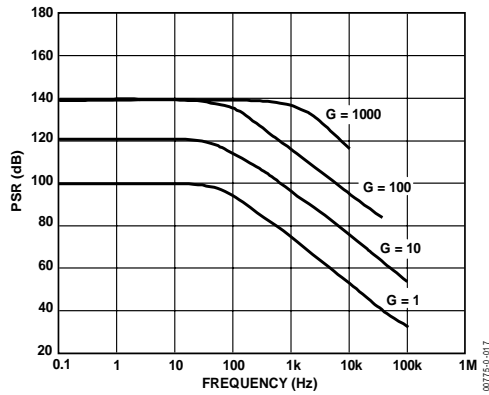


Figure 17. Positive PSR vs. Frequency, RTI (G=1-1000)

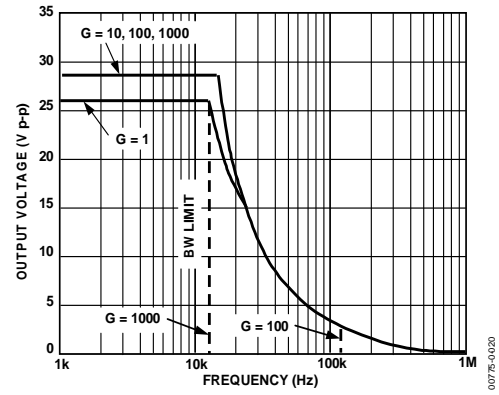


Figure 20. Large Signal Frequency Response

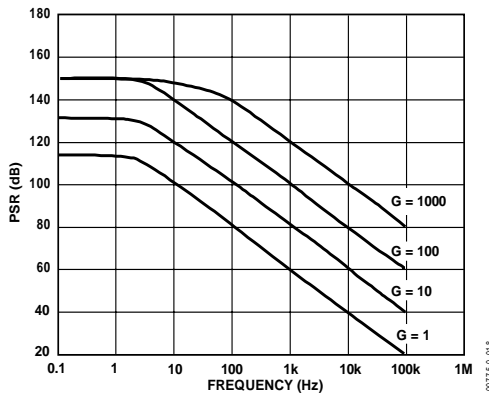


Figure 18. Negative PSR vs. Frequency, RTI (G=1-1000)

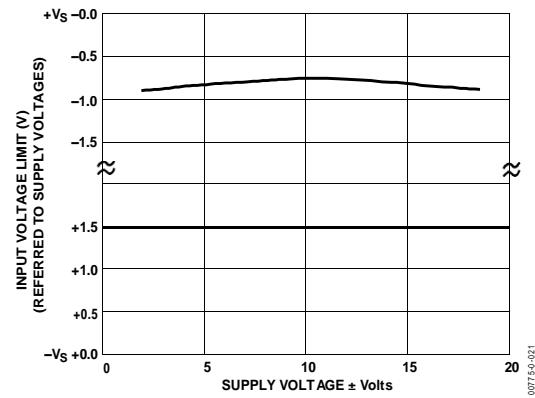


Figure 21. Input Voltage Range vs. Supply Voltage, G=1

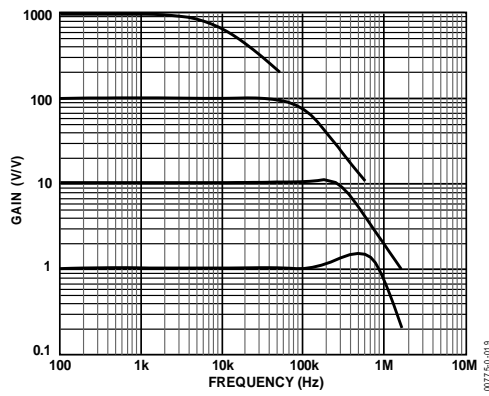


Figure 19. Gain vs. Frequency

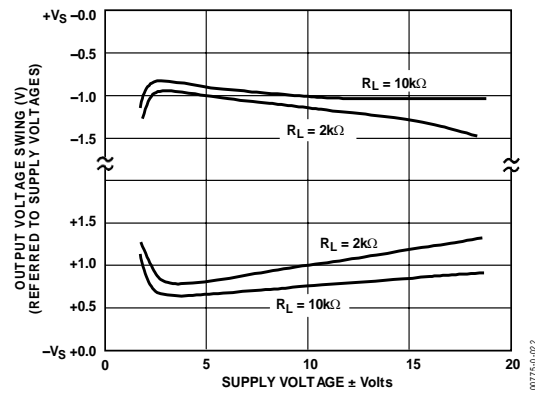


Figure 22. Output Voltage Swing vs. Supply Voltage, G=10

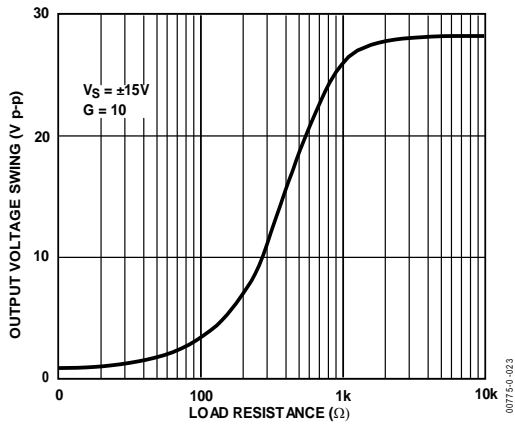


Figure 23. Output Voltage Swing vs. Load Resistance

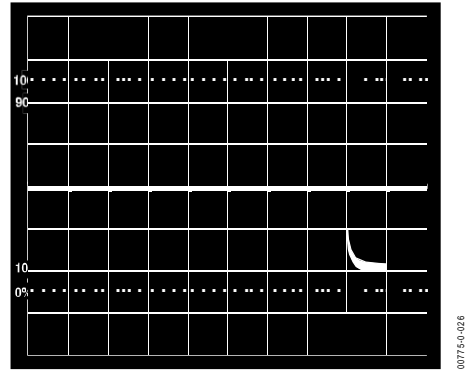


Figure 26. Large Signal Response and Settling Time,  $G = 10$  ( $0.5 \text{ mV} = 0.01\%$ )

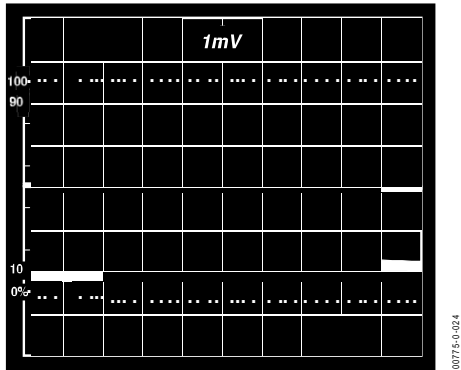


Figure 24. Large Signal Pulse Response and Settling Time  
 $G = 1$  ( $0.5 \text{ mV} = 0.01\%$ )

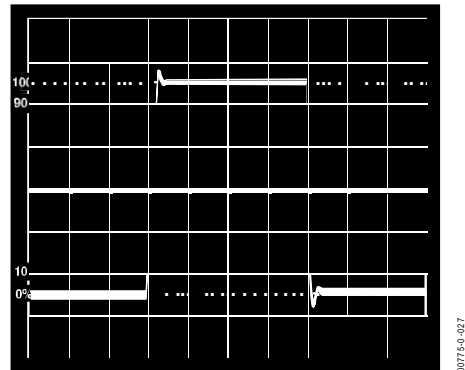


Figure 27. Small Signal Response,  $G = 10$ ,  $R_L = 2 \text{ k}\Omega$ ,  $C_L = 100 \text{ pF}$

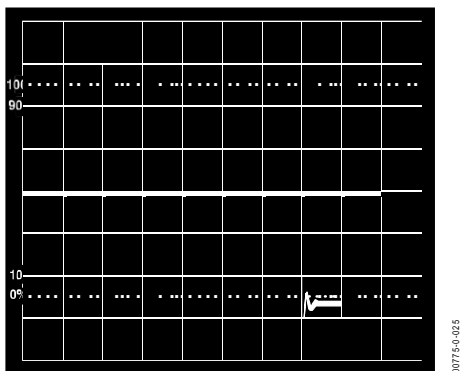


Figure 25. Small Signal Response,  $G = 1$ ,  $R_L = 2 \text{ k}\Omega$ ,  $C_L = 100 \text{ pF}$

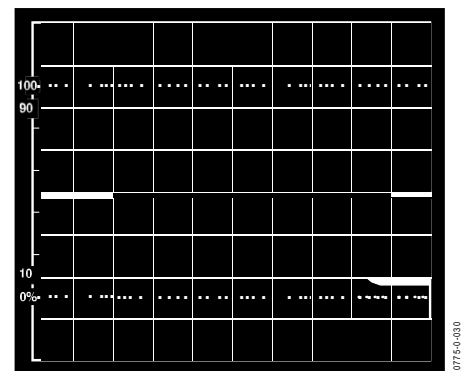


Figure 28. Large Signal Response and Settling Time,  $G = 100$  ( $0.5 \text{ mV} = 0.01\%$ )

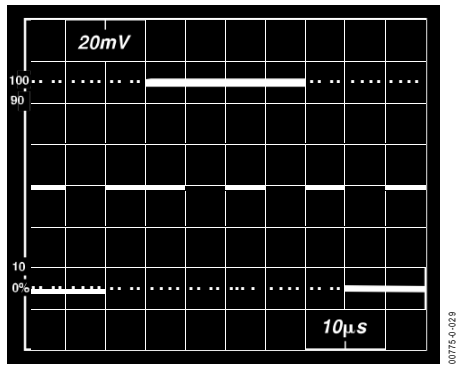


Figure 29. Small Signal Pulse Response,  $G = 100$ ,  $R_L = 2\text{ k}\Omega$ ,  $C_L = 100\text{ pF}$

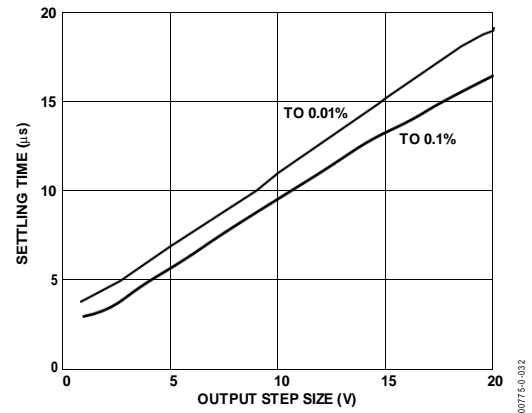


Figure 32. Settling Time vs. Step Size ( $G = 1$ )

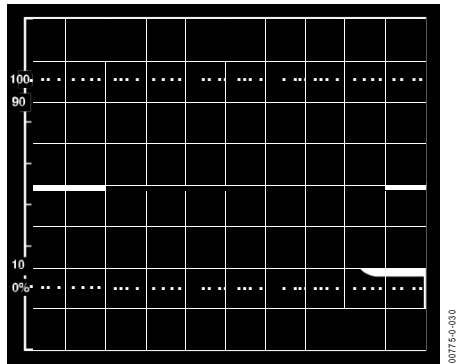


Figure 30. Large Signal Response and Settling Time,  
 $G = 1000$  ( $0.5\text{ mV} = 0.01\%$ )

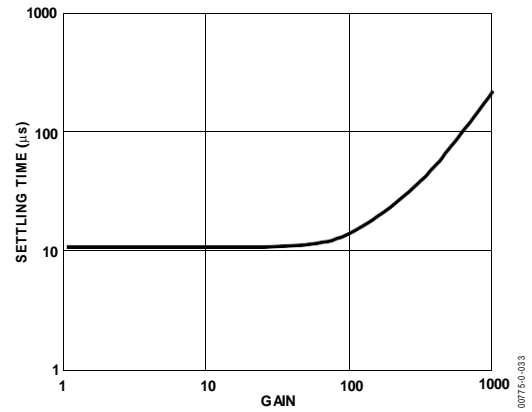


Figure 33. Settling Time to 0.01% vs. Gain, for a 10 V Step

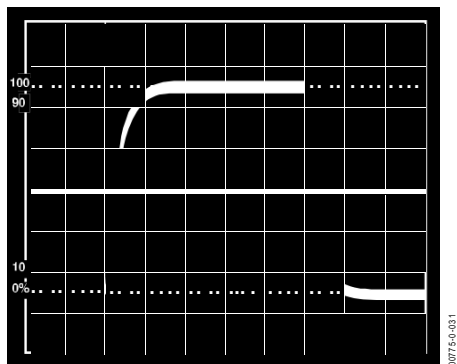


Figure 31. Small Signal Pulse Response,  $G = 1000$ ,  $R_L = 2\text{ k}\Omega$ ,  $C_L = 100\text{ pF}$

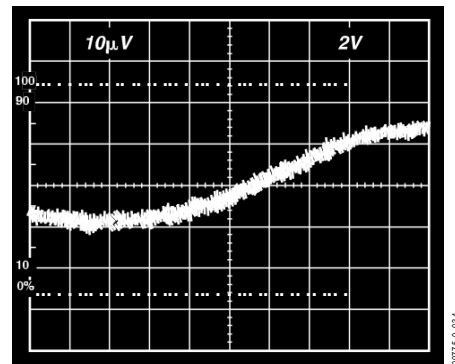


Figure 34. Gain Nonlinearity,  $G = 1$ ,  $R_L = 10\text{ k}\Omega$  ( $10\mu\text{V} = 1\text{ ppm}$ )



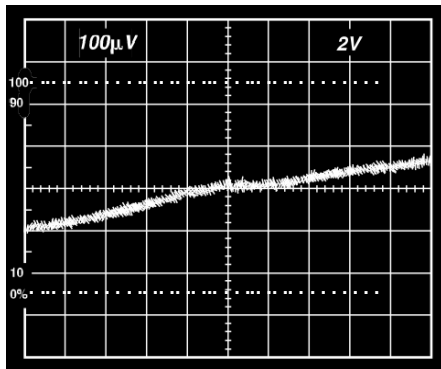


Figure 35. Gain Nonlinearity,  $G = 100$ ,  $R_L = 10\text{ k}\Omega$   
( $100\text{ }\mu\text{V} = 10\text{ ppm}$ )

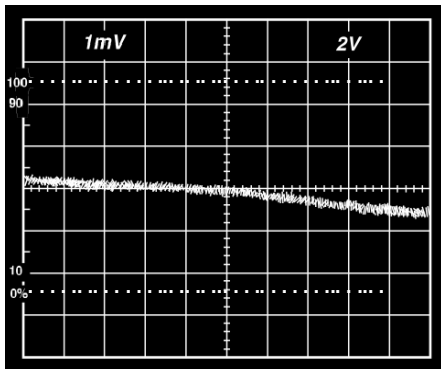


Figure 36. Gain Nonlinearity,  $G = 1000$ ,  $R_L = 10\text{ k}\Omega$   
( $1\text{ mV} = 100\text{ ppm}$ )

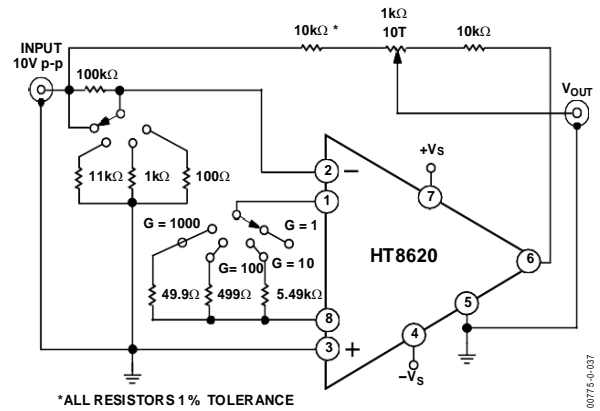


Figure 37. Settling Time Test Circuit

## THEORY OF OPERATION

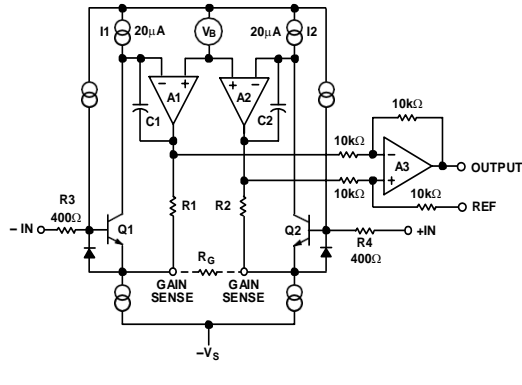


Figure 38. Simplified Schematic of HT8620

The HT8620 is a monolithic instrumentation amplifier based on a modification of the classic three op amp approach. Absolute value trimming allows the user to program gain accurately (to 0.15% at  $G = 100$ ) with only one resistor. Monolithic construction and laser wafer trimming allow the tight matching and tracking of circuit components, thus ensuring the high level of performance inherent in this circuit.

The input transistors Q1 and Q2 provide a single differential-pair bipolar input for high precision (Figure 38), yet offer  $10\times$  lower input bias current thanks to Superβ processing. Feedback through the Q1-A1-R1 loop and the Q2-A2-R2 loop maintains constant collector current of the input devices Q1 and Q2, thereby impressing the input voltage across the external gain setting resistor  $R_G$ . This creates a differential gain from the inputs to the A1/A2 outputs given by  $G = (R1 + R2)/R_G + 1$ . The unity-gain subtractor, A3, removes any common-mode signal, yielding a single-ended output referred to the REF pin potential.

The value of  $R_G$  also determines the transconductance of the preamp stage. As  $R_G$  is reduced for larger gains, the transconductance increases asymptotically to that of the input transistors. This has three important advantages: (a) Open-loop gain is boosted for increasing programmed gain, thus reducing gain related errors. (b) The gain-bandwidth product (determined by C1 and C2 and the preamp transconductance) increases with programmed gain, thus optimizing frequency response. (c) The input voltage noise is reduced to a value of  $9 \text{ nV}/\sqrt{\text{Hz}}$ , determined mainly by the collector current and base resistance of the input devices.

The internal gain resistors, R1 and R2, are trimmed to an absolute value of  $24.7 \text{ k}\Omega$ , allowing the gain to be programmed accurately with a single external resistor.

The gain equation is then

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

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

### Make vs. Buy: a Typical Bridge Application Error Budget

The HT8620 offers improved performance over “homebrew” three op amp IA designs, along with smaller size, fewer components, and  $10\times$  lower supply current. In the typical application, shown in Figure 39, a gain of 100 is required to amplify a bridge output of 20 mV full-scale over the industrial temperature range of  $-40^\circ\text{C}$  to  $+85^\circ\text{C}$ . Table 3 shows how to calculate the effect various error sources have on circuit accuracy.

Regardless of the system in which it is being used, the HT8620 provides greater accuracy at low power and price. In simple systems, absolute accuracy and drift errors are by far the most significant contributors to error. In more complex systems with an intelligent processor, an autogain/autozero cycle will remove all absolute accuracy and drift errors, leaving only the resolution errors of gain, nonlinearity, and noise, thus allowing full 14-bit accuracy.

Note that for the homebrew circuit, the OP07 specifications for input voltage offset and noise have been multiplied by  $\sqrt{2}$ . This is because a three op amp type in-amp has two op amps at its inputs, both contributing to the overall input error.

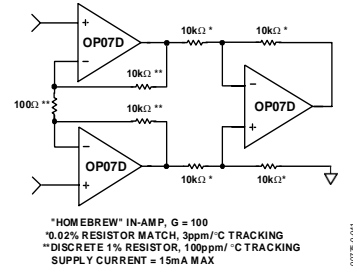
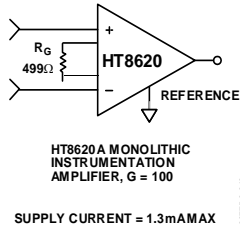
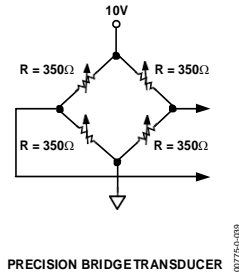


Figure 39. Make vs. Buy

Table 3. Make vs. Buy Error Budget

Error Source	HT8620 Circuit Calculation	“Homebrew” Circuit Calculation	Error, ppm of Full Scale	
			HT8620	Homebrew
ABSOLUTE ACCURACY at T <sub>A</sub> = 25°C				
Input Offset Voltage, μV	125 μV/20 mV	(150 μV × √2)/20 mV	6,250	10,607
Output Offset Voltage, μV	1000 μV/100 mV/20 mV	((150 μV × 2)/100)/20 mV	500	150
Input Offset Current, nA	2 nA ×350 Ω/20 mV	(6 nA ×350 Ω)/20 mV	18	53
CMR, dB	110 dB(3.16 ppm) x5 V/20 mV	(0.02% Match x 5 V)/20 mV/100	791	500
		Total Absolute Error	7,559	11,310
DRIFT TO 85°C				
Gain Drift, ppm/°C	(50 ppm + 10 ppm) x60°C	100 ppm/°C Track x 60°C	3,600	6,000
Input Offset Voltage Drift, μV/°C	1 μV/°C x 60°C/20 mV	(2.5 μV/°C x √2 x 60°C)/20 mV	3,000	10,607
Output Offset Voltage Drift, μV/°C	15 μV/°C x 60°C/100 mV/20 mV	(2.5 μV/°C x 2 x 60°C)/100 mV/20 mV	450	150
		Total Drift Error	7,050	16,757
RESOLUTION				
Gain Nonlinearity, ppm of Full Scale	40 ppm	40 ppm	40	40
Typ0.1 Hzto 10 Hz Voltage Noise, μVp-p	0.28 μV p-p/20 mV	(0.38 μV p-p x √2)/20 mV	14	27
		Total Resolution Error	54	67
		Grand Total Error	14,663	28,134

G = 100, V<sub>S</sub> =  $\pm 15 \text{ V}$ .

(All errors are min/max and referred to input.)

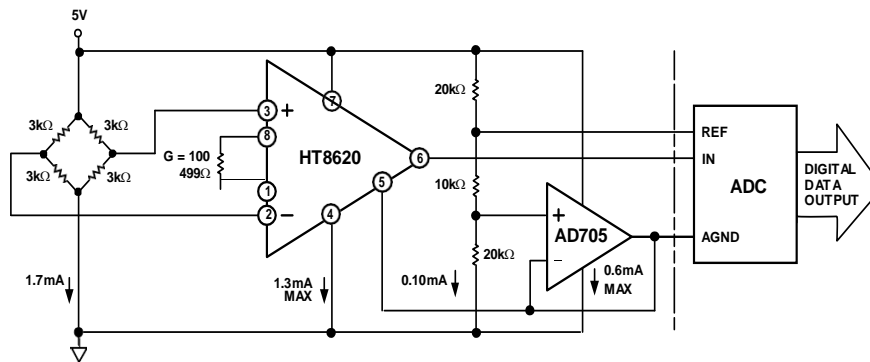


Figure 40. A Pressure Monitor Circuit that Operates on a 5 V Single Supply

### Pressure Measurement

Although useful in many bridge applications, such as weigh scales, the HT8620 is especially suitable for higher resistance pressure sensors powered at lower voltages where small size and low power become more significant.

Figure 40 shows a 3 kΩ pressure transducer bridge powered from 5 V. In such a circuit, the bridge consumes only 1.7 mA. Adding the HT8620 and a buffered voltage divider allows the signal to be conditioned for only 3.8 mA of total supply current.

Small size and low cost make the HT8620 especially attractive for voltage output pressure transducers. Since it delivers low noise and drift, it will also serve applications such as diagnostic noninvasive blood pressure measurement.

### Medical ECG

The low current noise of the HT8620 allows its use in ECG monitors (Figure 41) where high source resistances of 1 MΩ or higher are not uncommon. The HT8620's low power, low supply voltage requirements, and space-saving 8-lead mini-DIP and SOIC package offerings make it an excellent choice for battery-powered data recorders.

Furthermore, the low bias currents and low current noise, coupled with the low voltage noise of the HT8620, improve the dynamic range for better performance.

The value of capacitor C1 is chosen to maintain stability of the right leg drive loop. Proper safeguards, such as isolation, must be added to this circuit to protect the patient from possible harm.

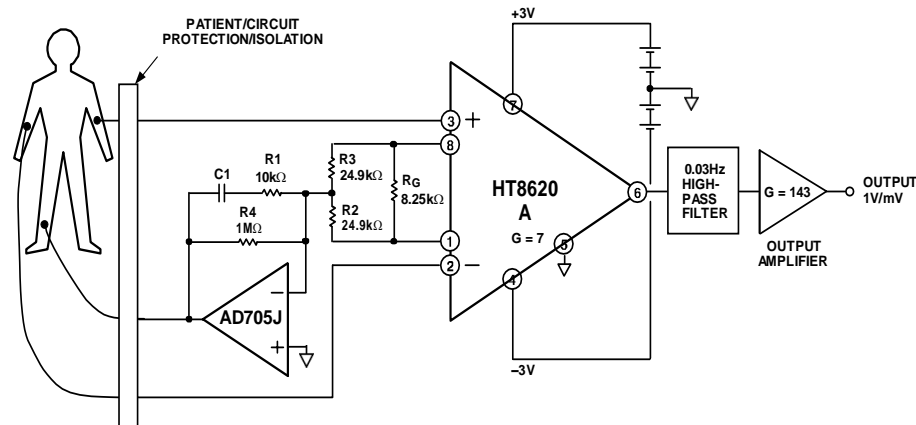


Figure 41. A Medical ECG Monitor Circuit

### Precision V-I Converter

The HT8620, along with another op amp and two resistors, makes a precision current source (Figure 42). The op amp buffers the reference terminal to maintain good CMR. The output voltage,  $V_x$ , of the HT8620 appears across  $R_1$ , which converts it to a current. This current, less only the input bias current of the op amp, then flows out to the load.

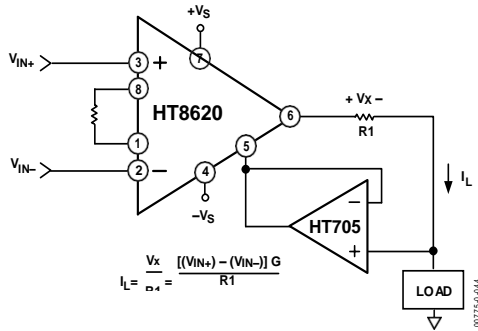


Figure 42. Precision Voltage-to-Current Converter (Operates on 1.8mA,  $\pm 3V$ )

### GAIN SELECTION

The HT8620's gain is resistor-programmed by  $R_G$ , or more precisely, by whatever impedance appears between Pins 1 and 8. The HT8620 is designed to offer accurate gains using 0.1% to 1% resistors. Table 4 shows required values of  $R_G$  for various gains. Note that for  $G = 1$ , the  $R_G$  pins are unconnected ( $R_G = \infty$ ). For any arbitrary gain,  $R_G$  can be calculated by using the formula:

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

To minimize gain error, avoid high parasitic resistance in series with  $R_G$ ; to minimize gain drift,  $R_G$  should have a low TC—less than 10 ppm/°C—for the best performance.

Table 4. Required Values of Gain Resistors

1% Std Table Value of $R_G(\Omega)$	Calculated Gain	0.1% Std Table Value of $R_G(\Omega)$	Calculated Gain
49.9 k	1.990	49.3 k	2.002
12.4 k	4.984	12.4 k	4.984
5.49 k	9.998	5.49 k	9.998
2.61 k	19.93	2.61 k	19.93
1.00 k	50.40	1.01 k	49.91
499	100.0	499	100.0
249	199.4	249	199.4
100	495.0	98.8	501.0
49.9	991.0	49.3	1,003.0

### INPUT AND OUTPUT OFFSET VOLTAGE

The low errors of the HT8620 are attributed to two sources, input and output errors. The output error is divided by  $G$  when referred to the input. In practice, the input errors dominate at high gains, and the output errors dominate at low gains. The total  $V_{OS}$  for a given gain is calculated as

$$\text{Total Error RTI} = \text{input error} + (\text{output error} / G)$$

$$\text{Total Error RTO} = (\text{input error} \times G) + \text{output error}$$

### REFERENCE TERMINAL

The reference terminal potential defines the zero output voltage and is especially useful when the load does not share a precise ground with the rest of the system. It provides a direct means of injecting a precise offset to the output, with an allowable range of 2 V within the supply voltages. Parasitic resistance should be kept to a minimum for optimum CMR.

### INPUT PROTECTION

The HT8620 features 400  $\Omega$  of series thin film resistance at its inputs and will safely withstand input overloads of up to  $\pm 15V$  or  $\pm 60 \text{ mA}$  for several hours. This is true for all gains and power on and off, which is particularly important since the signal source and amplifier may be powered separately. For longer time periods, the current should not exceed 6 mA ( $I_{IN} \leq V_{IN}/400 \Omega$ ). For input overloads beyond the supplies, clamping the inputs to the supplies (using a low leakage diode such as an FD333) will reduce the required resistance, yielding lower noise.

### RF INTERFERENCE

All instrumentation amplifiers rectify small out of band signals. The disturbance may appear as a small dc voltage offset. High frequency signals can be filtered with a low pass R-C network placed at the input of the instrumentation amplifier. Figure 43 demonstrates such a configuration. The filter limits the input signal according to the following relationship:

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

$$\text{FilterFreq}_{CM} = \frac{1}{2\pi R C_C}$$

where  $C_D \geq 10C_C$ .

$C_D$  affects the difference signal.  $C_C$  affects the common-mode signal. Any mismatch in  $R \times C_C$  will degrade the HT8620's CMRR. To avoid inadvertently reducing CMRR-bandwidth performance, make sure that  $C_C$  is at least one magnitude smaller than  $C_D$ . The effect of mismatched  $C_C$ s is reduced with a larger  $C_D:C_C$  ratio.

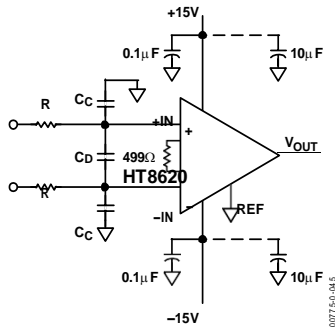


Figure 43. Circuit to Attenuate RF Interference

### COMMON-MODE REJECTION

Instrumentation amplifiers, such as the HT8620, offer high CMR, which is a measure of the change in output voltage when both inputs are changed by equal amounts. These specifications are usually given for a full-range input voltage change and a specified source imbalance.

For optimal CMR, the reference terminal should be tied to a low impedance point, and differences in capacitance and resistance should be kept to a minimum between the two inputs. In many applications, shielded cables are used to minimize noise; for best CMR over frequency, the shield should be properly driven. Figure 44 and Figure 45 show active data guards that are configured to improve ac common-mode rejections by “bootstrapping” the capacitances of input cable shields, thus minimizing the capacitance mismatch between the inputs.

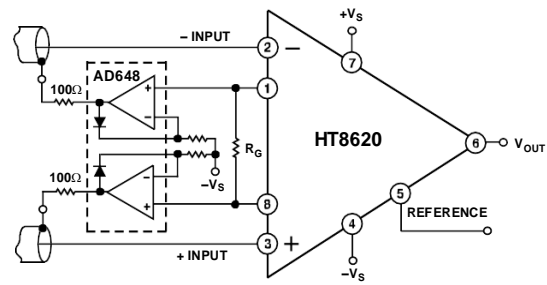


Figure 44. Differential Shield Driver

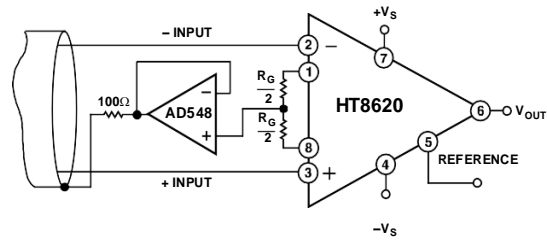


Figure 45. Common-Mode Shield Driver

### GROUNDING

Since the HT8620 output voltage is developed with respect to the potential on the reference terminal, it can solve many grounding problems by simply tying the REF pin to the appropriate “local ground.”

To isolate low level analog signals from a noisy digital environment, many data-acquisition components have separate analog and digital ground pins (Figure 46). It would be convenient to use a single ground line; however, current through ground wires and PC runs of the circuit card can cause hundreds of millivolts of error. Therefore, separate ground returns should be provided to minimize the current flow from the sensitive points to the system ground. These ground returns must be tied together at some point, usually best at the ADC package shown in Figure 46.

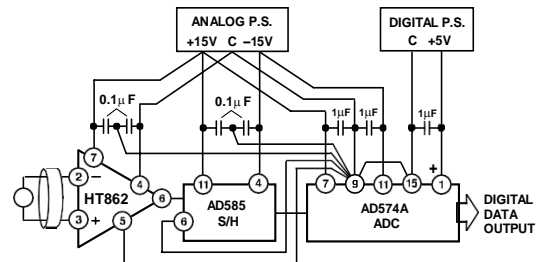


Figure 46. Basic Grounding Practice

## GROUND RETURNS FOR INPUT BIAS CURRENTS

Input bias currents are those currents necessary to bias the input transistors of an amplifier. There must be a direct return path for these currents. Therefore, when amplifying “floating” input sources, such as transformers or ac-coupled sources, there must be a dc path from each input to ground, as shown in Figure 47, Figure 48, and Figure 49. Refer to *A Designer's Guide to Instrumentation Amplifiers* (free from Analog Devices) for more information regarding in-amp applications.

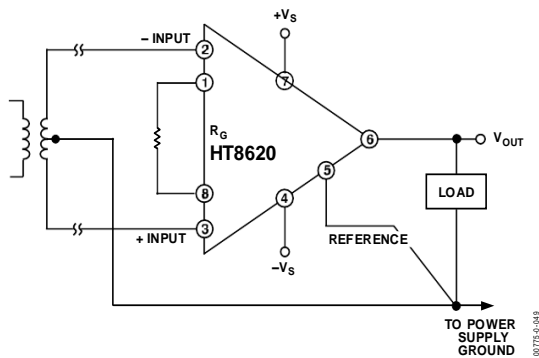


Figure 47. Ground Returns for Bias Currents with Transformer-Coupled Inputs

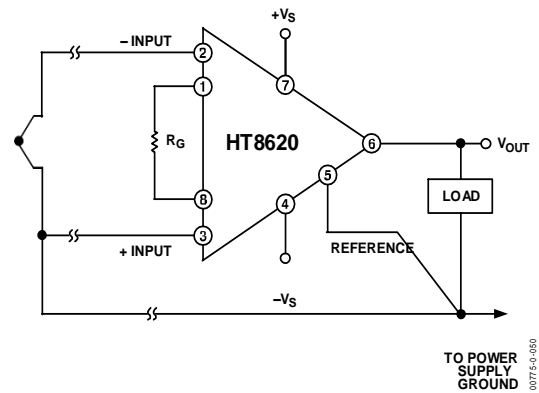


Figure 48. Ground Returns for Bias Currents with Thermocouple Inputs

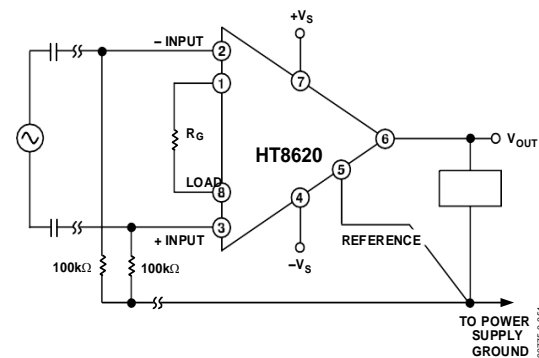
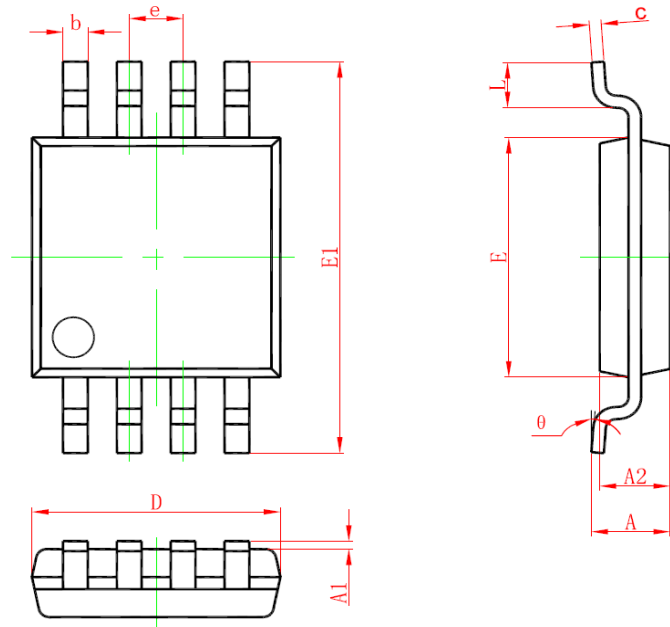


Figure 49. Ground Returns for Bias Currents with AC-Coupled Inputs

## Package Outline Dimensions

MSOP-8

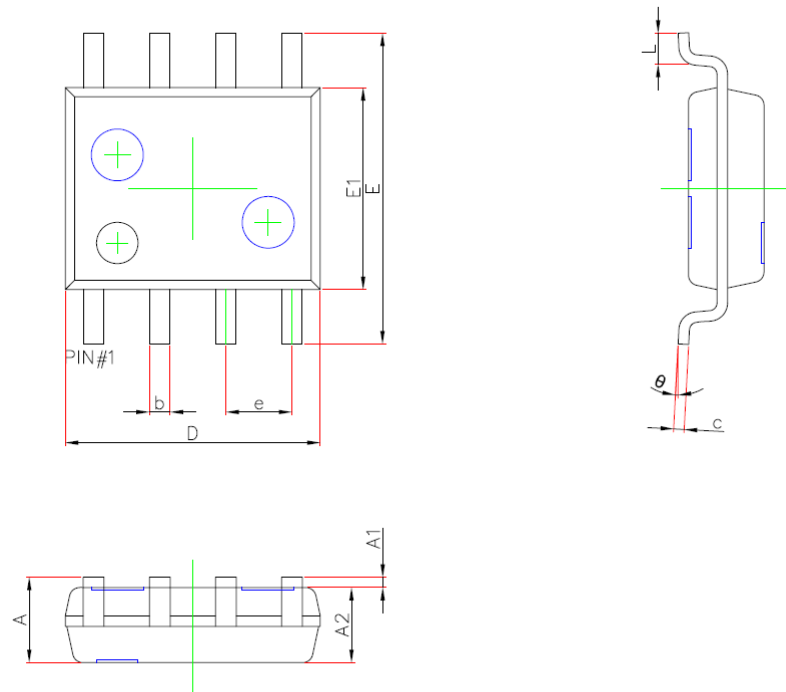


Symbol	Dimensions In Millimeters		Dimensions In Inches	
	Min	Max	Min	Max
A	0.820	1.100	0.032	0.043
A1	0.020	0.150	0.001	0.006
A2	0.750	0.950	0.030	0.037
b	0.250	0.380	0.010	0.015
c	0.090	0.230	0.004	0.009
D	2.900	3.100	0.114	0.122
e	0.650(BSC)		0.026(BSC)	
E	2.900	3.100	0.114	0.122
E1	4.750	5.050	0.187	0.199
L	0.400	0.800	0.016	0.031
θ	0°	6°	0°	6°

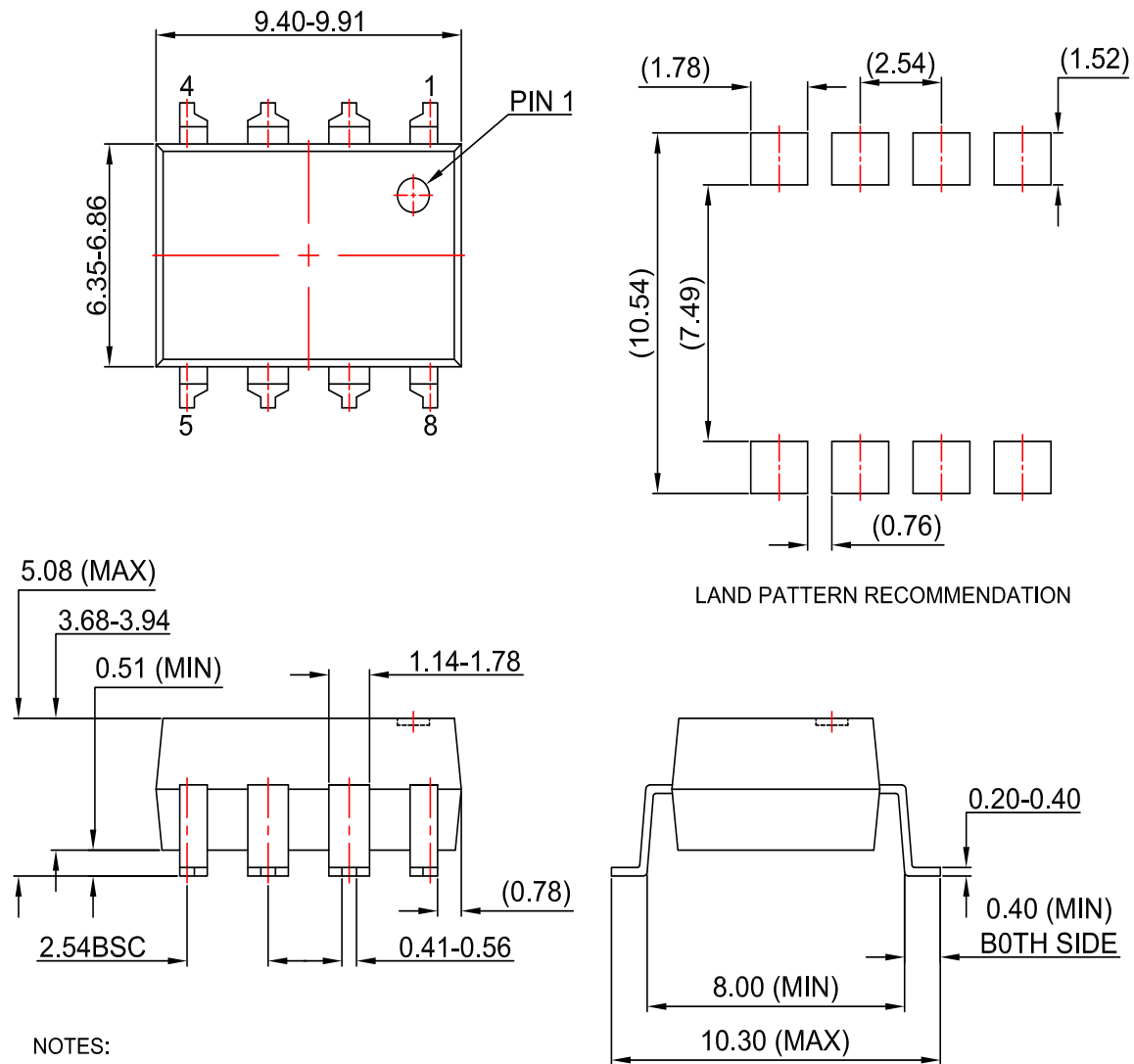


## Package Outline Dimensions

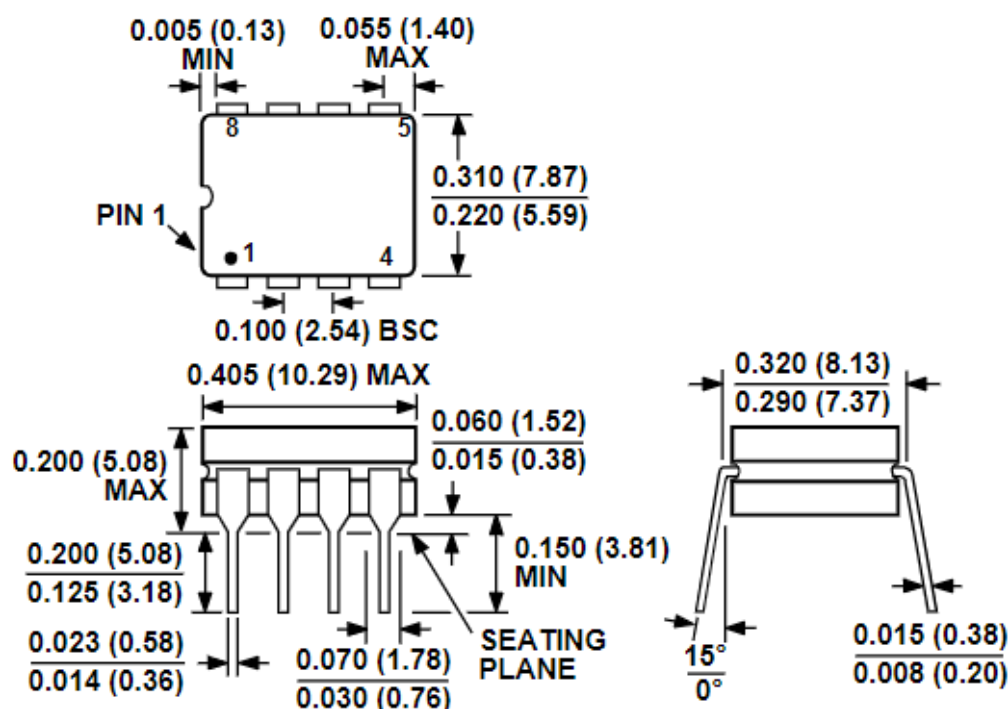
SOP-8 (SOIC-8)



Symbol	Dimensions In Millimeters		Dimensions In Inches	
	Min.	Max.	Min.	Max.
A	1.350	1.750	0.053	0.069
A1	0.100	0.250	0.004	0.010
A2	1.350	1.550	0.053	0.061
b	0.330	0.510	0.013	0.020
c	0.170	0.250	0.007	0.010
D	4.700	5.100	0.185	0.201
E	5.800	6.200	0.228	0.244
E1	3.800	4.000	0.150	0.157
e	1.270(BSC)		0.050(BSC)	
L	0.400	0.800	0.016	0.031
θ	0°	8°	0°	8°

**DUB8 (7.5\*6.5) package information**

**NOTES:**

- A) NO STANDARD APPLIES TO THIS PACKAGE
- B) ALL DIMENSIONS ARE IN MILLIMETERS.
- C) DIMENSIONS ARE EXCLUSIVE OF BURRS, MOLD FLASH, AND TIE BAR EXTRUSION
- D) DRAWING FILENAME AND REVISION: MKT-N08Hrev7.



CONTROLLING DIMENSIONS ARE IN INCHES; MILLIMETER DIMENSIONS (IN PARENTHESES) ARE ROUNDED-OFF INCH EQUIVALENTS FOR REFERENCE ONLY AND ARE NOT APPROPRIATE FOR USE IN DESIGN

Figure 51. 8-Lead Ceramic Dual In-Line Package [CERDIP] (Q-8)  
Dimensions shown in inches and (millimeters)