



## FEATURES

**Multistage Demodulating Logarithmic Amplifier**  
**Voltage Output, Rise Time <15 ns**  
**High Current Capacity: 25 mA into Grounded RL**  
**95 dB Dynamic Range: -91 dBV to +4 dBV**  
**Single Supply of 2.7 V Min at 8 mA Typ**  
**DC-440 MHz Operation, 60.4 dB Linearity**  
**Slope of +24 mV/dB, Intercept of -108 dBV**  
**Highly Stable Scaling over Temperature**  
**Fully Differential DC-Coupled Signal Path**  
**100 ns Power-Up Time, 1 mA Sleep Current**

## APPLICATIONS

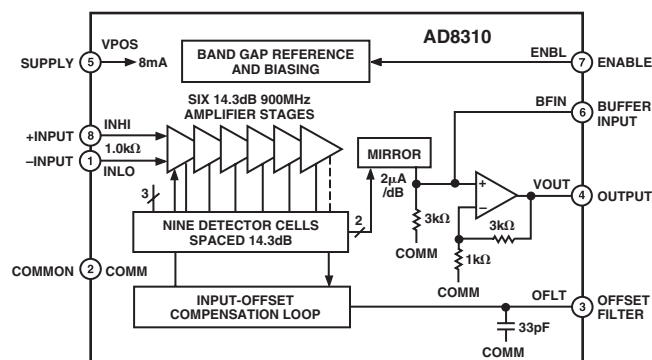
**Conversion of Signal Level to Decibel Form**  
**Transmitter Antenna Power Measurement**  
**Receiver Signal Strength Indication (RSSI)**  
**Low Cost Radar and Sonar Signal Processing**  
**Network and Spectrum Analyzers**  
**Signal-Level Determination Down to 20 Hz**  
**True-Decibel AC Mode for Multimeters**

## GENERAL DESCRIPTION

The AD8310 is a complete, dc–440 MHz demodulating logarithmic amplifier (log amp) with a very fast voltage mode output capable of driving up to 25 mA into a grounded load in under 15 ns. It uses the progressive compression (successive detection) technique to provide a dynamic range of up to 95 dB to  $\pm 3$  dB law conformance, or 90 dB to a  $\pm 1$  dB error bound up to 100 MHz. It is extremely stable and easy to use, requiring no significant external components. A single-supply voltage of 2.7 V to 5.5 V at 8 mA is needed, corresponding to a power consumption of only 24 mW at 3 V. A fast-acting CMOS compatible enable pin is provided.

Each of the six cascaded amplifier/limiter cells has a small-signal gain of 14.3 dB, with a  $-3$  dB bandwidth of 900 MHz. A total of nine detector cells are used to provide a dynamic range that extends from  $-91$  dBV (where 0 dBV is defined as the amplitude of a 1 V rms sine wave), that is, an amplitude of about  $\pm 40$   $\mu$ V, up to  $+4$  dBV (or  $\pm 2.2$  V). The demodulated output is accurately scaled, with a log slope of 24 mV/dB and an intercept of  $-108$  dBV; the scaling parameters are supply- and temperature-independent. The fully differential input offers a moderately high impedance (1 k $\Omega$  in parallel with about 1 pF). A simple network

### FUNCTIONAL BLOCK DIAGRAM



can match the input to 50  $\Omega$  and provide a power sensitivity of  $-78$  dBm to  $+17$  dBm. The logarithmic linearity is typically within  $\pm 0.4$  dB up to 100 MHz over the central portion of the range, but it is somewhat greater at 440 MHz. There is no minimum frequency limit; the AD8310 may be used down to low audio frequencies. Special filtering features are provided to support this wide range.

The output voltage runs from a noise-limited lower boundary of 400 mV to an upper limit within 200 mV of the supply voltage for light loads. The slope and intercept can be readily altered using external resistors. The output is tolerant of a wide variety of load conditions and is stable with capacitive loads of 100 pF.

The AD8310 provides a unique combination of low cost, small size, small power consumption, high accuracy and stability, high dynamic range, a frequency range encompassing audio to UHF, fast response time, and good load-driving capabilities, making this product useful in numerous applications that require the reduction of a signal to its decibel equivalent.

The AD8310 is available in the industrial temperature range of  $-40^{\circ}\text{C}$  to  $+85^{\circ}\text{C}$  in an 8-lead MSOP package.

## REV. B

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# AD8310—SPECIFICATIONS (@ T<sub>A</sub> = 25°C, V<sub>S</sub> = 5 V, unless otherwise noted.)

Parameter	Conditions	Min	Typ	Max	Unit
<b>INPUT STAGE</b>					
Maximum Input <sup>1</sup>	(Inputs INHI, INLO) Single-Ended, p-p	±2.0	±2.2		V
Equivalent Power in 50 Ω	Termination Resistor of 52.3 Ω		4		dBV
Noise Floor	Differential Drive, p-p		17		dBm
Equivalent Power in 50 Ω	Terminated 50 Ω Source		20		dBm
Input Resistance	440 MHz Bandwidth		1.28		nV/√Hz
Input Capacitance	From INHI to INLO	800	1000	1200	Ω
DC Bias Voltage	From INHI to INLO		1.4		pF
	Either Input		3.2		V
<b>LOGARITHMIC AMPLIFIER</b>					
±3 dB Error Dynamic Range	(Output VOUT) From Noise Floor to Maximum Input		95		dB
Transfer Slope	10 MHz ≤ f ≤ 200 MHz	22	24	26	mV/dB
	Over Temperature -40°C < T <sub>A</sub> < +85°C	20		26	mV/dB
Intercept (Log Offset) <sup>2</sup>	10 MHz ≤ f ≤ 200 MHz	-115	-108	-99	dBV
	Equivalent dBm (re 50 Ω)	-102	-95	-86	dBm
	Over Temperature -40°C ≤ T <sub>A</sub> ≤ +85°C	-120		-96	dBV
	Equivalent dBm (re 50 Ω)	-107		-83	dBm
	Temperature Sensitivity		-0.04		dB/°C
Linearity Error (Ripple)	Input from -88 dBV (-75 dBm) to +2 dBV (+15 dBm)		±0.4		dB
Output Voltage	Input = -91 dBV (-78 dBm)		0.4		V
	Input = 9 dBV (22 dBm)		2.6		V
Minimum Load Resistance, R <sub>L</sub>			100		Ω
Maximum Sink Current			0.5		mA
Output Resistance			0.05		Ω
Video Bandwidth			25		MHz
Rise Time (10%–90%)	Input Level = -43 dBV (-30 dBm), R <sub>L</sub> ≥ 402 Ω, C <sub>L</sub> ≤ 68 pF		15		ns
	Input Level = -3 dBV (+10 dBm), R <sub>L</sub> ≥ 402 Ω, C <sub>L</sub> ≤ 68 pF		20		ns
Fall Time (90%–10%)	Input Level = -43 dBV (-30 dBm), R <sub>L</sub> ≥ 402 Ω, C <sub>L</sub> ≤ 68 pF		30		ns
	Input Level = -3 dBV (+10 dBm), R <sub>L</sub> ≥ 402 Ω, C <sub>L</sub> ≤ 68 pF		40		ns
Output Settling Time to 1%	Input Level = -13 dBV (0 dBm), R <sub>L</sub> ≥ 402 Ω, C <sub>L</sub> ≤ 68 pF		40		ns
<b>POWER INTERFACES</b>					
Supply Voltage, V <sub>POS</sub>		2.7		5.5	V
Quiescent Current	Zero-Signal	6.5	8.0	9.5	mA
Over Temperature	-40°C < T <sub>A</sub> < +85°C	5.5	8.5	10	mA
Disable Current			0.05		μA
Logic Level to Enable Power	HI Condition, -40°C < T <sub>A</sub> < +85°C		2.3		V
Input Current when HI	3 V at ENBL		35		μA
Logic Level to Disable Power	LO Condition, -40°C < T <sub>A</sub> < +85°C		0.8		V

## NOTES

<sup>1</sup>The input level is specified in dBV because logarithmic amplifiers respond strictly to voltage, not power. 0 dBV corresponds to a sinusoidal single-frequency input of 1 V rms. A power level of 0 dBm (1 mW) in a 50 Ω termination corresponds to an input of 0.2236 V rms. Therefore, the relationship between dBV and dBm is a fixed offset of 13 dBm in the special case of a 50 Ω termination.

<sup>2</sup>Guaranteed but not tested; limits are specified at six sigma levels.

Specifications subject to change without notice.

## ABSOLUTE MAXIMUM RATINGS\*

Supply Voltage $V_S$	7.5 V
Input Power (re 50 $\Omega$ ), Single-Ended	18 dBm
Differential Drive	22 dBm
Internal Power Dissipation	200 mW
$\theta_{JA}$	200°C/W
Maximum Junction Temperature	125°C
Operating Temperature Range	−40°C to +85°C
Storage Temperature Range	−65°C to +150°C
Lead Temperature Range (Soldering 60 sec)	300°C

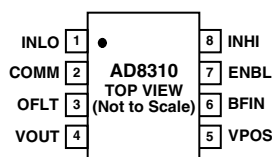
\*Stresses above those listed under Absolute Maximum Ratings may cause permanent damage to the device. This is a stress rating only; functional operation of the device at these or any other conditions above those indicated in the operational section of this specification is not implied. Exposure to absolute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

## ORDERING GUIDE

Model	Package Description	Package Option
AD8310ARM*	RM-8 Tube	RM-8
AD8310ARM-REEL	RM-8 13" Tape and Reel	RM-8
AD8310ARM-REEL7	RM-8 7" Tape and Reel	RM-8
AD8310-EVAL	Evaluation Board	

\*Device branded as J6A.

## PIN CONFIGURATION



## PIN FUNCTION DESCRIPTIONS

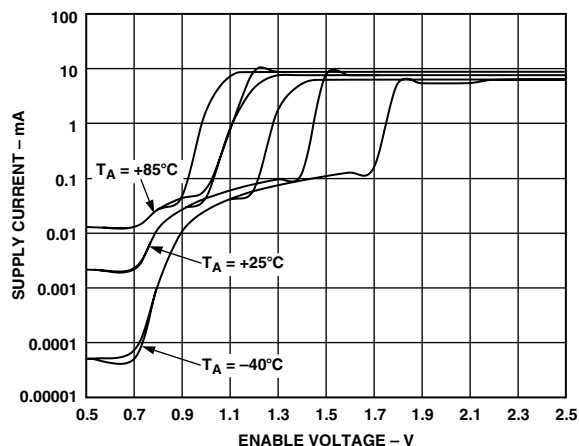
Pin No.	Mnemonic	Function
1	INLO	One of Two Balanced Inputs, Biased Roughly to VPOS/2
2	COMM	Common Pin (Usually Grounded)
3	OFLT	Offset Filter Access, Nominally at about 1.75 V
4	VOUT	Low Impedance Output Voltage, 25 mA Max Load
5	VPOS	Positive Supply, 2.7 V to 5.5 V at 8 mA Quiescent Current
6	BFIN	Buffer Input; Used to Lower Post-Detection Bandwidth
7	ENBL	CMOS Compatible Chip Enable (active when HI)
8	INHI	Second of Two Balanced Inputs

## CAUTION

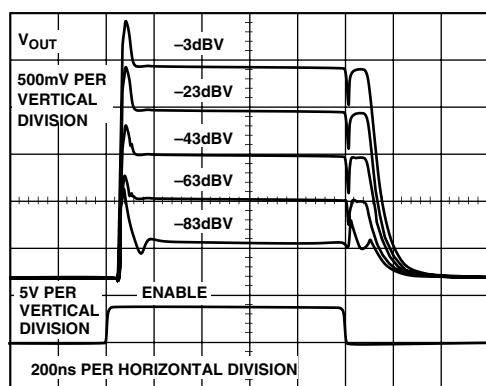
ESD (electrostatic discharge) sensitive device. Electrostatic charges as high as 4000V readily accumulate on the human body and test equipment and can discharge without detection. Although the AD8310 features proprietary ESD protection circuitry, permanent damage may occur on devices subjected to high energy electrostatic discharges. Therefore, proper ESD precautions are recommended to avoid performance degradation or loss of functionality.



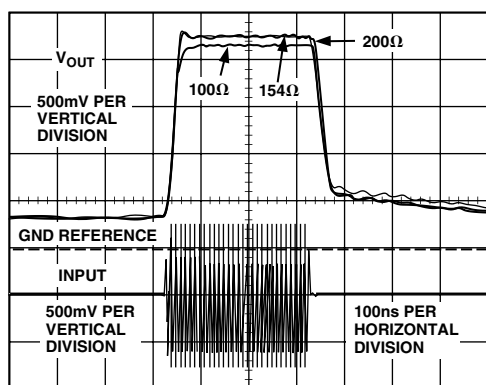
# AD8310—Typical Performance Characteristics



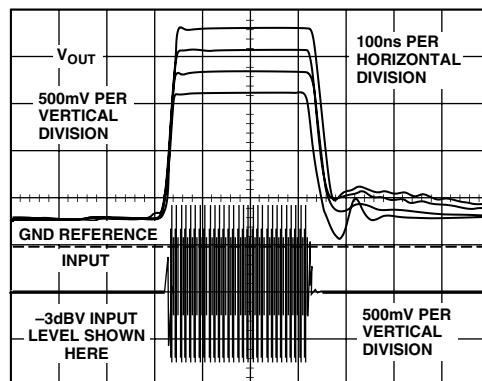
TPC 1. Supply Current vs. Enable Voltage  
@  $T_A = -40^\circ\text{C}$ ,  $+25^\circ\text{C}$ , and  $+85^\circ\text{C}$



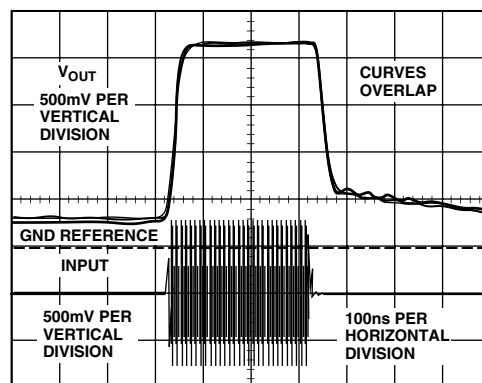
TPC 2. Power On/Off Response Time with RF Input  
of  $-83\text{ dBV}$  to  $-3\text{ dBV}$



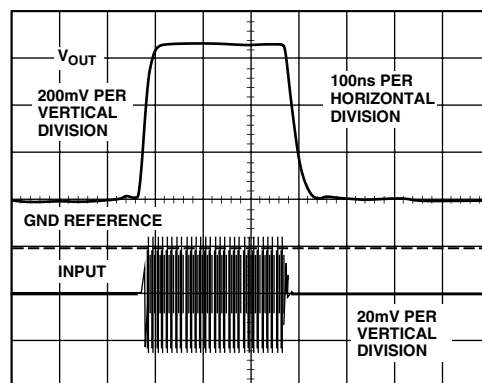
TPC 3. Large Signal RSSI Pulse Response with  
 $C_L = 100\text{ pF}$  and  $R_L = 100\ \Omega$ ,  $154\ \Omega$ , and  $200\ \Omega$



TPC 4. RSSI Pulse Response with  $R_L = 402\ \Omega$  and  
 $C_L = 68\text{ pF}$ , for Inputs Stepped from  $0\text{ dBV}$  to  $-33\text{ dBV}$ ,  
 $-23\text{ dBV}$ ,  $-13\text{ dBV}$ , and  $-3\text{ dBV}$

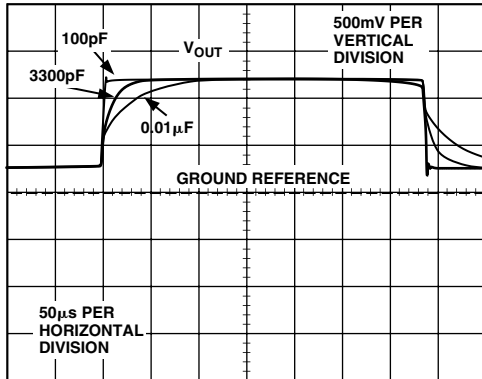


TPC 5. Large Signal RSSI Pulse Response with  
 $R_L = 100\ \Omega$  and  $C_L = 33\text{ pF}$ ,  $68\text{ pF}$ , and  $100\text{ pF}$

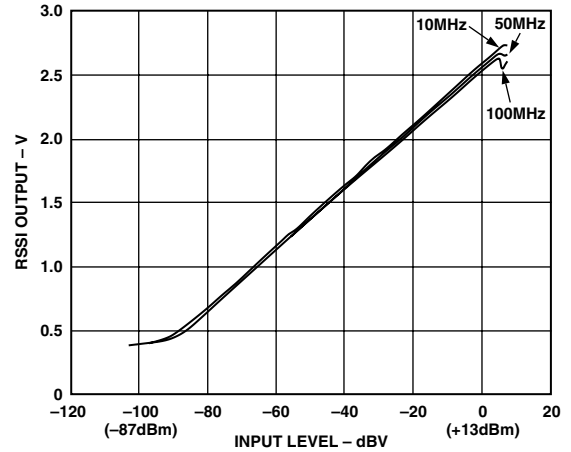


TPC 6. Small Signal RSSI Pulse Response  
with  $R_L = 50\ \Omega$  and Back Termination of  $50\ \Omega$   
(Total Load =  $100\ \Omega$ )

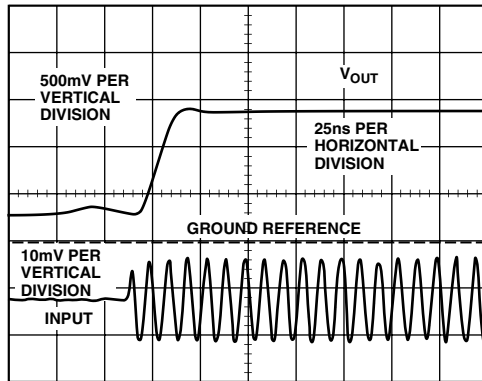
# Typical Performance Characteristics—AD8310



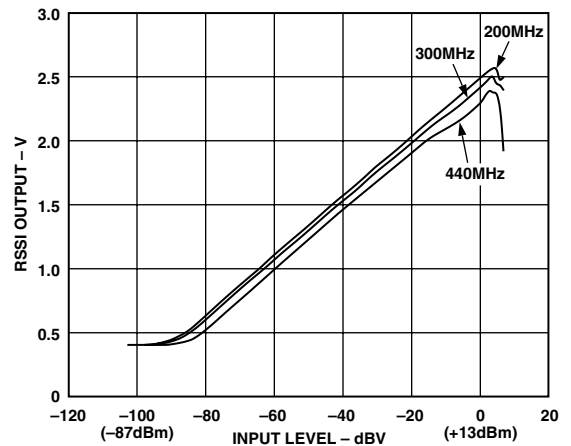
TPC 7. Small Signal AC Response of RSSI Output with External BFIN Capacitance of 100 pF, 3300 pF, and 0.01  $\mu$ F



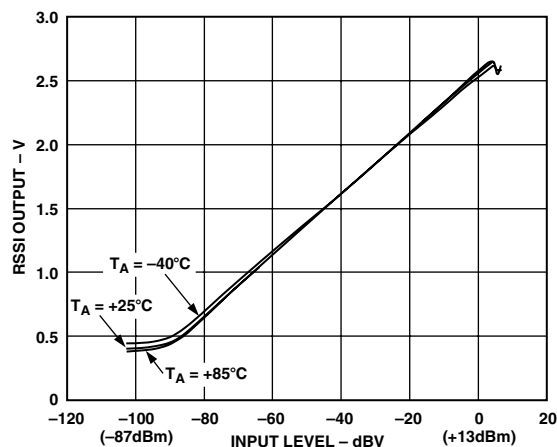
TPC 10. RSSI Output vs. Input Level at  $T_A = 25^\circ\text{C}$  for Frequencies of 10 MHz, 50 MHz, and 100 MHz



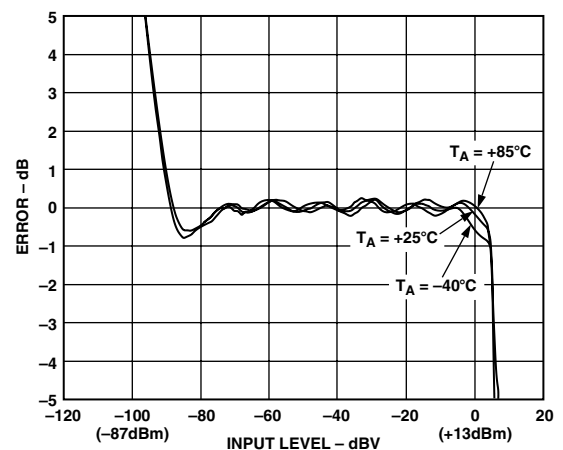
TPC 8. Small Signal RSSI Pulse Response with  $R_L = 402\ \Omega$  and  $C_L = 68\ \text{pF}$



TPC 11. RSSI Output vs. Input Level at  $T_A = 25^\circ\text{C}$  for Frequencies of 200 MHz, 300 MHz, and 440 MHz

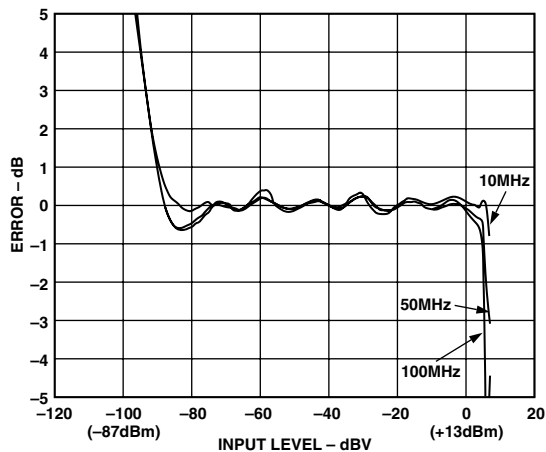


TPC 9. RSSI Output vs. Input Level, 100 MHz Sine Input at  $T_A = -40^\circ\text{C}$ ,  $+25^\circ\text{C}$ , and  $+85^\circ\text{C}$ , Single-Ended Input

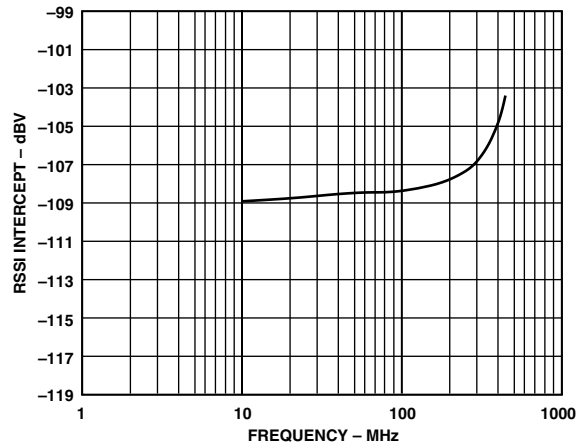


TPC 12. Log Linearity of RSSI Output vs. Input Level, 100 MHz Sine Input at  $T_A = -40^\circ\text{C}$ ,  $+25^\circ\text{C}$ , and  $+85^\circ\text{C}$

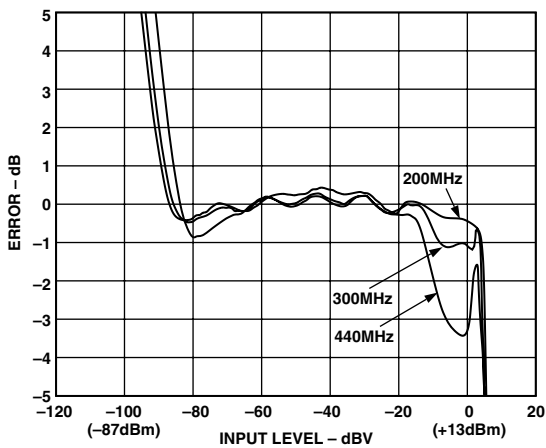
# AD8310—Typical Performance Characteristics



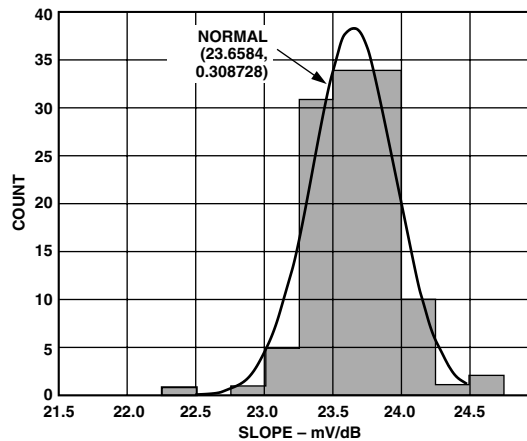
TPC 13. Log Linearity of RSSI Output vs. Input Level, at  $T_A = 25^\circ\text{C}$ , for Frequencies of 10 MHz, 50 MHz, and 100 MHz



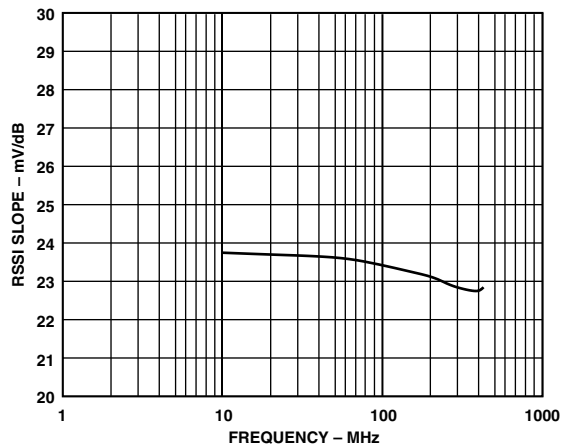
TPC 16. RSSI Intercept vs. Frequency



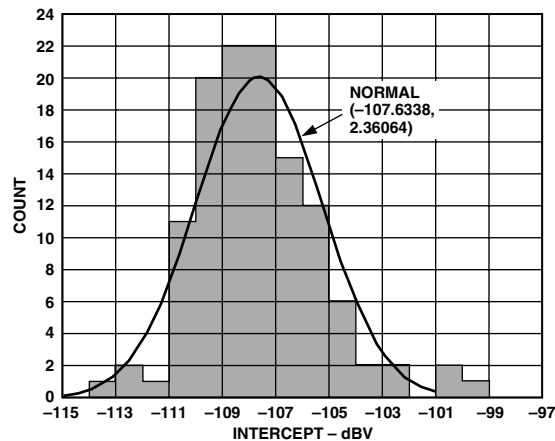
TPC 14. Log Linearity of RSSI Output vs. Input Level at  $T_A = 25^\circ\text{C}$  for Frequencies of 200 MHz, 300 MHz, and 440 MHz



TPC 17. Transfer Slope Distribution,  $V_S = 5\text{ V}$ , Frequency = 100 MHz,  $25^\circ\text{C}$



TPC 15. RSSI Slope vs. Frequency



TPC 18. Intercept Distribution  $V_S = 5\text{ V}$ , Frequency = 100 MHz,  $25^\circ\text{C}$



## GENERAL THEORY

Logarithmic amplifiers perform a more complex operation than that of classical linear amplifiers, and their circuitry is significantly different. A good grasp of what log amps do, and how they do it, will avoid many pitfalls in their application. For a complete discussion of the theory, refer to the AD8307 data sheet.

The essential purpose of a log amp is not to amplify, though amplification is needed internally, but to compress a signal of wide dynamic range to its decibel equivalent. It is thus a *measurement device*. A better term might be *logarithmic converter*, since the function is the conversion of a signal from one domain of representation to another, via a precise nonlinear transformation:

$$V_{OUT} = V_Y \log (V_{IN}/V_X) \quad (1)$$

where  $V_{OUT}$  is the output voltage,  $V_Y$  is the slope voltage, the logarithm is usually taken to base-ten (in which case  $V_Y$  is also the volts-per-decade),  $V_{IN}$  is the input voltage, and  $V_X$  is the intercept voltage. Log amps implicitly require two references, here  $V_X$  and  $V_Y$ , that determine the *scaling* of the circuit. The accuracy of a log amp cannot be any better than the accuracy of its *scaling references*. In the AD8310, these are provided by a band gap reference.

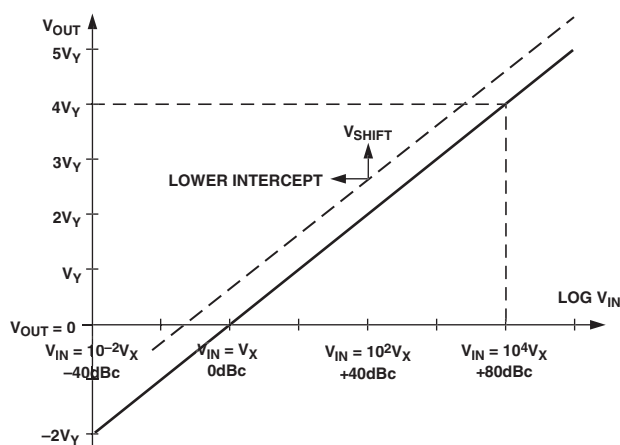


Figure 1. General Form of the Logarithmic Function

While Equation 1, plotted in Figure 1, is fundamentally correct, a different formula is appropriate for specifying the calibration attributes or demodulating log amps like the AD8310, operating in RF applications with a sine wave input:

$$V_{OUT} = V_{SLOPE} (P_{IN} - P_O) \quad (2)$$

Here,  $V_{OUT}$  is the demodulated and filtered baseband (video or RSSI) output,  $V_{SLOPE}$  is the logarithmic slope, now expressed in volts/dB (25 mV/dB for the AD8310),  $P_{IN}$  is the input power, expressed in decibels relative to some reference power level, and is  $P_O$  the logarithmic intercept, expressed in decibels relative to the same reference level. A widely used reference in RF systems is decibels above 1 mW in 50  $\Omega$ , a level of 0 dBm. Note that the quantity  $(P_{IN} - P_O)$  is just dB. The logarithmic function disappears from the formula because the conversion has already been implicitly performed in stating the input in decibels. This is strictly a concession to popular convention: log amps manifestly do *not* respond to *power* (tacitly “power absorbed at the input”), but rather to input *voltage*. The input is specified in dBV (decibels with respect to 1 V rms) throughout this data sheet. This is more precise, although still incomplete, since the signal waveform is

also involved. Since many users specify RF signals in terms of power—usually in dBm/50  $\Omega$ —we also use this convention in specifying the performance of the AD8310.

## Progressive Compression

High speed, high dynamic range log amps use a cascade of non-linear amplifier cells to generate the logarithmic function as a series of contiguous segments, a type of piecewise linear technique. The AD8310 employs six cells in its main signal path, each having a small-signal gain of 14.3 dB ( $\times 5.2$ ) and a  $-3$  dB bandwidth of about 900 MHz; the overall gain is about 20,000 (86 dB) and the overall bandwidth of the chain is some 500 MHz, resulting in a gain-bandwidth product (GBW) of 10,000 GHz, about a million times that of a typical op amp. This very high GBW is essential to accurate operation under small-signal conditions and at high frequencies. The AD8310 exhibits a logarithmic response down to inputs as small as 40  $\mu$ V at 440 MHz.

Progressive compression log amps either provide a baseband video response or they accept an RF input and demodulate this signal to develop an output that is essentially the envelope of the input represented on a logarithmic or decibel scale. The AD8310 is the latter kind. Demodulation is performed in a total of nine detector cells, six are associated with the amplifier stages and three are passive detectors that receive a progressively-attenuated fraction of the full input. The maximum signal frequency can be 440 MHz but, since all the gain stages are dc-coupled, operation at very low frequencies is possible.

## Slope and Intercept Calibration

All monolithic log amps from Analog Devices use precision design techniques to control the logarithmic slope and intercept. The primary source of this calibration is a pair of accurate voltage references, that provide supply- and temperature-independent scaling. The slope is set to 24 mV/dB by the bias chosen for the detector cells and the subsequent gain of the post-detector output interface. With this slope, the full 95 dB dynamic range can easily be accommodated within the output swing capacity when operating from a 2.7 V supply. Intercept positioning at  $-108$  dBV ( $-95$  dBm re 50  $\Omega$ ) has likewise been chosen to provide an output centered in the available voltage range.

Precise control of the slope and intercept results in a log amp having stable scaling parameters, making it a true measurement device as, for example, a calibrated Received Signal Strength Indicator (RSSI). In this application, the input waveform is invariably sinusoidal. The input level is correctly specified in dBV. It may alternatively be stated as an equivalent power, in dBm, but here we must be careful since it is essential to specify the impedance in which this power is presumed to be measured. In RF practice, it is common to assume a reference impedance of 50  $\Omega$ , in which 0 dBm (1 mW) corresponds to a sinusoidal *amplitude* of 316.2 mV (223.6 mV rms). However, the power metric is correct only when the input impedance is lowered to 50  $\Omega$ , either by a termination resistor added across INHI and INLO, or by the use of a narrow-band matching network.

It cannot be stated too strongly that log amps do not inherently respond to power, but to the *voltage* applied to their input. The AD8310 presents a nominal input impedance much higher than 50  $\Omega$  (typically 1 k $\Omega$  at low frequencies). A simple input matching network can considerably improve the *power sensitivity* of this type of log amp. This increases the voltage applied to the input and

# AD8310

therefore alters the intercept. For a  $50\ \Omega$  reactive match, the voltage gain is about 4.8, and the whole dynamic range moves down by 13.6 dB. Finally, note that the effective intercept is a function of waveform. For example, a square-wave input will read 6 dB higher than a sine wave of the same *amplitude*, and a Gaussian noise input 0.5 dB higher than a sine wave of the same *rms value*.

## Offset Control

In a monolithic log amp, direct coupling is used between the stages for several reasons. First, it avoids the need for coupling capacitors, which may typically have a chip area at least as large as that of a basic gain cell, thus considerably increasing die size. Second, the capacitor values predetermine the lowest frequency at which the log amp can operate; for moderate values, this may be as high as 30 MHz, limiting the application range. Third, the parasitic back-plate capacitance lowers the bandwidth of the cell, further limiting the scope of applications.

However, the very high dc gain of a direct-coupled amplifier raises a practical issue. An offset voltage in the early stages of the chain is indistinguishable from a real signal. If it were as high as, say,  $400\ \mu\text{V}$ , it would be 18 dB larger than the smallest ac signal ( $50\ \mu\text{V}$ ), potentially reducing the dynamic range by this amount. This problem is averted by using a global feedback path from the last stage to the first, which corrects this offset in a similar fashion to the dc negative feedback applied around an op amp. The high frequency components of the feedback signal must, of course, be removed to prevent a reduction of the HF gain in the forward path.

An on-chip filter capacitor of 33 pF provides sufficient suppression of HF feedback to allow operation above 1 MHz. (The  $-3\ \text{dB}$  point in the high-pass response is at 2 MHz, but the usable range extends well below this frequency). To further lower the frequency range, an external capacitor may be added at OFLT (Pin 3). For example, 300 pF lowers it by a factor of 10; operation at low audio frequencies requires a capacitor of about  $1\ \mu\text{F}$ . Note that this filter has no effect for input levels well above the offset voltage, where the frequency range would extend down to dc (for a signal applied directly to the input pins). The dc offset can optionally be nulled by adjusting the voltage on the OFLT pin (see the Applications section).

## PRODUCT OVERVIEW

The AD8310 comprises six main amplifier/limiter stages. These six cells, and their associated  $g_m$  styled full-wave detectors, handle the lower two-thirds of the dynamic range. Three top-end detectors, placed at 14.3 dB taps on a passive attenuator, handle the upper third of the 95 dB range. The first amplifier stage provides a low noise spectral density ( $1.28\ \text{nV}/\sqrt{\text{Hz}}$ ). Biasing for these cells is provided by two references: one determines their gain; the other is a band gap circuit that determines the logarithmic slope and stabilizes it against supply and temperature variations. The AD8310 may be enabled/disabled by a CMOS compatible level at ENBL (Pin 7).

The differential current-mode outputs of the nine detectors are summed and then converted to single-sided form, nominally scaled  $2\ \mu\text{A}/\text{dB}$ . The output voltage is developed by applying this current to a  $3\ \text{k}\Omega$  load resistor followed by a high speed gain-of-four

buffer amplifier, resulting in a logarithmic slope of  $24\ \text{mV}/\text{dB}$  (i.e.,  $480\ \text{mV}/\text{decade}$ ) at VOUT (Pin 4). The unbuffered voltage can be accessed at BFIN (Pin 6), allowing certain functional modifications, including the addition of an external post-demodulation filter capacitor and the alteration or adjustment of slope and intercept.

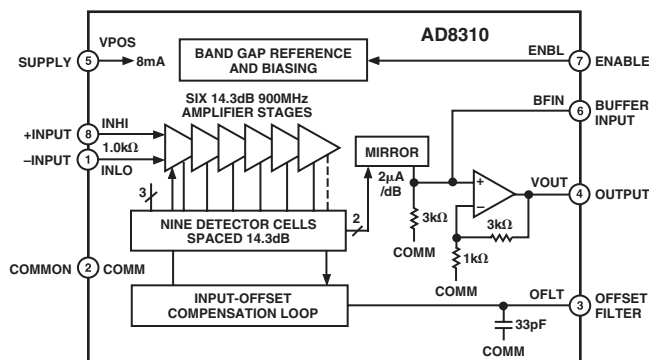


Figure 2. Main Features of AD8310

The last gain stage also includes an offset-sensing cell. This generates a bipolarity output current should the main signal path exhibit an imbalance due to accumulated dc offsets. This current is integrated by an on-chip capacitor that may be increased in value by an off-chip component, at OFLT (Pin 3). The resulting voltage is used to null the offset at the output of the first stage. Since it does not involve the signal input connections, whose ac-coupling capacitors otherwise introduce a second pole into the feedback path, the stability of the offset correction loop is assured.

The AD8310 is built on an advanced, dielectrically isolated, complementary bipolar process. In the following interface diagrams, resistors denoted with an uppercase "R" are thin-film resistors having a low temperature coefficient of resistance (TCR) and high linearity under large-signal conditions. Their absolute tolerance will typically be within  $\pm 20\%$ . Similarly, capacitors denoted using an uppercase "C," have a typical tolerance of  $\pm 15\%$  and essentially zero temperature or voltage sensitivity. Most interfaces have additional small junction capacitances associated with them, due to active devices or ESD protection; these may be neither accurate nor stable. Component numbering in each of these interface diagrams is local.

## Enable Interface

The chip-enable interface is shown in Figure 3. The currents in the diode-connected transistors control the turn-on and turn-off states of the band-gap reference and the bias generator, and are a maximum of  $100\ \mu\text{A}$  when ENBL is taken to 5 V, under worst-case conditions. For voltages below 1 V, the AD8310 will be disabled and will consume a sleep current of under  $1\ \mu\text{A}$ ; tied to the supply, or a voltage above 2 V, it will be fully enabled. The internal bias circuitry is very fast (typically  $<100\ \text{ns}$  for either OFF or ON). In practice, however, the latency period before the log amp exhibits its full dynamic range is more likely to be limited by factors relating to the use of ac-coupling at the input or the settling of the offset-control loop (see the following sections).



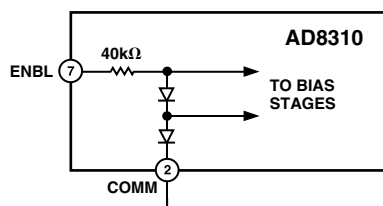


Figure 3. ENABLE Interface

### Input Interface

Figure 4 shows the essentials of the input interface.  $C_P$  and  $C_M$  are parasitic capacitances;  $C_D$  is the differential input capacitance, largely due to Q1 and Q2. In most applications, both input pins are ac-coupled. The switches S close when Enable is asserted. When disabled, bias current  $I_E$  is shut off and the inputs float; thus, the coupling capacitors remain charged. If the log amp is disabled for long periods, small leakage currents will discharge these capacitors. Then, if they are poorly matched, charging currents at power-up can generate a transient input voltage that may block the lower reaches of the dynamic range until it becomes much less than the signal.

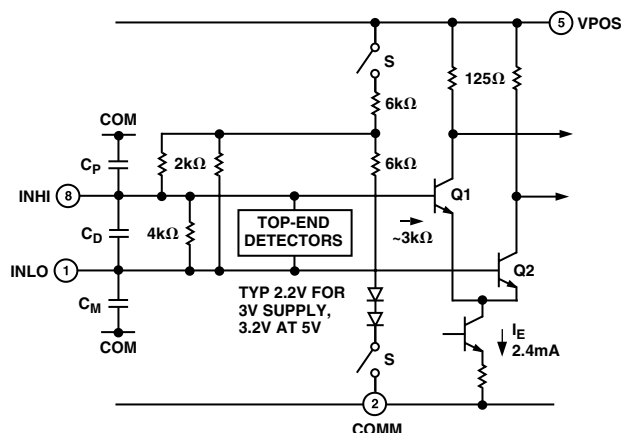


Figure 4. Signal Input Interface

A single-sided signal may be applied via a blocking capacitor to either Pin 1 or 8, with the other pin ac-coupled to ground. Under these conditions, the largest input signal that can be handled is 0 dBV (a sine amplitude of 1.4 V) when using a 3 V supply; a +5 dBV input (2.5 V amplitude) may be handled with a 5 V supply. When using a fully-balanced drive, this maximum input level is permissible for supply voltages as low as 2.7 V. Above 10 MHz, this is easily achieved using an LC matching network. Such a

network, having an inductor at the input, usefully eliminates the input transient noted above.

Occasionally, it may be desirable to use the dc-coupled potential of the AD8310 in baseband applications. The main challenge here is to present the signal at the elevated common-mode input level, which may require the use of low noise, low offset buffer amplifiers. In some cases, it may be possible to use dual supplies of  $\pm 3$  V, which allows the input pins to operate at ground potential. The output, which is internally referenced to the COMM pin (now at  $-3$  V), may be positioned back to ground level, with essentially no sensitivity to the particular value of the negative supply.

### Offset Interface

The input-referred dc offsets in the signal path are nulled via the interface associated with Pin 3, shown in Figure 5. Q1 and Q2 are the first-stage input transistors, having slightly unbalanced load resistors, resulting in a deliberate offset voltage of about 1.5 mV referred to the input pins. Q3 generates a small current to null this error, dependent on the voltage at the OFLT pin. When Q1 and Q2 are perfectly matched this voltage is about 1.75 V; in practice, it will range from approximately 1 V to 2.5 V for an input-referred offset of  $\pm 1.5$  mV.

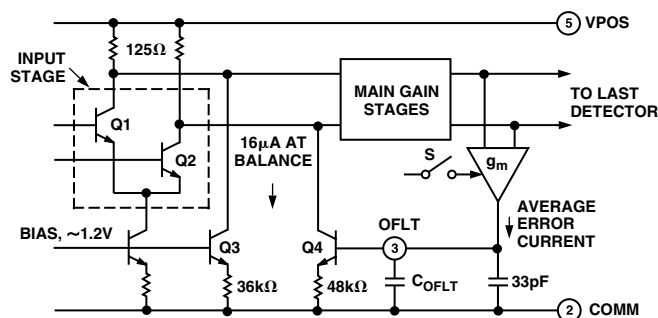


Figure 5. Offset Interface and Offset-Nulling Path

In normal operation using an ac-coupled input signal, the OFLT pin should be left unconnected. The  $g_m$  cell, which is gated off when the chip is disabled, converts a residual offset (sensed at a point near the end of the cascade of amplifiers) to a current. This is integrated by the on-chip capacitor  $C_{HP}$ , plus any added external capacitance  $C_{OFLT}$ , to generate the voltage that is applied back to the input stage in the polarity needed to null the output offset. From a small-signal perspective, this feedback alters the response of the amplifier, which exhibits a zero in its ac transfer function, resulting in a closed-loop *high-pass*  $-3$  dB corner at about 2 MHz. An external capacitor will lower the high-pass corner to arbitrarily low frequencies; using 1  $\mu$ F, the 3 dB corner is at 60 Hz.

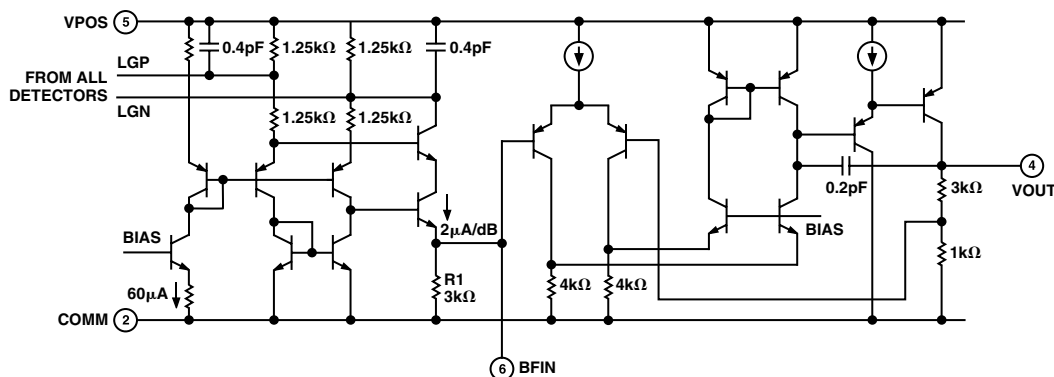


Figure 6. Simplified Output Interface

## Output Interface

The nine detectors generate differential currents, having an average value that is dependent on the signal input level, plus a fluctuation at twice the input frequency. These are summed at nodes LGP and LGN in Figure 6. Further currents are added at these nodes, to position the intercept, by slightly raising the output for zero input, and to provide temperature compensation.

For zero-signal conditions, all the detector output currents are equal. For a finite input of either polarity, their difference is converted by the output interface to a single-sided unipolar current, nominally scaled 2  $\mu\text{A}/\text{dB}$  (40  $\mu\text{A}/\text{decade}$ ), at the output pin BFIN. An on-chip resistor, R1, of  $\sim 3 \text{ k}\Omega$ , converts this current to a voltage of 6 mV/dB. This is then amplified by a factor of 4 in the output buffer, which can drive a current of up to 25 mA in a grounded load resistor. The overall rise-time of the AD8310 is under 15 ns; there is also a delay time of about 6 ns when the log amp is driven by an RF burst, starting at zero amplitude. When driving capacitive loads, it is desirable to add a low value of load resistor to speed up the return to the baseline; the buffer is stable for loads of a least 100 pF. The output bandwidth may be lowered by adding a grounded capacitor at BFIN. The time-constant of the resulting single-pole filter is formed with the 3 k $\Omega$  internal load resistor (having a tolerance of 20%); thus, to set the  $-3 \text{ dB}$  frequency to 20 kHz, use a capacitor of 2.7 nF. Using 2.7  $\mu\text{F}$ , the filter corner is at 20 Hz.

## USING THE AD8310

The AD8310 has very high gain and bandwidth. Consequently, it is susceptible to all signals that appear at the input terminals within a very broad frequency range. Without the benefit of filtering, these will be quite indistinguishable from the desired signal, and will have the effect of raising the apparent noise floor (that is, lowering the useful dynamic range). For example, while the signal of interest may be an IF of 50 MHz, any of the following could easily be larger than the IF signal at the lower extremities of its dynamic range: a few hundred microvolts of 60 Hz hum, picked up due to poor grounding techniques; spurious coupling from a digital clock source on the same PC board, local radio stations, and so on. Careful shielding and supply decoupling is therefore essential. A ground plane should be used to provide a low impedance connection to the common pin COMM, for the decoupling capacitor(s) used at VPOS and for the output ground.

## Basic Connections

Figure 7 shows the connections needed for most applications. A supply voltage between 2.7 V and 5.5 V is applied to VPOS and is decoupled using a 0.01  $\mu\text{F}$  capacitor close to the pin. Optionally, a small series resistor can be placed in the power line to give additional filtering of power supply noise. The ENBL input, which has a threshold of approximately 1.3 V (see Figure 1), should be tied to VPOS when this feature is not needed.

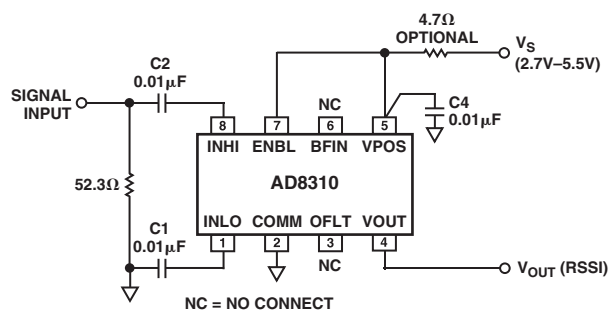
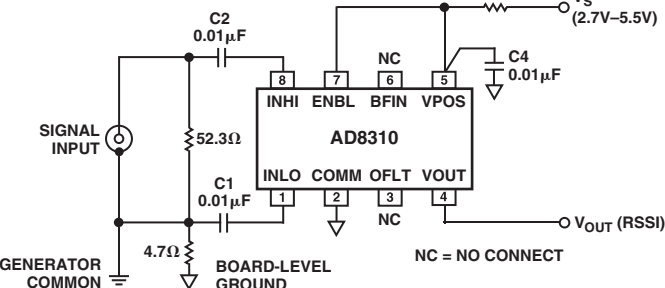


Figure 7. Basic Connections

While the AD8310's input can be driven differentially, the input signal will, in general, be single-ended. C1 is tied to ground and the input signal is coupled in through C2. Capacitors C1 and C2 should have the same value to minimize start-up transients when the enable feature is used; otherwise, their values need not be equal.

The 52.3  $\Omega$  resistor combines with the 1.1 k $\Omega$  input impedance of the AD8310 to yield a simple broadband 50  $\Omega$  input match. An input matching network can also be used (see the Input Matching section).

The coupling time constant  $50 \times C_C/2$ , forms a high-pass corner with a 3 dB attenuation at  $f_{HP} = 1/(\pi \times 50 \times C_C)$ , where  $C_1 = C_2 = C_C$ . In high frequency applications,  $f_{HP}$  should be as large as possible in order to minimize the coupling of unwanted low frequency signals. In low frequency applications, a simple RC network forming a low-pass filter should be added at the input for similar reasons. This should generally be placed at the generator side of the coupling capacitors, thus lowering the required capacitance value for a given high-pass corner frequency.



*Figure 8. Connections for Isolation of Source Ground from Device Ground*

In applications where the ground plane may not be an equipotential (possibly due to noise in the ground plane), the low input of an unbalanced source should generally be ac-coupled through a separate connection of the low associated with the source. Furthermore, it is good practice in such situations to break the ground loop by inserting a small resistance to ground in the low side of the input connector (see Figure 8).

Figure 9 shows the output versus the input level for sine inputs at 10 MHz, 50 MHz, and 100 MHz; Figure 10 shows the logarithmic conformance under the same conditions.

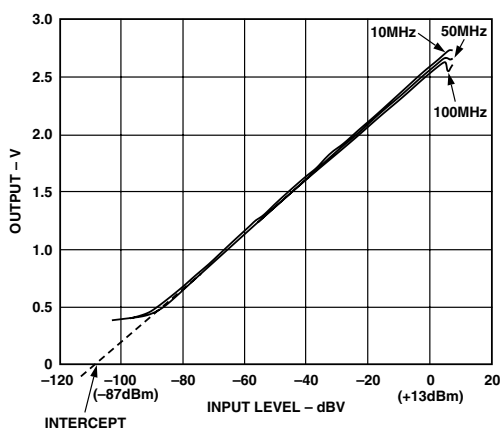


Figure 9. Output vs. Input Level at 10 MHz, 50 MHz, and 100 MH

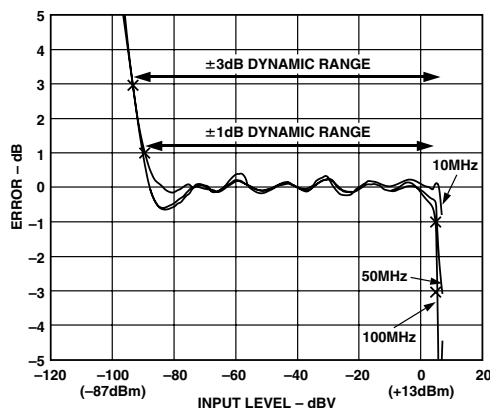


Figure 10. Log Conformance Errors vs. Input Level  
at 10 MHz, 50 MHz, and 100 MHz

### Transfer Function in Terms of Slope and Intercept

The transfer function of the AD8310 is characterized in terms of its slope and intercept. The logarithmic slope is defined as the change in the RSSI output voltage for a 1 dB change at the input. For the AD8310, slope is nominally 24 mV/dB. Therefore, a 10 dB change at the input results in a change at the output of approximately 240 mV. The plot of Log Conformance shows the range over which the device maintains its constant slope. The dynamic range of the log amp is defined as the range over which the slope remains within a certain error band, usually  $\pm 1$  dB or  $\pm 3$  dB. In Figure 10, for example, the  $\pm 1$  dB dynamic range is approximately 95 dB (from +4 dBV to -91 dBV).

The intercept is the point at which the extrapolated linear response would intersect the horizontal axis (see Figure 9). For the AD8310, the intercept is calibrated to be  $-108\text{ dBV}$  ( $-95\text{ dBm}$ ). Using the slope and intercept, the output voltage can be calculated for any input level within the specified input range using the equation:

$$V_{OUT} = V_{SLOPE} \times (P_{IN} - P_O)$$

where  $V_{OUT}$  is the demodulated and filtered RSSI output,  $V_{SLOPE}$  is the logarithmic slope expressed in V/dB,  $P_{IN}$  is the input signal expressed in decibels relative to some reference level (either dBm or dBV in this case), and  $P_O$  is the logarithmic intercept expressed in decibels relative to the same reference level.

For example, for an input level of  $-33$  dBV ( $-20$  dBm), the output voltage will be

$$V_{OUT} = 0.024 \text{ V/dB} \times (-33 \text{ dBV} - (-108 \text{ dBV})) = 1.8 \text{ V}$$

## dBV vs. dBm

The most widely used convention in RF systems is to specify power in dBm, that is, decibels above 1 mW in 50  $\Omega$ . Specification of log amp input level in terms of power is strictly a concession to popular convention; they do *not* respond to power (tacitly “power absorbed at the input”), but to the input voltage. The use of dBV, defined as *decibels with respect to a 1 V rms sine wave*, is more precise, although this is still not unambiguous because waveform is also involved in the response of a log amp, which, for a complex input (such as a CDMA signal) will not follow the rms value exactly. Since most users specify RF signals in terms of power—more specifically, in dBm/50  $\Omega$ —we use both dBV and dBm in specifying the performance of the AD8310, showing equivalent dBm levels for the special case of a 50  $\Omega$  environment. Values in dBV are converted to dBm re 50  $\Omega$  by adding 13 dB.

### Effect of Waveform Type on Intercept

Input signals of equal rms power, but differing crest factors, will produce different results at the log amp's output.

Differing signal waveforms shift the effective value of the intercept. Graphically, this looks like a vertical shift in the log amp's transfer function. The logarithmic slope, however, is not affected. For example, consider the case of the AD8310 being alternately fed by an unmodulated sine wave and by a single CDMA channel of the same rms power. The output voltage will differ by the equivalent of 3.55 dB (71 mV) over the complete dynamic range of the device (the output for the CDMA input being lower).

# AD8310

Table I shows the correction factors that should be applied to measure the rms signal strength of various signal types. A *sine wave* input is used as a reference. To measure the rms power of a square wave, for example, the mV equivalent of the dB value given in the table ( $24 \text{ mV/dB} \times 3.01 \text{ dB}$ ) should be subtracted from the output voltage of the AD8310.

**Table I. Correction for Signals with Differing Crest Factors**

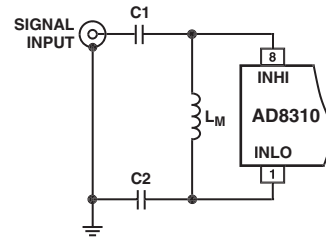
Signal Type	Correction Factor (Add to Measured Input Level) (dB)
Sine Wave	0
Square Wave or DC	-3.01
Triangular Wave	0.9
GSM Channel (All Time Slots On)	0.55
CDMA Channel (Forward Link, Nine Channels On)	3.55
CDMA Channel (Reverse Link)	0.5
PDC Channel (All Time Slots On)	0.58

## Input Matching

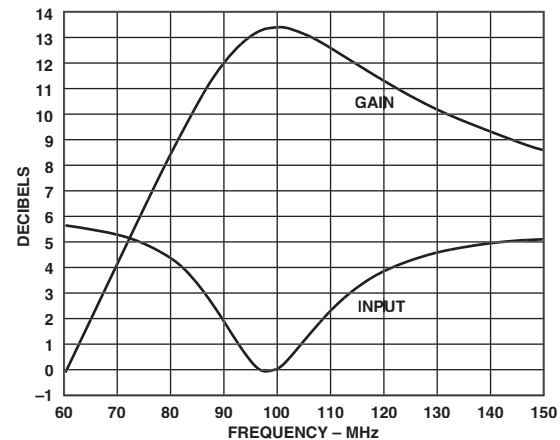
Where higher sensitivity is required, an input matching network is useful. Using a transformer to achieve the impedance transformation also eliminates the need for coupling capacitors, lowers the offset voltage generated directly at the input, and balances the drive amplitude to INLO and INHI. The choice of turns ratio will depend somewhat on the frequency. At frequencies below 50 MHz, the reactance of the input capacitance is much higher than the real part of the input impedance. In this frequency range, a turns ratio of about 1:4.8 will lower the input impedance to  $50 \Omega$  while raising the input voltage, and thus lowering the effect of the short circuit noise voltage by the same factor. The intercept will also be lowered by the turns ratio; for a  $50 \Omega$  match, it will be reduced by  $20 \log_{10}(4.8)$  or 13.6 dB. The total noise will be reduced by a somewhat smaller factor because there will be a small contribution from the input noise current.

## Narrow-Band Matching

Transformer coupling is useful in broadband applications. However, a magnetically coupled transformer may not be convenient in some situations. At high frequencies, it is often preferable to use a narrow-band matching network, as shown in Figure 11. This has several advantages. The same voltage gain is achieved, providing increased sensitivity, but now a measure of selectivity is also introduced. The component count is low: two capacitors and an inexpensive chip inductor. Further, by making these capacitors unequal, the amplitudes at INP and INM may be equalized when driving from a single-sided source; that is, the network also serves as a balun. Figure 12 shows the response for a center frequency of 100 MHz; note the very high attenuation at low frequencies. The high frequency attenuation is due to the input capacitance of the log amp.



**Figure 11. Reactive Matching Network**



**Figure 12. Response of 100 MHz Matching Network**

**Table II. Narrow-Band Matching Values**

$F_C$ (MHz)	$Z_{IN}$ ( $\Omega$ )	$C1$ (pF)	$C2$ (pF)	$L_M$ (nH)	Voltage Gain (dB)
10	45	160	150	3300	13.3
20	44	82	75	1600	13.4
50	46	30	27	680	13.4
100	50	15	13	270	13.4
150	57	10	8.2	220	13.2
200	57	7.5	6.8	150	12.8
250	50	6.2	5.6	100	12.3
500	54	3.9	3.3	39	10.9
10	103	100	91	5600	10.4
20	102	51	43	2700	10.4
50	99	22	18	1000	10.6
100	98	11	9.1	430	10.5
150	101	7.5	6.2	260	10.3
200	95	5.6	4.7	180	10.3
250	92	4.3	3.9	130	9.9
500	114	2.2	2.0	47	6.8



## General Matching Procedure

For other center frequencies and source impedances, the following method can be used to calculate the basic matching parameters.

### Step 1: Tune Out $C_{IN}$

At a center frequency  $f_C$ , the shunt impedance of the input capacitance  $C_{IN}$  can be made to disappear by resonating with a temporary inductor  $L_{IN}$ , whose value is given by

$$L_{IN} = \frac{1}{\omega^2 C_{IN}}$$

when  $C_{IN} = 1.4$  pF. For example, at  $f_C = 100$  MHz,  $L_{IN} = 1.8$   $\mu$ H.

### Step 2: Calculate $C_O$ and $L_O$

Now, having a purely resistive input impedance, we can calculate the nominal coupling elements  $C_O$  and  $L_O$ , using

$$C_O = \frac{1}{2\pi f_C \sqrt{R_{IN} R_M}}; L_O = \frac{\sqrt{(R_{IN} R_M)}}{2\pi f_C}$$

For the AD8310,  $R_{IN}$  is 1 k $\Omega$ . Thus, if a match to 50  $\Omega$  is needed, at  $f_C = 100$  MHz,  $C_O$  must be 7.12 pF and  $L_O$  must be 356 nH.

### Step 3: Split $C_O$ Into Two Parts

Since we want to provide the fully balanced form of network shown in Figure 11, two capacitors  $C_1 = C_2$ , each of nominally twice  $C_O$  and shown as  $C_M$  in the figure, can be used. This requires a value of 14.24 pF in this example. Under these conditions, the voltage amplitudes at INHI and INLO will be similar. A somewhat better balance in the two drives may be achieved when  $C_1$  is made slightly larger than  $C_2$ , which also allows a wider range of choices in selecting from standard values. For example, capacitors of  $C_1 = 15$  pF and  $C_2 = 13$  pF may be used (making  $C_O = 6.96$  pF).

### Step 4: Calculate $L_M$

The matching inductor required to provide both  $L_{IN}$  and  $L_O$  is just the parallel combination of these:

$$L_M = L_{IN} L_O / (L_{IN} + L_O)$$

With  $L_{IN} = 1.8$   $\mu$ H and  $L_O = 356$  nH, the value of  $L_M$  to complete this example of a match of 50  $\Omega$  at 100 MHz is 297.2 nH. The nearest standard value of 270 nH may be used with only a slight loss of matching accuracy. The voltage gain at resonance depends only on the ratio of impedances, as given by

$$GAIN = 20 \log \left( \sqrt{\frac{R_{IN}}{R_S}} \right) = 10 \log \left( \frac{R_{IN}}{R_S} \right)$$

## Slope and Intercept Adjustments

Where system (i.e., software) calibration is not available, the adjustments shown in Figure 13 can be used, either singly or in combination, to trim the absolute accuracy of the AD8310. The log slope may be raised or lowered by VR1; the values shown provide a calibration range of  $\pm 10\%$  (22.6 mV/dB to 27.4 mV/dB), which includes full allowance for the variability in the value of the internal resistances. The adjustment may be made by alternately applying two fixed input levels, provided by an accurate signal

generator, spaced over the central portion of the dynamic range, for example,  $-60$  dBV and  $-20$  dBV.

Alternatively, an AM-modulated signal at about the center of the dynamic range may be used. For a modulation depth  $M$ , expressed as a fraction, the decibel range between the peaks and troughs over one cycle of the modulation period is given by

$$\Delta dB = 20 \log_{10} \frac{1+M}{1-M} \quad (3)$$

For example., using a generator output of  $-40$  dBm with a 70% modulation depth ( $M = 0.7$ ), the decibel range is 15 dB because the signal varies from  $-47.5$  dBm to  $-32.5$  dBm.

The log intercept is adjustable by VR2 over a  $-3$  dB range with the component values shown. VR2 is adjusted while applying an accurately known CW signal, preferably near the lower end of the dynamic range, in order to minimize the effect of any residual uncertainty in the slope. For example, to position the intercept to  $-80$  dBm, a test level of  $-65$  dBm may be applied and VR2 adjusted to produce a dc output of 15 dB above zero at 24 mV/dB, which is 360 mV.

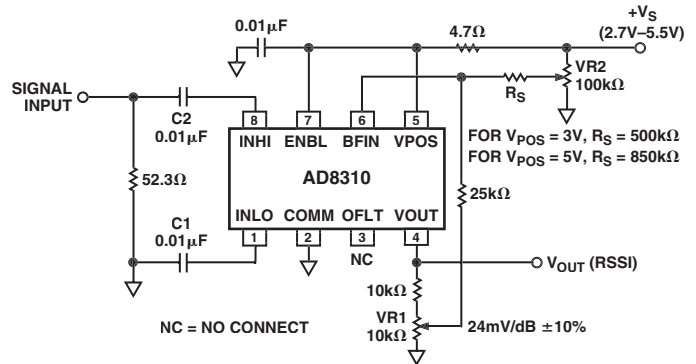


Figure 13. Slope and Intercept Adjustments

## Increasing the Slope to a Fixed Value

It is also possible to increase the slope to a new fixed value and thus increase the change in output for each decibel of input change. A common example of this is the need to map the output swing of the AD8310 into the input range of an analog-to-digital converter (ADC) with a rail-to-rail input swing. Alternatively, a situation might arise when only a part of the total dynamic range is required—say, just 20 dB—in an application where the nominal input level is more tightly constrained and a higher sensitivity to a change in this level is required. Of course, the maximum output will be limited either by the load resistance and the maximum output current rating of 25 mA, or by the supply voltage (see the Specifications table). The slope may easily be raised by adding a resistor from VOUT to BFIN, as shown in Figure 14. This alters the gain of the output buffer, by means of stable positive feedback, from its normal value of 4 to an effective value that may be as high as 16, corresponding to a slope of 100 mV/dB. The resistor  $R_{SLOPE}$  is set according to the equation

$$R_{SLOPE} = \frac{9.22 k\Omega}{1 - \frac{24 mV / dB}{Slope}}$$

# AD8310

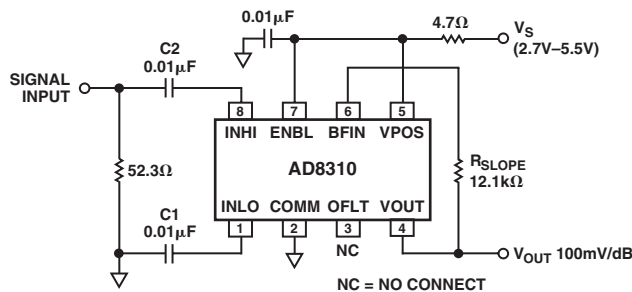


Figure 14. Raising the Slope to 100 mV/dB

## Output Filtering

In applications where maximum video bandwidth (and consequently fast rise time) is desired, it is essential that the BFIN pin be left unconnected and free of any stray capacitance.

The nominal output video bandwidth of 25 MHz can be reduced by connecting a ground-referenced capacitor ( $C_{FILT}$ ) to the BFIN pin, as shown in Figure 15. This is generally done to reduce output ripple (at twice the input frequency for a symmetric input waveform such as sinusoidal signals).

$C_{FILT}$  is selected using the equation

$$C_{FILT} = 1/(2\pi \times 3 \text{ k}\Omega \times \text{Video Bandwidth}) - 2.1 \text{ pF}$$

The *Video Bandwidth* should typically be set at a frequency equal to about one-tenth the minimum input frequency. This will ensure that the output ripple of the demodulated log output, which is at twice the input frequency, will be well filtered.

In many applications of log amps, it may be necessary to lower the corner frequency of the post-demodulation filtering in order to achieve low output ripple while maintaining a rapid response time to changes in signal level. An example of a four-pole active filter is shown in the AD8307 data sheet.

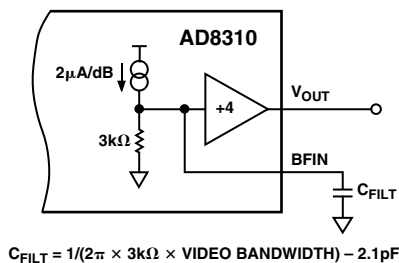


Figure 15. Lowering the Post-Demodulation Video Bandwidth

## Lowering the High-Pass Corner Frequency of the Offset Compensation Loop

In normal operation, using an ac-coupled input signal, the OFLT pin should be left unconnected. Input-referred dc offsets of about 1.5 mV in the signal path are nulled via an internal offset control loop. This loop has a *high-pass* -3 dB corner at about 2 MHz. In low frequency ac-coupled applications, it is necessary to lower this corner frequency to prevent input signals from being misinterpreted as offsets. An external capacitor on OFLT will lower the high-pass corner to arbitrarily low frequencies (Figure 16). For example, by using 1 μF capacitor, the 3 dB corner will be reduced to 60 Hz.

The corner frequency is set by the equation

$$F_{CORNER} = 1/(2\pi \times 2625 \times C_{OFLT})$$

where  $C_{OFLT}$  is the capacitor connected to OFLT.

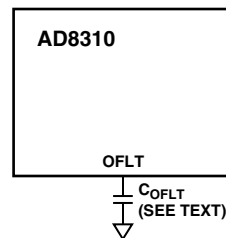


Figure 16. Lowering the High-Pass Corner Frequency of the Offset Control Loop

## APPLICATIONS

The AD8310 is highly versatile and easy to use. Being complete, it needs only a few external components, most of which can be immediately accommodated by using the simple connections shown in the preceding section. A few examples of more specialized applications are provided here; see also the AD8307 data sheet for further applications; note the slightly different pinout.

### Cable-Driving

The AD8310 is capable of driving a grounded 100 Ω load to 2.5 V, for a supply voltage of 3 V or greater. If reverse-termination is required when driving a 50 Ω cable, it should be included in series with the output, as shown in Figure 17. The slope at the load will then be 12 mV/dB. In some cases, it may be permissible to operate the cable without a termination at the far end, in which case the slope will not be lowered. Where a further increase in slope is desirable, the scheme shown in Figure 14 may be used.

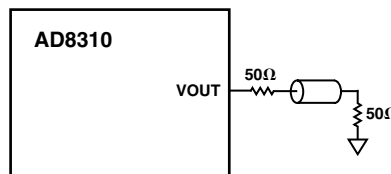


Figure 17. Output Response of Cable-Driver Application

### DC-Coupled Input

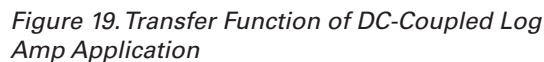
It may occasionally be necessary to provide response to dc inputs. Since the AD8310 is internally dc-coupled, there is no fundamental reason why this is precluded. However, there is a practical constraint, which is that its differential inputs must be positioned at least 2 V above the COM potential for proper biasing of the first stage. Usually, the source will be a single-sided ground-referenced signal, so it will thus be necessary to provide level shifting and a single-ended-to-differential conversion to correctly drive the AD8310's inputs.

Figure 18 shows how a level-shift to midsupply (2.5 V in this example) and a single-ended-to-differential conversion can be accomplished using the AD8138 differential amplifier. The four 499 Ω resistors set up a gain of unity. An output common-mode (or bias) voltage of 2.5 is achieved by applying 2.5 V (from a supply-referenced resistive divider) to the AD8138's VOCM pin. The differential outputs of the AD8138 directly drive the 1.1 kΩ input impedance of the AD8310.





After trimming, the lower end of the dynamic range is limited by the broadband noise at the output of the AD8138, which is approximately 425  $\mu\text{V}$  p-p. A differential low-pass filter may be added between the AD8138 and the AD8310 when the very fast pulse response of the circuit is not required.



An evaluation board, carefully laid out and tested to demonstrate the specified high speed performance of the AD8310, is available. Figure 20 shows the schematic of the evaluation board, which fairly closely follows the basic connections schematic shown in Figure 7. Connectors INHI, INLO, and VOUT are SMA type. Supply and ground are connected to vector pins TP1 and TP2; switches and component settings for different setups are described in Table III. The layout and silkscreen for the component side of the board are shown in Figure 21 and Figure 22. For ordering information, refer to the Ordering Guide.



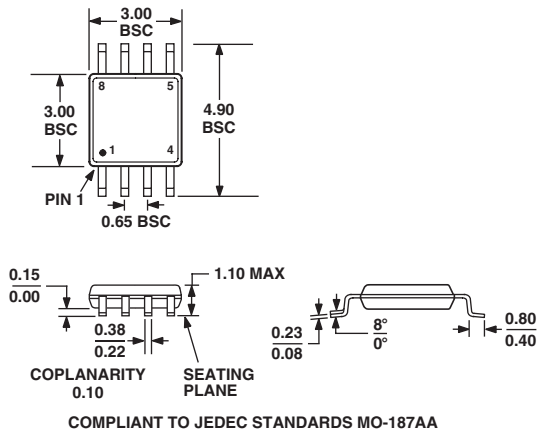
Table III. Evaluation Boards Setup Options

Component	Function	Default Condition
TP1, TP2 SW1	Supply and Ground Vector Pins Device Enable. When in Position A, the ENBL pin is connected to +V <sub>S</sub> and the AD8310 is in normal operating mode. In Position B, the ENBL pin is connected to ground, putting the device into sleep mode.	Not Applicable SW1 = A
R1/R4	SMA Connector Grounds. Connects common of INHI and INLO SMA connectors to ground. Can be used to isolate the generator ground from the evaluation board ground (see Figure 26).	R1 = R4 = 0 Ω
C1, C2, R3	Input Interface. R3 (52.3 Ω) combines with the AD8310's 1 kΩ input impedance to give an overall broadband input impedance of 50 Ω. C1, C2, and the AD8310's input impedance combine to set a high-pass input corner of 32 kHz. Alternatively, R3, C1, and C2 can be replaced by an inductor and matching capacitors to form an input matching network. See Input Matching section for more detail.	R3 = 52.3 Ω C1 = C2 = 0.01 μF
C3	RSSI (Video) Bandwidth Adjust. The addition of C3 (Farads) will lower the RSSI bandwidth of the VLOG output according to the equation: $C_{FILT} = 1/(2\pi \times 3\text{ k}\Omega \times \text{Video Bandwidth}) - 2.1\text{ pF}$ .	C3 = Open
C4, C5, R5	Supply Decoupling. The nominal supply decoupling of 0.01 μF (C4) can be augmented by a larger cap in C5. An inductor or small resistor can be placed in R5 for additional decoupling.	C4 = 0.01 μF C5 = Open, R5 = 0 Ω
R6	Output Source Impedance. In cable-driving applications, a resistor (typically 50 Ω or 75 Ω) can be placed in R6 to give the circuit a back-terminated output impedance.	R6 = 0 Ω
W1, W2, C6, R7	Output Loading. Resistors and capacitors can be placed in C6 and R7 to load test V <sub>OUT</sub> . Jumpers W1 and W2 are used to connect/disconnect the loads.	C6 = R7 = Open W1 = W2 = Installed
C7	Offset Compensation Loop. A capacitor in C7 will reduce the corner frequency of the offset control loop in low frequency applications.	C7 = Open

OUTLINE DIMENSIONS

8-Lead MSOP Package  
(RM-8)

Dimensions shown in millimeters



REVISION HISTORY

Location	Page
2/03—Data Sheet changed from REV. A to REV. B.	
Change to Evaluation Board section	15
Change to Table III	16
Updated OUTLINE DIMENSIONS	16