Wideband, Unity-Gain Stable, JFET-Input OPERATIONAL AMPLIFIER

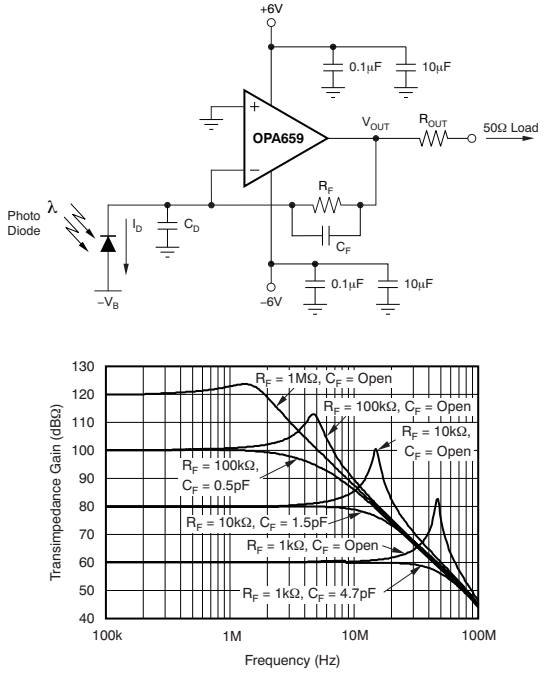
FEATURES

- HIGH BANDWIDTH: 650MHz (G = +1V/V)
- HIGH SLEW RATE: 2550V/μs (4V Step)
- EXCELLENT THD: –78dBc at 10MHz
- LOW INPUT VOLTAGE NOISE: 8.9nV/√Hz
- FAST OVERDRIVE RECOVERY: 8ns
- FAST SETTLING TIME (1% 4V Step): 8ns
- LOW INPUT OFFSET VOLTAGE: ±1mV
- LOW INPUT BIAS CURRENT: ±10pA
- HIGH OUTPUT CURRENT: 70mA

APPLICATIONS

- HIGH-IMPEDANCE DATA ACQUISITION INPUT AMPLIFIER
- HIGH-IMPEDANCE OSCILLOSCOPE INPUT AMPLIFIER
- WIDEBAND PHOTODIODE TRANSIMPEDANCE AMPLIFIER
- WAFER SCANNING EQUIPMENT

TRANSIMPEDANCE GAIN
vs FREQUENCY (C_D = 22pF)



DESCRIPTION

The OPA659 combines a very wideband, unity-gain stable, voltage-feedback operational amplifier with a JFET-input stage to offer an ultra-high dynamic range amplifier for high impedance buffering in data acquisition applications such as oscilloscope front-end amplifiers and machine vision applications such as photodiode transimpedance amplifiers used in wafer inspection.

The wide 650MHz unity-gain bandwidth is complemented by a very high 2550V/μs slew rate.

The high input impedance and low bias current provided by the JFET input are supported by the low 8.9nV/√Hz input voltage noise to achieve a very low integrated noise in wideband photodiode transimpedance applications.

Broad transimpedance bandwidths are possible with the high 350MHz gain bandwidth product of this device.

Where lower speed with lower quiescent current is required, consider the OPA656. Where unity-gain stability is not required, consider the OPA657.

RELATED OPERATIONAL AMPLIFIER PRODUCTS

DEVICE	V_S (V)	BW (MHz)	SLEW RATE (V/μs)	VOLTAGE NOISE (nV/√Hz)	AMPLIFIER DESCRIPTION
OPA356	+5	200	300	5.80	Unity-Gain Stable CMOS
OPA653	±6	500	2675	6.1	Fixed Gain of +2V/V JFET-Input
OPA656	±5	500	290	7	Unity-Gain Stable JFET-Input
OPA657	±5	1600	700	4.8	Gain of +7 Stable JFET-Input
OPA627	±15	16	55	4.5	Unity-Gain Stable DI-FET-Input
THS4631	±15	105	900	7	Unity-Gain Stable JFET-Input

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PRODUCTION DATA information is current as of publication date. Products conform to specifications per the terms of the Texas Instruments standard warranty. Production processing does not necessarily include testing of all parameters.

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This integrated circuit can be damaged by ESD. Texas Instruments recommends that all integrated circuits be handled with appropriate precautions. Failure to observe proper handling and installation procedures can cause damage.

ESD damage can range from subtle performance degradation to complete device failure. Precision integrated circuits may be more susceptible to damage because very small parametric changes could cause the device not to meet its published specifications.

ORDERING INFORMATION(1)

PRODUCT	PACKAGE-LEAD	PACKAGE DESIGNATOR	SPECIFIED TEMPERATURE RANGE	PACKAGE MARKING	ORDERING NUMBER	TRANSPORT MEDIA, QUANTITY
OPA659	SOT23-5	DBV	–40°C to +85°C	BZX	OPA659IDBVT	Tape and reel, 250
					OPA659IDBVR	Tape and reel, 3000
OPA659	VSON-8	DRB	–40°C to +85°C	OBFI	OPA659IDRBT	Tape and reel, 250
					OPA659IDRBR	Tape and reel, 3000

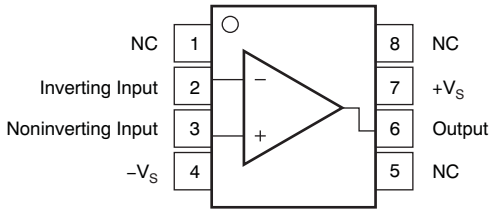
(1) For the most current package and ordering information see the Package Option Addendum at the end of this document, or see the TI web site at www.ti.com.

ABSOLUTE MAXIMUM RATINGS

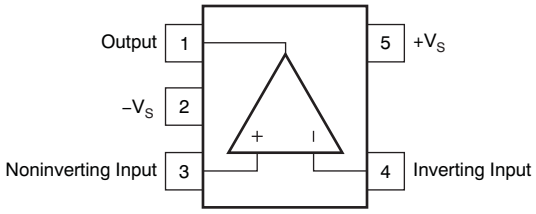
Over operating free-air temperature range (unless otherwise noted).

		OPA653	UNIT
Power Supply Voltage V_{S+} to $V_{S–}$		±6.5	V
Input Voltage		$\pm V_S$	V
Input Current		100	mA
Output Current		100	mA
Continuous Power Dissipation		See Thermal Characteristics	
Operating Free Air Temperature Range, T_A		–40 to +85	°C
Storage Temperature Range		–65 to +150	°C
Maximum Junction Temperature, T_J		+150	°C
Maximum Junction Temperature, T_J (continuous operation for long term reliability)		+125	°C
ESD Rating:	Human Body Model (HBM)	4000	V
	Charge Device Model (CDM)	1000	V
	Machine Model	200	V

DRB PACKAGE
VSON-8
(TOP VIEW)



DRV PACKAGE
SOT23-5
(TOP VIEW)



Note: NC: Not connected.

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ELECTRICAL CHARACTERISTICS: V_S = ±6V

At R_F = 0Ω, G = +1V/V, and R_L = 100Ω, T_A = +25°C, unless otherwise noted.

PARAMETER	CONDITIONS	OPA659			UNIT	TEST LEVEL(1)
		MIN	TYP	MAX		
AC PERFORMANCE						
Small-Signal Bandwidth	V_O = 200mV$_{PP}$, G = +1V/V		650		MHz	C
	V_O = 200mV$_{PP}$, G = +2V/V		335		MHz	C
	V_O = 200mV$_{PP}$, G = +5V/V		75		MHz	C
	V_O = 200mV$_{PP}$, G = +10V/V		35		MHz	C
Gain Bandwidth Product	G > +10V/V		350		MHz	C
Bandwidth for 0.1dB Flatness	G = +2V/V, V_O = 2V$_{PP}$		55		MHz	C
Large-Signal Bandwidth	V_O = 2V$_{PP}$, G = +1V/V		575		MHz	B
Slew Rate	V_O = 4V Step, G = +1V/V		2550		V/µs	B
Rise and Fall Time	V_O = 4V Step, G = +1V/V		1.3		ns	C
Settling Time to 1%	V_O = 4V Step, G = +1V/V		8		ns	C
Pulse Response Overshoot	V_O = 4V Step, G = +1V/V		12		%	C
Harmonic Distortion	V_O = 2V$_{PP}$, G = +1V/V, f = 10MHz					
2nd harmonic			–79		dBc	C
3rd harmonic			–100		dBc	C
Intermodulation Distortion	V_O= 2V$_{PP}$ Envelope (each tone 1V$_{PP}$), G = +2V/V, f_1 = 10MHz, f_2 = 11MHz					
2nd intermodulation			–72		dBc	C
3rd intermodulation			–96		dBc	C
Input Voltage Noise	f > 100kHz		8.9		nV/√Hz	C
Input Current Noise	f < 10MHz		1.8		fA/√Hz	C
DC PERFORMANCE						
Open-Loop Voltage Gain (A_{OL})	T_A = +25°C, V_{CM} = 0V, R_L = 100Ω	52	58		dB	A
	T_A = –40°C to +85°C, V_{CM} = 0V, R_L = 100Ω	49	55		dB	B
Input Offset Voltage	T_A = +25°C, V_{CM} = 0V		±1	±5	mV	A
	T_A = –40°C to +85°C, V_{CM} = 0V, DRB package		±1.5	±7.6	mV	B
	T_A = –40°C to +85°C, V_{CM} = 0V, DBV package		±1.5	±8.9	mV	B
Average offset voltage drift(2)	T_A = –40°C to +85°C, V_{CM} = 0V, DRB package		±10	±40	µV/°C	B
	T_A = –40°C to +85°C, V_{CM} = 0V, DBV package		±10	±60	µV/°C	B
Input Bias Current	T_A = +25°C, V_{CM} = 0V		±10	±50	pA	A
	T_A = 0°C to +70°C, V_{CM} = 0V		±240	±1200	pA	B
	T_A = –40°C to +85°C, V_{CM} = 0V		±640	±3200	pA	B
Average input bias current drift	T_A = 0°C to +70°C, V_{CM} = 0V		±5	±26	pA/°C	B
	T_A = –40°C to +85°C, V_{CM} = 0V		±7	±34	pA/°C	B
Input Offset Current	T_A = +25°C, V_{CM} = 0V		±5	±25	pA	A
	T_A = 0°C to +70°C, V_{CM} = 0V		±120	±600	pA	B
	T_A = –40°C to +85°C, V_{CM} = 0V		±320	±1600	pA	B

(1) Test levels: **(A)** 100% tested at +25°C. Over temperature limits set by characterization and simulation. **(B)** Limits set by characterization and simulation. **(C)** Typical value only for information.
(2) DRB package only.

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ELECTRICAL CHARACTERISTICS: V_S = ±6V (continued)

At R_F = 0Ω, G = +1V/V, and R_L = 100Ω, T_A = +25°C, unless otherwise noted.

PARAMETER	CONDITIONS	OPA659 MIN	OPA659 TYP	OPA659 MAX	UNIT	TEST LEVEL(1)
INPUT						
Common-Mode Input Range(3)	T_A = +25°C	±3	±3.5		V	A
	T_A = –40°C to +85°C	±2.87	±3.37		V	B
Common-Mode Rejection Ratio	T_A = +25°C, V_{CM} = ±0.5V	68	70		dB	A
	T_A = –40°C to +85°C, V_{CM} = ±0.5V	64	66		dB	B
Input Impedance						
Differential			10^{12} ‖ 1		Ω ‖ pF	C
Common-mode			10^{12} ‖ 2.5		Ω ‖ pF	C
OUTPUT						
Output Voltage Swing	T_A = +25°C, No Load	±4.6	±4.8		V	A
	T_A = +25°C, R_L = 100Ω	±3.8	±4.0		V	A
	T_A = –40°C to +85°C, No Load	±4.45	±4.65		V	B
	T_A = –40°C to +85°C, R_L = 100Ω	±3.65	±3.85		V	B
Output Current, Sourcing, Sinking	T_A = +25°C	±60	±70		mA	A
	T_A = –40°C to +85°C	±56	±65		mA	B
Closed-Loop Output Impedance	G = +1V/V, f = 100kHz		0.04		Ω	C
POWER SUPPLY						
Operating Voltage		±3.5	±6	±6.5	V	B
Quiescent Current	T_A = +25°C	30.5	32	33.5	mA	A
	T_A = –40°C to +85°C	28.3		35.7	mA	B
Power-Supply Rejection Ratio (PSRR)	T_A = 25°C, V_S = ±5.5V to ±6.5V	58	62		dB	A
	T_A = –40°C to 85°C, V_S = ±5.5V to ±6.5V	56	60		dB	A
THERMAL CHARACTERISTICS						
Specified Operating Range DRB and DRV Packages		–40		+85	°C	C
Thermal Resistance, θ_{JA}	Junction-to-ambient					
DRB VSON-8			55		°C/W	C
DRV SOT23-5			105		°C/W	C

(3) Tested <6dB below minimum specified CMRR at ±CMIR limits.

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TYPICAL CHARACTERISTICS

Table of Graphs

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Noninverting Small-Signal Frequency Response	V_O = 200mV$_{PP}$	Figure 1
Noninverting Large-Signal Frequency Response	V_O = 2V$_{PP}$	Figure 2
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Transimpedance Gain vs Frequency	C_D = 100pF	Figure 31
Maximum/Minimum ±V_{OUT} vs R_{LOAD}		Figure 32
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TYPICAL CHARACTERISTICS

At V_S = ±6V, R_F = 0Ω, G = +1V/V, and R_L = 100Ω, unless otherwise noted.

NONINVERTING SMALL-SIGNAL FREQUENCY RESPONSE (V_O = 200mV$_{PP}$)

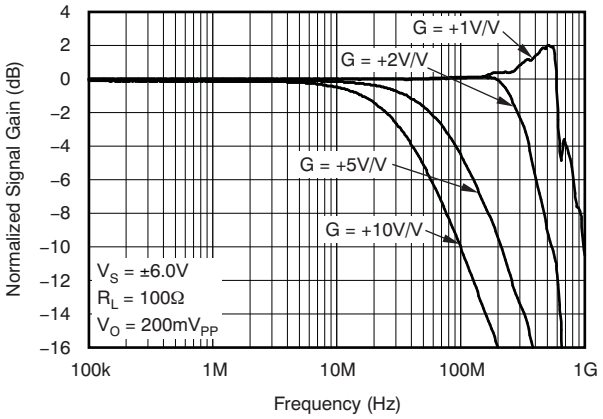


Figure 1.

NONINVERTING LARGE-SIGNAL FREQUENCY RESPONSE (V_O = 2V$_{PP}$)

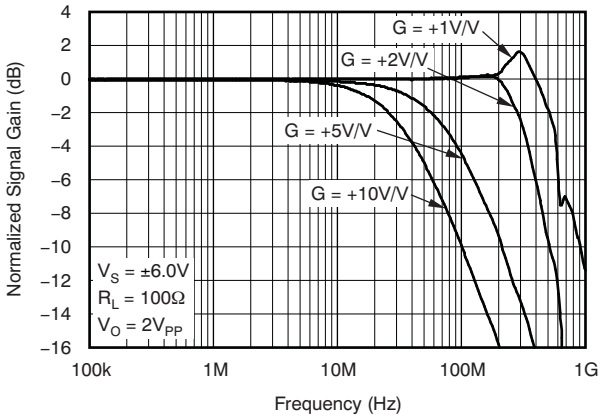


Figure 2.

NONINVERTING LARGE-SIGNAL FREQUENCY RESPONSE (V_O = 6V$_{PP}$)

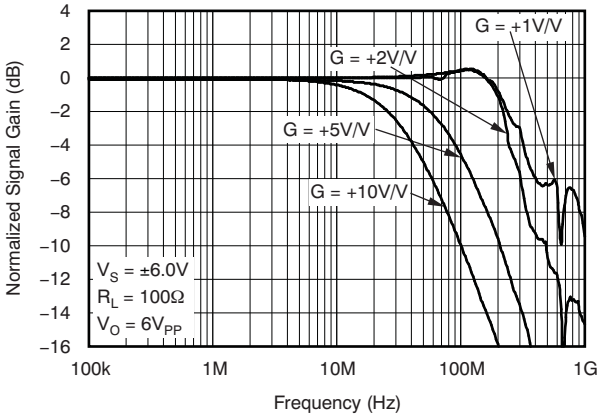


Figure 3.

INVERTING SMALL-SIGNAL FREQUENCY RESPONSE (V_O = 200mV$_{PP}$)

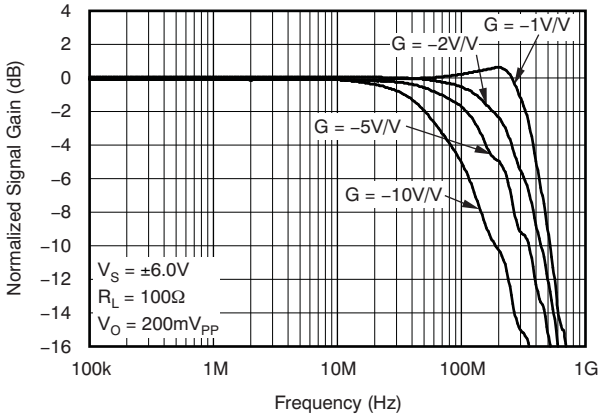


Figure 4.

INVERTING LARGE-SIGNAL FREQUENCY RESPONSE (V_O = 2V$_{PP}$)

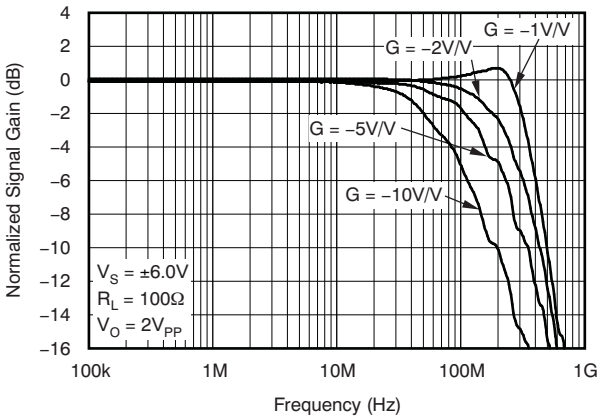


Figure 5.

INVERTING LARGE-SIGNAL FREQUENCY RESPONSE (V_O = 6V$_{PP}$)

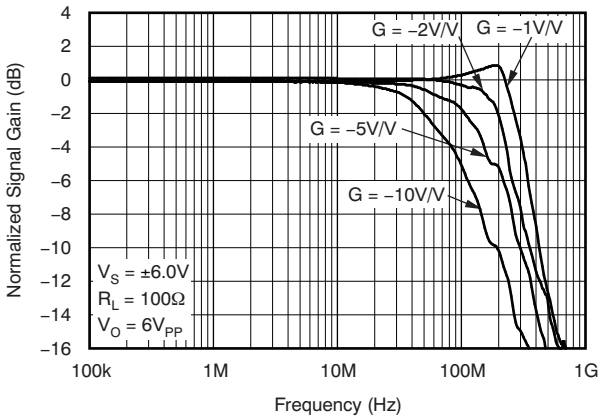


Figure 6.

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TYPICAL CHARACTERISTICS (continued)

At V_S = ±6V, R_F = 0Ω, G = +1V/V, and R_L = 100Ω, unless otherwise noted.

NONINVERTING TRANSIENT RESPONSE (0.5V STEP)

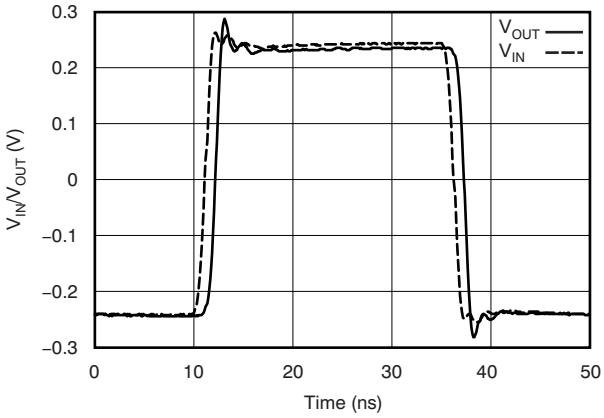


Figure 7.

NONINVERTING TRANSIENT RESPONSE (2V STEP)

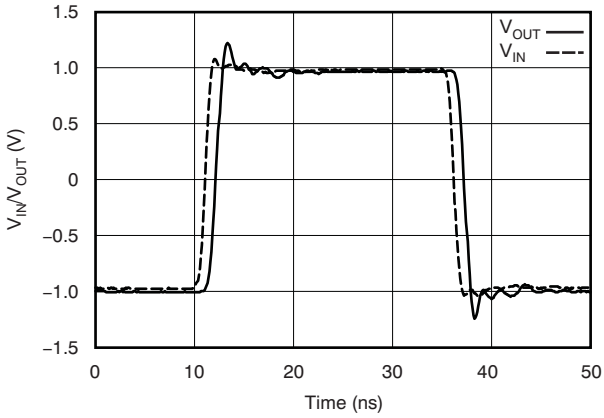


Figure 8.

NONINVERTING TRANSIENT RESPONSE (5V STEP)

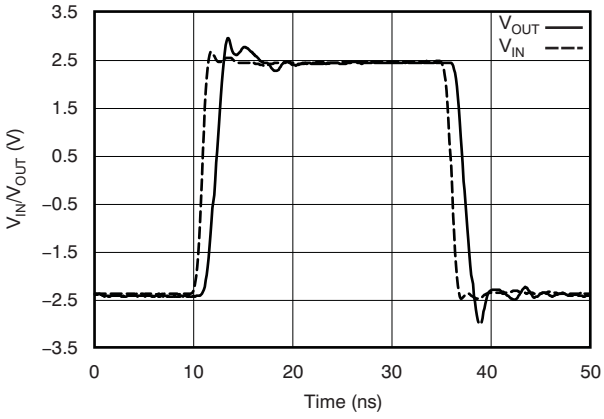


Figure 9.

INVERTING TRANSIENT RESPONSE (0.5V STEP)

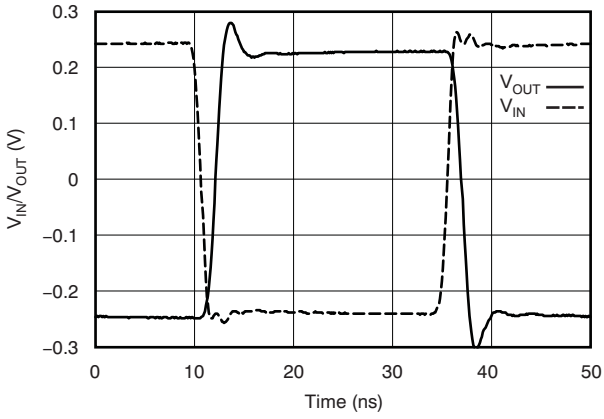


Figure 10.

INVERTING TRANSIENT RESPONSE (2V STEP)

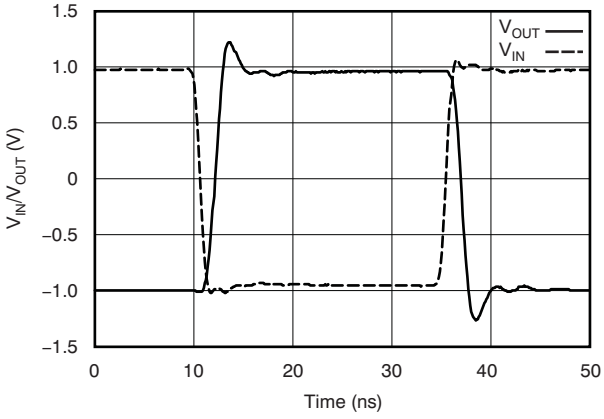


Figure 11.

INVERTING TRANSIENT RESPONSE (5V STEP)

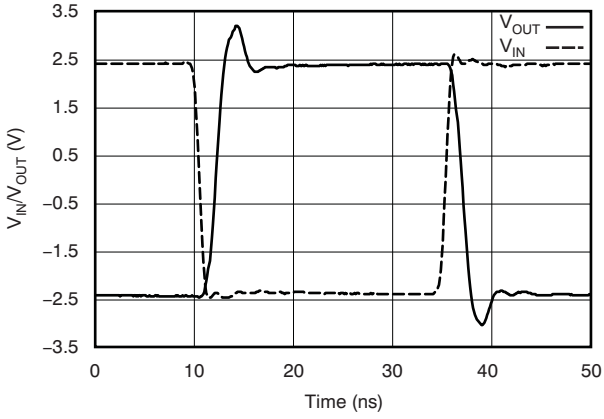


Figure 12.

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TYPICAL CHARACTERISTICS (continued)

At V_S = ±6V, R_F = 0Ω, G = +1V/V, and R_L = 100Ω, unless otherwise noted.

HARMONIC DISTORTION vs FREQUENCY

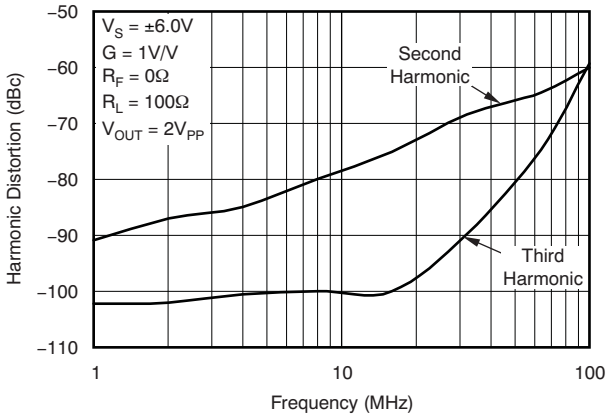


Figure 13.

HARMONIC DISTORTION vs NONINVERTING GAIN AT 10MHz

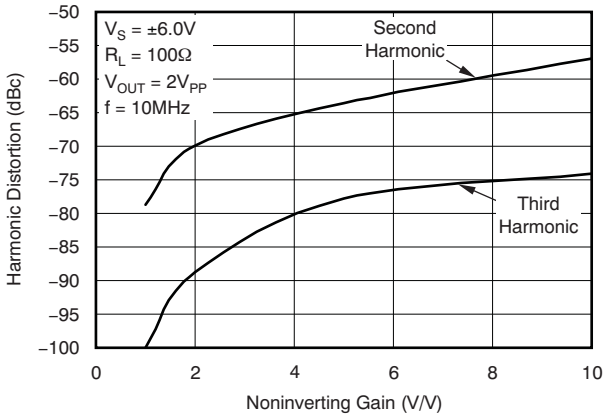


Figure 14.

HARMONIC DISTORTION vs INVERTING GAIN AT 10MHz

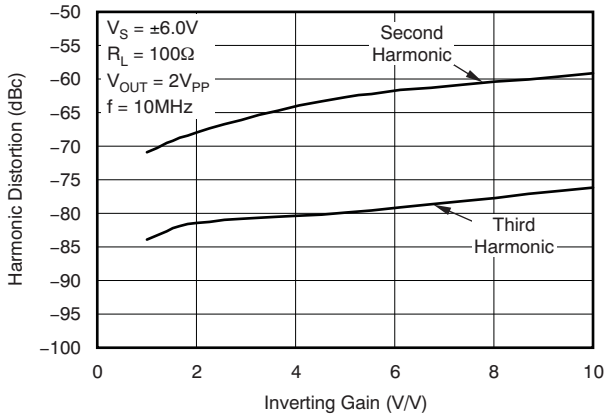


Figure 15.

HARMONIC DISTORTION vs LOAD RESISTANCE AT 10MHz

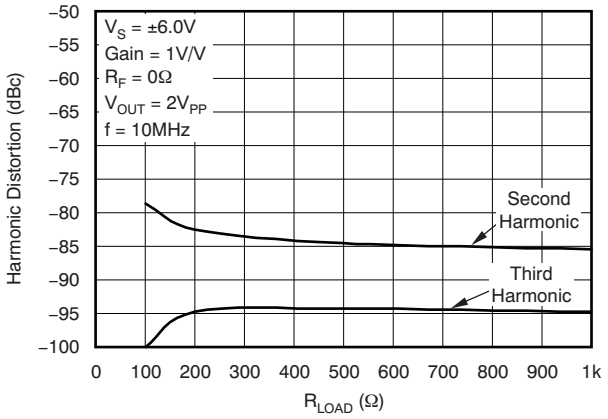


Figure 16.

HARMONIC DISTORTION vs OUTPUT VOLTAGE

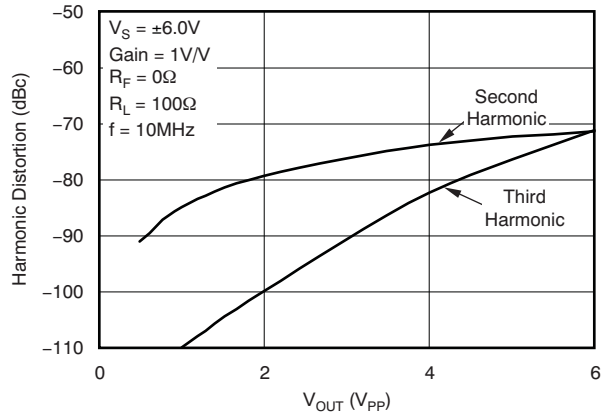


Figure 17.

HARMONIC DISTORTION vs ±SUPPLY VOLTAGE

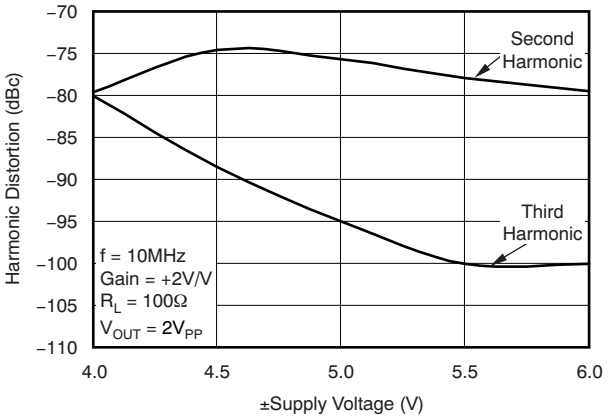


Figure 18.

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TYPICAL CHARACTERISTICS (continued)

At V_S = ±6V, R_F = 0Ω, G = +1V/V, and R_L = 100Ω, unless otherwise noted.

TWO-TONE, SECOND- AND THIRD-ORDER IMD vs FREQUENCY

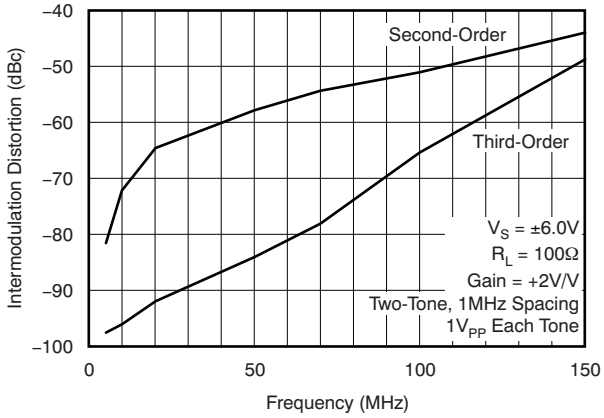


Figure 19.

OVERDRIVE RECOVERY (GAIN = +2V/V)

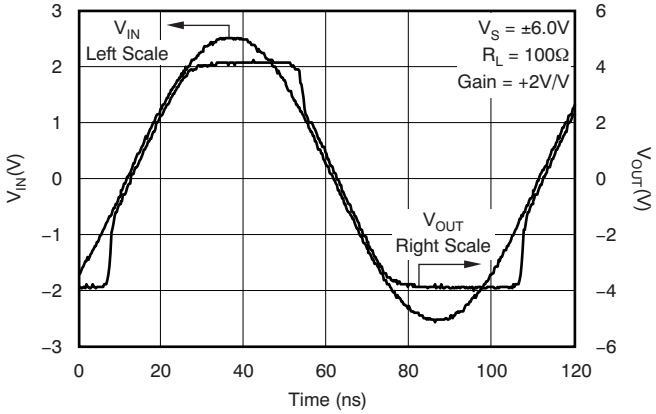


Figure 20.

OVERDRIVE RECOVERY (GAIN = –2V/V)

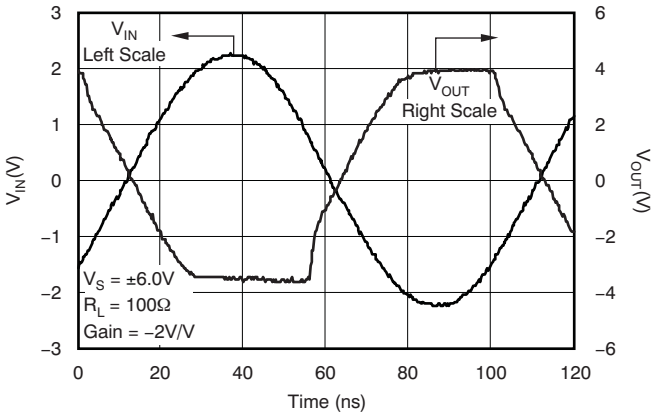


Figure 21.

INPUT-REFERRED VOLTAGE AND CURRENT NOISE DENSITY

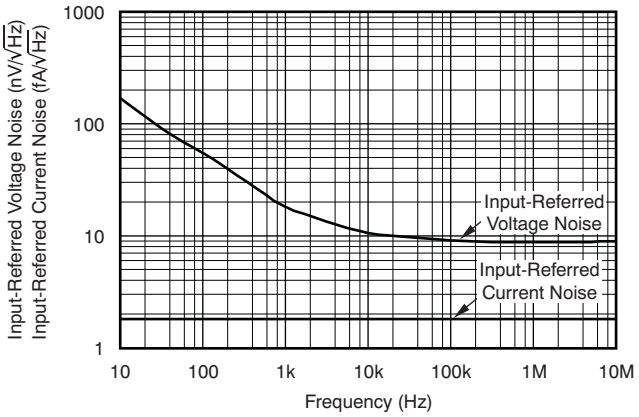


Figure 22.

COMMON-MODE REJECTION RATIO AND POWER-SUPPLY REJECTION RATIO vs FREQUENCY

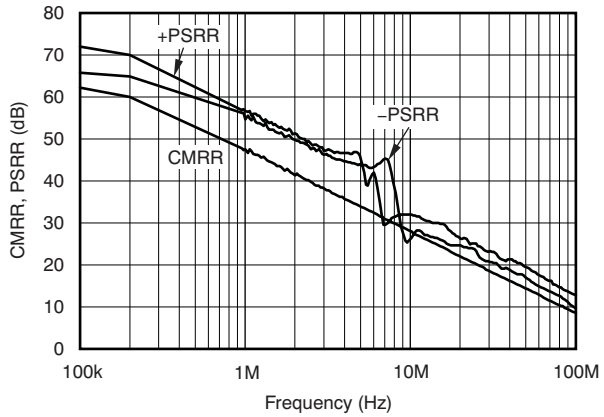


Figure 23.

RECOMMENDED R_{ISO} vs CAPACITIVE LOAD (R_{LOAD} = 1kΩ)

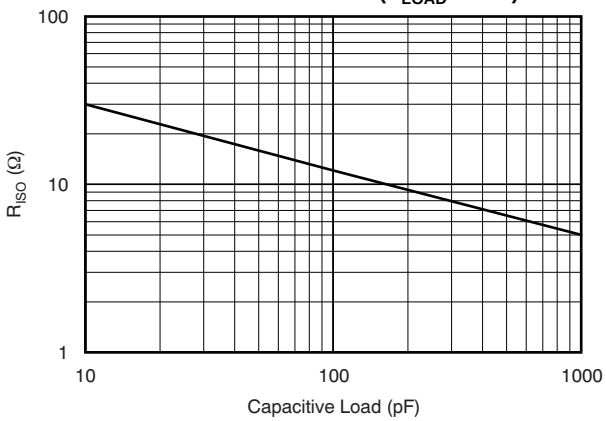


Figure 24.

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TYPICAL CHARACTERISTICS (continued)

At V_S = ±6V, R_F = 0Ω, G = +1V/V, and R_L = 100Ω, unless otherwise noted.

FREQUENCY RESPONSE vs CAPACITIVE LOAD (R_{LOAD} = 1kΩ)

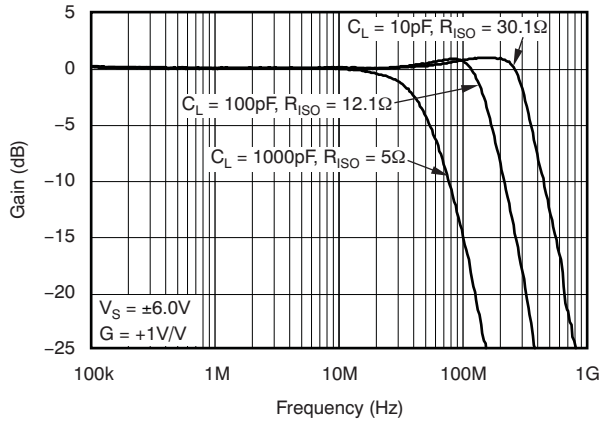


Figure 25.

OPEN-LOOP GAIN AND PHASE

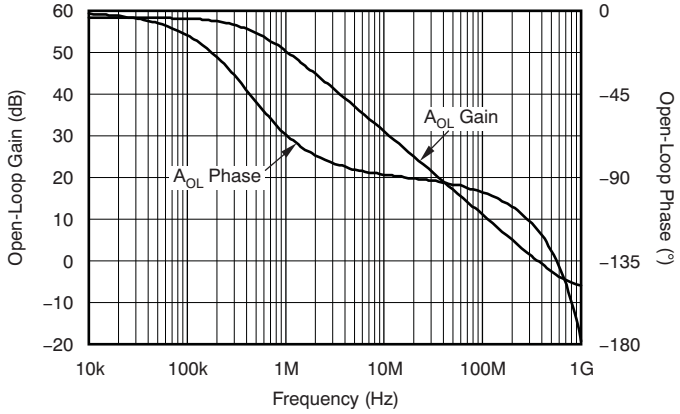


Figure 26.

CLOSED-LOOP OUTPUT IMPEDANCE vs FREQUENCY

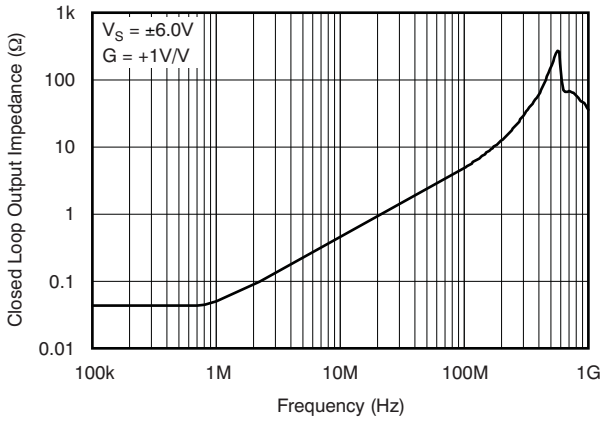


Figure 27.

TRANSIMPEDANCE GAIN vs FREQUENCY (C_D = 10pF)

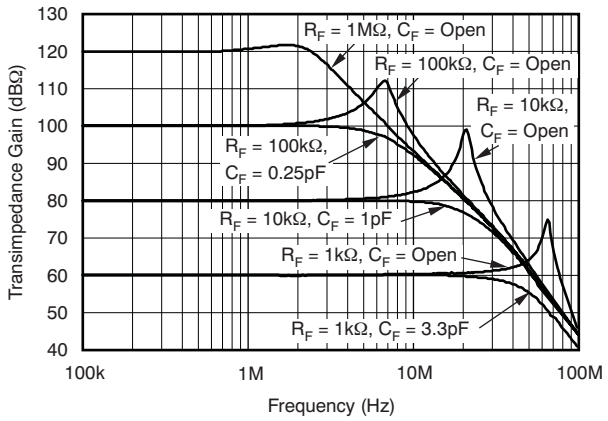


Figure 28.

TRANSIMPEDANCE GAIN vs FREQUENCY (C_D = 22pF)

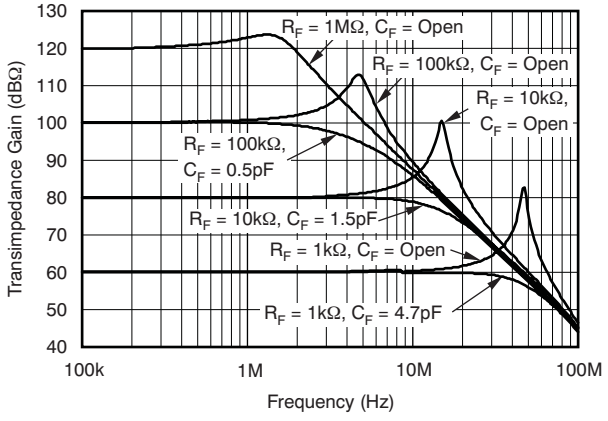


Figure 29.

TRANSIMPEDANCE GAIN vs FREQUENCY (C_D = 47pF)

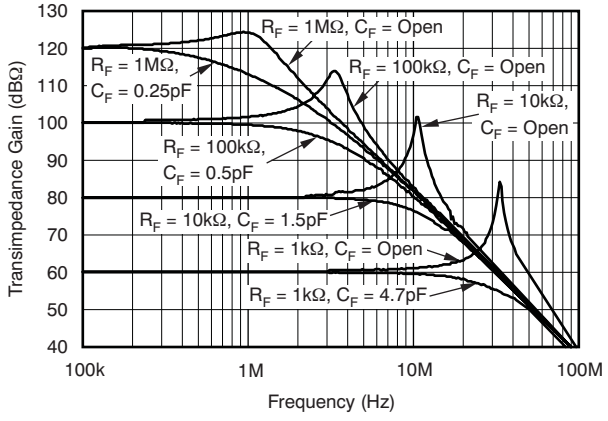


Figure 30.

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TYPICAL CHARACTERISTICS (continued)

At V_S = ±6V, R_F = 0Ω, G = +1V/V, and R_L = 100Ω, unless otherwise noted.

TRANSIMPEDANCE GAIN vs FREQUENCY (C_D = 100pF)

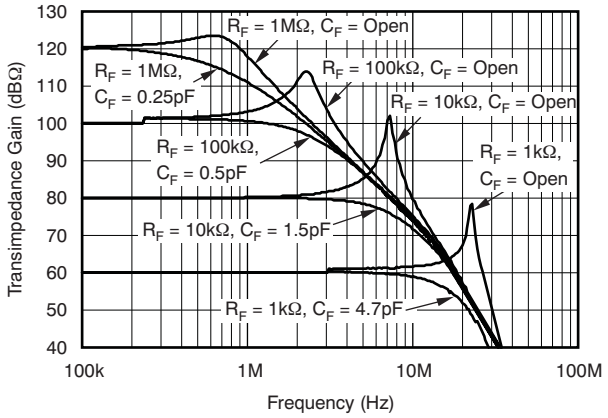


Figure 31.

MAXIMUM/MINIMUM ±V_{OUT} vs R_{LOAD}

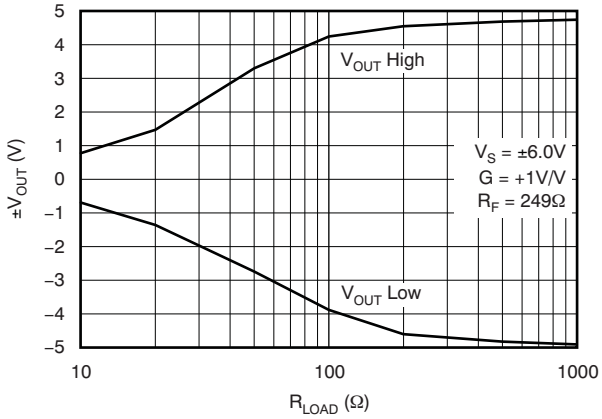


Figure 32.

SLEW RATE vs V_{OUT} STEP

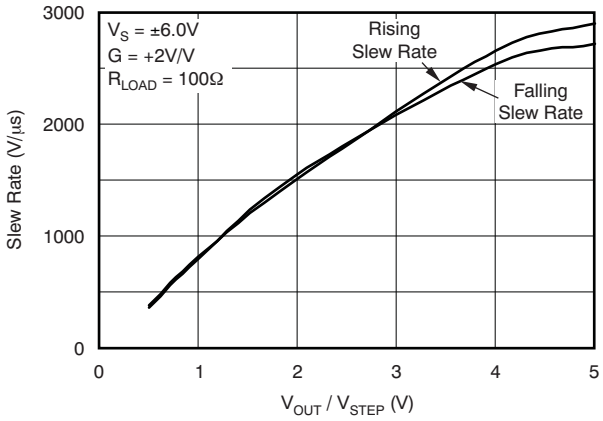


Figure 33.

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APPLICATION INFORMATION

Wideband, Noninverting Operation

The OPA659 is a very broadband, unity-gain stable, voltage-feedback amplifier with a high impedance JFET-input stage. Its very high gain bandwidth product (GBP) of 350MHz can be used to either deliver high signal bandwidths for low-gain buffers, or to deliver broadband, low-noise, transimpedance bandwidth to photodiode-detector applications. The OPA659 is designed to to provide very low distortion and accurate pulse response with low overshoot and ringing. To achieve the full performance of the OPA659, careful attention to printed circuit board (PCB) layout and component selection are required, as discussed in the remaining sections of this data sheet.

Figure 34 shows the noninverting gain of +1 circuit; Figure 35 shows the more general circuit used for other noninverting gains. These circuits are used as the basis for most of the noninverting gain Typical Characteristics graphs. Most of the graphs were characterized using signal sources with 50Ω driving impedance, and with measurement equipment presenting a 50Ω load impedance. In Figure 34, the shunt resistor R_T at V_{IN} should be set to 50Ω to match the source impedance of the test generator and cable, while the series output resistor, R_{OUT}, at V_{OUT} should also be set to 50Ω to provide matching impedance for the measurement equipment load and cable. Generally, data sheet voltage swing specifications are measured at the output pin, V_{OUT}, in Figure 34 and Figure 35.

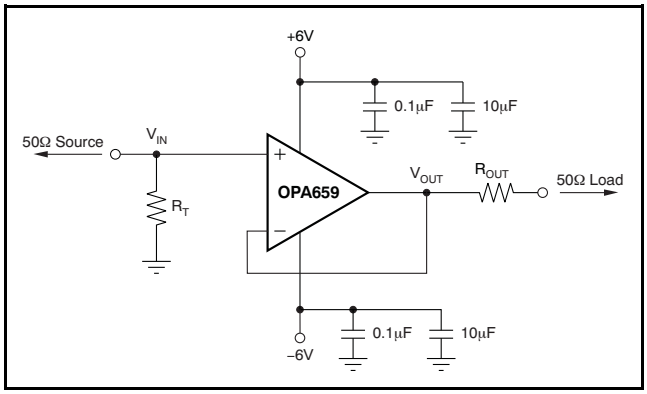


Figure 34. Noninverting Gain of +1 Test Circuit

Voltage-feedback op amps can use a wide range of resistor values to set the gain. To retain a controlled frequency response for the noninverting voltage amplifier of Figure 35, the parallel combination of R_F || R_G should always be less than 200Ω. In the noninverting configuration, the parallel combination of R_F || R_G forms a pole with the parasitic input and board layout capacitance at the inverting input of the OPA659. For best performance, this pole should be at a frequency greater than the closed-loop bandwidth for the OPA659. For this reason, a direct short from the output to the inverting input is recommended for the unity-gain follower application. Table 1 lists several recommended resistor values for noninverting gains with a 50Ω input/output match.

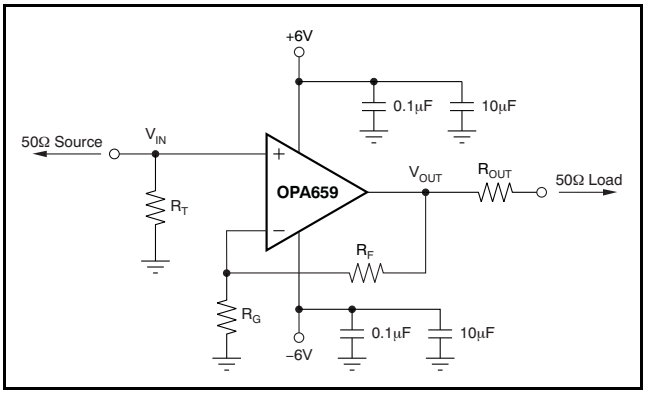


Figure 35. General Noninverting Test Circuit

Table 1. Resistor Values for Noninverting Gains with 50Ω Input/Output Match

NONINVERTING GAIN	R_F	R_G	R_T	R_{OUT}
+1	0	Open	49.9	49.9
+2	249	249	49.9	49.9
+5	249	61.9	49.9	49.9
+10	249	27.4	49.9	49.9

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Wideband, Inverting Gain Operation

The circuit of Figure 36 shows the inverting gain test circuit used for most of the inverting Typical Characteristics graphs. As with the noninverting applications, most of the curves were characterized using signal sources with 50Ω driving impedance, and with measurement equipment that presents a 50Ω load impedance. In Figure 36, the shunt resistor R_T at V_{IN} should be set so the parallel combination of the shunt resistor and R_G equals 50Ω to match the source impedance of the test generator and cable, while the series output resistor R_{OUT} at V_{OUT} should also be set to 50Ω to provide matching impedance for the measurement equipment load and cable. Generally, data sheet voltage swing specifications are measured at the output pin, V_{OUT}, in Figure 36.

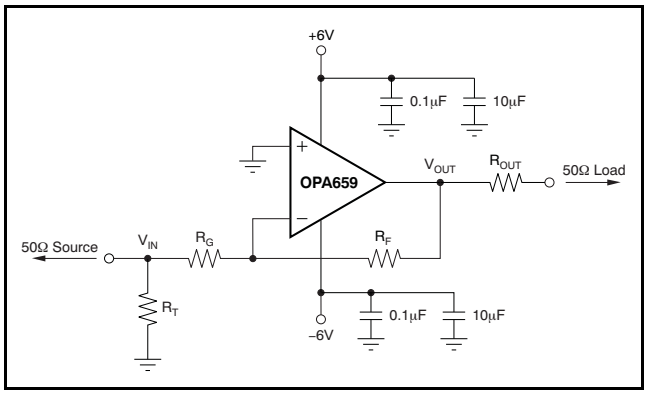


Figure 36. General Inverting Test Circuit

The inverting circuit can also use a wide range of resistor values to set the gain; Table 2 lists several recommended resistor values for inverting gains with a 50Ω input/output match.

Table 2. Resistor Values for Inverting Gains with 50Ω Input/Output Match

INVERTING GAIN	R_F	R_G	R_T	R_{OUT}
–1	249	249	61.9	49.9
–2	249	124	84.5	49.9
–5	249	49.9	Open	49.9
–10	499	49.9	Open	49.9

Figure 36 shows the noninverting input tied directly to ground. Often, a bias current-cancelling resistor to ground is included here to nullify the dc errors caused by input bias current effects. For a JFET input op amp such as the OPA659, the input bias currents are so low that dc errors caused by input bias currents are negligible. Thus, no bias current-cancelling resistor is recommended at the noninverting input.

Wideband, High-Sensitivity, Transimpedance Design

The high GBP and low input voltage and current noise for the OPA659 make it an ideal wideband, transimpedance amplifier for low to moderate transimpedance gains. Higher transimpedance gains (above 100kΩ) can benefit from the low input noise current of a JFET input op amp such as the OPA659. Designs that require high bandwidth from a large area detector can benefit from the low input voltage noise for the OPA659. This input voltage noise is peaked up over frequency by the diode source capacitance, and in many cases, may become the limiting factor to input sensitivity. The key elements to the design are the expected diode capacitance (C_D) with the reverse bias voltage ($–V_B$) applied, the desired transimpedance gain, R_F, and the GBP for the OPA659 (350MHz). Figure 37 shows a general transimpedance amplifier circuit, or TIA, using the OPA659. Given the source diode capacitance plus parasitic input capacitance for the OPA659, the transimpedance gain, and known GBP, the feedback capacitor value, C_F, may be calculated to avoid excessive peaking in the frequency response.

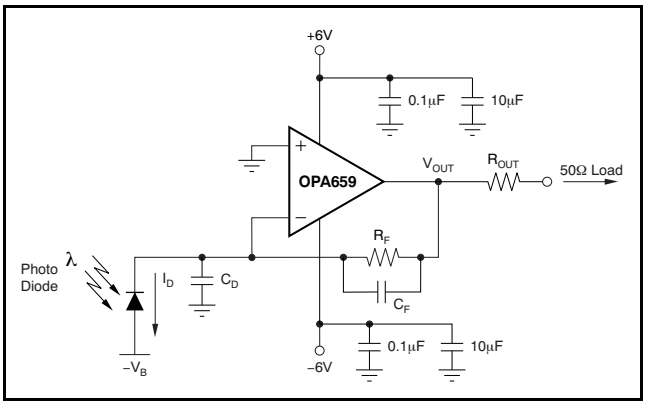


Figure 37. Wideband, Low-Noise, Transimpedance Amplifier (TIA)

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To achieve a maximally flat second-order Butterworth frequency response, the feedback pole should be set to:

$$\frac{1}{2\pi R_F C_F} = \sqrt{\frac{GBP}{4\pi R_F C_D}} \quad (1)$$

For example, adding the common mode and differential mode input capacitance (0.7 + 2.8 = 3.5)pF to the diode source with the 20pF capacitance, and targeting a 100kΩ transimpedance gain using the 350MHz GBP for the OPA659, requires a feedback pole set to 3.44MHz. This pole in turn requires a total feedback capacitance of 0.46pF. Typical surface mount resistors have a parasitic capacitance of 0.2pF, leaving the required 0.26pF value to achieve the required feedback pole. This calculation gives an approximate 4.9MHz, –3dB bandwidth computed by:

$$f_{-3dB} = \sqrt{\frac{GBP}{2\pi R_F C_D}} \quad (2)$$

Table 3 lists the calculated component values and –3dB bandwidths for various TIA gains and diode capacitance.

Table 3. OPA659 TIA Component Values and Bandwidth for Various Diode Capacitance and Gains

C_{DIODE} = 10pF			
C_D	R_F	C_F	f_{-3dB}
13.5 pF	1kΩ	3.50pF	64.24MHz
13.5 pF	10kΩ	1.11pF	20.31MHz
13.5 pF	100kΩ	0.35pF	6.42MHz
13.5 pF	1MΩ	0.11pF	2.03MHz
C_{DIODE} = 20pF			
23.5 pF	1kΩ	4.62pF	48.69MHz
23.5 pF	10kΩ	1.46pF	15.40MHz
23.5 pF	100kΩ	0.46pF	4.87MHz
23.5 pF	1MΩ	0.15pF	1.54MHz
C_{DIODE} = 50pF			
53.5 pF	1kΩ	6.98pF	32.27MHz
53.5 pF	10kΩ	2.21pF	10.20MHz
53.5 pF	100kΩ	0.70pF	3.23MHz
53.5 pF	1MΩ	0.22pF	1.02MHz
C_{DIODE} = 100pF			
103.5 pF	1kΩ	9.70pF	23.20MHz
103.5 pF	10kΩ	3.07pF	7.34MHz
103.5 pF	100kΩ	0.97pF	2.32MHz
103.5 pF	1MΩ	0.31pF	0.73MHz

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OPERATING SUGGESTIONS

Setting Resistor Values to Minimize Noise

The OPA659 provides a very low input noise voltage. To take full advantage of this low input noise, designers must pay careful attention to other possible noise contributors. Figure 38 shows the op amp noise analysis model with all the noise terms included. In this model, all the noise terms are taken to be noise voltage or current density terms in either nV/√Hz or pA/√Hz.

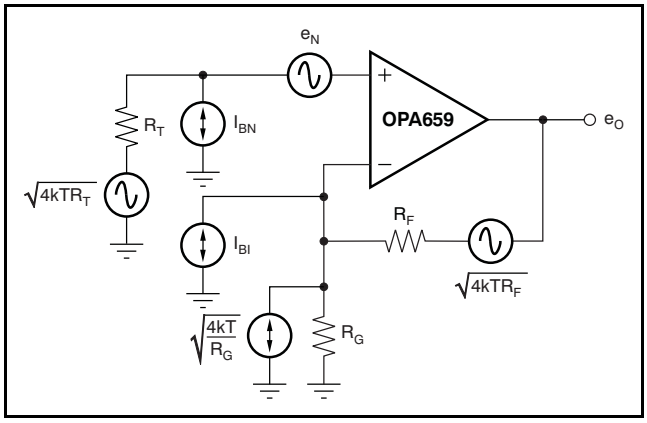


Figure 38. Op Amp Noise Analysis Model

The total output spot noise voltage can be computed as the square root of the squared contributing terms to the output noise voltage. This computation adds all the contributing noise powers at the output by superposition, then takes the square root to arrive at a spot noise voltage. Equation 3 shows the general form for this output noise voltage using the terms shown in Figure 38.

$$e_O = \sqrt{\left[4kTR_T + (I_{BN}R_T)^2 + e_N^2\right]\left(1 + \frac{R_F}{R_G}\right)^2 + (I_{BI}R_F)^2 + 4kTR_F\left(1 + \frac{R_F}{R_G}\right)} \quad (3)$$

Dividing this expression by the noise gain ($G_N = 1 + R_F/R_G$) gives the equivalent input-referred spot noise voltage at the noninverting input, as Equation 4 shows.

$$e_{NI} = \sqrt{4kTR_T + (I_{BN}R_T)^2 + e_N^2 + \left(\frac{I_{BI}R_F}{\text{Noise Gain}}\right)^2 + \left(\frac{4kTR_F}{\text{Noise Gain}}\right)} \quad (4)$$

Putting high resistor values into Equation 4 can quickly dominate the total equivalent input-referred noise. A source impedance on the noninverting input of 5kΩ adds a Johnson voltage noise term equal to that of the amplifier alone (8.9nV/√Hz). While the JFET input of the OPA659 is ideal for high source impedance applications in the noninverting configuration of Figure 34 or Figure 35, both the overall bandwidth and noise are limited by high source impedances.

Frequency Response Control

Voltage-feedback op amps such as the OPA659 exhibit decreasing signal bandwidth as the signal gain increases. In theory, this relationship is described by the gain bandwidth product (GBP) shown in the Electrical Characteristics. Ideally, dividing the GBP by the noninverting signal gain (also called the Noise Gain, or NG) can predict the closed-loop bandwidth. In practice, this guideline is valid only when the phase margin approaches 90 degrees, as it does in high gain configurations. At low gains (with increased feedback factors), most high-speed amplifiers exhibit a more complex response with lower phase margins. The OPA659 is compensated to give a maximally-flat frequency response at a noninverting gain of +1 (see Figure 34). This compensation results in a typical gain of +1 bandwidth of 650MHz, far exceeding that predicted by dividing the 350MHz GBP by 1. Increasing the gain causes the phase margin to approach 90 degrees and the bandwidth to more closely approach the predicted value of (GBP/NG). At a gain of +10, the OPA659 shows the 35MHz bandwidth predicted using the simple formula and the typical GBP of 350MHz. Unity-gain stable op amps such as the OPA659 can also be band-limited in gains other than +1 by placing a capacitor across the feedback resistor. For the noninverting configuration of Figure 35, a capacitor across the feedback resistor decreases the gain with frequency down to a gain of +1. For instance, to band-limit a gain of +2 design to 20MHz, a 32pF capacitor can be placed in parallel with the 249Ω feedback resistor. This configuration, however, only decreases the gain from 2 to 1. Using a feedback capacitor to limit the signal bandwidth is more effective in the inverting configuration of Figure 36. Adding that same capacitance to the feedback of Figure 36 sets a pole in the signal frequency response at 20MHz, but in this case it continues to attenuate the signal gain to less than 1. Note, however, that the noise gain of the circuit is only reduced to a gain of 1 with the addition of the feedback capacitor.

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Driving Capacitive Loads

One of the most demanding, and yet very common, load conditions for an op amp is capacitive loading. The OPA659 is very robust, but care should be taken with light loading scenarios so that output capacitance does not decrease stability and increase closed-loop frequency response peaking when a capacitive load is placed directly on the output pin. When the amplifier open-loop output resistance is considered, this capacitive load introduces an additional pole in the signal path that can decrease the phase margin. Several external solutions to this problem have been suggested. When the primary considerations are frequency response flatness, pulse response fidelity, and/or distortion, the simplest and most effective solution is to isolate the capacitive load from the feedback loop by inserting a series isolation resistor, R_{ISO}, between the amplifier output and the capacitive load. In effect, this resistor isolates the phase shift from the loop gain of the amplifier, thus increasing the phase margin and improving stability. The Typical Characteristics show the recommended R_{ISO} versus capacitive load and the resulting frequency response with a 1kΩ load (see Figure 24). Note that larger R_{ISO} values are required for lower capacitive loading. In this case, a design target of a maximally-flat frequency response was used. Lower values of R_{ISO} may be used if some peaking can be tolerated. Also, operating at higher gains (instead of the +1 gain used in the Typical Characteristics) requires lower values of R_{ISO} for a minimally-peaked frequency response. Parasitic capacitive loads greater than 2pF can begin to degrade the performance of the OPA659. Moreover, long PCB traces, unmatched cables, and connections to multiple devices can easily cause this value to be exceeded. Always consider this effect carefully, and add the recommended series resistor as close as possible to the OPA659 output pin (see the Board Layout section).

With heavier loads (for example, the 100Ω load presented in the test circuits and used for testing typical characteristic performance), the OPA659 is very robust; R_{ISO} can be as low as 10Ω with capacitive loads less than 5pF and continue to show a flat frequency response.

Distortion Performance

The OPA659 is capable of delivering a low distortion signal at high frequencies over a wide range of gains. The distortion plots in the Typical Characteristics show the typical distortion under a wide variety of conditions. Generally, until the fundamental signal reaches very high frequencies or powers, the second harmonic dominates the distortion with a negligible third harmonic component. Focusing then on the second harmonic, increasing the load impedance improves distortion directly. Remember that the total load includes the feedback network: in the noninverting configuration, this network is the sum of $R_F + R_G$, while in the inverting configuration the network is only R_F (see Figure 35). Increasing the output voltage swing directly increases harmonic distortion. A 6dB increase in output swing generally increases the second harmonic by 12dB and the third harmonic by 18dB. Increasing the signal gain also increases the second-harmonic distortion. Again, a 6dB increase in gain increases the second and third harmonics by about 6dB, even with a constant output power and frequency. Finally, the distortion increases as the fundamental frequency increases because of the rolloff in the loop gain with frequency. Conversely, the distortion improves going to lower frequencies, down to the dominant open-loop pole at approximately 300kHz.

Note that power-supply decoupling is critical for harmonic distortion performance. In particular, for optimal second-harmonic performance, the power-supply high-frequency 0.1μF decoupling capacitors to the positive and negative supply pins should be brought to a single point ground located away from the input pins.

The OPA659 has an extremely low third-order harmonic distortion. This characteristic also shows up in the two-tone, third-order intermodulation spurious (IMD3) response curves (see Figure 19). The third-order spurious levels are extremely low (less than –100dBc) at low output power levels and frequencies below 10MHz. The output stage continues to hold these levels low even as the fundamental power reaches higher levels. As with most op amps, the spurious intermodulation powers do not increase as predicted by a traditional intercept model. As the fundamental power level increases, the dynamic range does not decrease significantly. For two tones centered at 10MHz, with –2dBm/tone into a matched 50Ω load (that is, $0.5V_{PP}$ for each tone at the load, which requires $2V_{PP}$ for the overall two-tone envelope at the output pin), the Typical Characteristics show a 96dBc difference between the test tones and the third-order intermodulation spurious levels. This exceptional performance improves further when operating at lower frequencies and/or higher load impedances.

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Board Layout

Achieving optimum performance with a high-frequency amplifier such as the OPA659 requires careful attention to PCB layout parasitics and external component types. Recommendations that can optimize device performance include the following.

a) Minimize parasitic capacitance to any ac ground for all of the signal input/output (I/O) pins. Parasitic capacitance on the output and inverting input pins can cause instability: on the noninverting input, it can react with the source impedance to cause unintentional band-limiting. To reduce unwanted capacitance, a window around the signal I/O pins should be opened in all of the ground and power planes around those pins. Otherwise, ground and power planes should be unbroken elsewhere on the board.

b) Minimize the distance (less than 0.25in, or 6,35mm) from the power-supply pins to the high-frequency, 0.1µF decoupling capacitors. At the device pins, the ground and power plane layout should not be in close proximity to the signal I/O pins. Use a single point ground, located away from the input pins, for the positive and negative supply high-frequency, 0.1µF decoupling capacitors. Avoid narrow power and ground traces to minimize inductance between the pins and the decoupling capacitors. The power-supply connections should always be decoupled with these capacitors. Larger (2.2µF to 10µF) decoupling capacitors, effective at lower frequencies, should also be used on the supply pins. These larger capacitors may be placed somewhat farther from the device and may be shared among several devices in the same area of the PCB.

c) Careful selection and placement of external components preserves the high-frequency performance of the OPA659. Resistors should be a very low reactance type. Surface-mount resistors work best and allow a tighter overall layout. Metal film and carbon composition, axially-leaded resistors can also provide good high-frequency performance. Again, keep the leads and PCB trace length as short as possible. Never use wirewound-type resistors in a high-frequency application. The inverting input pin is the most sensitive to parasitic capacitance; consequently, always position the feedback resistor as close to the negative input as possible. The output is also sensitive to parasitic capacitance; therefore, position a series output resistor (in this case, R_{ISO}) as close to the output pin as possible.

Other network components, such as noninverting input termination resistors, should also be placed close to the package. Even with a low parasitic capacitance, excessively high resistor values can create significant time constants that can degrade device performance. Good axial metal film or surface-mount resistors have approximately 0.2pF in shunt with the resistor. For resistor values greater than 1.5kΩ, this parasitic capacitance can add a pole and/or zero below 500MHz that can affect circuit operation. Keep resistor values as low as possible, consistent with load driving considerations. It is recommended to keep $R_F \parallel R_G$ less than 250Ω. This low value ensures that the resistor noise terms remain low, and minimizes the effects of the parasitic capacitance. Transimpedance applications (for example, see Figure 37) can use the feedback resistor required by the application as long as the feedback compensation capacitor is set given consideration to all parasitic capacitance terms on the inverting node.

d) Connections to other wideband devices on the board may be made with short direct traces or through onboard transmission lines. For short connections, consider the trace and the input to the next device as a lumped capacitive load. Relatively wide traces (50mils to 100mils, or 1,27cm to 2,54cm) should be used. Estimate the total capacitive load and set R_{ISO} from the plot of *Recommended R_{ISO} vs Capacitive Load* (Figure 24). Low parasitic capacitive loads (less than 5pF) may not need an R_{ISO} because the OPA659 is nominally compensated to operate with a 2pF parasitic load.

Higher parasitic capacitive loads without an R_{ISO} are allowed as the signal gain increases (increasing the unloaded phase margin). If a long trace is required, and the 6dB signal loss intrinsic to a doubly-terminated transmission line is acceptable, implement a matched impedance transmission line using microstrip or stripline techniques (consult an ECL design handbook for microstrip and stripline layout techniques). A 50Ω environment is normally not necessary onboard, and in fact a higher impedance environment improves distortion as shown in the distortion versus load plots. With a characteristic board trace impedance defined based on board material and trace dimensions, a matching series resistor into the trace from the output of the OPA659 is used as well as a terminating shunt resistor at the input of the destination device. Remember also that the terminating impedance is the parallel combination of the shunt resistor and the input impedance of the destination device: this total effective impedance should be set to match the trace impedance. If the 6dB attenuation of a doubly-terminated transmission line is unacceptable, a long trace can be series-terminated at the source end only. Treat the trace as a capacitive load in this case, and set the series resistor value as shown in the plot of *R_{ISO} vs Capacitive Load* (Figure 24). This configuration does not preserve signal integrity as

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well as a doubly-terminated line. If the input impedance of the destination device is low, there will be some signal attenuation as a result of the voltage divider formed by the series output into the terminating impedance.

e) Socketing a high-speed part such as the OPA659 is not recommended. The additional lead length and pin-to-pin capacitance introduced by the socket can create an extremely troublesome parasitic network that can make it almost impossible to achieve a smooth, stable frequency response. Best results are obtained by soldering the OPA659 directly onto the board.

f) The thermal slug on bottom of OPA659 DRB package must be tied to the most negative supply. The DRB package is a PowerPAD thermally-enhanced package. Best results are obtained by soldering the exposed metal tab on the bottom of the OPA659 DRB directly to a metal plane on the PCB that is connected to the most negative supply voltage of the op amp. General layout guidelines can be found in the application reports, *PowerPAD™ Layout Guidelines* (SLOA120) and *PowerPAD™ Made Easy* (SLMA004) , both available for download from the TI web site at www.ti.com. It may also be helpful to copy the EVM layout shown in the Evaluation Module section.

PowerPAD™ Information

The DRB package option is a PowerPAD™ package that includes a thermal pad for increased thermal performance. When using this package, it is recommended to distribute the negative supply as a power plane, and tie the PowerPAD to this supply with multiple vias for proper power dissipation. For proper operation, the PowerPAD must be tied to the most negative supply voltage. It is recommended to use five evenly-spaced vias under the device as shown in the EVM layer views (see Figure 41). For more general data and detailed information about the PowerPAD package, refer to the *PowerPAD™ Thermally Enhanced Package* application note (SLMA002).

Input and ESD Protection

The OPA659 is built using a very high-speed complementary bipolar process. The internal junction breakdown voltages are relatively low for these very small geometry devices. These breakdowns are reflected in the Absolute Maximum Ratings table. All device pins are protected with internal ESD protection diodes to the power supplies, as Figure 39 shows.

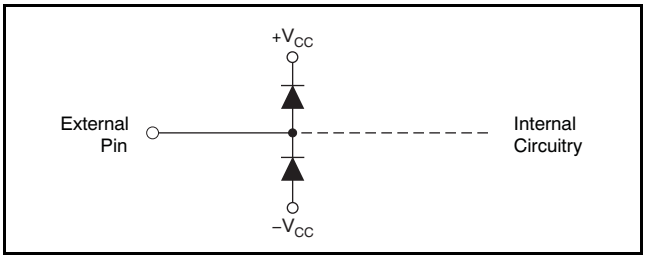


Figure 39. Internal ESD Protection

These diodes provide moderate protection to input overdrive voltages above the supplies as well. The protection diodes can typically support 30mA continuous current. Where higher currents are possible (for example, in systems with ±12V supply parts driving into the OPA659), current limiting series resistors should be added into the two inputs. Keep these resistor values as low as possible because high values degrade both noise performance and frequency response.

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EVALUATION MODULE

Schematic and PCB Layout

Figure 40 is the OPA659EVM schematic. Layers 1 through 4 of the PCB are shown in Figure 41. It is recommended to follow the layout of the external components near to the amplifier, ground plane construction, and power routing as closely as possible.

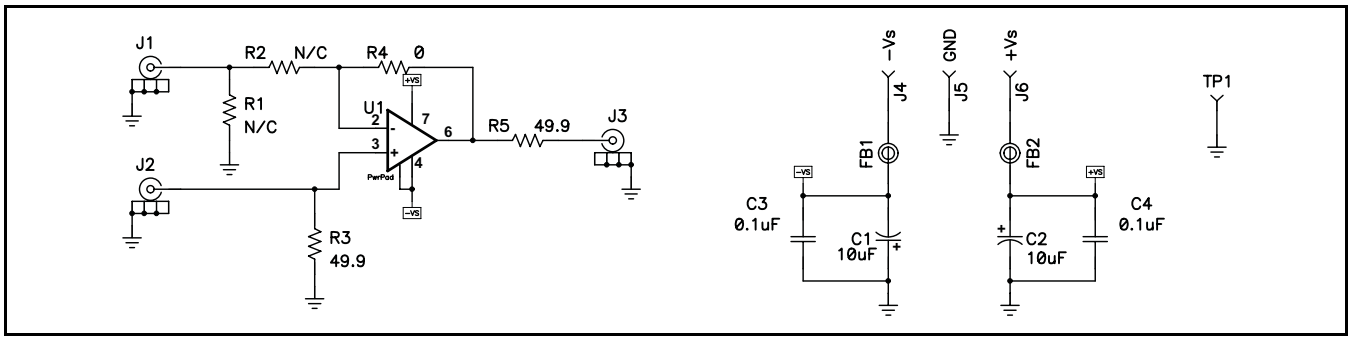


Figure 40. OPA659EVM Schematic

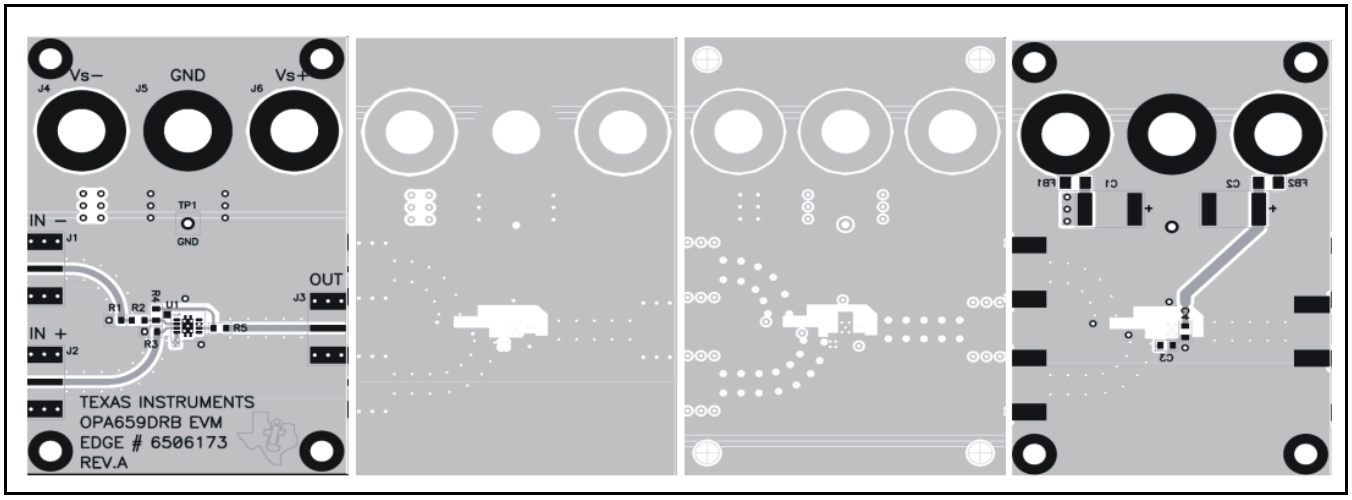


Figure 41. OPA659EVM Layers 1 through 4

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Bill of Materials

Table 4 lists the bill of material for the OPA659EVM as supplied from TI.

Table 4. OPA659EVM Parts List

ITEM	DESCRIPTION	SMD SIZE	REFERENCE DESIGNATOR	QUANTITY	MANUFACTURER PART NUMBER
1	Cap, 10.0µF, Tantalum, 10%, 35V	D	C1, C2	2	(AVX) TAJ106K035R
2	Cap, 0.1µF, Ceramic, X7R, 16V	0603	C3, C4	2	(AVX) 0603YC104KAT2A
3	Open	0603	R1, R2	2	
4	Resistor, 0Ω	0603	R4	1	(ROHM) MCR03EZPJ000
5	Resistor, 49.9Ω, 1/10W, 1%	0603	R3, R5	2	(ROHM) MCR03EZPFX49R9
6	Jack, Banana Receptance, 0.25in diameter hole		J4, J5, J8	3	(SPC) 813
7	Connector, Edge, SMA PCB Jack		J1, J2, J3	3	(JOHNSON) 142-0701-801
8	Test Point, Black		TP1	1	(KEYSTONE) 5001
9	IC, OPA659		U1	1	(TI) OPA659DRB
10	Standoff, 4-40 HEX, 0.625in length			4	(KEYSTONE) 1808
11	Screw, Phillips, 4-40, .250in			4	SHR-0440-016-SN
12	Board, Printed Circuit			1	(TI) EDGE# 6506173
13	Bead, Ferrite, 3A, 80Ω	1206	FB1, FB2	2	(STEWARD) HI1206N800R-00

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EVM WARNINGS AND RESTRICTIONS

It is important to operate this EVM within the input voltage range of ±3.5V to ±6.5V split-supply and the output voltage range of ±3.5V to ±6.5V power-supply voltage; do not exceed ±6.5V power-supply voltage.

Exceeding the specified input range may cause unexpected operation and/or irreversible damage to the EVM. If there are questions concerning the input range, please contact a TI field representative prior to connecting the input power.

Applying loads outside of the specified output range may result in unintended operation and/or possible permanent damage to the EVM. Please consult the EVM User's Guide prior to connecting any load to the EVM output. If there is uncertainty as to the load specification, please contact a TI field representative.

During normal operation, some circuit components may have case temperatures greater than +85°C. The EVM is designed to operate properly with certain components above +85°C as long as the input and output ranges are maintained. These components include but are not limited to linear regulators, switching transistors, pass transistors, and current sense resistors. These types of devices can be identified using the EVM schematic located in the EVM User's Guide. When placing measurement probes near these devices during operation, please be aware that these devices may be very warm to the touch.

Mailing Address: Texas Instruments, Post Office Box 655303, Dallas, Texas 75265
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Revision History

NOTE: Page numbers for previous revisions may differ from page numbers in the current version.

Changes from Revision A (March, 2009) to Revision B **Page**

- Removed footnote and table markings regarding SOT23-5 package availabilty ... 2
- Removed lead temperature specification from *Absolute Maximum Ratings* table ... 2
- Added *DRB package* to test condition for Input Offset Voltage parameter, T_A = –40°C to +85°C ... 3
- Added performance specifications for Input Offset Voltage parameter, DBV package ... 3
- Added performance specifications for Average Offset Voltage Drift parameter, DBV package ... 3
- Added footnote (2) to *Electrical Characteristics (V_S = ±6V)* table ... 3
- Added paragraph (f) to the *Board Layout* section ... 18
- Added *PowerPAD™ Information* section ... 18

Changes from Original (December, 2008) to Revision A **Page**

- Changed ordering information for SOT23-5 (DBV) package and added footnote; availability expected 2Q 2009 ... 2

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PACKAGING INFORMATION

Orderable Device	Status (1)	Package Type	Package Drawing	Pins	Package Qty	Eco Plan (2)	Lead/Ball Finish	MSL Peak Temp (3)	Op Temp (°C)	Top-Side Markings (4)	Samples
OPA659IDBVR	ACTIVE	SOT-23	DBV	5	3000	Green (RoHS & no Sb/Br)	CU NIPDAU	Level-2-260C-1 YEAR	-40 to 85	BZX	Samples
OPA659IDBVT	ACTIVE	SOT-23	DBV	5	250	Green (RoHS & no Sb/Br)	CU NIPDAU	Level-2-260C-1 YEAR	-40 to 85	BZX	Samples
OPA659IDRBR	ACTIVE	SON	DRB	8	3000	Green (RoHS & no Sb/Br)	CU NIPDAU	Level-2-260C-1 YEAR	-40 to 85	OBFI	Samples
OPA659IDRBT	ACTIVE	SON	DRB	8	250	Green (RoHS & no Sb/Br)	CU NIPDAU	Level-2-260C-1 YEAR	-40 to 85	OBFI	Samples

(1) The marketing status values are defined as follows:
ACTIVE: Product device recommended for new designs.
LIFEBUY: TI has announced that the device will be discontinued, and a lifetime-buy period is in effect.
NRND: Not recommended for new designs. Device is in production to support existing customers, but TI does not recommend using this part in a new design.
PREVIEW: Device has been announced but is not in production. Samples may or may not be available.
OBSOLETE: TI has discontinued the production of the device.

(2) Eco Plan - The planned eco-friendly classification: Pb-Free (RoHS), Pb-Free (RoHS Exempt), or Green (RoHS & no Sb/Br) - please check http://www.ti.com/productcontent for the latest availability information and additional product content details.
TBD: The Pb-Free/Green conversion plan has not been defined.
Pb-Free (RoHS): TI's terms "Lead-Free" or "Pb-Free" mean semiconductor products that are compatible with the current RoHS requirements for all 6 substances, including the requirement that lead not exceed 0.1% by weight in homogeneous materials. Where designed to be soldered at high temperatures, TI Pb-Free products are suitable for use in specified lead-free processes.
Pb-Free (RoHS Exempt): This component has a RoHS exemption for either 1) lead-based flip-chip solder bumps used between the die and package, or 2) lead-based die adhesive used between the die and leadframe. The component is otherwise considered Pb-Free (RoHS compatible) as defined above.
Green (RoHS & no Sb/Br): TI defines "Green" to mean Pb-Free (RoHS compatible), and free of Bromine (Br) and Antimony (Sb) based flame retardants (Br or Sb do not exceed 0.1% by weight in homogeneous material)

(3) MSL, Peak Temp. -- The Moisture Sensitivity Level rating according to the JEDEC industry standard classifications, and peak solder temperature.

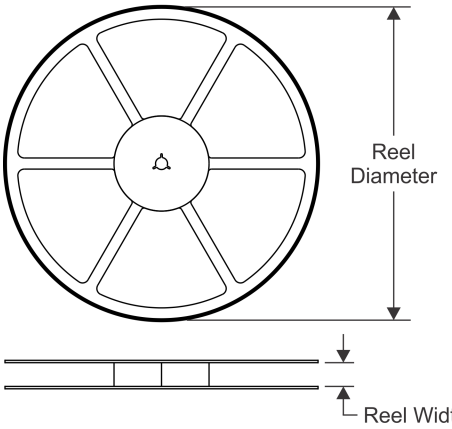
(4) Multiple Top-Side Markings will be inside parentheses. Only one Top-Side Marking contained in parentheses and separated by a "~" will appear on a device. If a line is indented then it is a continuation of the previous line and the two combined represent the entire Top-Side Marking for that device.

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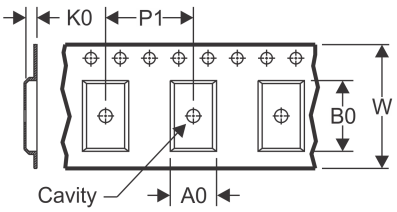
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TAPE AND REEL INFORMATION

REEL DIMENSIONS

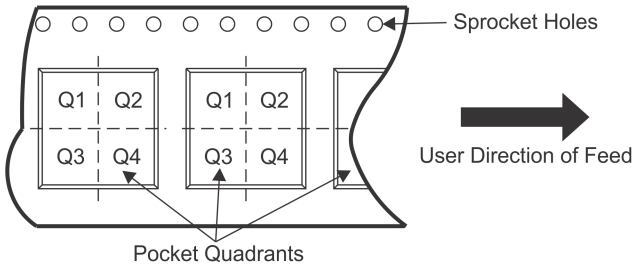


TAPE DIMENSIONS



A0	Dimension designed to accommodate the component width
B0	Dimension designed to accommodate the component length
K0	Dimension designed to accommodate the component thickness
W	Overall width of the carrier tape
P1	Pitch between successive cavity centers

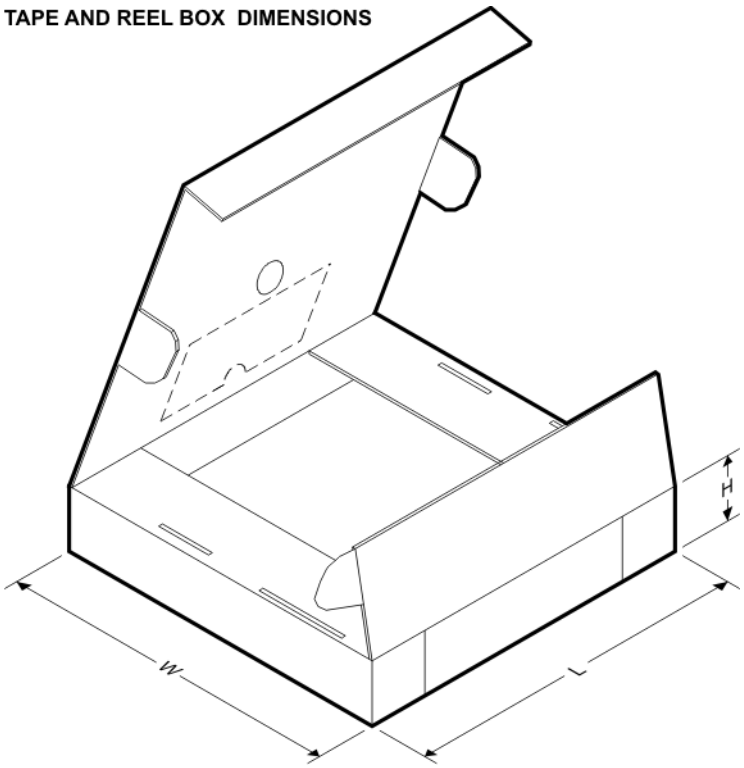
QUADRANT ASSIGNMENTS FOR PIN 1 ORIENTATION IN TAPE



*All dimensions are nominal

Device	Package Type	Package Drawing	Pins	SPQ	Reel Diameter (mm)	Reel Width W1 (mm)	A0 (mm)	B0 (mm)	K0 (mm)	P1 (mm)	W (mm)	Pin1 Quadrant
OPA659IDBVR	SOT-23	DBV	5	3000	180.0	8.4	3.2	3.1	1.39	4.0	8.0	Q3
OPA659IDBVT	SOT-23	DBV	5	250	180.0	8.4	3.2	3.1	1.39	4.0	8.0	Q3
OPA659IDRBR	SON	DRB	8	3000	330.0	12.4	3.3	3.3	1.1	8.0	12.0	Q2
OPA659IDRBT	SON	DRB	8	250	180.0	12.4	3.3	3.3	1.1	8.0	12.0	Q2

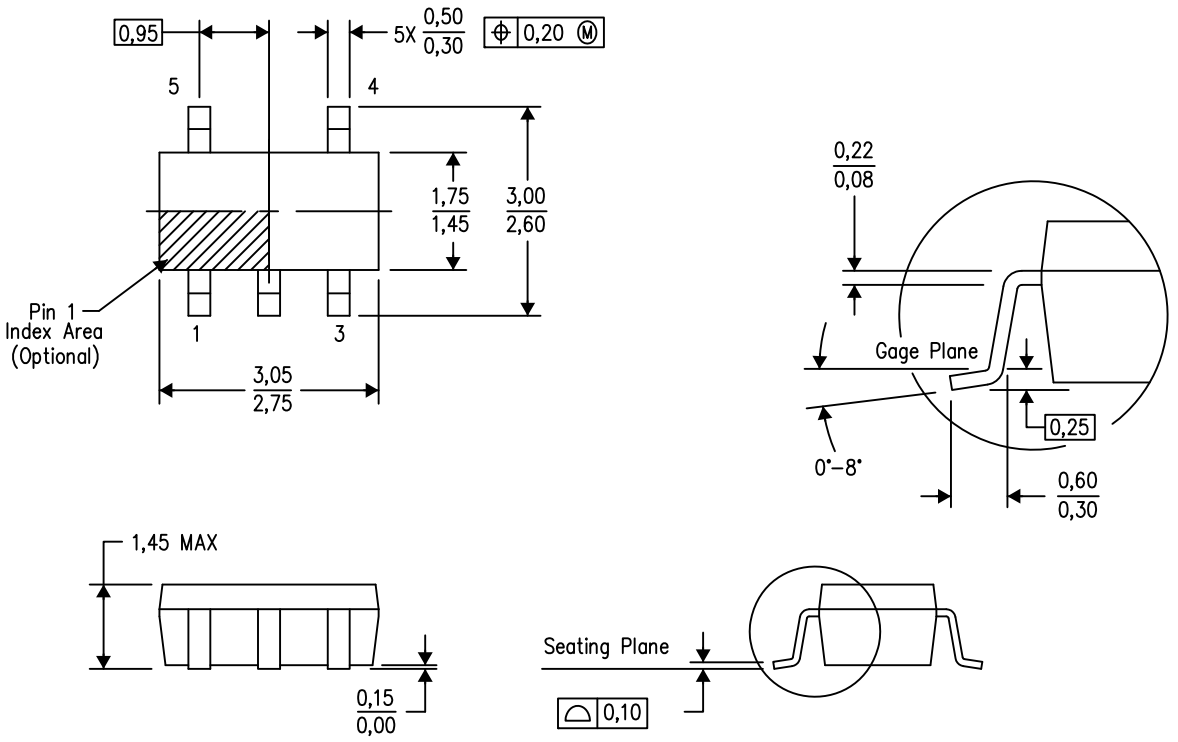
TAPE AND REEL BOX DIMENSIONS



*All dimensions are nominal

Device	Package Type	Package Drawing	Pins	SPQ	Length (mm)	Width (mm)	Height (mm)
OPA659IDBVR	SOT-23	DBV	5	3000	210.0	185.0	35.0
OPA659IDBVT	SOT-23	DBV	5	250	210.0	185.0	35.0
OPA659IDRBR	SON	DRB	8	3000	367.0	367.0	35.0
OPA659IDRBT	SON	DRB	8	250	210.0	185.0	35.0

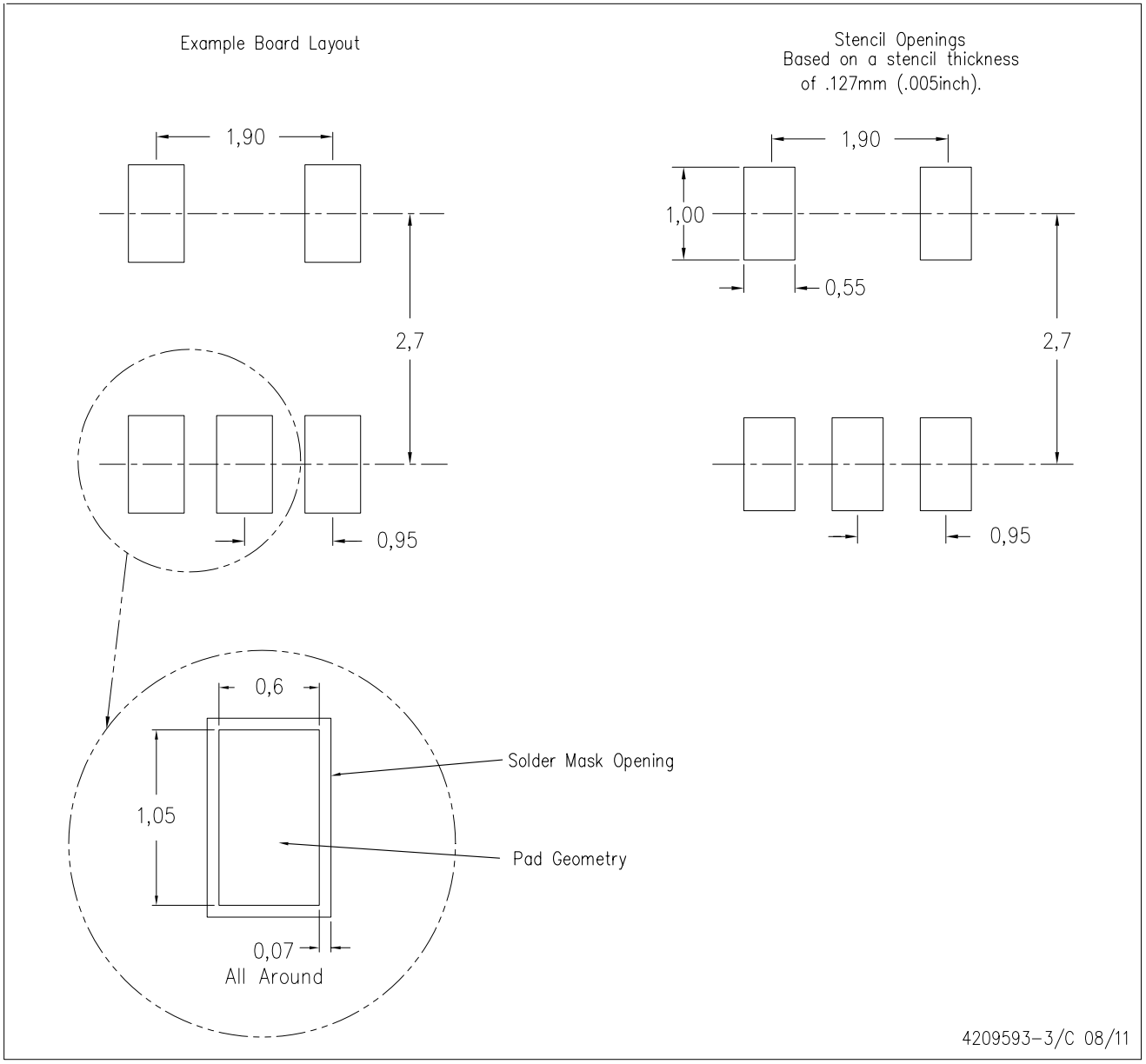
DBV (R–PDSO–G5) PLASTIC SMALL–OUTLINE PACKAGE



4073253–4/N 12/14

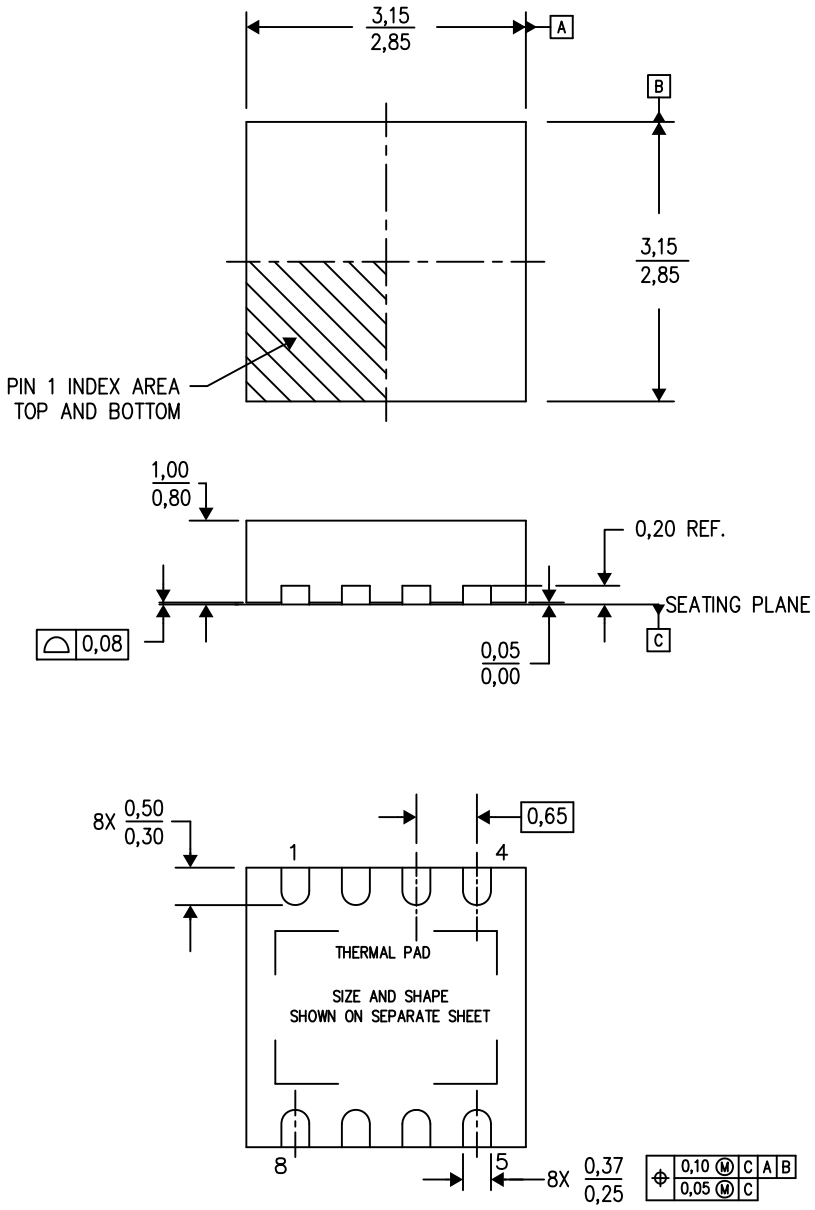
NOTES:
A. All linear dimensions are in millimeters.
B. This drawing is subject to change without notice.
C. Body dimensions do not include mold flash or protrusion. Mold flash and protrusion shall not exceed 0.15 per side.
D. Falls within JEDEC MO–178 Variation AA.

DBV (R–PDSO–G5) PLASTIC SMALL OUTLINE



NOTES:
A. All linear dimensions are in millimeters.
B. This drawing is subject to change without notice.
C. Customers should place a note on the circuit board fabrication drawing not to alter the center solder mask defined pad.
D. Publication IPC–7351 is recommended for alternate designs.
E. Laser cutting apertures with trapezoidal walls and also rounding corners will offer better paste release. Customers should contact their board assembly site for stencil design recommendations. Example stencil design based on a 50% volumetric metal load solder paste. Refer to IPC–7525 for other stencil recommendations.

DRB (S–PVSON–N8) PLASTIC SMALL OUTLINE NO–LEAD



4203482–2/K 06/12

NOTES:
A. All linear dimensions are in millimeters. Dimensioning and tolerancing per ASME Y14.5M–1994.
B. This drawing is subject to change without notice.
C. Small Outline No–Lead (SON) package configuration.
D. The package thermal pad must be soldered to the board for thermal and mechanical performance.
E. See the additional figure in the Product Data Sheet for details regarding the exposed thermal pad features and dimensions.

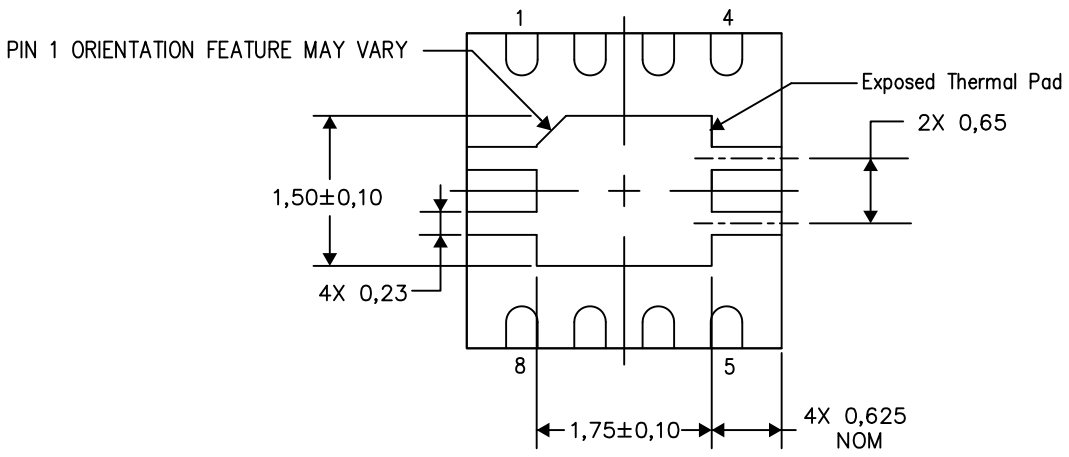
DRB (S–PVSON–N8) PLASTIC SMALL OUTLINE NO–LEAD

THERMAL INFORMATION

This package incorporates an exposed thermal pad that is designed to be attached directly to an external heatsink. The thermal pad must be soldered directly to the printed circuit board (PCB). After soldering, the PCB can be used as a heatsink. In addition, through the use of thermal vias, the thermal pad can be attached directly to the appropriate copper plane shown in the electrical schematic for the device, or alternatively, can be attached to a special heatsink structure designed into the PCB. This design optimizes the heat transfer from the integrated circuit (IC).

For information on the Quad Flatpack No–Lead (QFN) package and its advantages, refer to Application Report, QFN/SON PCB Attachment, Texas Instruments Literature No. SLUA271. This document is available at www.ti.com.

The exposed thermal pad dimensions for this package are shown in the following illustration.



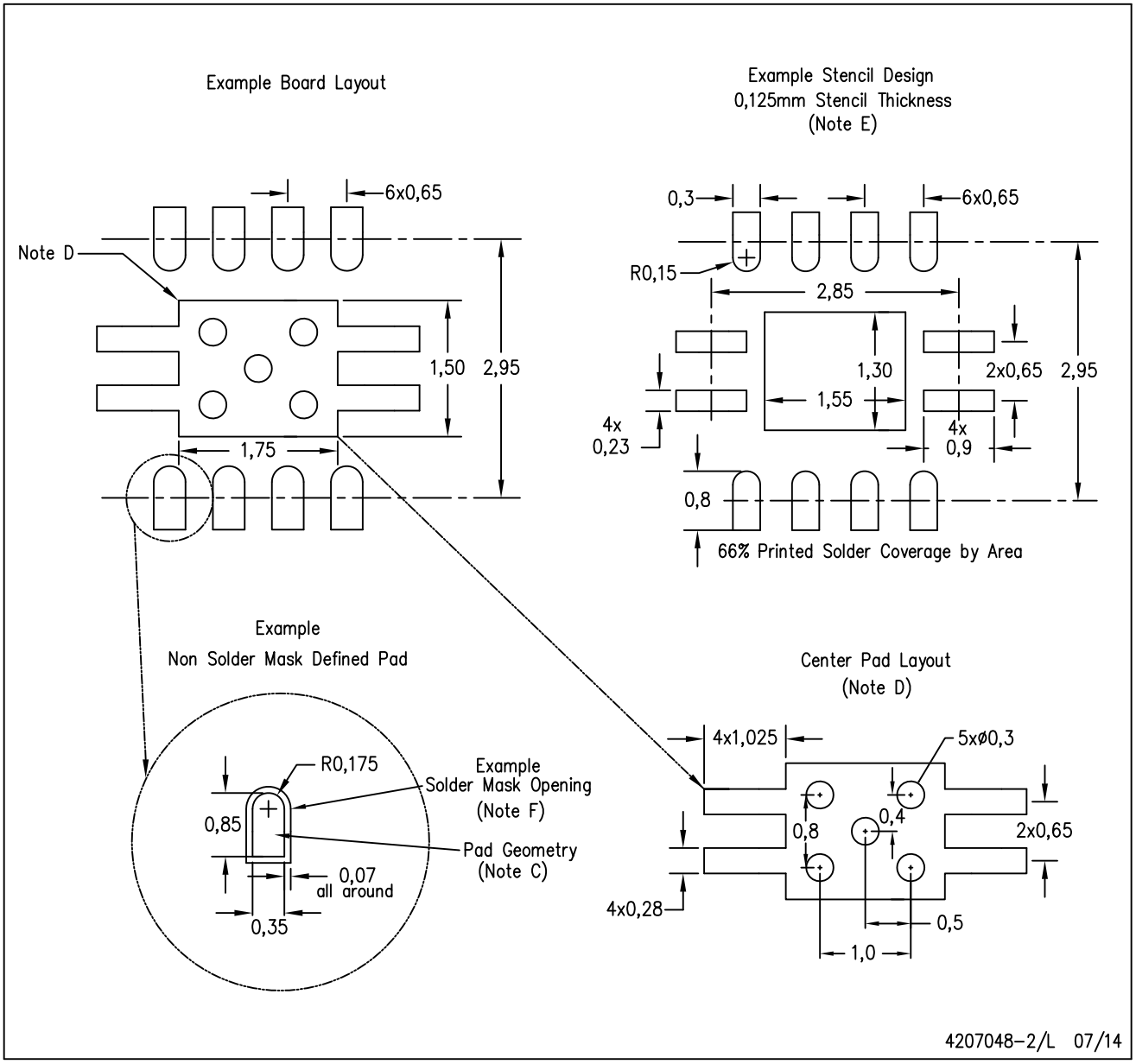
Bottom View

Exposed Thermal Pad Dimensions

4206340–2/P 07/14

NOTE: All linear dimensions are in millimeters

DRB (S–PVSON–N8) PLASTIC SMALL OUTLINE NO–LEAD



NOTES:
A. All linear dimensions are in millimeters.
B. This drawing is subject to change without notice.
C. Publication IPC–7351 is recommended for alternate designs.
D. This package is designed to be soldered to a thermal pad on the board. Refer to Application Note, QFN Packages, Texas Instruments Literature No. SLUA271, and also the Product Data Sheets for specific thermal information, via requirements, and recommended board layout. These documents are available at www.ti.com <http://www.ti.com>.
E. Laser cutting apertures with trapezoidal walls and also rounding corners will offer better paste release. Customers should contact their board assembly site for stencil design recommendations. Refer to IPC 7525 for stencil design considerations.
F. Customers should contact their board fabrication site for solder mask tolerances.

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DLP® Products	www.dlp.com	Consumer Electronics	www.ti.com/consumer-apps
DSP	dsp.ti.com	Energy and Lighting	www.ti.com/energy
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Interface	interface.ti.com	Medical	www.ti.com/medical
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Power Mgmt	power.ti.com	Space, Avionics and Defense	www.ti.com/space-avionics-defense
Microcontrollers	microcontroller.ti.com	Video and Imaging	www.ti.com/video
RFID	www.ti-rfid.com		
OMAP Applications Processors	www.ti.com/omap	**TI E2E Community**	e2e.ti.com
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