

FEATURES

Low noise

Voltage noise = 2.2 nV/√Hz

Current noise = 4.8 pA/√Hz (positive input)

Wide bandwidth (−3 dB) = 280 MHz

Nominal gain range: 0 dB to 24 dB (preamp gain = 6 dB)

Gain scaling: 19.7 dB/V

DC-coupled

Single-ended input and output

High speed uncommitted op amp input

Supplies: +5 V, ±2.5 V, or ±5 V

Low power: 78 mW with ±2.5 V supplies

APPLICATIONS

Gain trim

PET scanners

High performance AGC systems

I/Q signal processing

Video

Industrial and medical ultrasound

Radar receivers

GENERAL DESCRIPTION

The AD8337 is a low noise, single-ended, linear-in-dB, general-purpose variable gain amplifier (VGA) usable at frequencies from dc to 100 MHz; the −3 dB bandwidth is 280 MHz.

Excellent bandwidth uniformity across the entire gain range and low output-referred noise makes the AD8337 ideal for gain trim applications and for driving high speed analog-to-digital converters (ADCs).

Excellent dc characteristics combined with high speed make the AD8337 particularly suited for industrial ultrasound, PET scanners, and video applications. Dual-supply operation enables gain control of negative-going pulses such as generated by photodiodes or photomultiplier tubes.

The AD8337 uses the popular and versatile X-AMP® architecture, exclusively from Analog Devices, Inc., with a gain range of 24 dB. The gain control interface provides precise linear-in-dB scaling of 19.7 dB/V, referenced to VCOM.

FUNCTIONAL BLOCK DIAGRAM

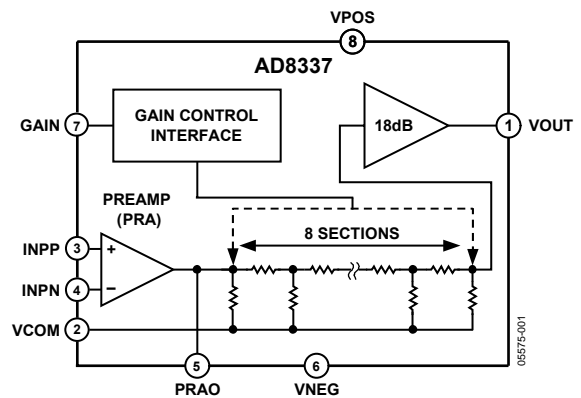


Figure 1.

The AD8337 includes an uncommitted operational current-feedback preamplifier (PrA) that operates in inverting or noninverting configurations. Using external resistors, the device can be configured for gains of 6 dB or greater. The AD8337 is characterized by a noninverting PrA gain of 2× using two external 100 Ω resistors. The attenuator has a range of 24 dB, and the output amplifier has a fixed gain of 8× (18.06 dB). The lowest nominal gain range is 0 dB to 24 dB and can be shifted up or down by adjusting the preamp gain. Multiple AD8337s can be connected in series for larger gain ranges, and for interstage filtering to suppress noise and distortion, and for nulling offset voltages.

The operating temperature range is −40°C to +85°C, and it is available in an 8-lead, 3 mm × 3 mm LFCSP.

Rev. B

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REVISION HISTORY

2/07—Rev. A to Rev. B		6/06—Rev. 0 to Rev. A	
Changes to Figure 30, Figure 31, and Figure 32	11	Updated Format.....	Universal
Changes to Single-Supply Operation and		Changes to Table 3.....	6
AC Coupling Section	19	Changes to Figure 22, Figure 25, and Figure 26	10
Moved Noise Section to Page.....	19	Changes to Figure 39 and Figure 40.....	13
Changes to Ordering Guide	24	Changes to Figure 74 and Figure 75.....	23
		Updated Outline Dimensions.....	25
		Changes to Ordering Guide	25
		9/05—Revision 0: Initial Version	

SPECIFICATIONS

$V_S = \pm 2.5\text{ V}$, $T_A = 25^\circ\text{C}$, PrA Gain = +2, $V_{\text{COM}} = \text{GND}$, $f = 10\text{ MHz}$, $C_L = 5\text{ pF}$, $R_L = 500\ \Omega$, including a $20\ \Omega$ snubbing resistor, unless otherwise specified.

Table 1.

Parameter	Conditions	Min	Typ	Max	Unit
GENERAL PARAMETERS					
–3 dB Small Signal Bandwidth	$V_{\text{OUT}} = 10\text{ mV p-p}$		280		MHz
–3 dB Large Signal Bandwidth	$V_{\text{OUT}} = 1\text{ V p-p}$		100		MHz
Slew Rate	$V_{\text{OUT}} = 2\text{ V p-p}$		625		V/ μs
	$V_{\text{OUT}} = 1\text{ V p-p}$		490		V/ μs
Input Voltage Noise	$f = 10\text{ MHz}$		2.15		nV/ $\sqrt{\text{Hz}}$
Input Current Noise	$f = 10\text{ MHz}$		4.8		pA/ $\sqrt{\text{Hz}}$
Noise Figure	$V_{\text{GAIN}} = 0.7\text{ V}$, $R_S = 50\ \Omega$, unterminated		8.5		dB
	$V_{\text{GAIN}} = 0.7\text{ V}$, $R_S = 50\ \Omega$, shunt terminated with $50\ \Omega$		14		dB
Output-Referred Noise	$V_{\text{GAIN}} = 0.7\text{ V}$ (Gain = 24 dB)		34		nV/ $\sqrt{\text{Hz}}$
	$V_{\text{GAIN}} = -0.7\text{ V}$ (Gain = 0 dB)		21		nV/ $\sqrt{\text{Hz}}$
Output Impedance	DC to 10 MHz		1		Ω
Output Signal Range	$R_L \geq 500\ \Omega$, $V_S = \pm 2.5\text{ V}$, +5 V		$V_{\text{COM}} \pm 1.3$		V
	$R_L \geq 500\ \Omega$, $V_S = \pm 5\text{ V}$		$V_{\text{COM}} \pm 3.8$		V
Output Offset Voltage	$V_{\text{GAIN}} = 0.7\text{ V}$ (Gain = 24 dB)	–25	± 5	+25	mV
DYNAMIC PERFORMANCE					
Harmonic Distortion	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 1\text{ V p-p}$				
HD2	$f = 1\text{ MHz}$		–72		dBc
HD3			–66		dBc
HD2	$f = 10\text{ MHz}$		–62		dBc
HD3			–63		dBc
HD2	$f = 45\text{ MHz}$		–58		dBc
HD3			–56		dBc
Input 1 dB Compression Point	$V_{\text{GAIN}} = -0.7\text{ V}$, $f = 10\text{ MHz}$ (preamp limited)		8.2		dBm
	$V_{\text{GAIN}} = +0.7\text{ V}$, $f = 10\text{ MHz}$ (VGA limited)		–9.4		dBm
Two-Tone Intermodulation Distortion (IMD3)	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 1\text{ V p-p}$, $f_1 = 10\text{ MHz}$, $f_2 = 11\text{ MHz}$		–71		dBc
	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 1\text{ V p-p}$, $f_1 = 45\text{ MHz}$, $f_2 = 46\text{ MHz}$		–57		dBc
	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 2\text{ V p-p}$, $f_1 = 10\text{ MHz}$, $f_2 = 11\text{ MHz}$		–58		dBc
	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 2\text{ V p-p}$, $f_1 = 45\text{ MHz}$, $f_2 = 46\text{ MHz}$		–45		dBc
Output Third-Order Intercept	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 1\text{ V p-p}$, $f = 10\text{ MHz}$		34		dBm
	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 1\text{ V p-p}$, $f = 45\text{ MHz}$		28		dBm
	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 2\text{ V p-p}$, $f = 10\text{ MHz}$		35		dBm
	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 2\text{ V p-p}$, $f = 45\text{ MHz}$		26		dBm
Overload Recovery	$V_{\text{GAIN}} = 0.75\text{ V}$, $V_{\text{IN}} = 50\text{ mV p-p}$ to 500 mV p-p		50		ns
Group Delay Variation	$1\text{ MHz} < f < 100\text{ MHz}$, full gain range		± 1		ns

AD8337

Parameter	Conditions	Min	Typ	Max	Unit
DYNAMIC PERFORMANCE	$V_S = \pm 5\text{ V}$				
Harmonic Distortion	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 1\text{ V p-p}$				
HD2	$f = 1\text{ MHz}$		–85		dBc
HD3			–75		dBc
HD2	$f = 10\text{ MHz}$		–90		dBc
HD3			–80		dBc
HD2	$f = 35\text{ MHz}$		–75		dBc
HD3			–76		dBc
Input 1 dB Compression Point	$V_{\text{GAIN}} = -0.7\text{ V}$, $f = 10\text{ MHz}$		14.5		dBm
	$V_{\text{GAIN}} = +0.7\text{ V}$, $f = 10\text{ MHz}$		–1.7		dBm
Two-Tone Intermodulation Distortion (IMD3)	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 1\text{ V p-p}$, $f_1 = 10\text{ MHz}$, $f_2 = 11\text{ MHz}$		–74		dBc
	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 1\text{ V p-p}$, $f_1 = 45\text{ MHz}$, $f_2 = 46\text{ MHz}$		–60		dBc
	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 2\text{ V p-p}$, $f_1 = 10\text{ MHz}$, $f_2 = 11\text{ MHz}$		–64		dBc
	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 2\text{ V p-p}$, $f_1 = 45\text{ MHz}$, $f_2 = 46\text{ MHz}$		–49		dBc
Output Third-Order Intercept	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 1\text{ V p-p}$, $f = 10\text{ MHz}$		35		dBm
	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 1\text{ V p-p}$, $f = 45\text{ MHz}$		28		dBm
	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 2\text{ V p-p}$, $f = 10\text{ MHz}$		36		dBm
	$V_{\text{GAIN}} = 0\text{ V}$, $V_{\text{OUT}} = 2\text{ V p-p}$, $f = 45\text{ MHz}$		28		dBm
Overload Recovery	$V_{\text{GAIN}} = 0.7\text{ V}$, $V_{\text{IN}} = 0.1\text{ V p-p}$ to 1 V p-p		50		ns
ACCURACY					
Absolute Gain Error	$-0.7\text{ V} < V_{\text{GAIN}} < -0.6\text{ V}$		0.7 to 3.5		dB
	$-0.6\text{ V} < V_{\text{GAIN}} < -0.5\text{ V}$	–1.25	± 0.35	+1.25	dB
	$-0.5\text{ V} < V_{\text{GAIN}} < +0.5\text{ V}$	–1.0	± 0.25	+1.0	dB
	$0.5\text{ V} < V_{\text{GAIN}} < 0.6\text{ V}$	–1.25	± 0.35	+1.25	dB
	$0.6\text{ V} < V_{\text{GAIN}} < 0.7\text{ V}$		–0.7 to –3.5		dB
GAIN CONTROL INTERFACE					
Gain Scaling Factor	$-0.6\text{ V} < V_{\text{GAIN}} < +0.6\text{ V}$		19.7		dB/V
Gain Range			24		dB
Intercept	$V_{\text{GAIN}} = 0\text{ V}$		12.65		dB
Input Voltage (V_{GAIN}) Range	No foldover	$-V_S$		$+V_S$	V
Input Impedance			70		M Ω
Bias Current	$-0.7\text{ V} < V_{\text{GAIN}} < +0.7\text{ V}$		0.3		μA
Response Time	24 dB gain change		200		ns
POWER SUPPLY					
Supply Voltage	V_{POS} to V_{NEG} (dual- or single-supply operation)	4.5	5	10	V
$V_S = \pm 2.5\text{ V}$					
Quiescent Current	Each supply (V_{POS} and V_{NEG})	10.5	15.5	23.5	mA
Power Dissipation	No signal, V_{POS} to $V_{\text{NEG}} = 5\text{ V}$		78		mW
PSRR	$V_{\text{GAIN}} = 0.7\text{ V}$, $f = 1\text{ MHz}$		–40		dB
$V_S = \pm 5\text{ V}$					
Quiescent Current	Each supply (V_{POS} and V_{NEG})	13.5	18.5	25.5	mA
Power Dissipation	No signal, V_{POS} to $V_{\text{NEG}} = 10\text{ V}$		185		mW
PSRR	$V_{\text{GAIN}} = 0.7\text{ V}$, $f = 1\text{ MHz}$		–40		dB

ABSOLUTE MAXIMUM RATINGS

Table 2.

Parameter	Rating
Voltage	
Supply Voltage (VPOS, VNEG)	± 6 V
Input Voltage (INPx)	VPOS, VNEG
GAIN Voltage	VPOS, VNEG
Power Dissipation (Exposed Pad Soldered to PC Board)	866 mW
Temperature	
Operating Temperature Range	-40°C to $+85^{\circ}\text{C}$
Storage Temperature Range	-65°C to $+150^{\circ}\text{C}$
Lead Temperature (Soldering, 60 sec)	300°C
Thermal Data—4-Layer JEDEC Board No Air Flow Exposed Pad Soldered to PC Board	
θ_{JA}	75.4°C/W
θ_{JB}	47.5°C/W
θ_{JC}	17.9°C/W
ψ_{JT}	2.2°C/W
ψ_{JB}	46.2°C/W

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.

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

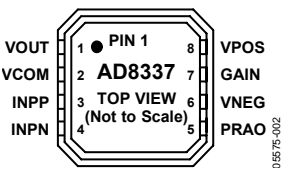


Figure 2. Pin Configuration

Table 3. Pin Function Descriptions

Pin No.	Mnemonic	Description
1	VOUT	VGA Output.
2	VCOM	Common Ground when using Plus and Minus Supply Voltages. For single-supply operation, provide half the positive supply voltage at Pin VPOS to Pin VCOM.
3	INPP	Positive Input to Preamplifier.
4	INPN	Negative Input to Preamplifier.
5	PRAO	Preamplifier Output.
6	VNEG	Negative Supply (–VPOS for Dual-Supply; GND for Single-Supply).
7	GAIN	Gain Control Input Centered at VCOM.
8	VPOS	Positive Supply.

TYPICAL PERFORMANCE CHARACTERISTICS

$V_S = \pm 2.5\text{ V}$, $T_A = 25^\circ\text{C}$, $R_L = 500\ \Omega$, including a $20\ \Omega$ snubbing resistor, $f = 10\text{ MHz}$, $C_L = 2\text{ pF}$, $V_{IN} = 10\text{ mV p-p}$, noninverting configuration, unless otherwise noted.

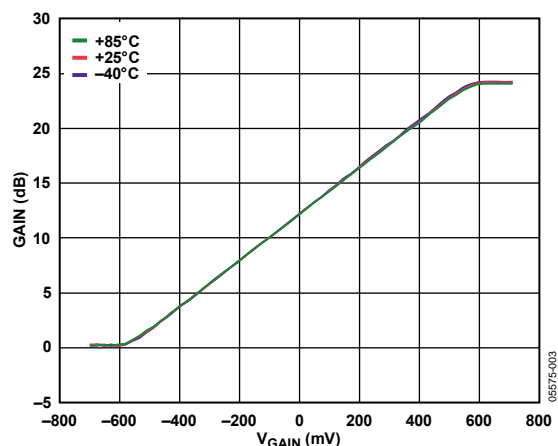


Figure 3. Gain vs. V_{GAIN} at Three Temperatures
See Figure 44

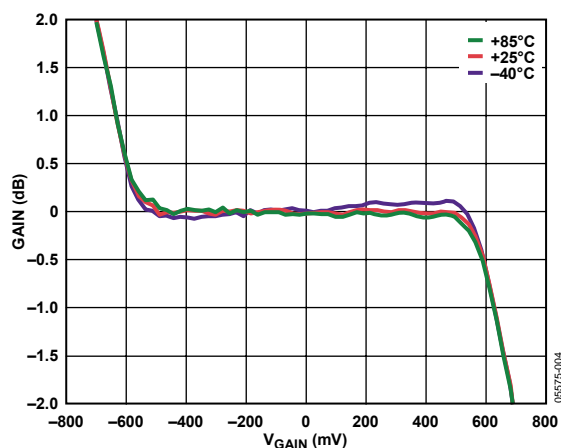


Figure 4. Gain Error vs. V_{GAIN} at Three Temperatures
See Figure 44

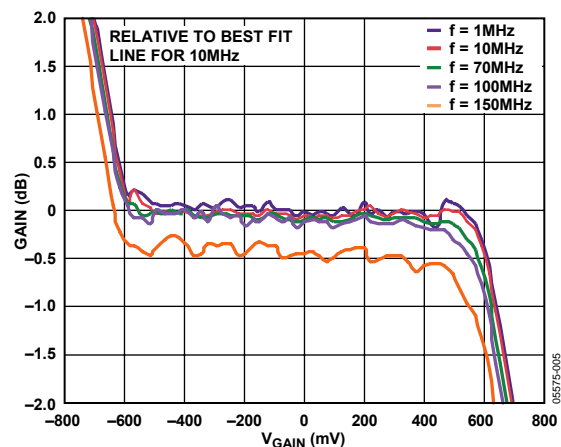


Figure 5. Gain Error vs. V_{GAIN} at Five Frequencies
See Figure 44

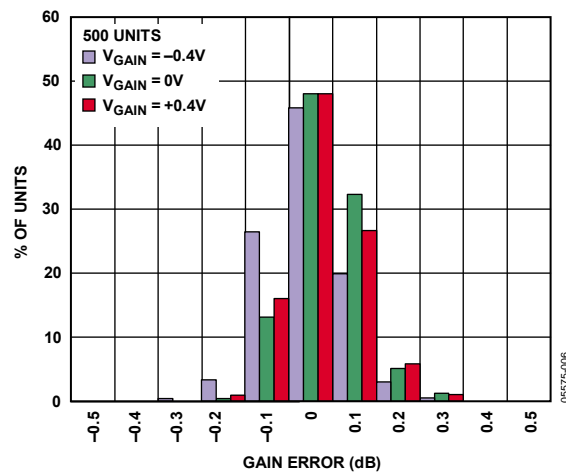


Figure 6. Gain Error Histogram for Three Values of V_{GAIN}

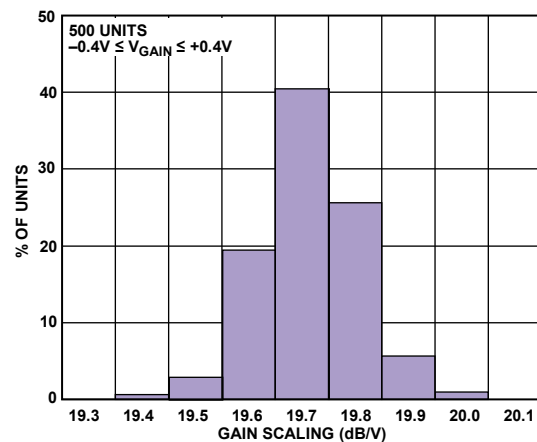


Figure 7. Gain Scaling Histogram

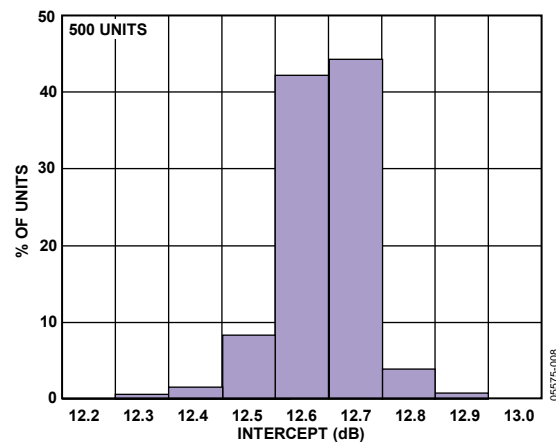


Figure 8. Intercept Histogram

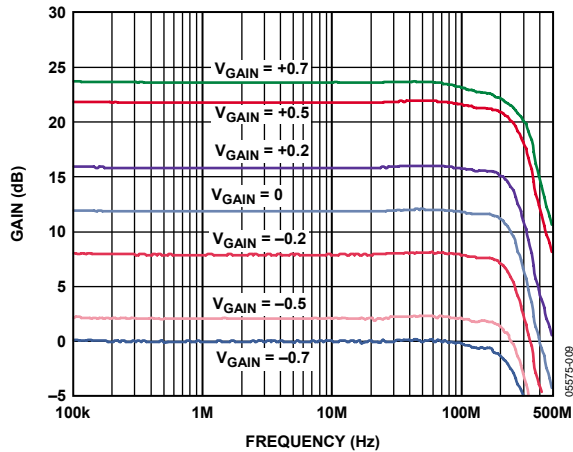


Figure 9. Frequency Response for Various Values of V_{GAIN}
See Figure 45

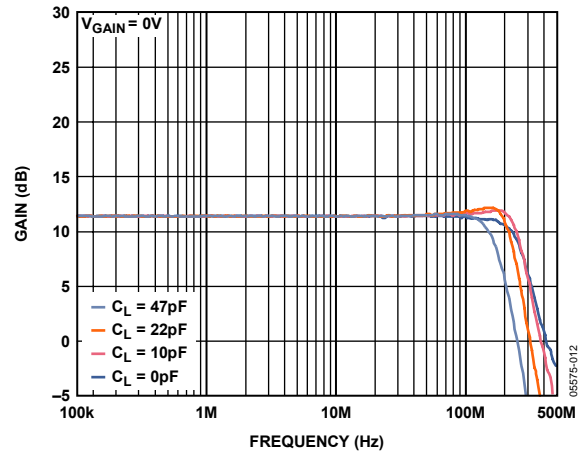


Figure 12. Frequency Response for Three Values of C_L
with a $20\ \Omega$ Snubbing Resistor
See Figure 45

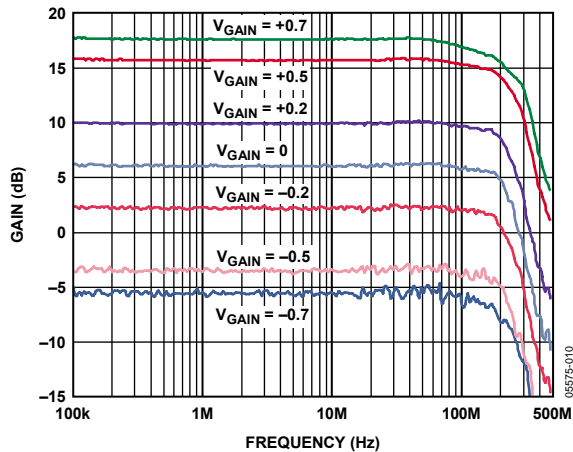


Figure 10. Frequency Response for Various Values of V_{GAIN} —Inverting Input
See Figure 58

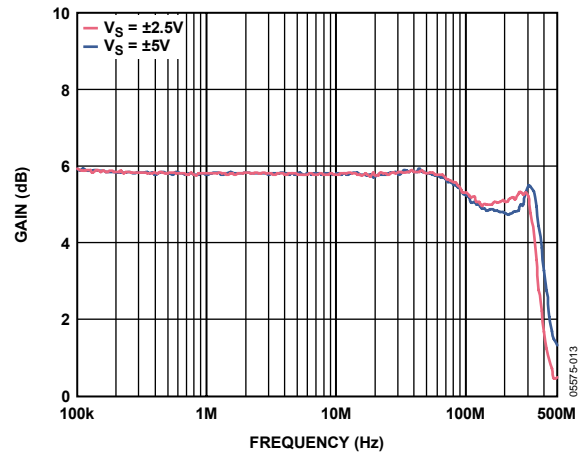


Figure 13. Frequency Response—Preamp
See Figure 46

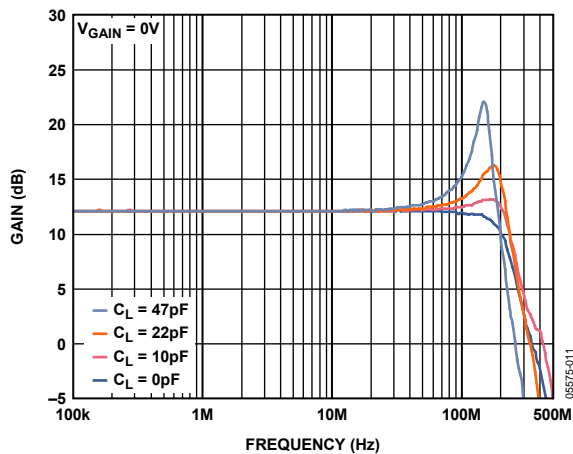


Figure 11. Frequency Response for Three Values of C_L
See Figure 45

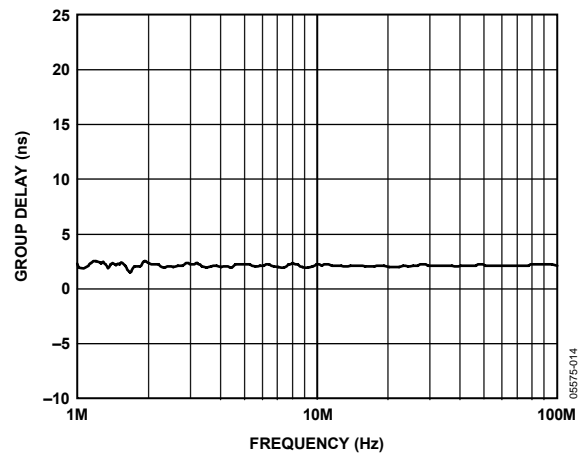


Figure 14. Group Delay vs. Frequency
See Figure 47

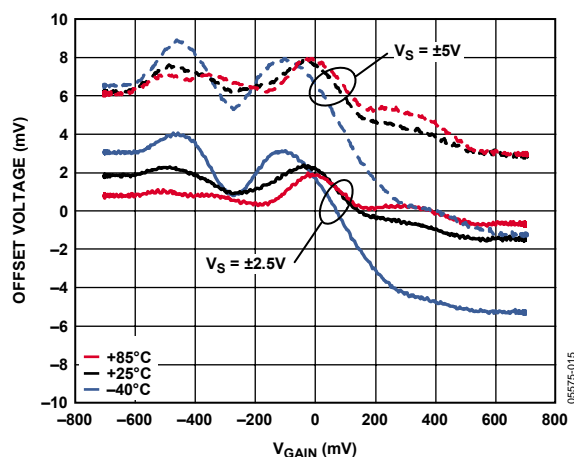


Figure 15. Offset Voltage vs. V_{GAIN} at Three Temperatures
See Figure 48

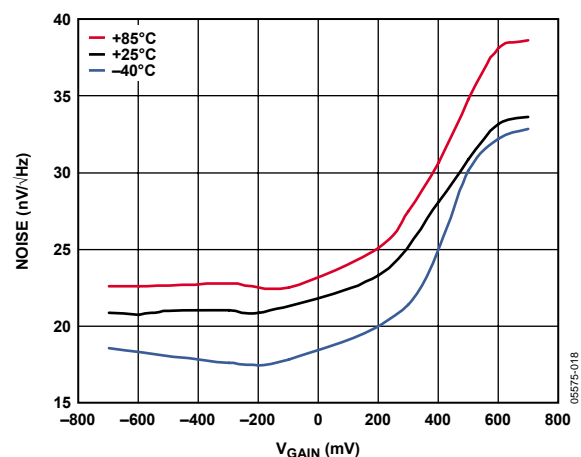


Figure 18. Output-Referred Noise vs. V_{GAIN} at Three Temperatures
See Figure 50

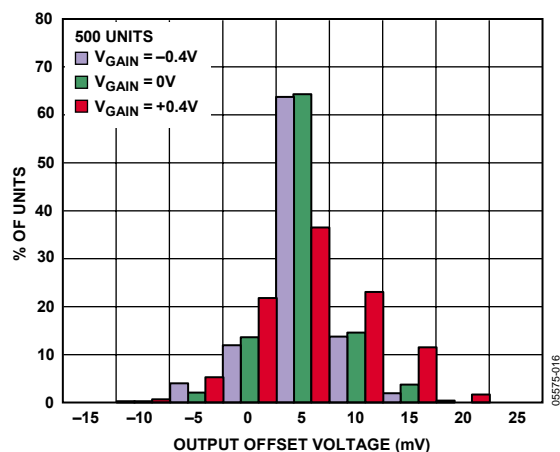


Figure 16. Output Offset Voltage Histogram for Three Values of V_{GAIN}

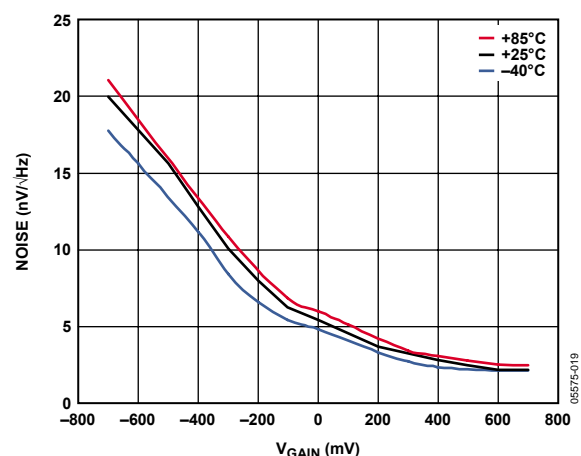


Figure 19. Short-Circuit, Input-Referred Noise at Three Temperatures
See Figure 50

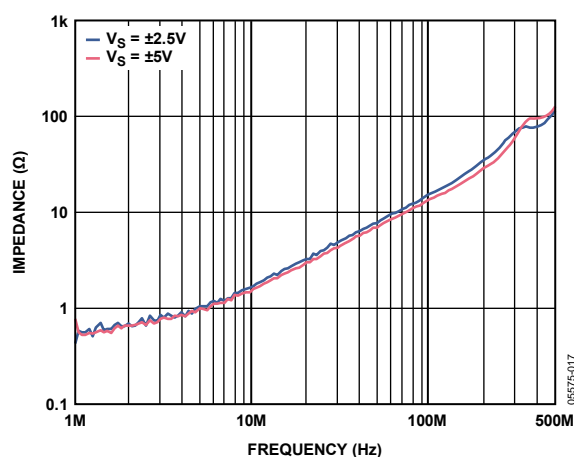


Figure 17. VGA Output Impedance vs. Frequency
See Figure 49

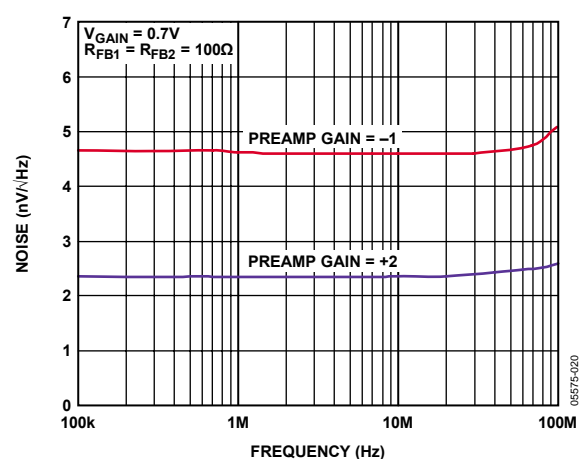


Figure 20. Short-Circuit, Input-Referred Noise vs. Frequency at Max Gain—
Inverting and Noninverting Preamp Gain = -1 and +2
See Figure 50

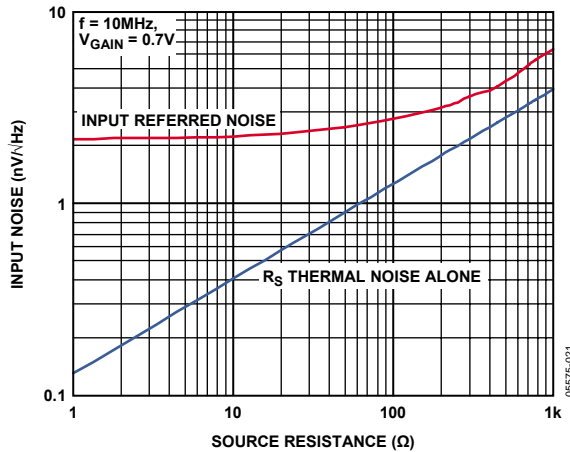


Figure 21. Input-Referred Noise vs. R_S
See Figure 61

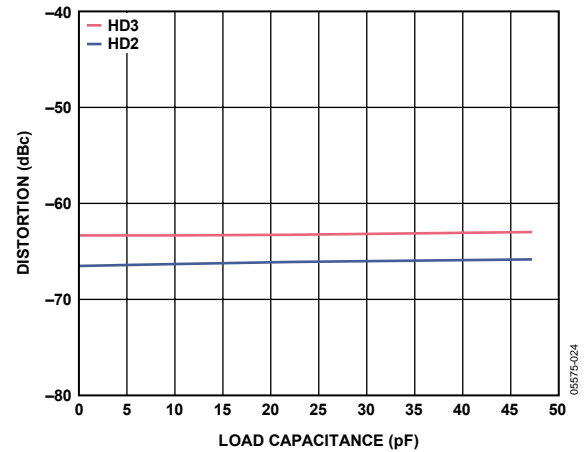


Figure 24. Harmonic Distortion vs. Load Capacitance
See Figure 52

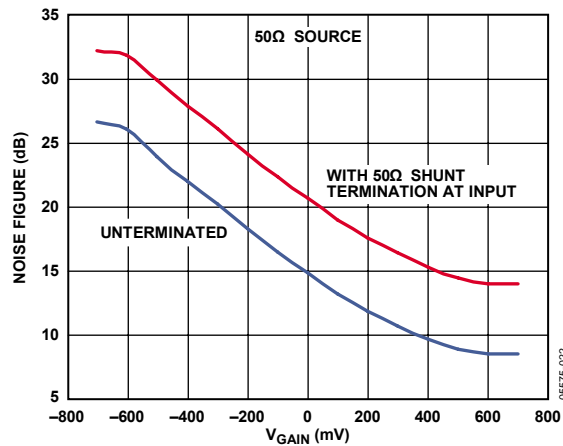


Figure 22. Noise Figure vs. V_{GAIN}
See Figure 51

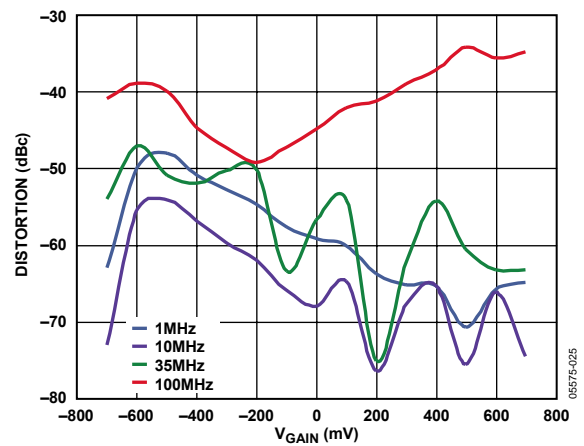


Figure 25. HD2 vs. V_{GAIN} at Four Frequencies
See Figure 52

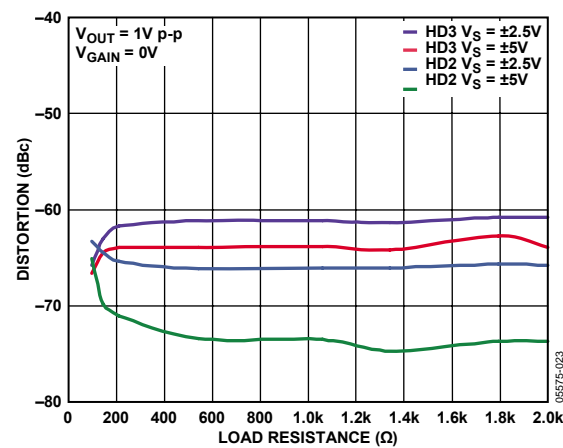


Figure 23. Harmonic Distortion vs. R_{LOAD} and Supply Voltage
See Figure 52

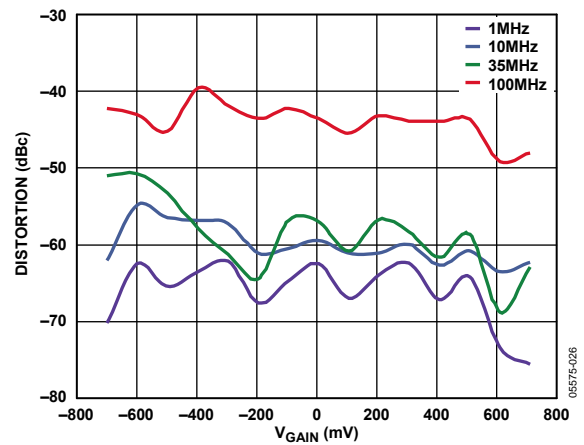


Figure 26. HD3 vs. V_{GAIN} at Four Frequencies
See Figure 52

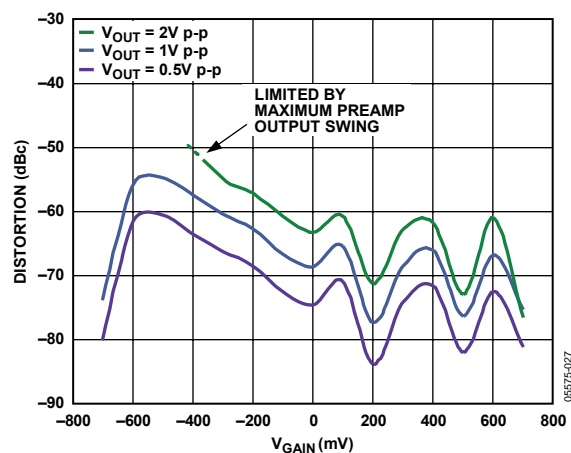


Figure 27. HD2 vs. V_{GAIN} for Three Levels of Output Voltage
See Figure 52

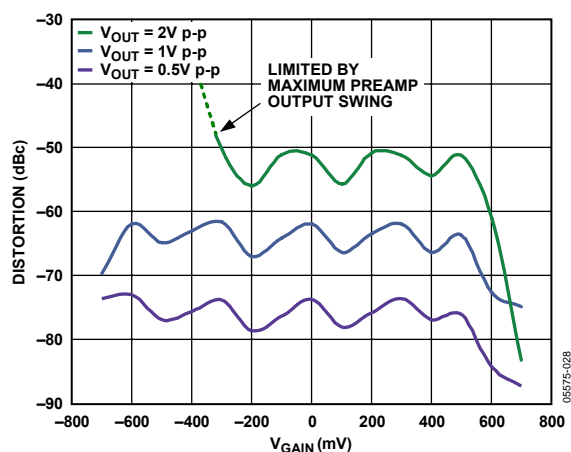


Figure 28. HD3 vs. V_{GAIN} for Three Levels of Output Voltage
See Figure 52

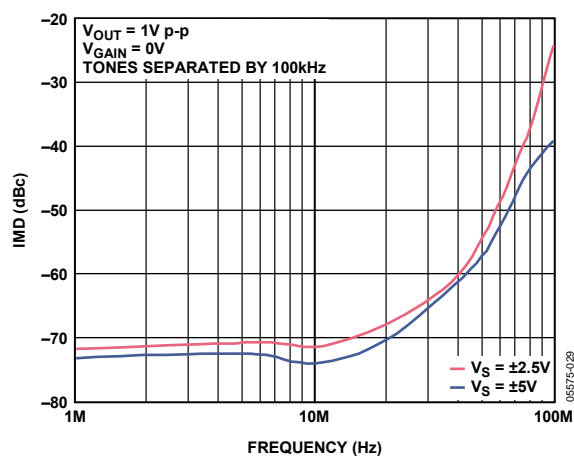


Figure 29. IMD3 vs. Frequency
See Figure 64

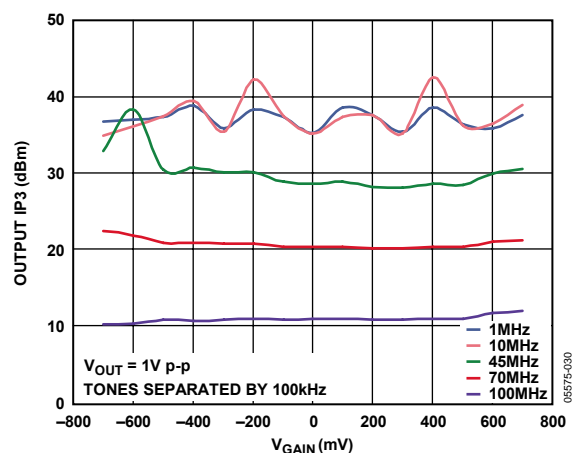


Figure 30. Output-Referred IP3 (OIP3) vs. V_{GAIN}
at Five Frequencies
See Figure 64

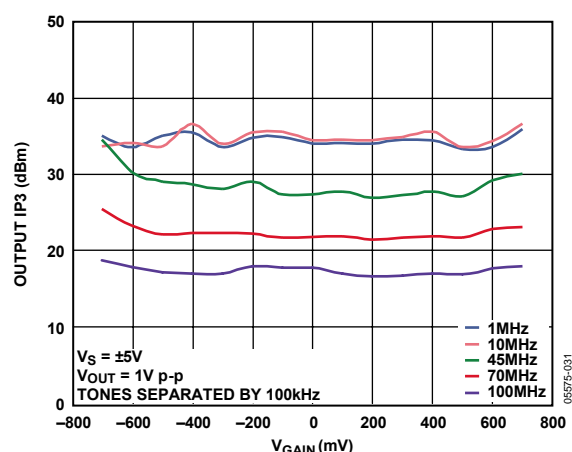


Figure 31. Output-Referred IP3 (OIP3) vs. V_{GAIN} , $V_S = \pm 5V$
at Five Frequencies
See Figure 64

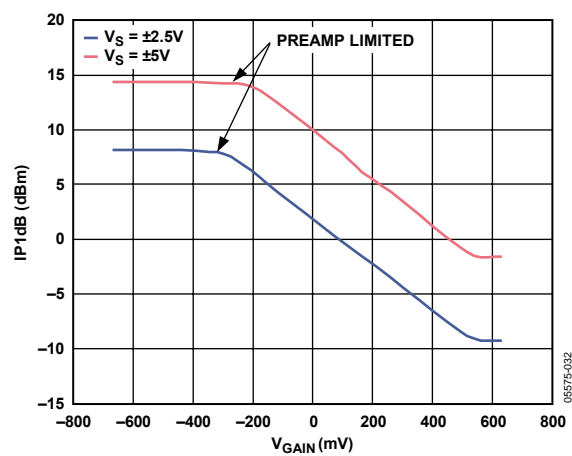


Figure 32. Input P1dB (IP1dB) vs. V_{GAIN}
See Figure 63

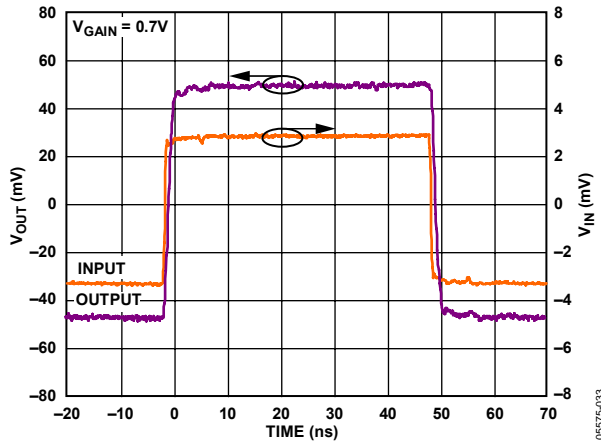


Figure 33. Small Signal Pulse Response
See Figure 53

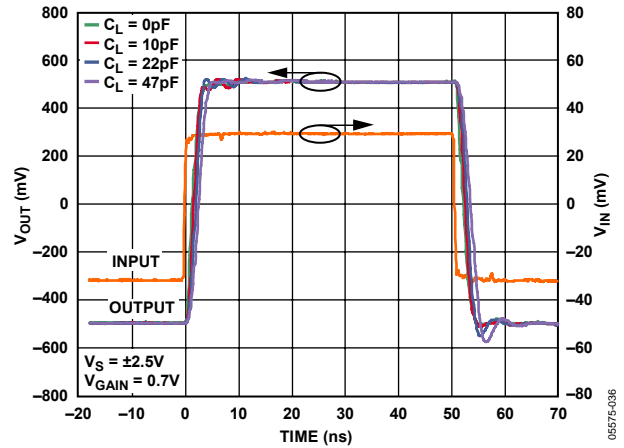


Figure 36. Large Signal Pulse Response for Three Capacitive Loads
See Figure 53

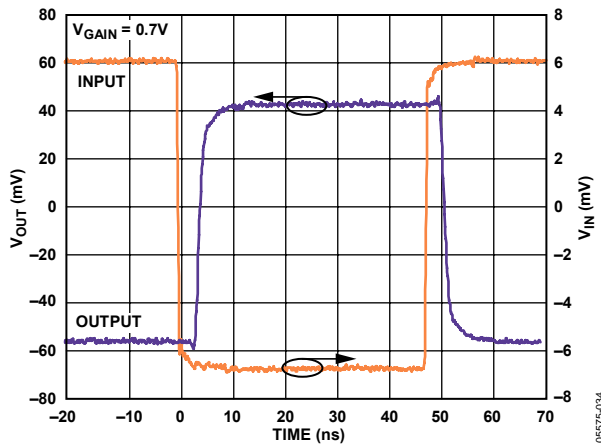


Figure 34. Small Signal Pulse Response—Inverting Feedback
See Figure 59

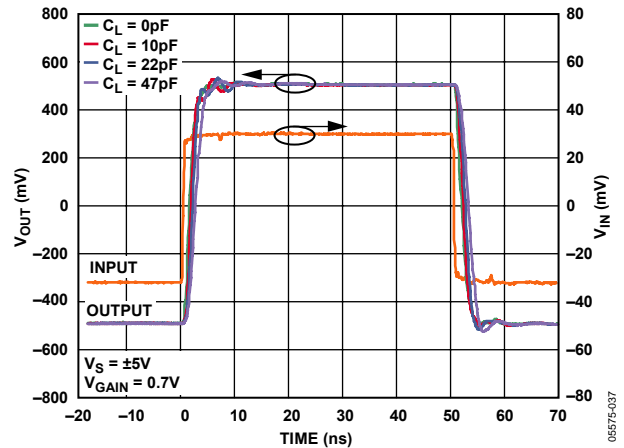


Figure 37. Large Signal Pulse Response for Three Capacitive Loads, $V_S = \pm 5V$
See Figure 53

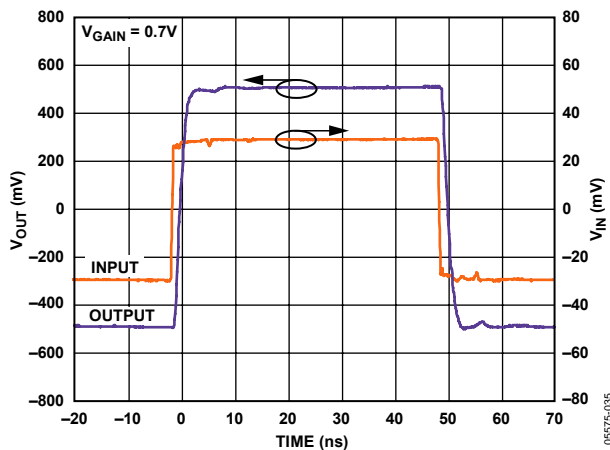


Figure 35. Large Signal Pulse Response
See Figure 53

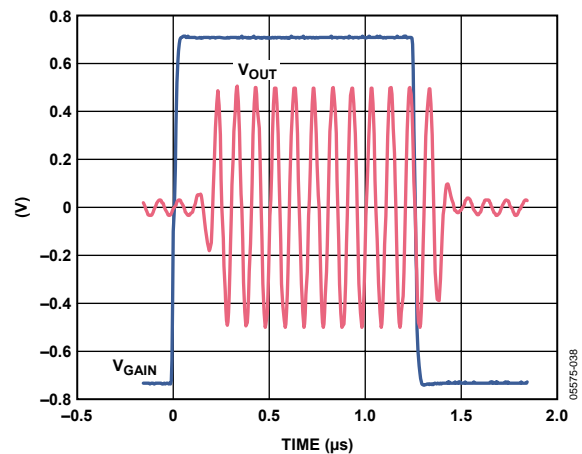


Figure 38. Gain Response
See Figure 54

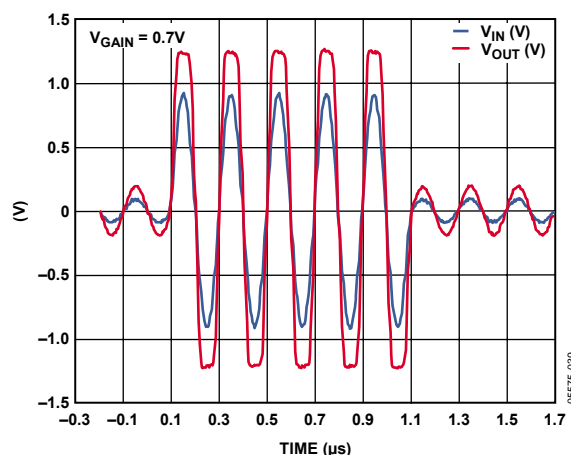


Figure 39. Preamplifier Overdrive Recovery
See Figure 55

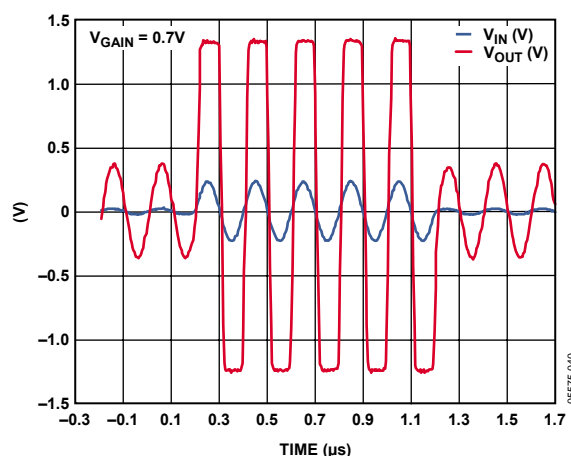


Figure 40. VGA Overdrive Recovery
See Figure 56

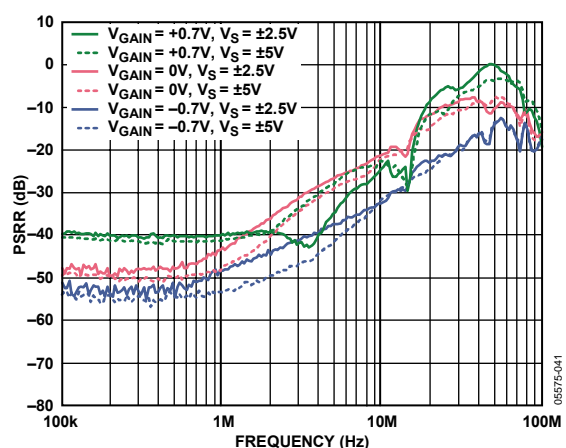


Figure 41. PSRR vs. Frequency of Positive Supply
See Figure 60

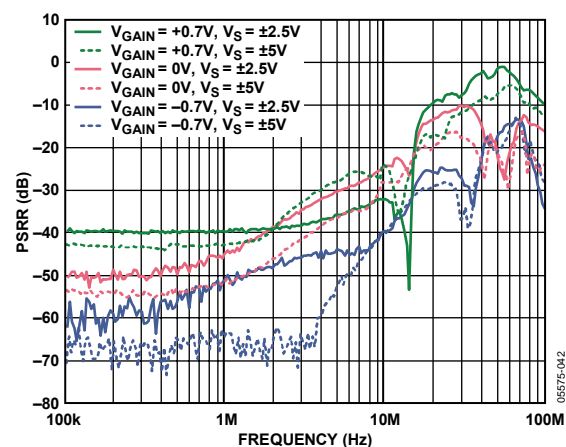


Figure 42. PSRR vs. Frequency of Negative Supply
See Figure 60

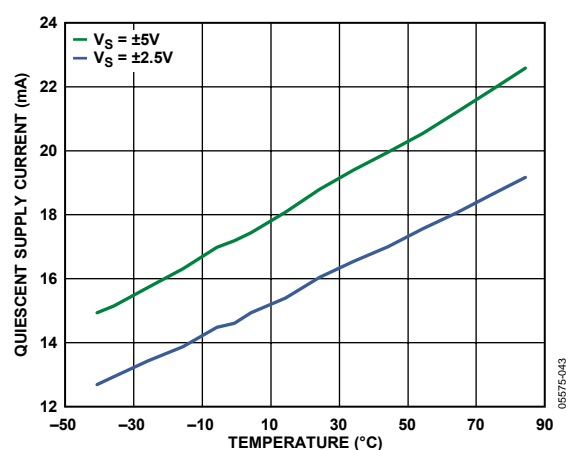


Figure 43. Quiescent Supply Current vs. Temperature
See Figure 57

TEST CIRCUITS

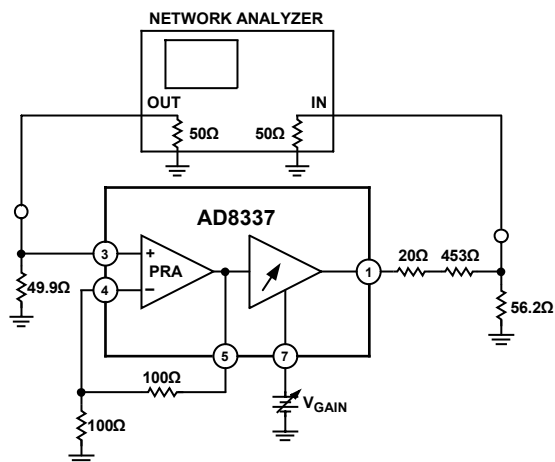


Figure 44. Gain and Gain Error vs. V_{GAIN}

05575-044

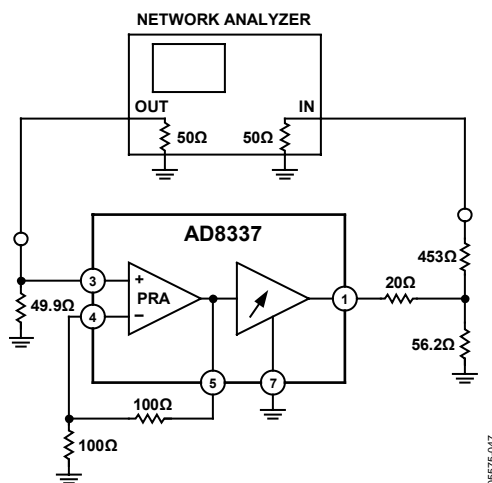


Figure 47. Group Delay

05575-047

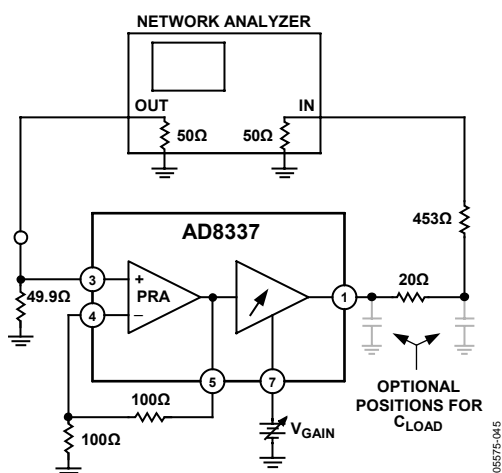


Figure 45. Frequency Response

05575-045

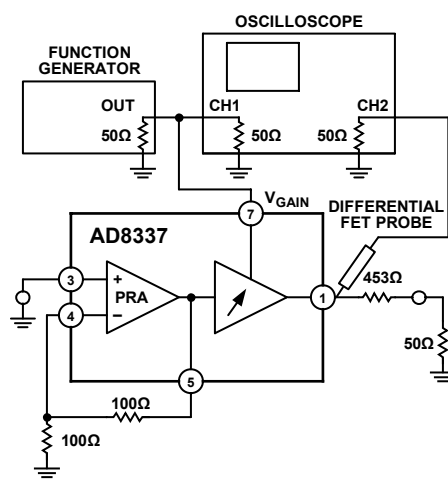


Figure 48. Offset Voltage

05575-048

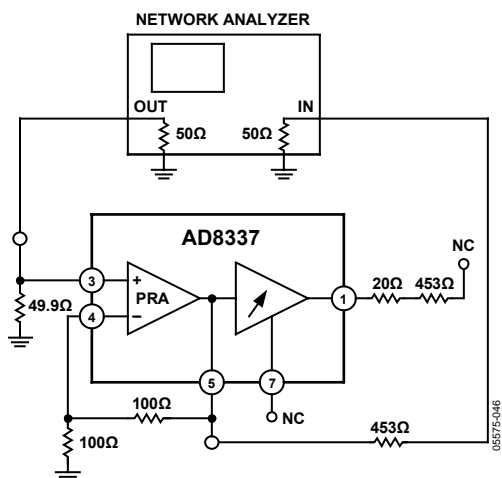


Figure 46. Frequency Response—Preamp

05575-046

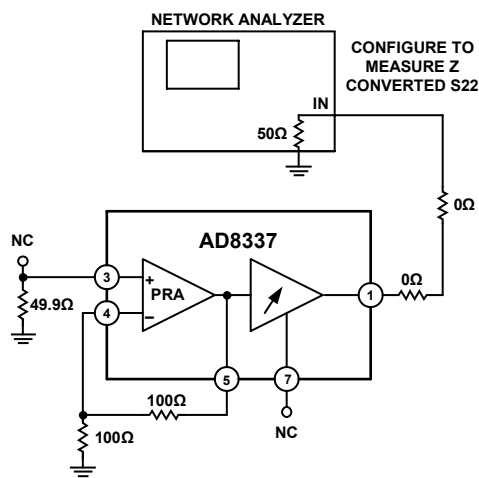


Figure 49. Output Resistance vs. Frequency

05575-049

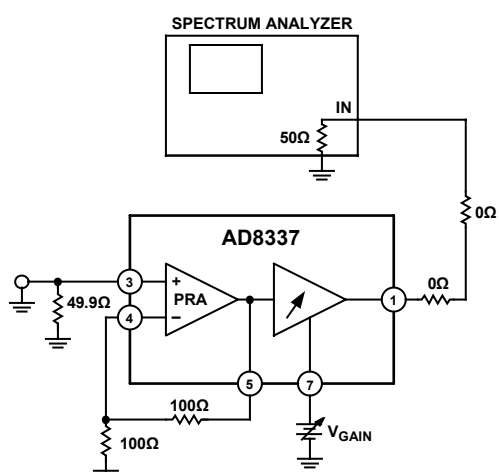


Figure 50. Input-Referred and Output-Referred Noise

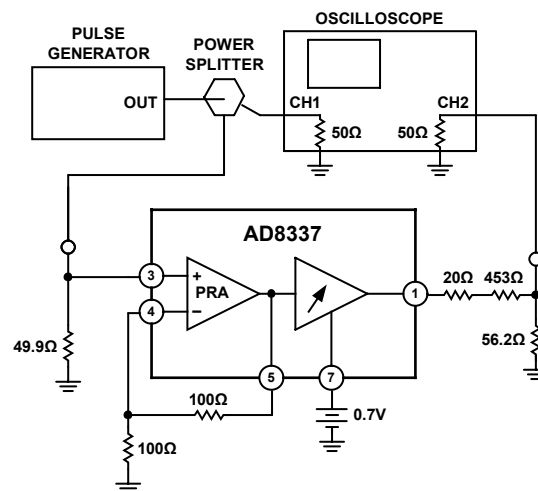


Figure 53. Pulse Response

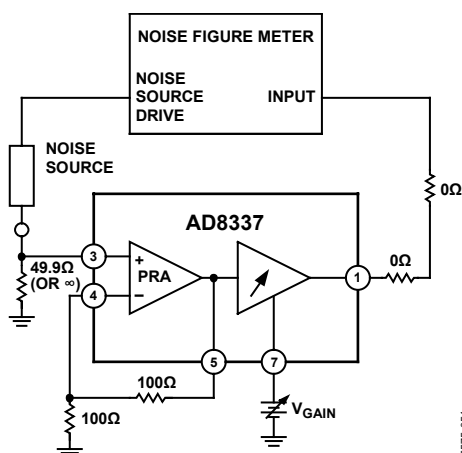
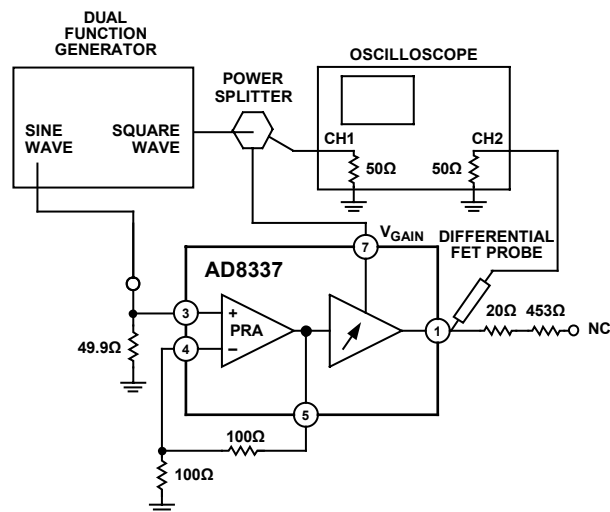
Figure 51. Noise Figure vs. V_{GAIN} 

Figure 54. Gain Response

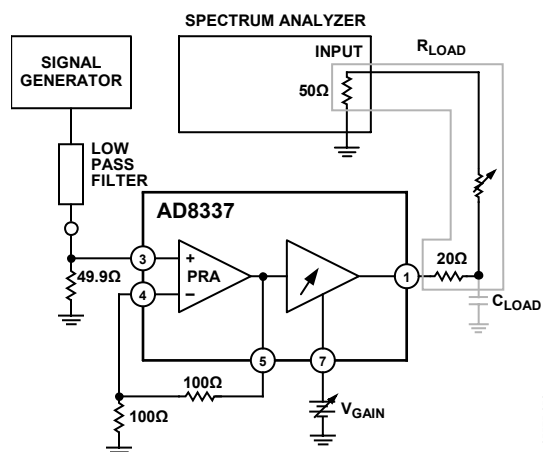


Figure 52. Harmonic Distortion

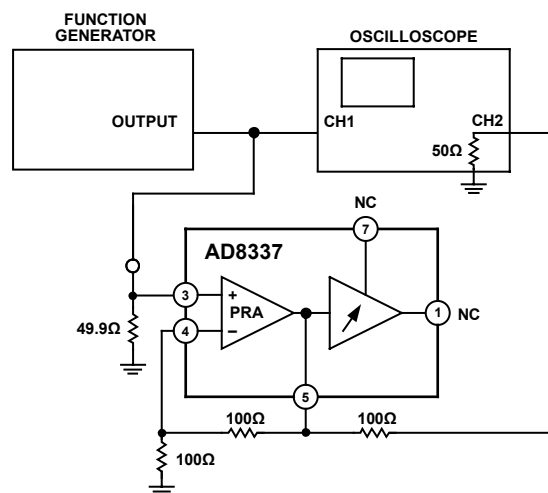


Figure 55. Preamp Overdrive Recovery

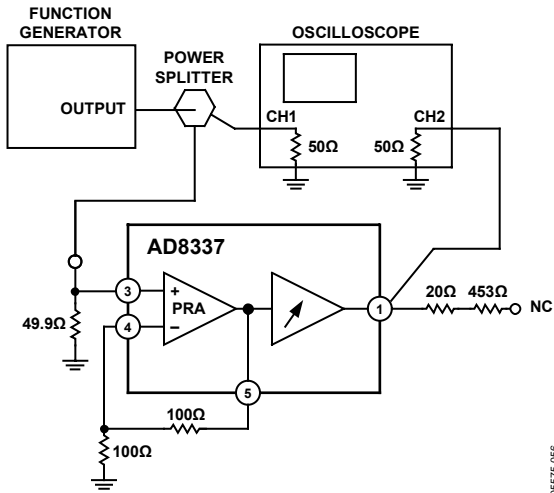


Figure 56. VGA Overdrive Recovery

05575-066

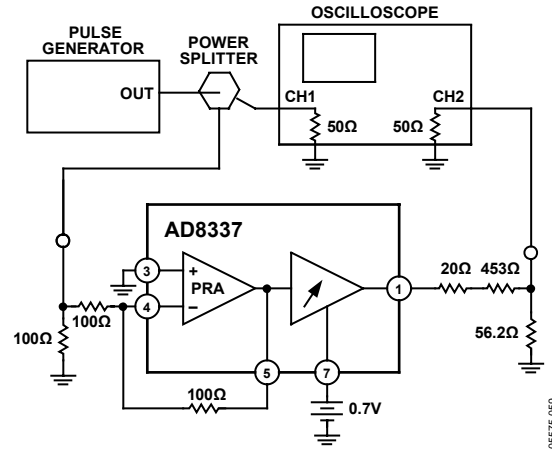


Figure 59. Pulse Response—Inverting Feedback

05575-069

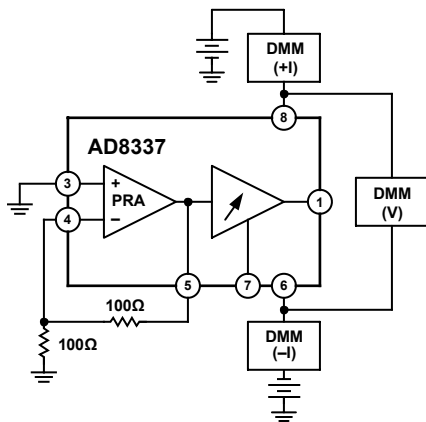


Figure 57. Supply Current

05575-067

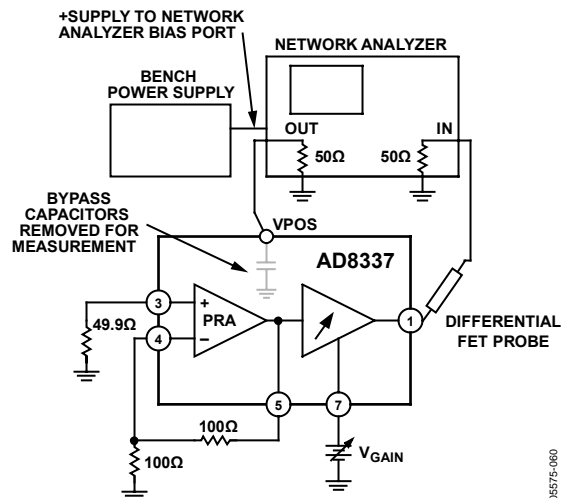


Figure 60. PSRR

05575-060

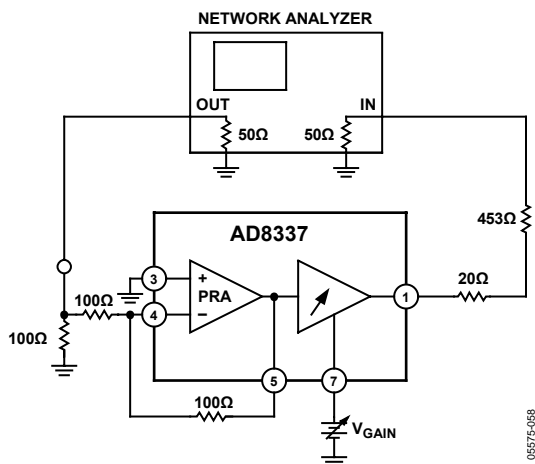


Figure 58. Frequency Response—Inverting Feedback

05575-068

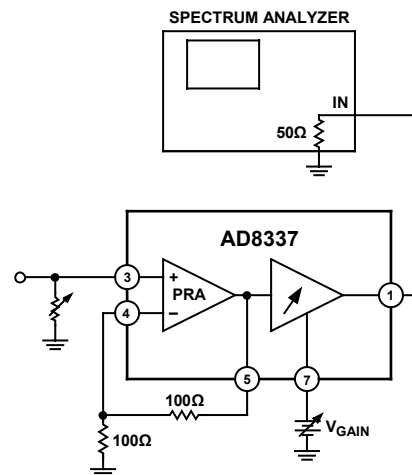


Figure 61. Input-Referred Noise vs. R_s

05575-061

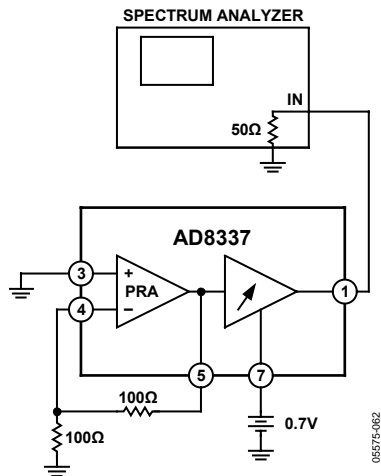


Figure 62. Short-Circuit Input Noise vs. Frequency

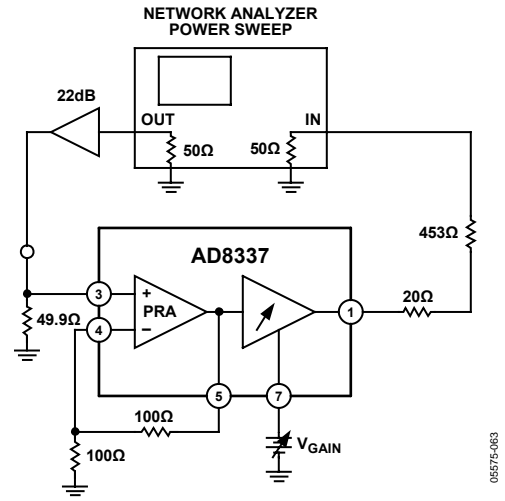


Figure 63. IP1dB vs. V_{GAIN}

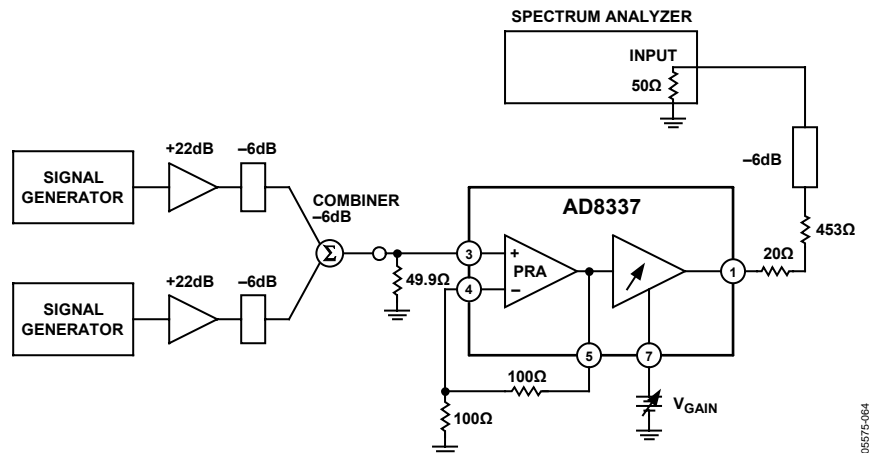


Figure 64. IMD and OIP3

THEORY OF OPERATION

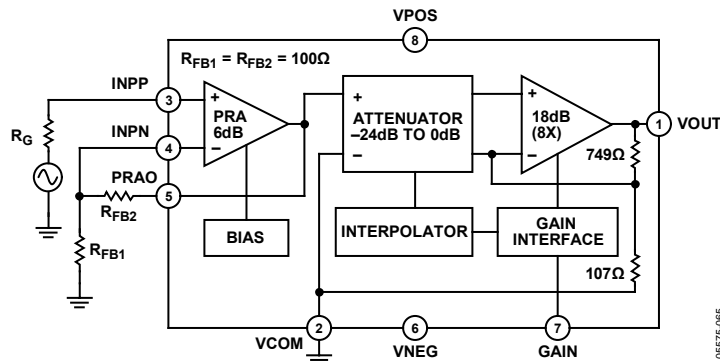


Figure 65. Block Diagram

OVERVIEW

The AD8337 is a low noise, single-ended, linear-in-dB, general-purpose, variable gain amplifier (VGA) usable at frequencies up to 100 MHz. It is fabricated using a proprietary Analog Devices dielectrically isolated, complementary bipolar process. The bandwidth is dc to 280 MHz and features low dc offset voltage and an ideal nominal gain range of 0 dB to 24 dB. Requiring about 15.5 mA, the power consumption is only 78 mW from either a single +5 V or a dual ± 2.5 V supply. Figure 65 is the circuit block diagram of the AD8337.

PREAMPLIFIER

An uncommitted, current-feedback op amp included in the AD8337 can be used as a preamplifier to buffer the ladder network attenuator of the X-AMP. As with any op amp, the gain is established using external resistors, and the preamplifier is specified with a noninverting gain of 6 dB (2 \times) and gain resistor values of 100 Ω . The preamplifier gain can be increased using larger values of R_{FB2} , trading off bandwidth and offset voltage. The value of R_{FB2} should be ≥ 100 Ω because it and an internal compensation capacitor determines the 3 dB bandwidth, and smaller values can compromise preamplifier stability.

Because the AD8337 is dc-coupled, larger preamp gains increase the offset voltage. The offset voltage can be compensated by connecting a resistor between the INPN input and the supply voltage. If the offset is negative, the resistor value connects to the negative supply. For ease of adjustment, a trimmer network can be used.

For larger gains, the overall noise is reduced if a low value of R_{FB1} is selected. For values of $R_{FB1} = 20$ Ω and $R_{FB2} = 301$ Ω , the preamp gain is 16 \times (24.1 dB), and the input-referred noise is approximately 1.5 nV/ $\sqrt{\text{Hz}}$. For this value of gain, the overall gain range increases by 18 dB; therefore, the gain range is 18 dB to 42 dB.

VGA

This X-AMP, with its linear-in-dB gain characteristic architecture, yields the optimum dynamic range for receiver applications. Referring to Figure 65, the signal path consists of a -24 dB variable attenuator followed by a fixed gain amplifier of 18 dB, for a total VGA gain range of -6 dB to +18 dB. With the preamplifier configured for a gain of 6 dB, the composite gain range is 0 dB to 24 dB.

The VGA plus preamp with 6 dB of gain implements the following exact gain law

$$\text{Gain(dB)} = \left[19.7 \frac{\text{dB}}{\text{V}} \times V_{\text{GAIN}} \right] + \text{ICPT(dB)}$$

where the nominal intercept (ICPT) is 12.65 dB.

The ICPT increases as the gain of the preamp is increased. For example, if the gain of the preamp is increased by 6 dB, ICPT increases to 18.65 dB. Although the previous equation shows the exact gain law as based on statistical data, a quick estimation of signal levels can be made using the default slope of 20 dB/V for a particular gain setting. For example, the change in gain for a V_{GAIN} change of 0.3 V is 6 dB using a slope of 20 dB/V and 5.91 dB using the exact slope of 19.6 dB/V. This is a difference of only 0.09 dB.

GAIN CONTROL

The gain control interface provides a high impedance input and is referenced to VCOM pin (in a single-supply application to midsupply at $[\text{VPOS} + \text{VNEG}]/2$ for optimum swing). When dual supplies are used, VCOM is connected to ground. The voltage on the VCOM pin determines the midpoint of the gain range. For a ground referenced design, the V_{GAIN} range is from -0.7 V to +0.7 V with the most linear-in-dB section of the gain control between -0.6 V and +0.6 V. In the center 80% of the V_{GAIN} range, the gain error is typically less than ± 0.2 dB. The gain control voltage can be increased or decreased to the positive or negative rails without gain foldover.

The gain scaling factor (gain slope) is designed for 20 dB/V; this relatively low slope ensures that noise on the GAIN input is not unduly amplified. Because a VGA functions as a multiplier, it is important to make sure that the GAIN input does not inadvertently modulate the output signal with unwanted noise. Because of its high input impedance, a simple low-pass filter can be added to the GAIN input to filter unwanted noise.

OUTPUT STAGE

The output stage is a Class AB, voltage-feedback, complementary emitter-follower with a fixed gain of 18 dB, similar to the preamplifier in speed and bandwidth. Because of the ac-beta roll-off of the output devices and the inherent reduction in feedback beyond the -3 dB bandwidth, the impedance looking into the output pin of the preamp and output stages appears to be inductive (increasing impedance with increasing frequency). The high speed output amplifier used in the AD8337 can drive large currents, but its stability is susceptible to capacitive loading. A small series resistor mitigates the effects of capacitive loading (see the Applications section).

ATTENUATOR

The input resistance of the VGA attenuator is nominally 265 Ω . Assuming the default preamplifier feedback network $R_{FB1} + R_{FB2}$ is 200 Ω , the effective preamplifier load is about 114 Ω . The attenuator is composed of eight 3.01 dB sections for a total attenuation range of -24.08 dB. Following the attenuator is a fixed gain amplifier with 8 \times (18.06 dB) gain. Because of this relatively low gain, the output offset is kept well below 20 mV over temperature; the offset is largest at maximum gain when the preamplifier offset is amplified. The VCOM pin defines the common-mode reference for the output, as shown in Figure 65.

SINGLE-SUPPLY OPERATION AND AC COUPLING

If the AD8337 is to be operated from a single 5 V supply, the bias supply for VCOM must be a very low impedance 2.5 V reference, especially if dc coupling is used. If the device is dc-coupled, the VCOM source must be able to handle the preamplifier and VGA dynamic load currents in addition to the bias currents.

When ac coupling the preamplifier input, a bias network and bypass capacitor must be connected to the opposite polarity input pin. The bias generator for Pin VCOM must provide the dynamic current to the preamplifier feedback network and the VGA attenuator. For many single 5 V applications, a reference, such as the [ADR391](#), and a good op amp provide an adequate VCOM source if a 2.5 V supply is unavailable.

NOISE

The total input-referred voltage and current noise of the positive input of the preamplifier are about 2.2 nV/ $\sqrt{\text{Hz}}$ and 4.8 pA/ $\sqrt{\text{Hz}}$. The VGA output-referred noise is about 21 nV/ $\sqrt{\text{Hz}}$ at low gains. This result is divided by the VGA fixed gain amplifier gain of 8 \times and results in a voltage noise density of 2.6 nV/ $\sqrt{\text{Hz}}$ referred to the VGA input. This value includes the noise of the VGA gain setting resistors as well. If this voltage is again divided by the preamp gain of 2, the VGA noise referred all the way to the preamp input is about 1.3 nV/ $\sqrt{\text{Hz}}$. From this, it is determined that the preamplifier, including the 100 Ω gain setting resistors, contributes about 1.8 nV/ $\sqrt{\text{Hz}}$. The two 100 Ω resistors contribute 1.29 nV/ $\sqrt{\text{Hz}}$ each at the output of the preamp. With the gain resistor noise subtracted, the preamplifier noise is about 1.55 nV/ $\sqrt{\text{Hz}}$.

Equation 2 shows the calculation that determines the output-referred noise at maximum gain (24 dB or 16 \times).

where:

- A_t is the total gain from preamp input to VGA output.
- R_S is the source resistance.
- e_{n-PrA} is the input-referred voltage noise of the preamp.
- i_{n-PrA} is the current noise of the preamp at the INPP pin.
- $e_{n-R_{FB1}}$ is the voltage noise of R_{FB1} .
- $e_{n-R_{FB2}}$ is the voltage noise of R_{FB2} .
- e_{n-VGA} is the input-referred voltage noise of the VGA (low gain, output-referred noise divided by a fixed gain of 8 \times).

Assuming $R_S = 0 \Omega$, $R_{FB1} = R_{FB2} = 100 \Omega$, $A_t = 16\times$, and $A_{VGA} = 8\times$, the noise simplifies to

$$e_{n-out} = \sqrt{(1.75 \times 16)^2 + 2(1.29 \times 8)^2 + (1.9 \times 8)^2} = 35 \text{ nV}/\sqrt{\text{Hz}} \quad (1)$$

Dividing the result by 16 gives the total input-referred noise with a short-circuited input as 2.2 nV/ $\sqrt{\text{Hz}}$. When the preamplifier is used in the inverting configuration with the same R_{FB1} and $R_{FB2} = 100 \Omega$ as previously noted, e_{n-out} does not change. However, because the gain dropped by 6 dB, the input-referred noise increases by a factor of 2 to about 4.4 nV/ $\sqrt{\text{Hz}}$. The reason for this increase is that the noise gain to the output of the noise generators stays the same, yet the preamp in the inverting configuration has a gain of -1 compared to the +2 in the noninverting configuration; this increases the input-referred noise by 2.

$$e_{n-out} = \sqrt{(R_S \times A_t)^2 + (e_{n-PrA} \times A_t)^2 + (i_{n-PrA} \times R_S)^2 + (e_{n-R_{FB1}} \times \frac{R_{FB2}}{R_{FB1}} \times A_{VGA})^2 + (e_{n-R_{FB2}} \times A_{VGA})^2 + (e_{n-VGA} \times A_{VGA})^2} \quad (2)$$

APPLICATIONS

PREAMPLIFIER CONNECTIONS

Noninverting Gain Configuration

The AD8337 preamplifier is an uncommitted, current-feedback op amp that is stable for values of $R_{FB2} \geq 100 \Omega$. See Figure 66 for the noninverting feedback connections.

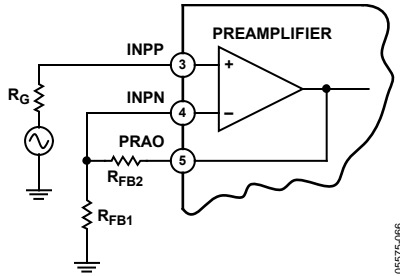


Figure 66. AD8337 Preamplifier Configured for Noninverting Gain

Two surface-mount resistors establish the preamplifier gain. Equal values of 100Ω configure the preamplifier for a 6 dB gain and the device for a default gain range of 0 dB to 24 dB.

For preamp gains ≥ 2 , select a value of $R_{FB2} \geq 100 \Omega$ and $R_{FB1} \leq 100 \Omega$. Higher values of R_{FB2} reduce the bandwidth and increase the offset voltage, but smaller values compromise stability. If $R_{FB1} \leq 100 \Omega$, the gain increases and the input-referred noise decreases.

Inverting Gain Configuration

For applications requiring polarity inversion of negative pulses, or for waveforms that require current sinking, the preamplifier can be configured as an inverting gain amplifier. When configured with bipolar supplies, the preamplifier amplifies positive or negative input voltages with no level shifting of the common-mode input voltage required. Figure 67 shows the AD8337 configured for inverting gain operation.

Because the AD8337 is a very high frequency device, stability issues can occur unless the circuit board on which it is used is carefully laid out. The stability of the preamp is affected by parasitic capacitance around the INPN pin. Position the preamp gain resistors, R_{FB1} and R_{FB2} , as close as possible to Pin 4, INPN, to minimize stray capacitance.

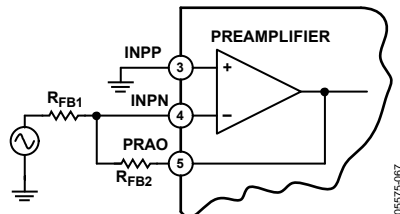


Figure 67. The AD8337 Preamplifier Configured for Inverting Gain

DRIVING CAPACITIVE LOADS

Because of the large bandwidth of the AD8337, stray capacitance at the output pin can induce peaking in the frequency response as the gain of the amplifier begins to roll off. Figure 68 shows peaking with two values of load capacitance using ± 2.5 V supplies and $V_{GAIN} = 0$ V.

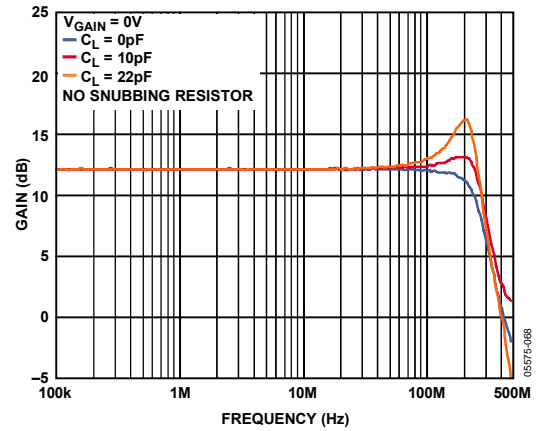


Figure 68. Peaking in the Frequency Response for Two Values of Output Capacitance with ± 2.5 V Supplies and No Snubbing Resistor

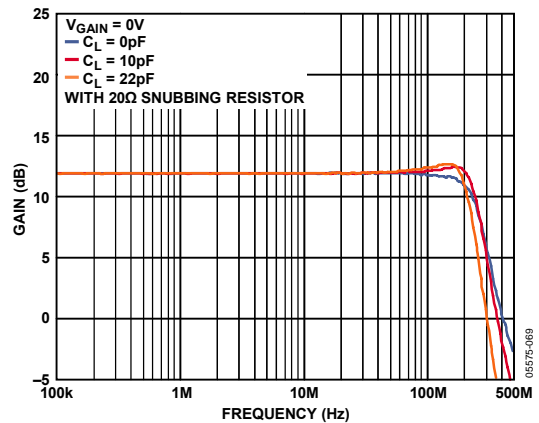


Figure 69. Frequency Response for Two Values of Output Capacitance with a 20Ω Snubbing Resistor

In the time domain, stray capacitance at the output pin can induce overshoot on the edges of transient signals, as seen in Figure 70 and Figure 72. The amplitude of the overshoot is also a function of the slewing of the transient (not shown). The transition time of the input pulses used for Figure 70 and Figure 72 was set deliberately high at 300 ps to demonstrate the fast response time of the amplifier. Signals with longer transition times generate less overshoot.

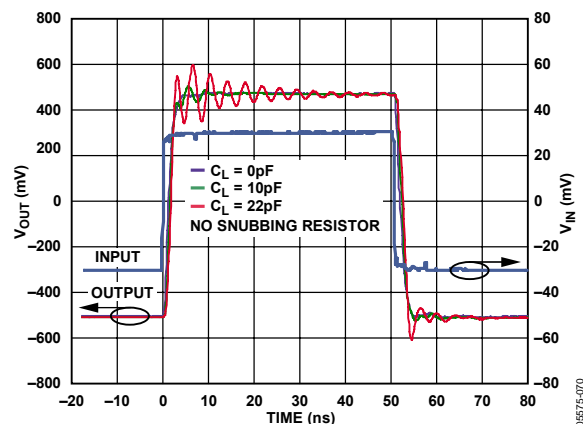


Figure 70. Pulse Response for Two Values of Output Capacitance with ± 2.5 V Supplies and No Snubbing Resistor

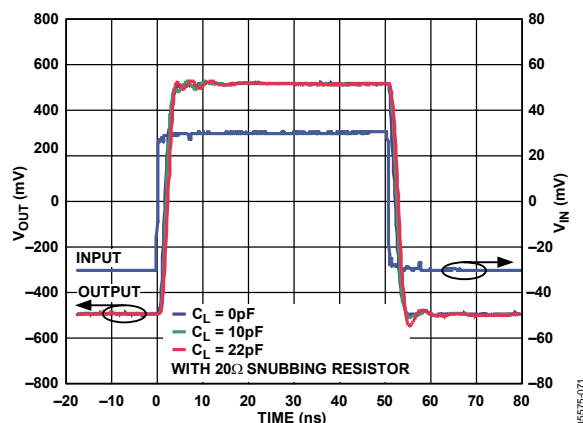


Figure 71. Pulse Response for Two Values of Output Capacitance with ± 2.5 V Supplies and a $20\ \Omega$ Snubbing Resistor

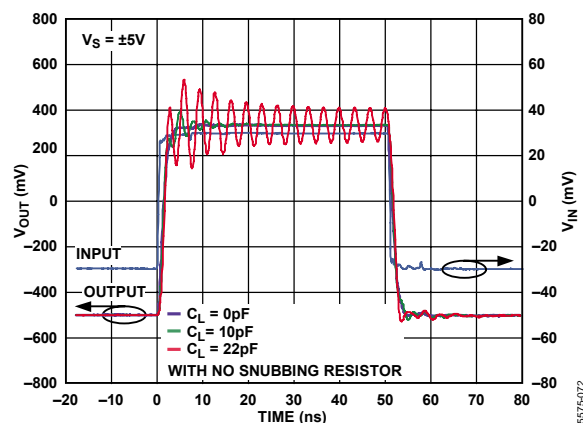


Figure 72. Large Signal Pulse Response for Two Values of Output Capacitance with ± 5 V Supplies and No Snubbing Resistor

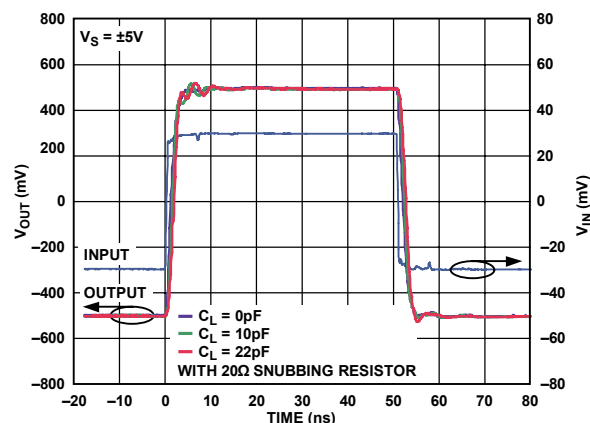


Figure 73. Pulse Response for Two Values of Output Capacitance with ± 5 V Supplies and a $20\ \Omega$ Snubbing Resistor

The effects of stray output capacitance are mitigated with a small value snubbing resistor, R_{SNUB} , placed in series with, and as near as possible to, the output pin. Figure 69, Figure 71, and Figure 73 show the improvement in dynamic performance with a $20\ \Omega$ snubbing resistor. R_{SNUB} reduces the gain slightly by the ratio of $R_{LOAD}/(R_{SNUB} + R_{LOAD})$, a very small loss when used with high impedance loads, such as ADCs. For other loads, alternate values of R_{SNUB} can be determined empirically. The data for the curves in the Typical Performance Characteristics section of this data sheet are derived using a $20\ \Omega$ snubbing resistor.

The best way to avoid the effects of stray capacitance is to exercise care in PC board layout. Locate the passive components or devices connected to the AD8337 output pins as close as possible to the package.

Although a nonissue, the preamplifier output is also sensitive to load capacitance. However, the series connection of R_{FB1} and R_{FB2} is typically the only load connected to the preamplifier. If overshoot appears, it can be mitigated in the same way as the VGA output, by inserting a snubbing resistor.

GAIN CONTROL CONSIDERATIONS

In typical applications, voltages applied to the GAIN input are dc or relatively low frequency signals. The high input impedance of the AD8337 enables several devices to be connected in parallel. This is useful for arrays of VGAs, such as those used for calibration adjustments.

Under dc or slowly changing ramp conditions, the gain tracks the gain control voltage as shown in Figure 3. However, it is often necessary to consider other effects influenced by the V_{GAIN} input.

The offset voltage effect of the AD8337, as with all VGAs, can appear as a complex waveform when observed across the range of V_{GAIN} voltage. Generated by multiple sources, each device has a unique V_{OS} profile while the GAIN input is swept through its voltage range. The offset voltage profile seen in Figure 15 is a typical example. If the V_{GAIN} input voltage is modulated, the output is the product of the V_{GAIN} and the dc profile of the offset voltage, and it can be observed on a scope as a small ac signal as shown in Figure 74. In Figure 74, the signal applied to the V_{GAIN} input is a 1 kHz ramp, and the output voltage signal is slightly less than 4 mV p-p.

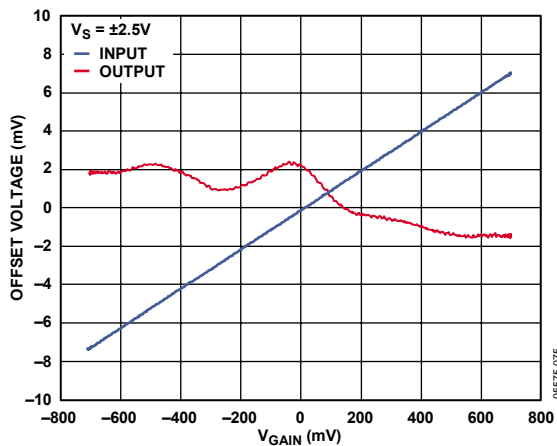


Figure 74. Offset Voltage vs. V_{GAIN} for a 1 kHz Ramp

The profile of the waveform shown in Figure 74 is consistent over a wide range of signals from dc to about 20 kHz. Above 20 kHz, secondary artifacts can be generated due to the effects of minor internal circuit tolerances, as seen in Figure 75. These artifacts are caused by settling and time constants of the interpolator circuit and appear at the output as the voltage spikes seen in Figure 75.

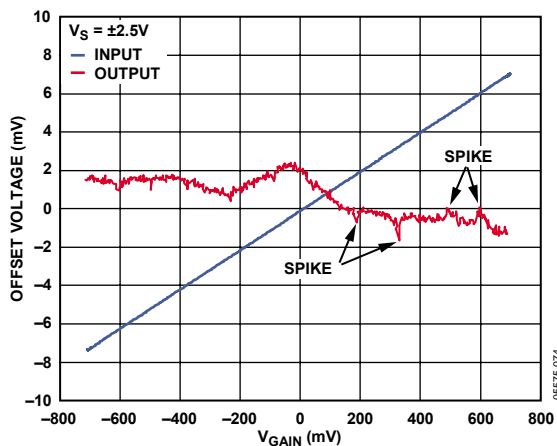


Figure 75. V_{OS} Profile for a 50 kHz Ramp

Under certain circumstances, the product of V_{GAIN} and the offset profile plus spikes is a coherent spurious signal within the signal band of interest and indistinguishable from desired signals. In general, the slower the ramp applied to the GAIN pin, the smaller the spikes are. In most applications, these effects are benign and not an issue.

THERMAL CONSIDERATIONS

The thermal performance of LFCSPs, such as the AD8337, departs significantly from that of leaded devices such as the larger TSSOP or QFSP. In larger packages, heat is conducted away from the die by the path provided by the bond wires and the device leads. In LFCSPs, the heat transfer mechanisms are surface-to-air radiation from the top and side surfaces of the package and conduction through the metal solder pad on the mounting surface of the device.

θ_{JC} is the traditional thermal metric found in the data sheets of integrated circuits. Heat transfer away from the die is a 3-dimensional dynamic, and the path is through the bond wires, leads, and the six surfaces of the package. Because of the small size of LFCSPs, the θ_{JC} is not measured conventionally; instead, it is calculated using thermodynamic rules.

The θ_{JC} value of the AD8837 listed in Table 2 assumes that the tab is soldered to the board and that there are three additional ground layers beneath the device connected by at least four vias. For a device with an unsoldered pad, the θ_{JC} nearly doubles, becoming 138°C/W.

PSI (Ψ)

Table 2 lists a subset of the classic theta specification, Ψ_{JT} (Psi junction to top). θ_{JC} is the metric of heat transfer from the die to the case, involving the six outside surfaces of the package. $\Psi_{\text{(XY)}}$ is a subset of the theta value and the thermal gradient from the junction (die) to each of the six surfaces. Ψ can be different for each of the surfaces, but since the top of the package is actually a fraction of a millimeter from the die, the surface temperature of the package is very close to the die temperature. The die temperature is calculated as the product of the power dissipation and Ψ_{JT} . Since the top surface temperature and power dissipation are easily measured, it follows that the die temperature is easily calculated. For example, for a dissipation of 180 mW and a Ψ_{JT} of 5.3°C/W, the die temperature is slightly less than 1°C higher than the surface temperature.

BOARD LAYOUT

Because the AD8337 is a high frequency device, board layout is critical. It is very important to have a good ground plane connection to the VCOM pin. Coupling through the ground plane, from the output to the input, can cause peaking at higher frequencies.

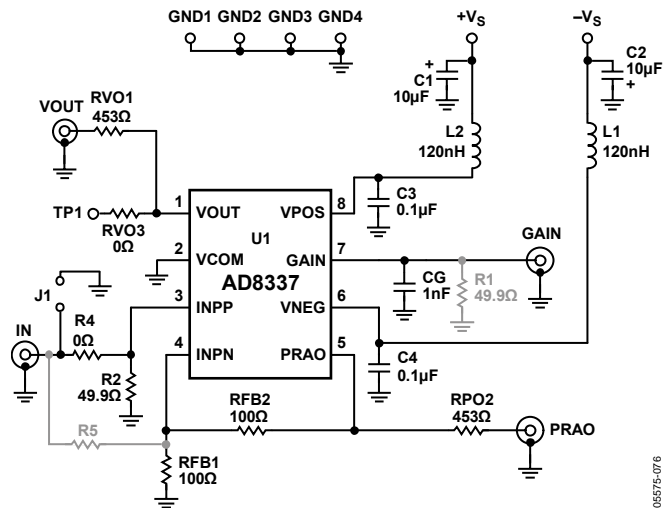


Figure 76. Evaluation Board Schematic—Noninverting Configuration

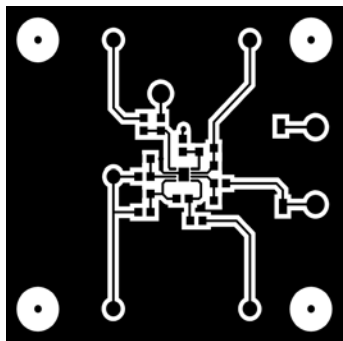


Figure 77. Evaluation Board—Component Side Copper

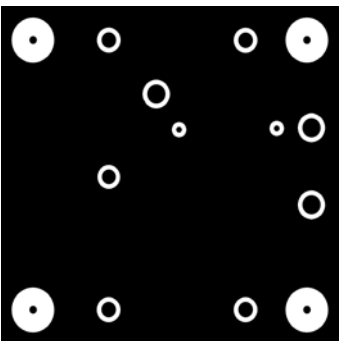


Figure 78. Evaluation Board—Wiring Side Copper

OUTLINE DIMENSIONS

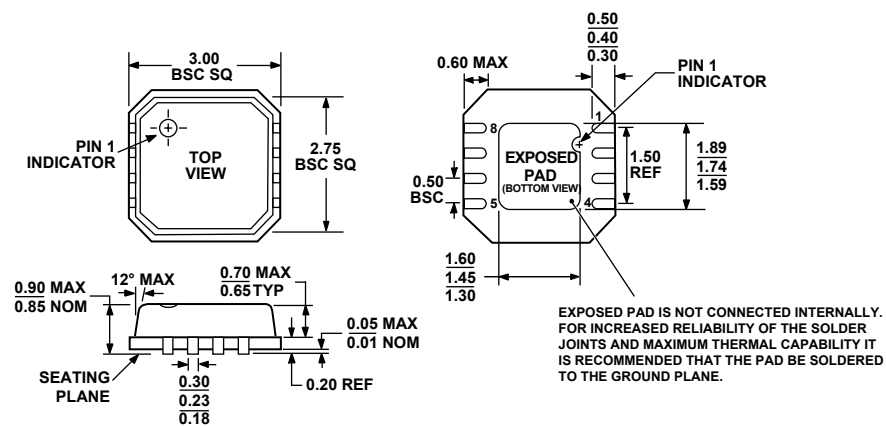


Figure 79. 8-Lead Lead Frame Chip Scale Package [LFCSP_VD]
3 mm × 3 mm Body, Very Thin, Dual Lead
(CP-8-2)
Dimensions shown in millimeters

ORDERING GUIDE

Model	Temperature Range	Package Description	Package Option	Branding
AD8337BCPZ-R2 ¹	−40°C to +85°C	8-Lead Lead Frame Chip Scale Package [LFCSP_VD]	CP-8-2	HVB
AD8337BCPZ-REEL ¹	−40°C to +85°C	8-Lead Lead Frame Chip Scale Package [LFCSP_VD]	CP-8-2	HVB
AD8337BCPZ-REEL7 ¹	−40°C to +85°C	8-Lead Lead Frame Chip Scale Package [LFCSP_VD]	CP-8-2	HVB
AD8337BCPZ-WP ¹	−40°C to +85°C	8-Lead Lead Frame Chip Scale Package [LFCSP_VD]	CP-8-2	HVB
AD8337-EVALZ ¹		Evaluation Board with Noninverting Gain Configuration		
AD8337-EVAL-INV		Evaluation Board with Inverting Gain Configuration		
AD8337-EVAL-SS		Evaluation Board with Single-Supply Operation		

¹ Z = Pb-free part.