Low Power Instrumentation Amplifie

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 μV max, and offset drift of 0.6 μV/℃ 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.

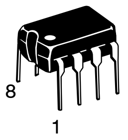
The low input bias current of 1.0 nA max is made possible with the use of Superϐeta processing in the input stage. The HT8620 works well as a preamplifier due to its low input voltage noise of 9 nV/√Hz at 1 kHz, 0.28 μV p-p in the 0.1 Hz to 10 Hz band, and 0.1 pA/√Hz input current noise. Also, the HT8620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

FEATURES

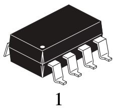
- Easy to use
  - Gain set with one external resistor
    - (Gain range 1 to 10,000)
  - Wide power supply range (±2.3 V to ±18 V)±10)
  - Higher performance than 3 op amp IA designs
  - Available in 8-lead DIP and SOIC packaging
  - Low power, 1.3 mA max supply current
- Excellent dc performance (B grade)
  - 50 μV max, input offset voltage
  - 0.6 μV/°C max, input offset drift
  - 1.0 nA max, input bias current
  - 100 dB min common-mode rejection ratio (G
- Low noise
  - 9 nV/√Hz @ 1 kHz, input voltage noise
  - 0.28 μV p-p noise (0.1 Hz to 10 Hz)
- Excellent ac specifications
  - 120 kHz bandwidth (G = 100)
  - 15 μs settling time to 0.01%

APPLICATIONS

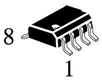
- Weigh scales
- ECG and medical instrumentation
- Transducer interface
- Data acquisition systems
- Industrial process controls
- Battery-powered and portable equipment



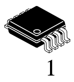
DIP8 N
SUFFIX
HT8620ANZ



DUB8 U
SUFFIX
HT8620ARUZ



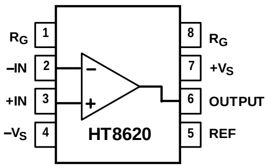
SOP8 R
SUFFIX
HT8620ARZ



MSOP8 M
SUFFIX
HT8620ARMZ

NOTE: If you need to customize the packaging shape, please contact the manufacturer.

CONNECTION DIAGRAM



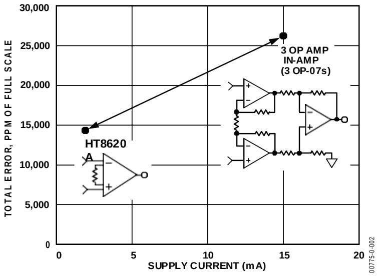


Figure 2. Three Op Amp IA Designs vs. HT8620

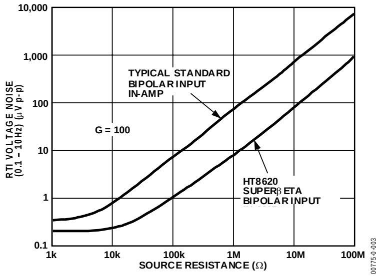


Figure 3. Total Voltage Noise vs. Source Resistance

SPECIFICATIONS

Typical @ 25 ℃, V_S = ±15 V, and R_L = 2 kΩ, unless otherwise noted.

Table 1.

		HT8620$_A$			HT8620$_B$			HT8620$_S$			
Parameter	**Conditions**	**Min**	**Typ**	**Max**	**Min**	**Typ**	**Max**	**Min**	**Typ**	**Max**	**Unit**
GAIN	G = 1 + (49.4 kΩ/R_G)										
Gain Range		1		10,000	1		10,000	1		10,000	
Gain Error^2	V_{OUT} = ±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 kΩ		10	40		10	40		10	40	ppm
G = 1–100	R_L = 2 kΩ		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
VOLTAGE OFFSET	(Total RTI Error = V_{OSI} + V_{OSO}/G)										
Input Offset, V_{OSI}	V_S = ±5 V to ± 15 V		30	125		15	50		30	125	µV
Overtemperature	V_S = ±5 V to ± 15 V			185			85			225	µV
Average TC	V_S = ±5 V to ± 15 V		0.3	1.0		0.1	0.6		0.3	1.0	µV/°C
Output Offset, V_{OSO}	V_S = ±15 V		400	1000		200	500		400	1000	µV
	V_S = ± 5 V			1500			750			1500	µV
Overtemperature	V_S = ±5 V to ± 15 V			2000			1000			2000	µV
Average TC	V_S = ±5 V to ± 15 V		5.0	15		2.5	7.0		5.0	15	µV/°C
Offset Referred to the Input vs. Supply (PSR)	V_S = ±2.3 V to ±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
INPUT CURRENT											
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
INPUT											
Input Impedance											
Differential			10‖2			10‖2			10‖2		GΩ_pF
Common-Mode			10‖2			10‖2			10‖2		GΩ_pF
Input Voltage Range^3	V_S = ±2.3 V to ±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		$-V_S + 2.1$		$+V_S - 1.3$	$-V_S + 2.1$		$+V_S - 1.3$	$-V_S + 2.1$		$+V_S - 1.3$	V
	V_S = ± 5 V to ±18 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 Min	HT8620A Typ	HT8620A Max	HT8620B Min	HT8620B Typ	HT8620B Max	HT8620S Min	HT8620S Typ	HT8620S Max	Unit
Common-Mode Rejection Ratio DC to 60 Hz with 1 kΩ Source Imbalance	V_{CM} = 0 V to ± 10 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 kΩ V_S = ±2.3 V to ± 5 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
Overtemperature		$-V_S + 1.4$		$+V_S - 1.3$	$-V_S + 1.4$		$+V_S - 1.3$	$-V_S + 1.6$		$+V_S - 1.3$	V
	V_S = ±5 V to ± 18 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
Overtemperature		$-V_S + 1.6$		$+V_S - 1.5$	$-V_S + 1.6$		$+V_S - 1.5$	$-V_S + 2.3$		$+V_S - 1.5$	V
Short Circuit Current			±18			±18			±18		mA
DYNAMIC RESPONSE											
Small Signal −3 dB Bandwidth											
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%	10 V Step										
G = 1–100			15			15			15		µs
G = 1000			150			150			150		µs
NOISE											
Voltage Noise, 1 kHz	$Total\ RTI\ Noise = \sqrt{(e^2_{ni}) + (e_{no}/G)^2}$										
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}			20			20			20		kΩ
I_{IN}	V_{IN+}, V_{REF} = 0		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 ± 0.0001			1 ± 0.0001			1 ± 0.0001			
POWER SUPPLY											
Operating Range[4]		±2.3		±18	±2.3		±18	±2.3		±18	V
Quiescent Current	V_S = ±2.3 V to ±18 V		0.9	1.3		0.9	1.3		0.9	1.3	mA
Overtemperature			1.1	1.6		1.1	1.6		1.1	1.6	mA
TEMPERATURE RANGE											
For Specified Performance		−40 to +85			−40 to +85			−55 to +125			°C

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[1] See Analog Devices military data sheet for 883B tested specifications.
[2] Does not include effects of external resistor R_G.
[3] One input grounded. G = 1.
[4] This is defined as the same supply range that is used to specify PSR.

TYPICAL PERFORMANCE CHARACTERISTICS

(@ 25 ℃, V_S = ±15 V, R_L = 2 kΩ, unless otherwise noted.)

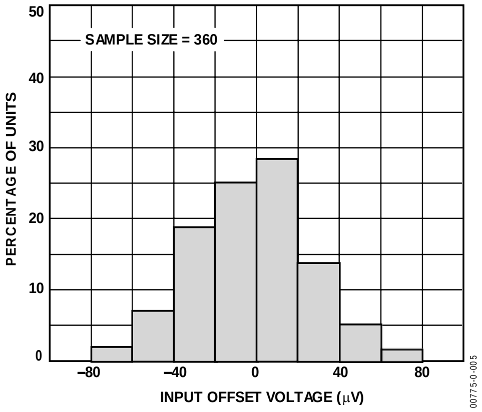


Figure 5. Typical Distribution of Input Offset Voltage

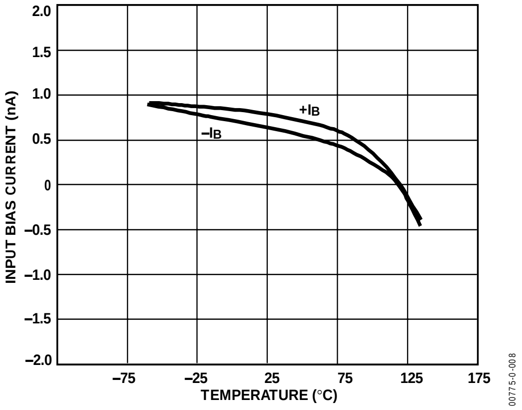


Figure 8. Input Bias Current vs. Temperature

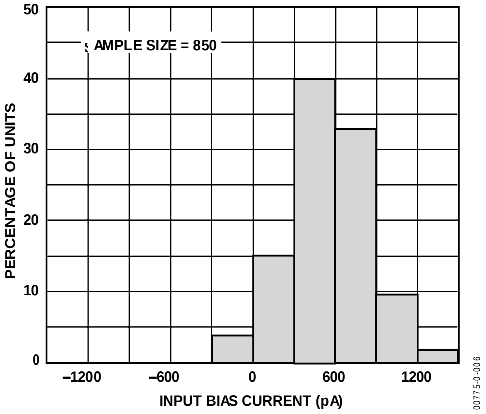


Figure 6. Typical Distribution of Input Bias Current

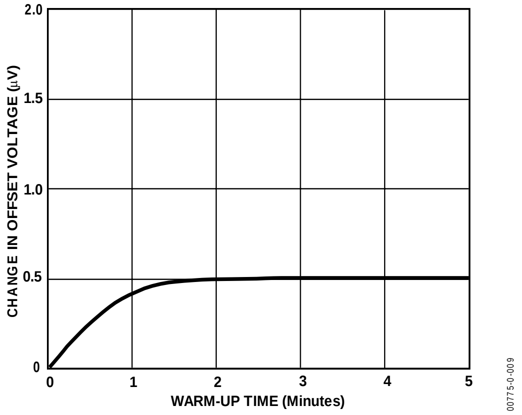


Figure 9. Change in Input Offset Voltage vs. Warm-Up Time

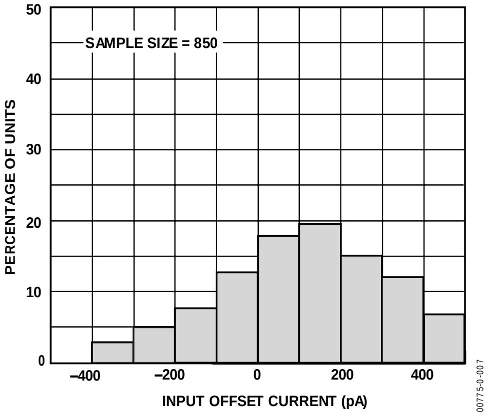


Figure 7. Typical Distribution of Input Offset Current

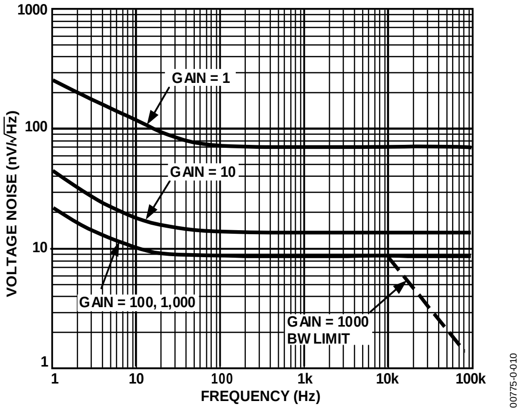


Figure 10. Voltage Noise Spectral Density vs. Frequency (G = 1–1000)

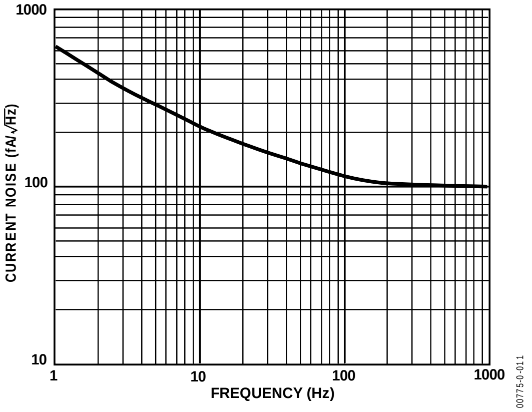


Figure 11. Current Noise Spectral Density vs. Frequency

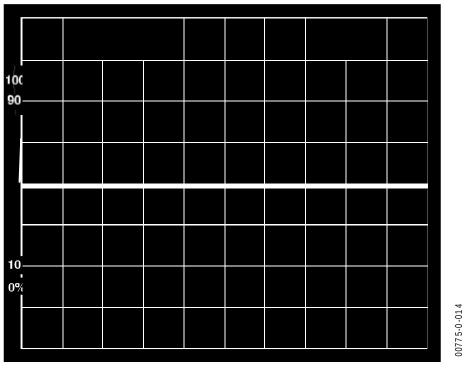


Figure 14. 0.1 Hz to 10 Hz Current Noise, 5 pA/Div

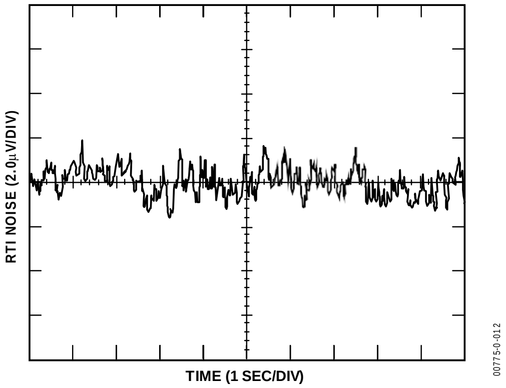


Figure 12. 0.1 Hz to 10 Hz RTI Voltage Noise (G = 1)

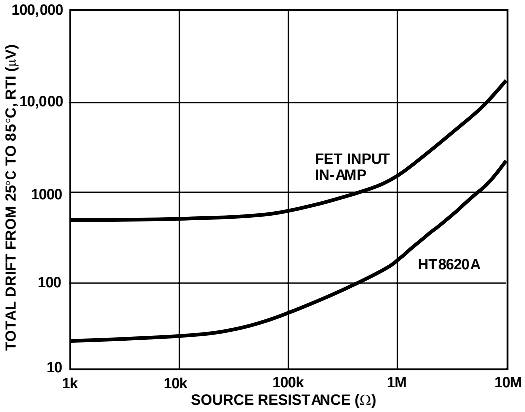


Figure 15. Total Drift vs. Source Resistance

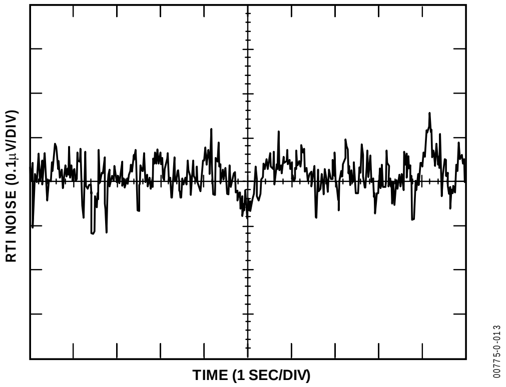


Figure 13. 0.1 Hz to 10 Hz RTI Voltage Noise (G = 1000)

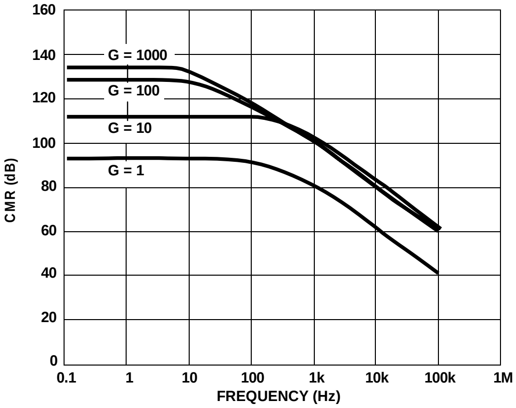


Figure 16. Typical CMR vs. Frequency, RTI, Zero to 1 kΩ Source Imbalance

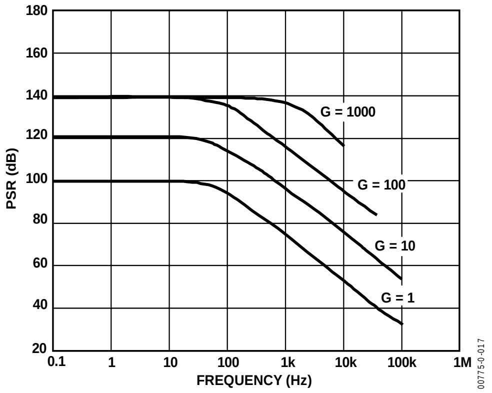


Figure 17. Positive PSR vs. Frequency, RTI (G = 1−1000)

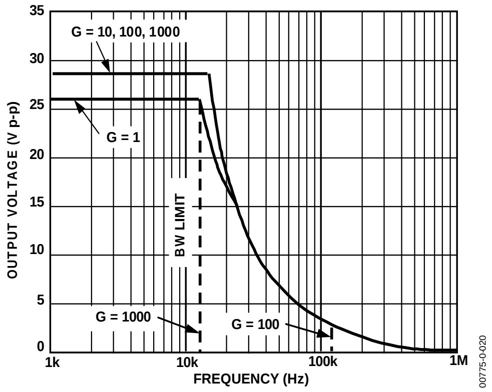


Figure 20. Large Signal Frequency Response

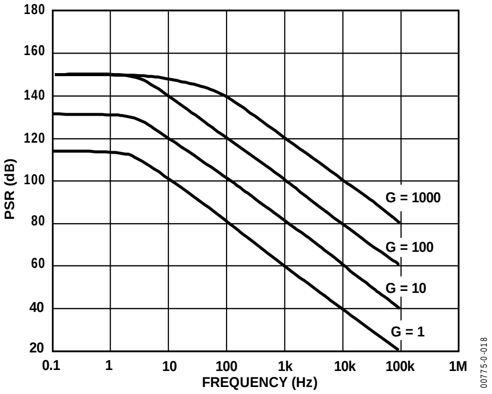


Figure 18. Negative PSR vs. Frequency, RTI (G = 1−1000)

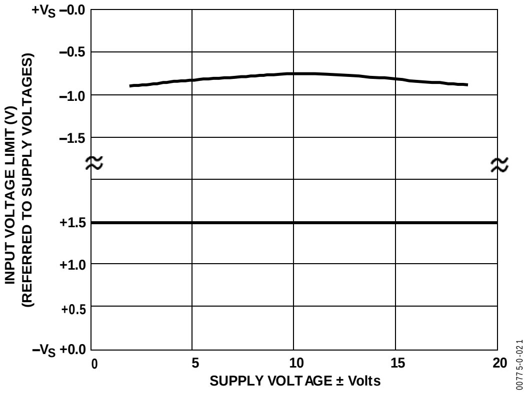


Figure 21. Input Voltage Range vs. Supply Voltage, G = 1

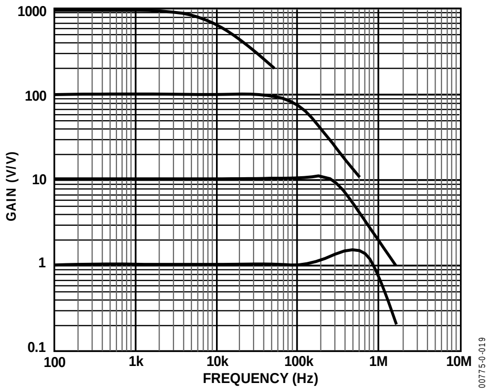


Figure 19. Gain vs. Frequency

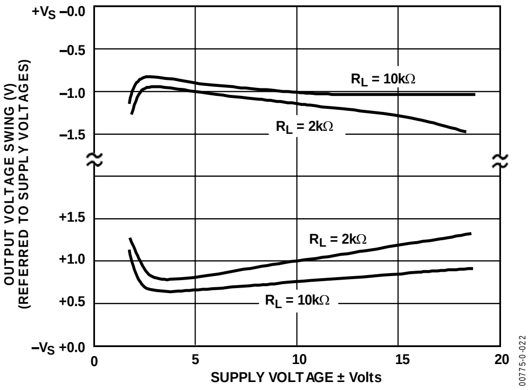


Figure 22. Output Voltage Swing vs. Supply Voltage, G = 10

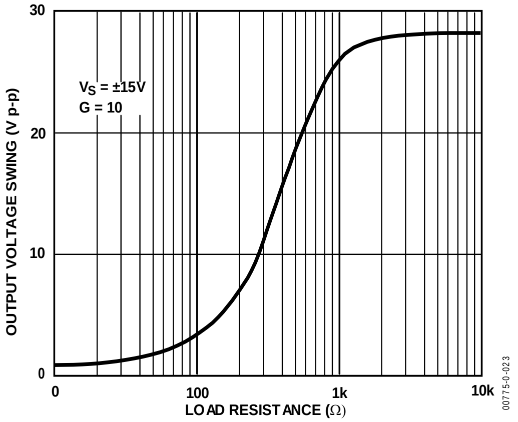


Figure 23. Output Voltage Swing vs. Load Resistance

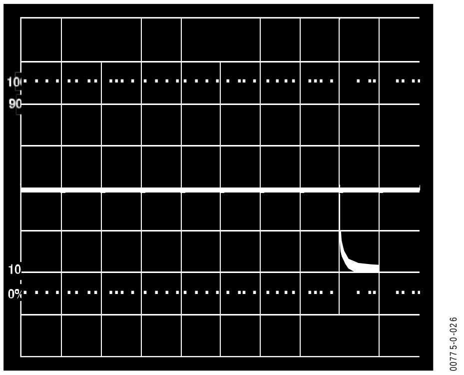


Figure 26. Large Signal Response and Settling Time, G = 10 (0.5 mV = 0.01%)

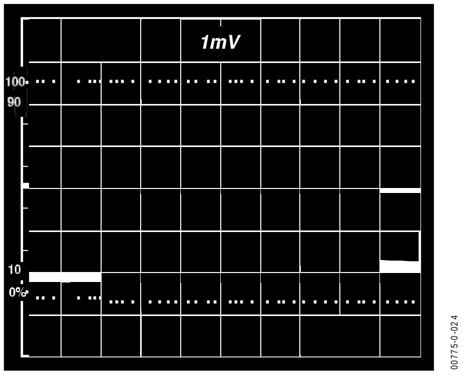


Figure 24. Large Signal Pulse Response and Settling Time G = 1 (0.5 mV = 0.01%)

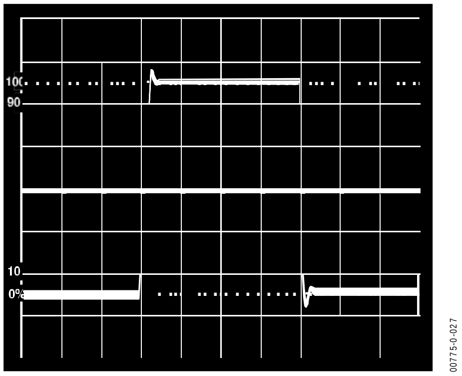


Figure 27. Small Signal Response, G = 10, R_L = 2 kΩ, C_L = 100 pF

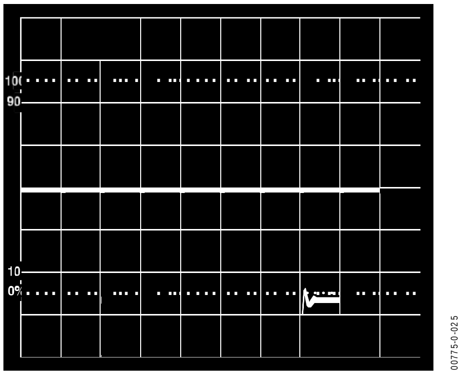


Figure 25. Small Signal Response, G = 1, R_L = 2 kΩ, C_L = 100 pF

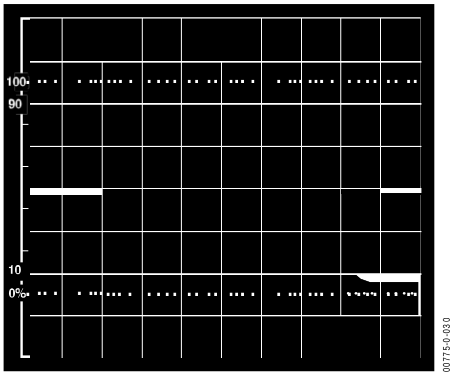


Figure 28. Large Signal Response and Settling Time, G = 100 (0.5 mV = 0.01%)

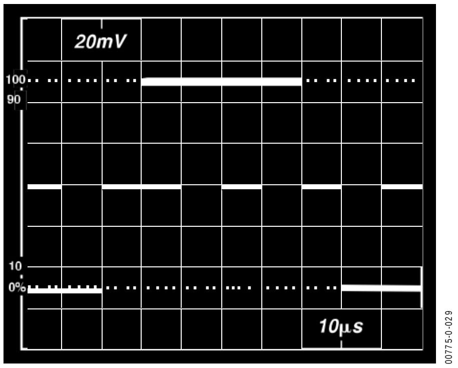


Figure 29. Small Signal Pulse Response, G = 100, R_L = 2 kΩ, C_L = 100 pF

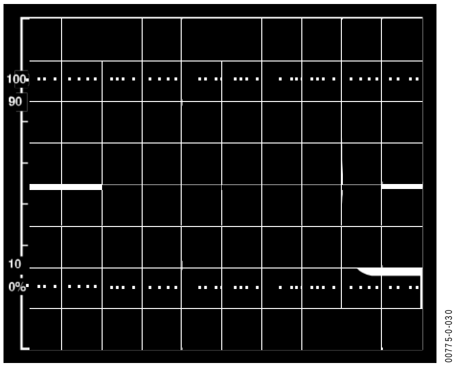


Figure 30. Large Signal Response and Settling Time, G = 1000 (0.5 mV = 0.01%)

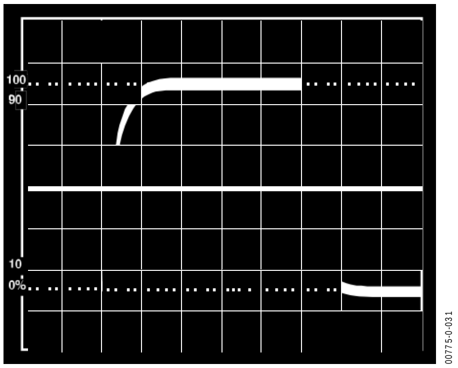


Figure 31. Small Signal Pulse Response, G = 1000, R_L = 2 kΩ, C_L = 100 pF

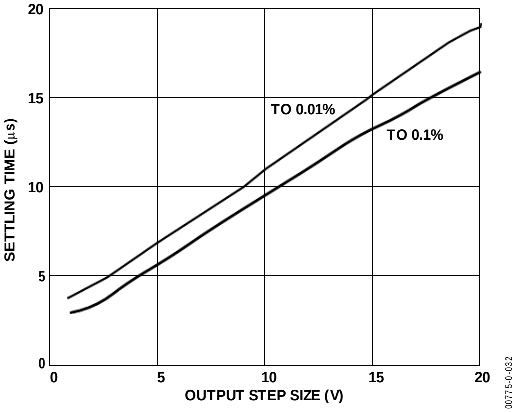


Figure 32. Settling Time vs. Step Size (G = 1)

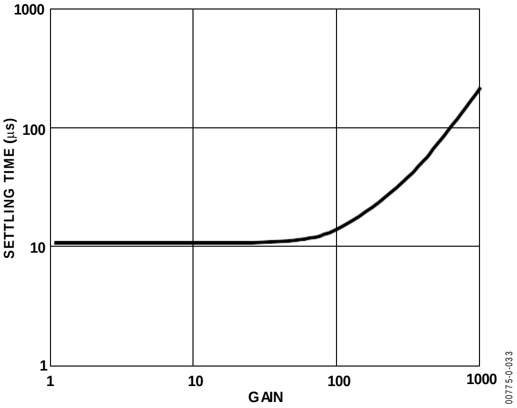


Figure 33. Settling Time to 0.01% vs. Gain, for a 10 V Step

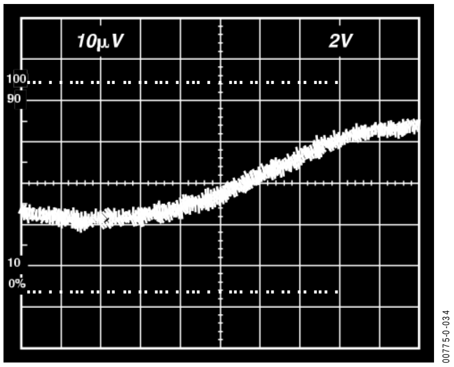


Figure 34. Gain Nonlinearity, G = 1, R_L = 10 kΩ (10 µV = 1 ppm)

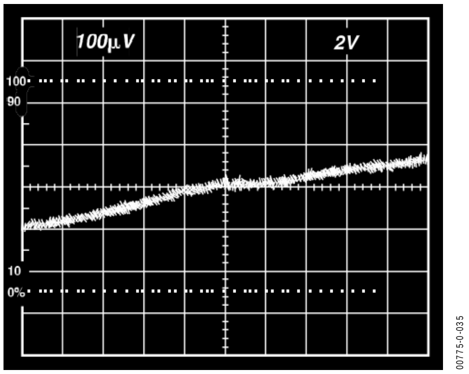


Figure 35. Gain Nonlinearity, G = 100, R_L = 10 kΩ (100 µV = 10 ppm)

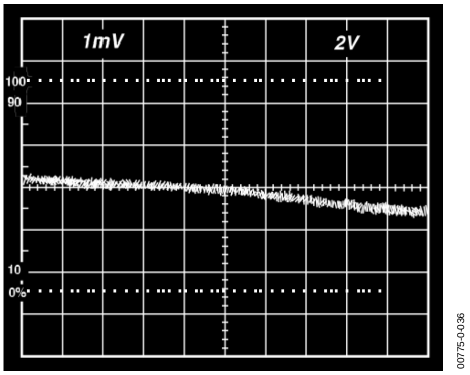
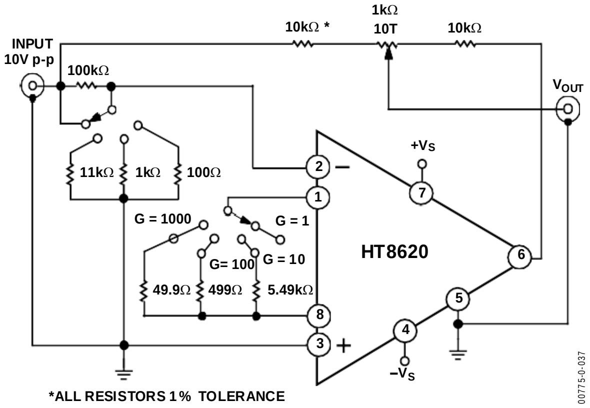


Figure 36. Gain Nonlinearity, G = 1000, R_L = 10 kΩ (1 mV = 100 ppm)



*ALL RESISTORS 1% TOLERANCE

Figure 37. Settling Time Test Circuit

THEORY OF OPERATION

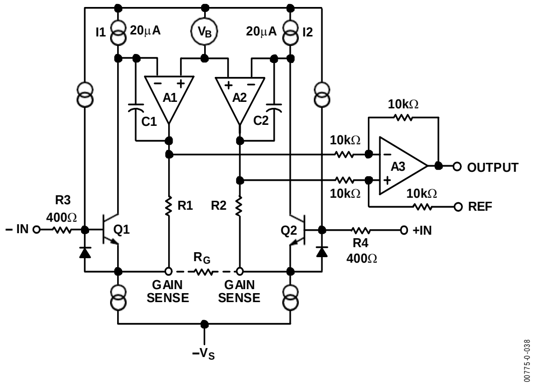


Figure 38. Simplified Schematic of HT8620

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× lower input bias current thanks to Superбeta 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 nV/√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 kΩ, allowing the gain to be programmed accurately with a single external resistor.

The gain equation is then

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

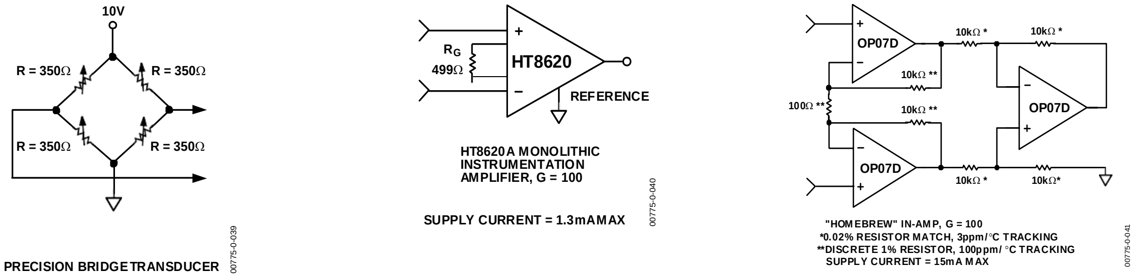
$$R_G = \frac{49.4k\Omega}{G-1}$$

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× 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°C to +85°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 √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.



PRECISION BRIDGE TRANSDUCER

HT8620A MONOLITHIC INSTRUMENTATION AMPLIFIER, G = 100

SUPPLY CURRENT = 1.3mA MAX

"HOMEBREW" IN-AMP, G = 100
*0.02% RESISTOR MATCH, 3ppm/°C TRACKING
**DISCRETE 1% RESISTOR, 100ppm/ °C TRACKING
SUPPLY CURRENT = 15mA MAX

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_A = 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) ×5 V/20 mV	(0.02% Match × 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) ×60°C	100 ppm/°C Track × 60°C	3,600	6,000
Input Offset Voltage Drift, µV/°C	1 µV/°C × 60°C/20 mV	(2.5 µV/°C × √2 × 60°C)/20 mV	3,000	10,607
Output Offset Voltage Drift, µV/°C	15 µV/°C × 60°C/100 mV/20 mV	(2.5 µV/°C × 2 × 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
Typ 0.1 Hz to 10 Hz Voltage Noise, µV p-p	0.28 µV p-p/20 mV	(0.38 µV p-p × √2)/20 mV	14	27
		Total Resolution Error	54	67
		Grand Total Error	14,663	28,134

G = 100, V_S = ±15 V.
(All errors are min/max and referred to input.)

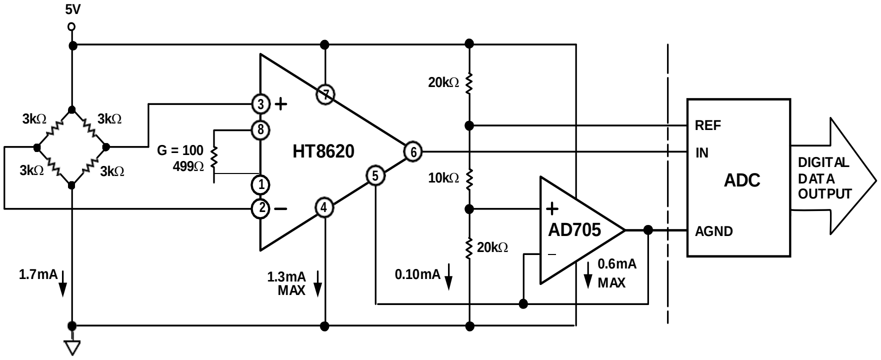


Figure 40. A Pressure Monitor Circuit that Operates on a 5 V Single Supply

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.

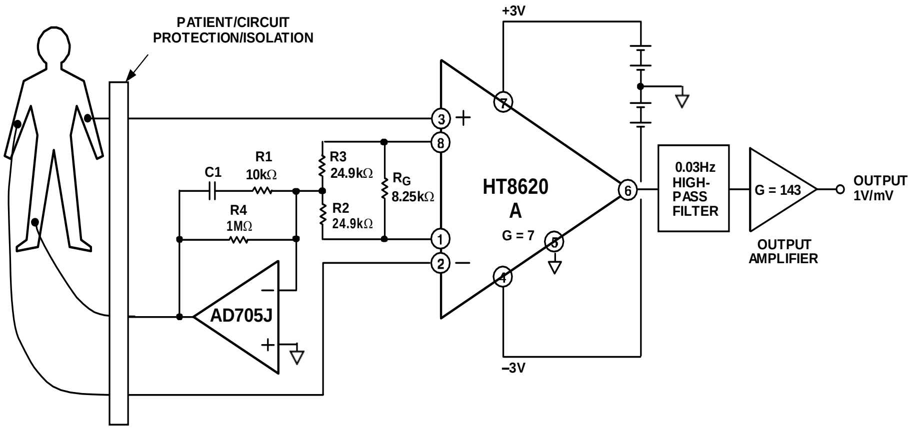


Figure 41. A Medical ECG Monitor Circuit

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 R1, which converts it to a current. This current, less only the input bias current of the op amp, then flows out to the load.

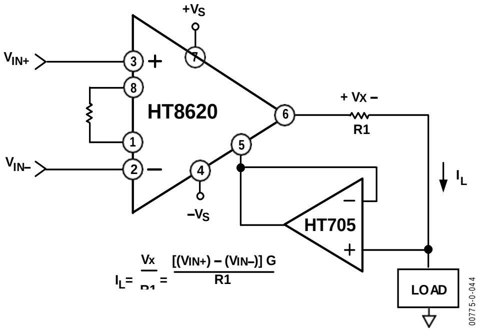


Figure 42. Precision Voltage-to-Current Converter (Operates on 1.8 mA, ±3 V)

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:

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

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/ ℃—for the best performance.

Table 4. Required Values of Gain Resistors

1% Std Table Value of R_G(Ω)	Calculated Gain	0.1% Std Table Value of R_G(Ω)	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

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

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

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 Ω of series thin film resistance at its inputs and will safely withstand input overloads of up to ±15 V or ±60 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 ($I_{IN} \le V_{IN}/400\ \Omega$). 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:

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

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

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_Cs is reduced with a larger C_D:C_C ratio.

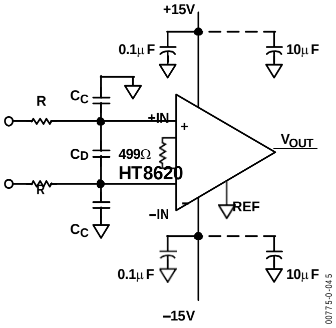


Figure 43. Circuit to Attenuate RF Interference

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.

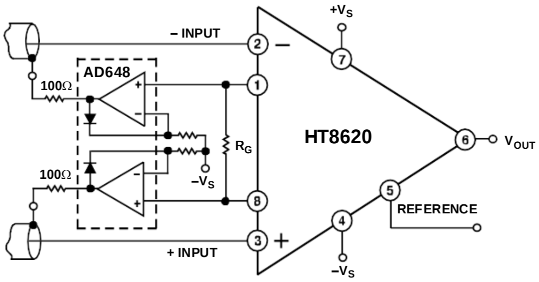


Figure 44. Differential Shield Driver

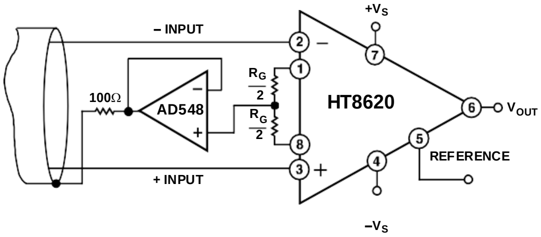


Figure 45. Common-Mode Shield Driver

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.

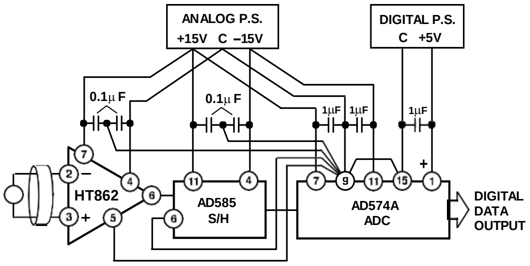


Figure 46. Basic Grounding Practice

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.

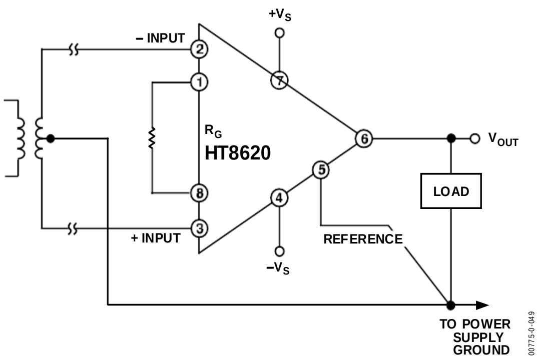


Figure 47. Ground Returns for Bias Currents with Transformer-Coupled Inputs

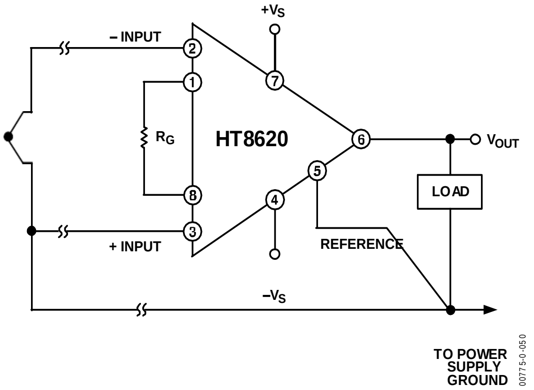


Figure 48. Ground Returns for Bias Currents with Thermocouple Inputs

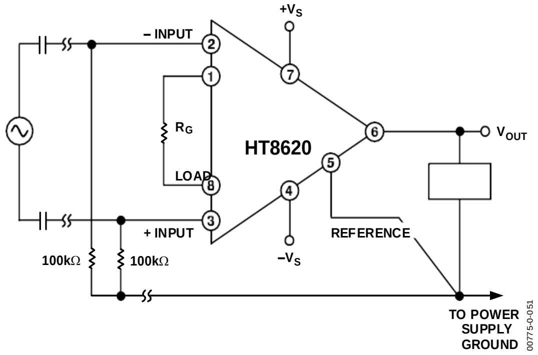
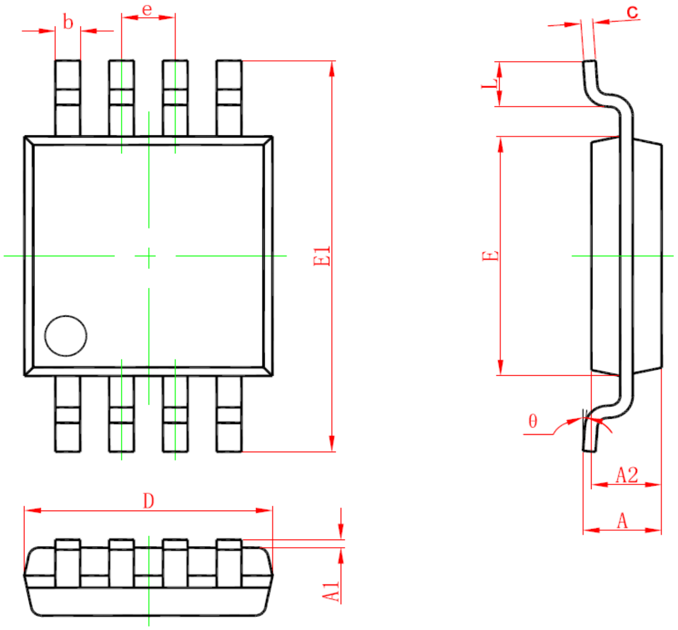


Figure 49. Ground Returns for Bias Currents with AC-Coupled Inputs

Package Outline Dimensions

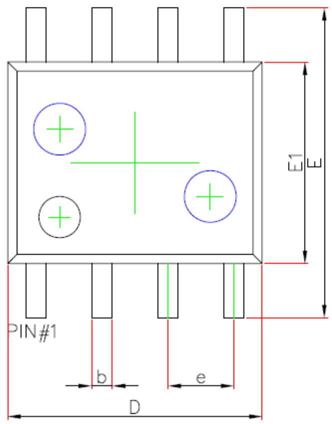
MSOP-8

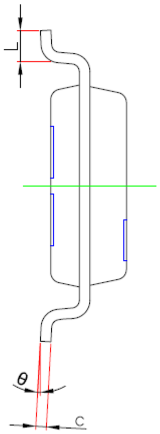


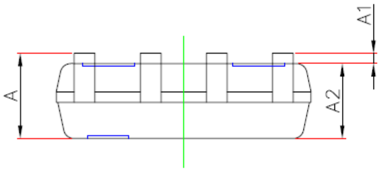
Symbol	Dimensions In Millimeters		Dimensions In Inches	
	Min	Max	Min	Max
A	0.820	1.100	0.032	0.043
A1	0.020	0.150	0.001	0.006
A2	0.750	0.950	0.030	0.037
b	0.250	0.380	0.010	0.015
c	0.090	0.230	0.004	0.009
D	2.900	3.100	0.114	0.122
e	0.650(BSC)		0.026(BSC)	
E	2.900	3.100	0.114	0.122
E1	4.750	5.050	0.187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

Package Outline Dimensions

SOP-8 (SOIC-8)

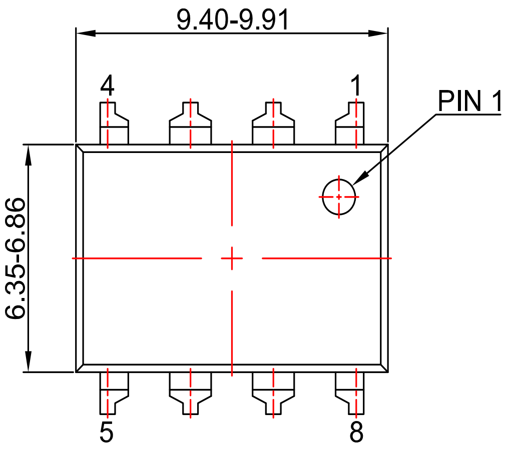


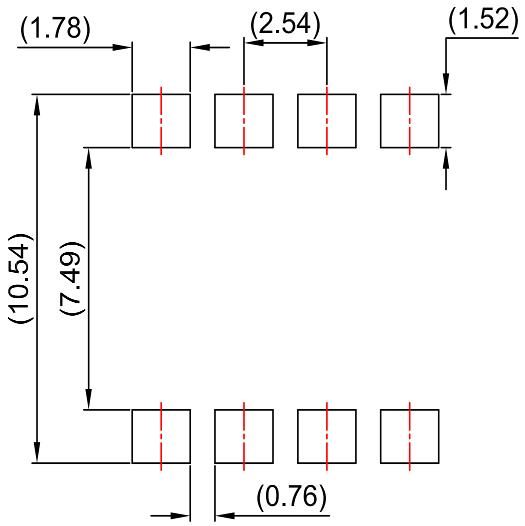




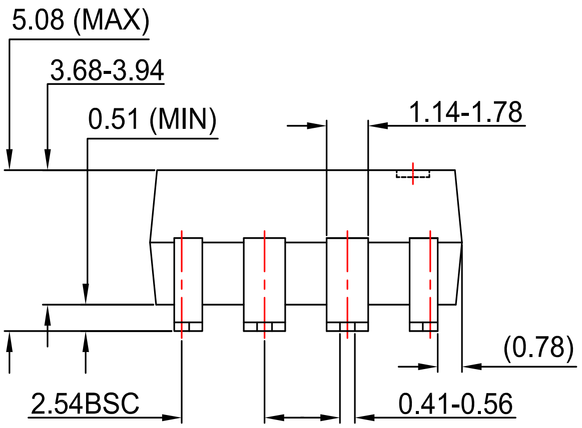
Symbol	Dimensions In Millimeters		Dimensions In Inches	
	Min.	Max.	Min.	Max.
A	1.350	1.750	0.053	0.069
A1	0.100	0.250	0.004	0.010
A2	1.350	1.550	0.053	0.061
b	0.330	0.510	0.013	0.020
c	0.170	0.250	0.007	0.010
D	4.700	5.100	0.185	0.201
E	5.800	6.200	0.228	0.244
E1	3.800	4.000	0.150	0.157
e	1.270(BSC)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

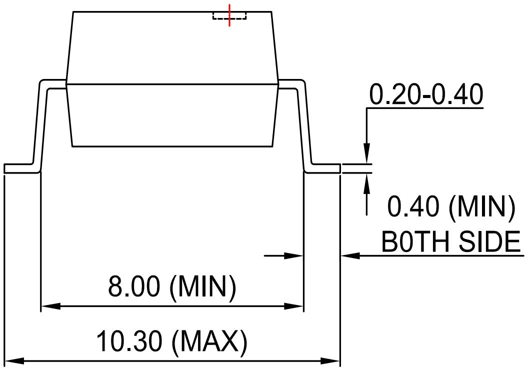
DUB8 (7.5*6.5) package information



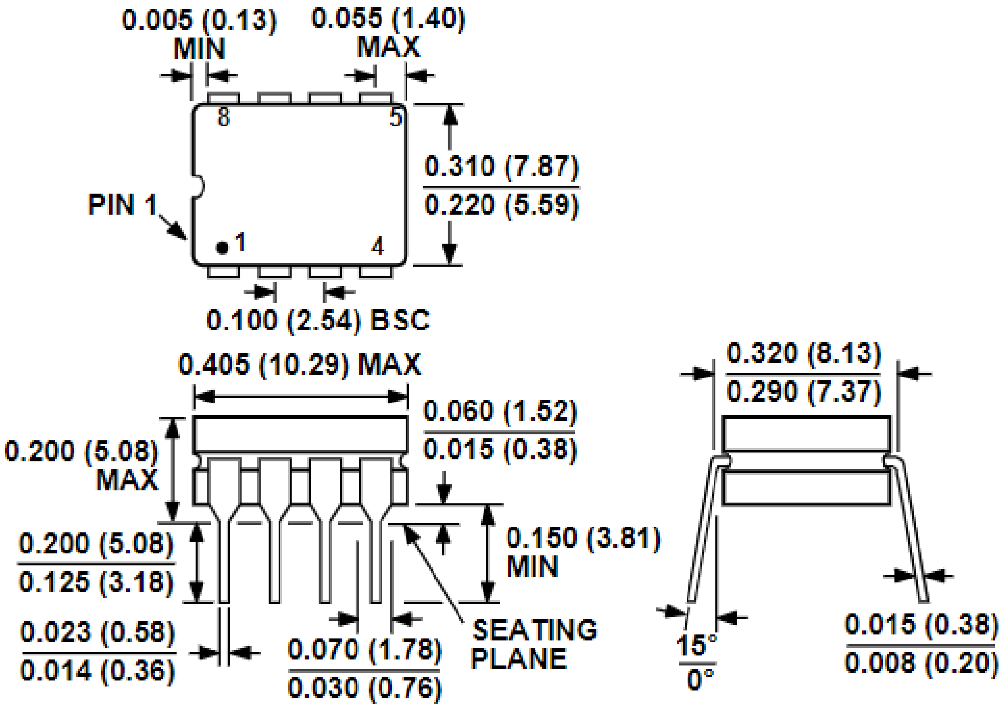


LAND PATTERN RECOMMENDATION





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 REVSION: 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)