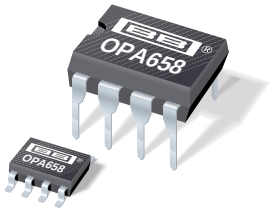
BURR - BROWN®



OPA658

Wideband, Low Power Current Feedback OPERATIONAL AMPLIFIER

FEATURES

- UNITY GAIN STABLE BANDWIDTH: 900MHz
- LOW POWER: 50mW
- LOW DIFFERENTIAL GAIN/PHASE ERRORS: 0.025%/0.02°
- HIGH SLEW RATE: 1700V/µs
- GAIN FLATNESS: 0.1dB to 135MHz
- HIGH OUTPUT CURRENT (80mA)

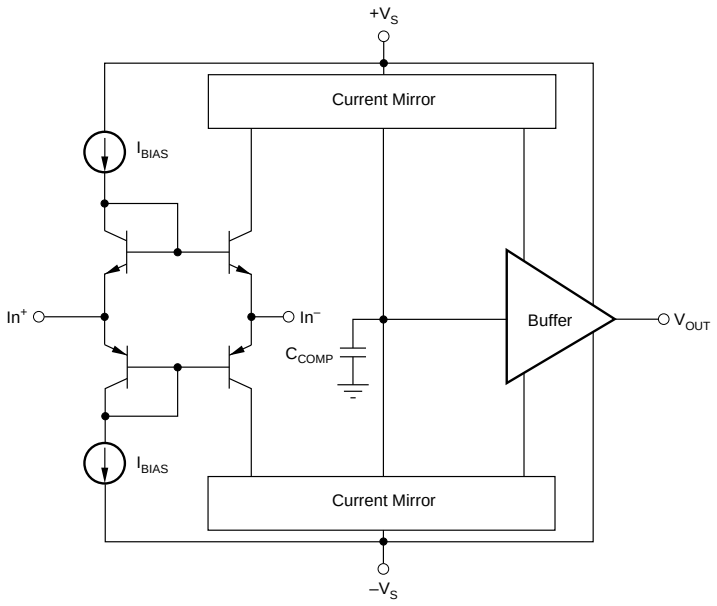
APPLICATIONS

- MEDICAL IMAGING
- HIGH-RESOLUTION VIDEO
- HIGH-SPEED SIGNAL PROCESSING
- COMMUNICATIONS
- PULSE AMPLIFIERS
- ADC/DAC GAIN AMPLIFIER
- MONITOR PREAMPLIFIER
- CCD IMAGING AMPLIFIER

DESCRIPTION

The OPA658 is an ultra-wideband, low power current feedback video operational amplifier featuring high slew rate and low differential gain/phase error. The current feedback design allows for superior large signal bandwidth, even at high gains. The low differential gain/phase errors, wide bandwidth and low quiescent current make the OPA658 a perfect choice for numerous video, imaging and communications applications.

The OPA658 is optimized for low gain operation and is also available in dual (OPA2658) and quad (OPA4658) configurations.



International Airport Industrial Park • Mailing Address: PO Box 11400 • Tucson, AZ 85734 • Street Address: 6730 S. Tucson Blvd. • Tucson, AZ 85706
Tel: (520) 746-1111 • Twx: 910-952-1111 • Cable: BBRCORP • Telex: 066-6491 • FAX: (520) 889-1510 • Immediate Product Info: (800) 548-6132

PDS-1268B
Printed in U.S.A. September, 1995

SPECIFICATIONS

T_A = +25°C, V_S = ±5V, R_L = 100Ω, and R_{FB} = 402Ω unless otherwise noted.

PARAMETER	CONDITION	OPA658P, U			OPA658PB, UB			UNITS
		MIN	TYP	MAX	MIN	TYP	MAX	
FREQUENCY RESPONSE								
Closed-Loop Bandwidth[2]	G = +1[4]		900			*[1]		MHz
	G = +2		680		400	*		MHz
	G = +5		370			*		MHz
	G = +10		200			*		MHz
Slew Rate[3]	G = +2, 2V Step		1700		1000	*		V/µs
At Minimum Specified Temperature			1500		900	*		V/µs
Settling Time: 0.01%	G = +2, 2V Step		15			*		ns
0.1%	G = +2, 2V Step		11.5			*		ns
1%	G = +2, 2V Step		6			*		ns
Spurious Free Dynamic Range	f = 5MHz, G = +2, V_O = 2Vp-p		68			*		dBc
	f = 20MHz, G= +2, V_O = 2Vp-p		56			*		dBc
Third Order Intercept Point	f = 10MHz, 4dBm Each Tone		40			*		dBm
Differential Gain	G = +2, NTSC, V_O = 1.4Vp-p, R_L = 150Ω		0.025			*		%
Differential Phase	G = +2, NTSC, V_O = 1.4Vp-p, R_L = 150Ω		0.02			*		degrees
Bandwidth for 0.1dB Flatness	G = +2		135[5]			*		MHz
OFFSET VOLTAGE								
Input Offset Voltage	V_{CM} = 0V		±3	±5.5		±2	±4.5	mV
Over Temperature Range			±5	±8		±4	±7	mV
Power Supply Rejection Ratio	V_S = ±4.7 to ±5.5V	55	64		58	67		dB
INPUT BIAS CURRENT								
Non-Inverting	V_{CM} = 0V		±5.7	±30		*	±18	µA
Over Temperature Range			±10	±80		*	±35	µA
Inverting	V_{CM} = 0V		±1.1	±35		*	*	µA
Over Temperature Range			±30	±75		*	*	µA
NOISE								
Input Voltage Noise Density								
f = 100Hz			11			*		nV/√Hz
f = 2kHz			3.3			*		nV/√Hz
f = 10kHz			2.7			*		nV/√Hz
f = 1MHz			2.7			*		nV/√Hz
f_B = 100Hz to 200MHz			38.2			*		µVrms
Input Bias Current Noise Density								
Inverting: f = 10MHz			12.6			*		pA/√Hz
Non-Inverting: f = 10MHz			12.6			*		pA/√Hz
Noise Figure (NF)	R_S = 50Ω		10			*		dB
	R_S = 100Ω		8			*		dB
INPUT VOLTAGE RANGE								
Common-Mode Input Range								
Over Temperature Range		±2.5	±2.9		*	*		V
Common-Mode Rejection	V_{CM} = ±1V	45	50		*	*		dB
INPUT IMPEDANCE								
Non-Inverting			500 || 1			*		kΩ || pF
Inverting			50			*		Ω
OPEN-LOOP TRANSRESISTANCE								
Open-Loop Transresistance	V_O = ±2V, R_L = 100Ω	150	190		200	250		kΩ
Over Temperature Range	V_O = ±2V, R_L = 100Ω	100			150			kΩ
OUTPUT								
Voltage Output	No Load	±2.7	±2.9		*	*		V
Over Temperature Range		±2.5	±2.75		*	*		V
Voltage Output	R_L = 250Ω	±2.7	±2.9		*	*		V
Over Temperature Range		±2.5	±2.7		*	*		V
Voltage Output	R_L = 100Ω	±2.2	±2.8		*	*		V
Over Temperature Range		±2.0	±2.5		*	*		V
Output Current, Sourcing		80	120		*	*		mA
Over Temperature		70			*			mA
Output Current, Sinking		60	80		*	*		mA
Over Temperature		35			*			mA
Short Circuit Current			150			*		mA
Output Resistance	0.1MHz, G = +2		0.02			*		Ω
POWER SUPPLY								
Specified Operating Voltage			±5			*		V
Operating Voltage Range		±4.5		±5.5	*		*	V
Quiescent Current	V_S = ±5V		±5	±7.75		±4.5	±5.75	mA
Over Temperature Range			±5.5	±8.5		±4.7	±6.5	mA
TEMPERATURE RANGE								
Specification: P, U, PB, UB		–40		+85	*		*	°C
Thermal Resistance, θ_{JA}								
P			100			*		°C/W
U			125			*		°C/W

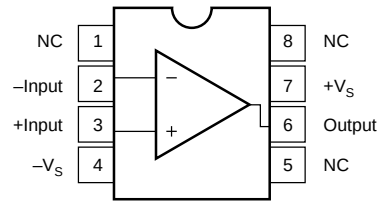
NOTES: (1) An asterisk (*) specifies the same value as the grade to the left. (2) Frequency response can be strongly influenced by PC board parasitics. The demonstration board, DEM-OPA65X shows a low parasitic layout for this part. Refer to the demonstration board layout for details. (3) Slew rate is rate of change from 10% to 90% of output voltage step. (4) At G = +1, R_{FB} = 560Ω for PDIP and 402Ω for SOIC. (5) This specification is PC board layout dependent.

ABSOLUTE MAXIMUM RATINGS

Supply	±5.5V
Internal Power Dissipation	See Thermal Considerations
Differential Input Voltage	±1.2V
Input Voltage Range	$\pm V_S$
Storage Temperature Range: P, PB, U, UB	–40°C to +125°C
Lead Temperature (soldering, 10s)	+300°C
(soldering, SOIC 3s)	+260°C
Junction Temperature (T_J)	+175°C

PIN CONFIGURATION

Top View **DIP/SO-8**



PACKAGE INFORMATION

MODEL	PACKAGE	PACKAGE DRAWING NUMBER(1)
OPA658P, PB	8-Pin Plastic DIP	006
OPA658U, UB	SO-8 Surface Mount	182

NOTE: (1) For detailed drawing and dimension table, please see end of data sheet, or Appendix D of Burr-Brown IC Data Book.

ORDERING INFORMATION(1)

MODEL	PACKAGE	TEMPERATURE RANGE
OPA658P, PB	8-Pin Plastic DIP	–40°C to +85°C
OPA658U, UB	SO-8 Surface Mount	–40°C to +85°C

NOTE: (1) The "B" grade of the SOIC package will be marked with a "B" by pin 8. Refer to mechanical section for the location.

ELECTROSTATIC DISCHARGE SENSITIVITY

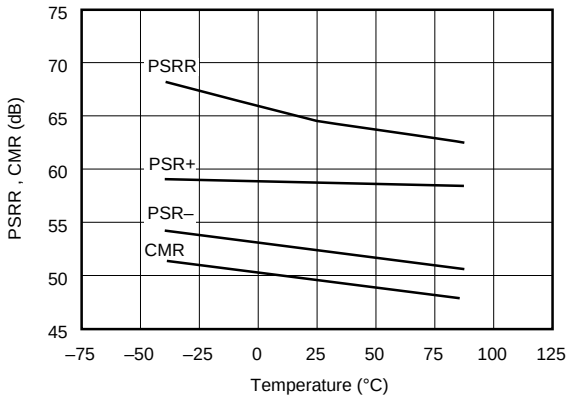
Electrostatic discharge can cause damage ranging from performance degradation to complete device failure. Burr-Brown Corporation recommends that all integrated circuits be handled and stored using appropriate ESD protection methods.

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 published specifications.

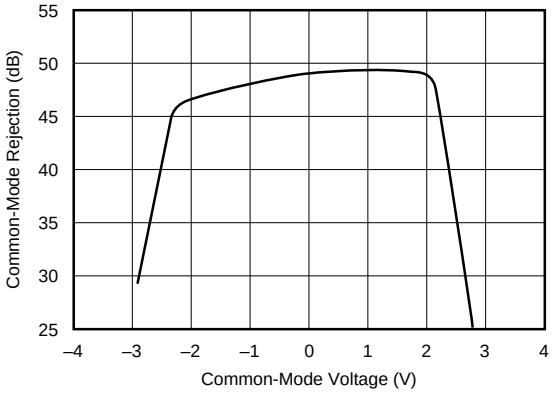
TYPICAL PERFORMANCE CURVES

T_A = +25°C, V_S = ±5V, R_L = 100Ω, and R_{FB} = 402Ω unless otherwise noted.

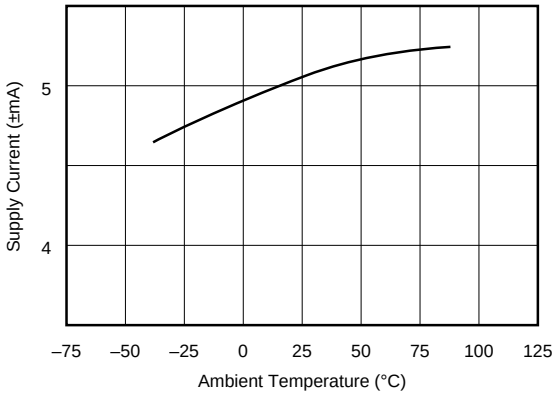
PSRR AND CMR vs TEMPERATURE



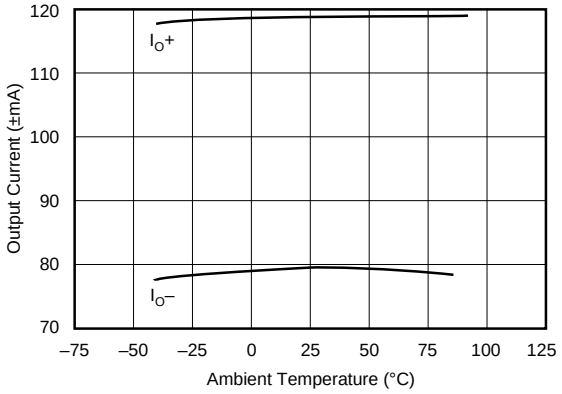
COMMON-MODE REJECTION
vs INPUT COMMON-MODE VOLTAGE



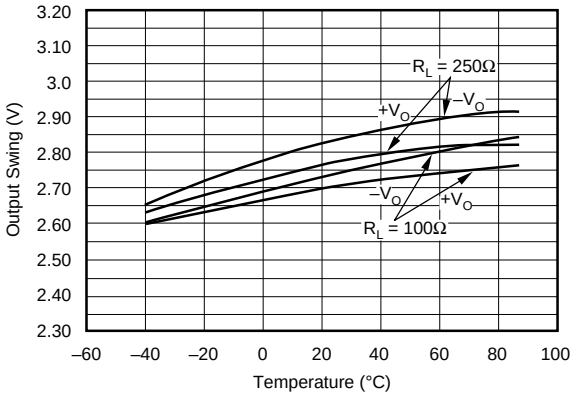
SUPPLY CURRENT vs TEMPERATURE



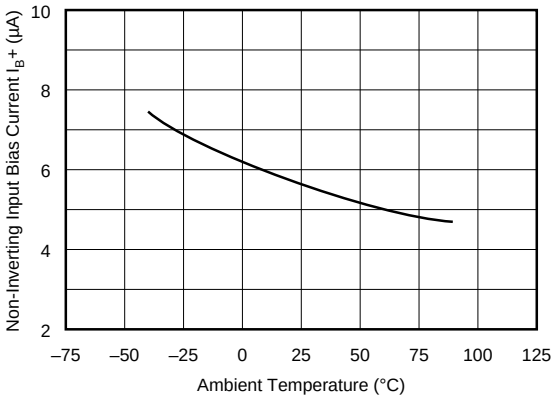
OUTPUT CURRENT vs TEMPERATURE



OUTPUT SWING vs TEMPERATURE



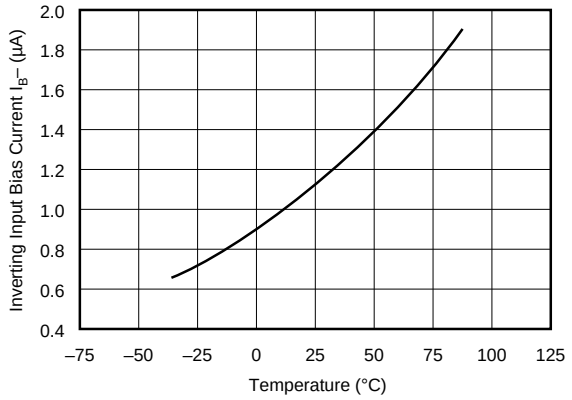
NON-INVERTING INPUT BIAS CURRENT
vs TEMPERATURE



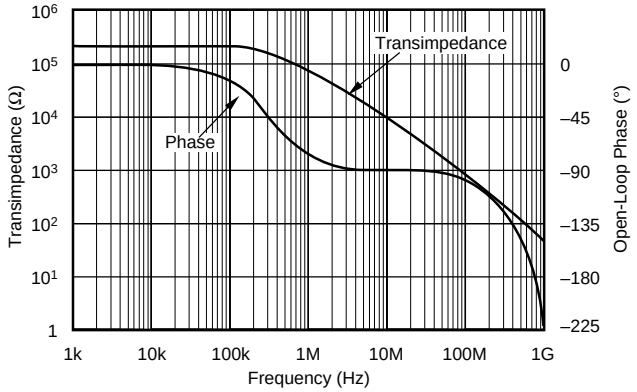
TYPICAL PERFORMANCE CURVES (CONT)

T_A = +25°C, V_S = ±5V, R_L = 100Ω, and R_{FB} = 402Ω unless otherwise noted.

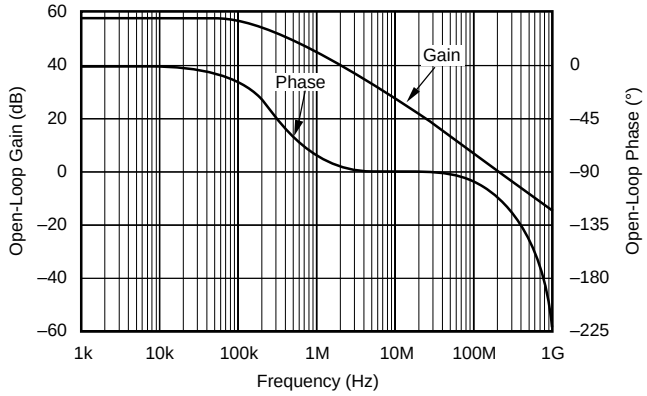
INVERTING INPUT BIAS CURRENT vs TEMPERATURE



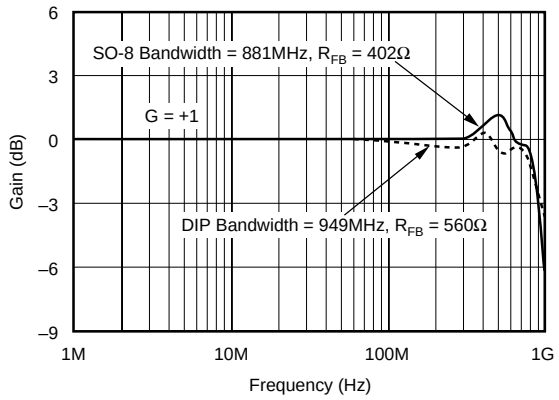
OPEN-LOOP TRANSIMPEDANCE AND PHASE vs FREQUENCY



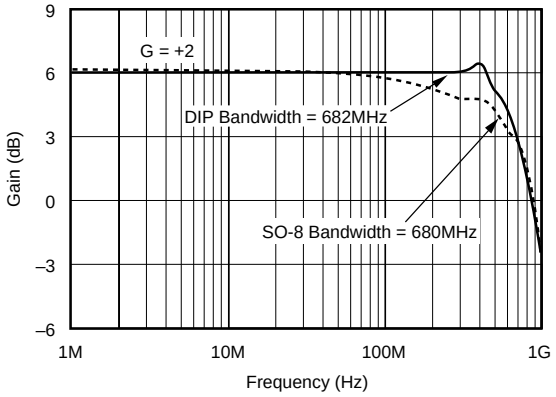
OPEN-LOOP GAIN AND PHASE vs FREQUENCY



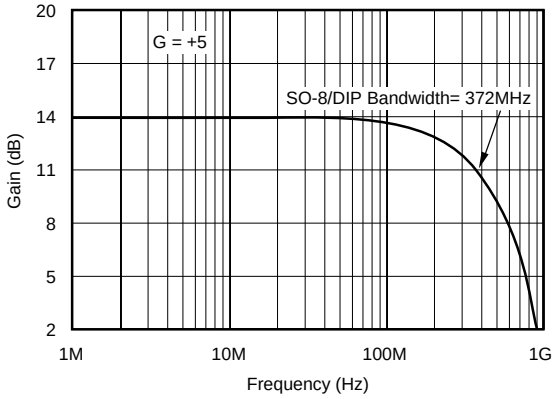
CLOSED-LOOP BANDWIDTH



CLOSED-LOOP BANDWIDTH



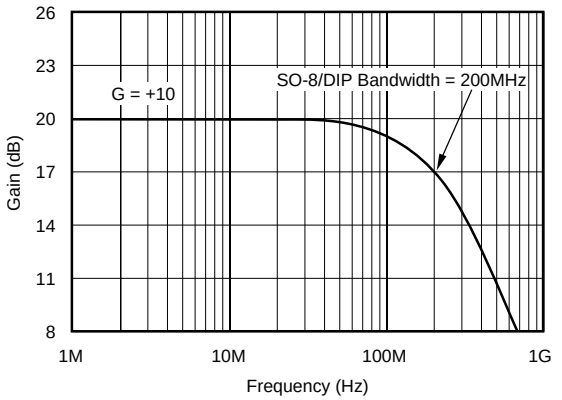
CLOSED-LOOP BANDWIDTH



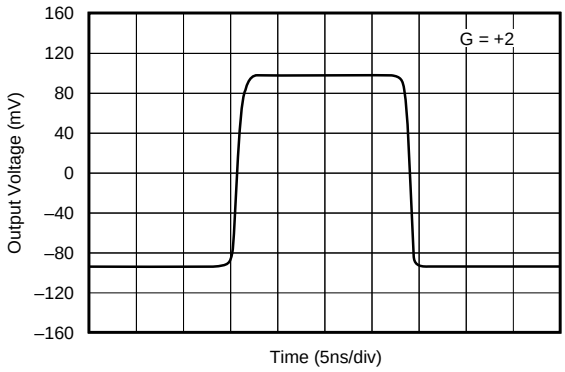
TYPICAL PERFORMANCE CURVES (CONT)

T_A = +25°C, V_S = ±5V, R_L = 100Ω, and R_{FB} = 402Ω unless otherwise noted.

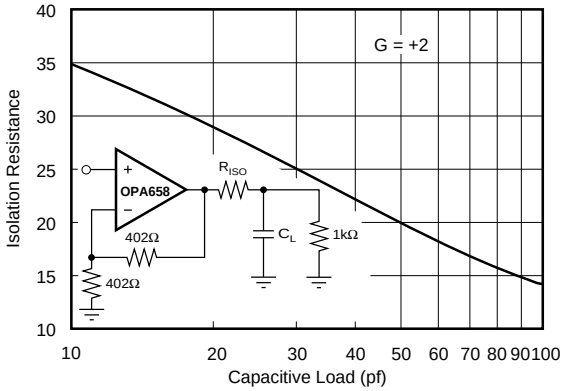
CLOSED-LOOP BANDWIDTH



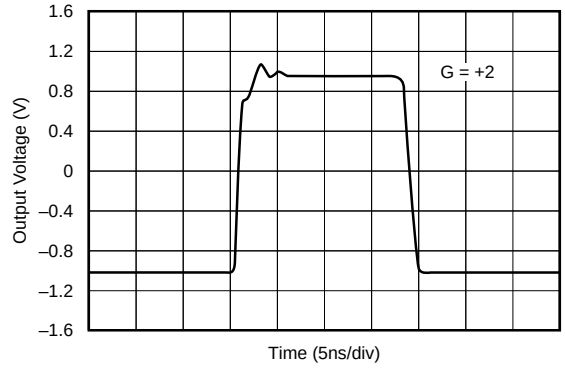
SMALL SIGNAL TRANSIENT RESPONSE



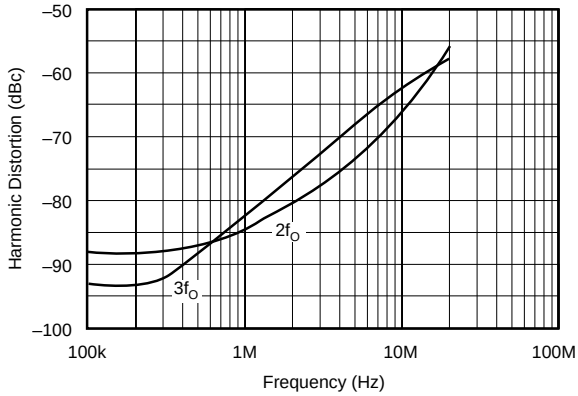
RECOMMENDED ISOLATION RESISTANCE vs CAPACITIVE LOAD



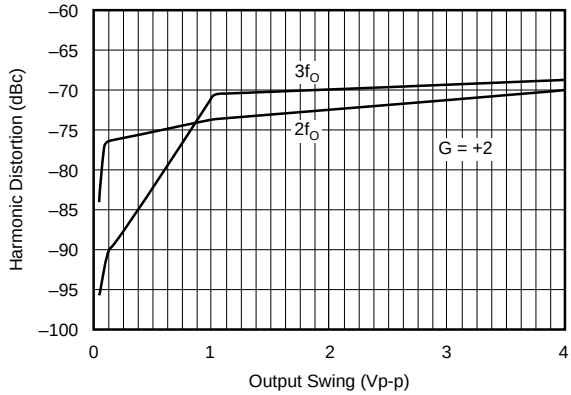
LARGE SIGNAL TRANSIENT RESPONSE



HARMONIC DISTORTION vs FREQUENCY



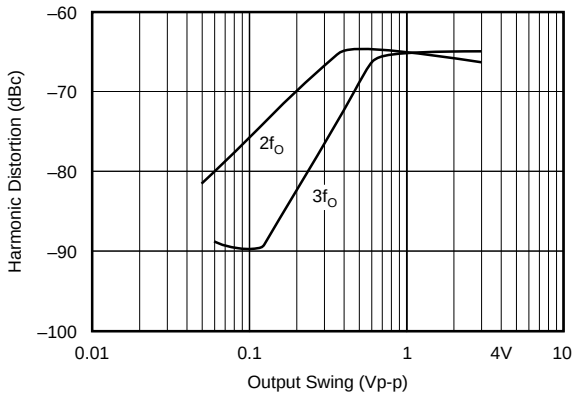
5MHz HARMONIC DISTORTION vs OUTPUT SWING



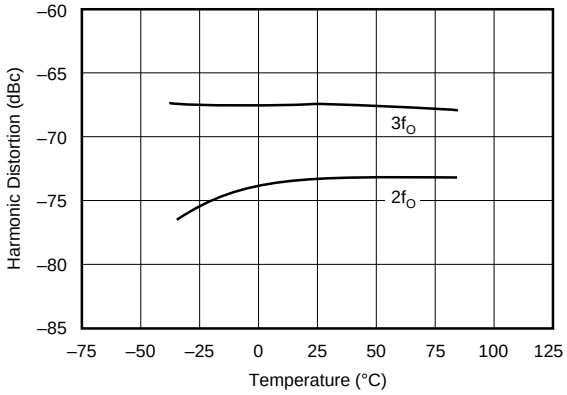
TYPICAL PERFORMANCE CURVES (CONT)

T_A = +25°C, V_S = ±5V, R_L = 100Ω, and R_{FB} = 402Ω unless otherwise noted.

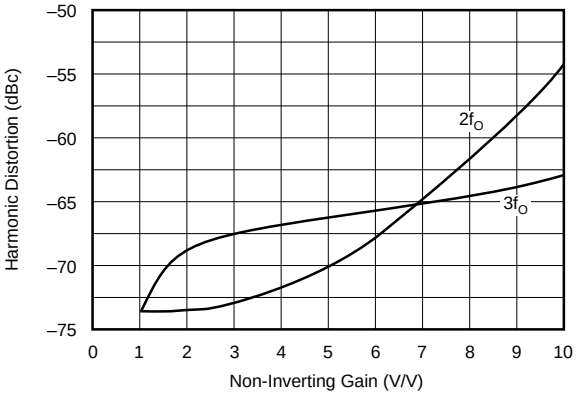
10MHz HARMONIC DISTORTION vs OUTPUT SWING



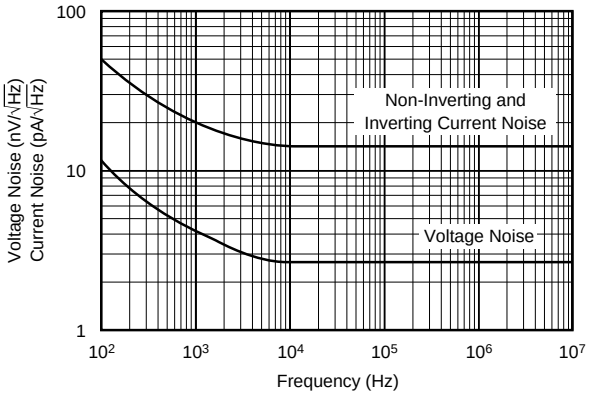
HARMONIC DISTORTION vs TEMPERATURE
(V_O = 2Vp-p, G = +2)



HARMONIC DISTORTION vs GAIN
(f_O = 5MHz, V_O = 2Vp-p)



INPUT VOLTAGE AND CURRENT NOISE
vs FREQUENCY



APPLICATIONS INFORMATION

THEORY OF OPERATION

Conventional op amps depend on feedback to drive their inputs to the same potential, however the current feedback op amp's inverting and non-inverting inputs are connected by a unity gain buffer, thus enabling the inverting input to automatically assume the same potential as the non-inverting input. This results in very low impedance at the inverting input to sense the feedback as an error current signal.

DISCUSSION OF PERFORMANCE

The OPA658 is a low-power, unity gain stable, current feedback operational amplifier which operates on ±5V power supply. The current feedback architecture offers the following important advantages over voltage feedback architectures: (1) the high slew rate allows the large signal performance to approach the small signal performance, and (2) there is very little bandwidth degradation at higher gain settings.

The current feedback architecture of the OPA658 provides the traditional strength of excellent large signal response plus wide bandwidth, making it a good choice for use in high resolution video, medical imaging and DAC I/V Conversion. The low power requirements make it an excellent choice for numerous portable applications.

DC GAIN TRANSFER CHARACTERISTICS

The circuit in Figure 1 shows the equivalent circuit for calculating the DC gain. When operating the device in the inverting mode, the input signal error current (I_E) is amplified by the open loop transimpedance gain (T_O). The output signal generated is equal to T_O x I_E. Negative feedback is applied through R_{FB} such that the device operates at a gain equal to $-R_{FB}/R_{FF}$.

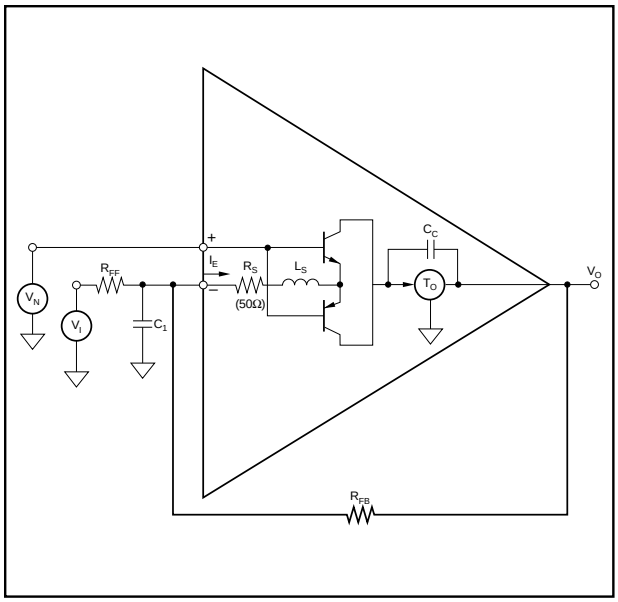


FIGURE 1. Equivalent Circuit.

For non-inverting operation, the input signal is applied to the non-inverting (high impedance buffer) input. The output (buffer) error current (I_E) is generated at the low impedance inverting input. The signal generated at the output is fed back to the inverting input such that the overall gain is $(1 + R_{FB}/R_{FF})$. Where a voltage-feedback amplifier has two symmetrical high impedance inputs, a current feedback amplifier has a low inverting (buffer output) impedance and a high non-inverting (buffer input) impedance.

The closed-loop gain for the OPA658 can be calculated using the following equations:

$$\text{Inverting Gain} = \frac{-\left(\frac{R_{FB}}{R_{FF}}\right)}{1+\frac{1}{\text{Loop Gain}}} \quad (1)$$

$$\text{Non–Inverting Gain} = \frac{\left[1+\frac{R_{FB}}{R_{FF}}\right]}{1+\frac{1}{\text{Loop Gain}}} \quad (2)$$

$$\text{where Loop Gain} = \left[\frac{T_O}{R_{FB}+R_S\left(1+\frac{R_{FB}}{R_{FF}}\right)}\right]$$

At higher gains the small value inverting input impedance causes an apparent loss in bandwidth. This can be seen from the equation:

$$f_{\text{ACTUAL}}\ \text{BW} \approx \frac{\left[f_{(A_V=+2)}\ \text{BW}\right] \text{x}\,(1.25)}{\left[1+\left(\frac{R_S}{R_{FB}}\right)\times\left(1+\frac{R_{FB}}{R_{FF}}\right)\right]} \quad (3)$$

This loss in bandwidth at high gains can be corrected without affecting stability by lowering the value of the feedback resistor from the specified value of 402Ω.

OFFSET VOLTAGE AND NOISE

The output offset is the algebraic sum of the input offset voltage and bias current errors. The output offset for non-inverting operation is calculated by the following equation:

$$\text{Output Offset Voltage} = \pm Ib_N \times R_N\left(1+\frac{R_{FB}}{R_{FF}}\right)\pm V_{IO}\left(1+\frac{R_{FB}}{R_{FF}}\right)\pm Ib_I \times R_{FB} \quad (4)$$

If all terms are divided by the gain $(1 + R_{FB}/R_{FF})$ it can be observed that input referred offsets improve as gain increases. The effective noise at the output can be determined by taking

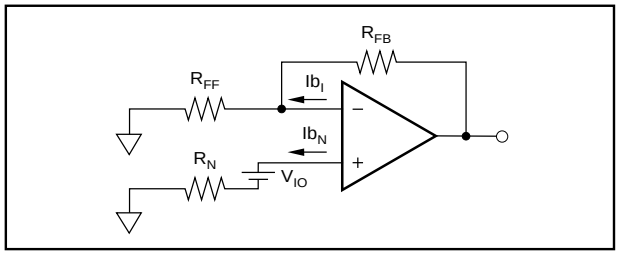


FIGURE 2. Output Offset Voltage Equivalent Circuit.

the root sum of the squares of equation (4) and applying the spectral noise values found in the Typical Performance Curve graph section. This applies to noise from the op amp only. Note that both the noise figure (NF) and the equivalent input offset voltages improve as the closed loop gain increases (by keeping R_{FB} fixed and reducing R_{FF} with $R_N = 0\Omega$).

INCREASING BANDWIDTH AT HIGH GAINS

The closed-loop bandwidth can be extended at high gains by reducing the value of the feedback resistor R_{FB}. This bandwidth reduction is caused by the feedback current being split between R_S and R_{FF} (refer to Figure 1). As the gain increases (for a fixed R_{FB}), more feedback current is shunted through R_{FF}, which reduces closed-loop bandwidth.

CIRCUIT LAYOUT AND BASIC OPERATION

Achieving optimum performance with a high frequency amplifier like the OPA658 requires careful attention to layout parasitics and selection of external components. Recommendations for PC board layout and component selection include:

a) Minimize parasitic capacitance to any ac ground for all of the signal 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 bandlimiting. To reduce unwanted capacitance, a window around the signal I/O pins should be opened in all of the ground and power planes. Otherwise, ground and power planes should be unbroken elsewhere on the board.

b) Minimize the distance (< 0.25") from the two power pins to high frequency 0.1µF decoupling capacitors. At the pins, the ground and power plane layout should not be in close proximity to the signal I/O pins. Avoid narrow power and ground traces to minimize inductance between the pins and the decoupling capacitors. Larger (2.2µF to 6.8µF) decoupling capacitors, effective at lower frequencies, should also be used. These may be placed somewhat farther from the device and may be shared among several devices in the same area of the PC board.

c) Careful selection and placement of external components will preserve the high frequency performance of the OPA658. Resistors should be a very low reactance type. Surface mount resistors work best and allow a tighter overall layout. Metal film or carbon composition axially-leaded resistors can also provide good high frequency performance. Again, keep their leads as short as possible. Never use wirewound type resistors in a high frequency application.

Since the output pin and the inverting input pin are most sensitive to parasitic capacitance, always position the feedback and series output resistor, if any, as close as possible to the package pins. Other network components, such as noninverting input termination resistors, should also be placed close to the package.

The feedback resistor value acts as the frequency response compensation element for a current feedback type amplifier. The 402Ω used in setting the specification achieves a nominal maximally flat butterworth response while assuming a 2pF output pin parasitic. Increasing the feedback resistor will over compensate the amplifier, rolling off the frequency response, while decreasing it will decrease phase margin, peaking up the frequency response. Note that a non-inverting, unity gain buffer application still requires a feedback resistor for stability (560Ω for SO-8 and 402Ω for PDIP).

d) Connections to other wideband devices on the board may be made with short direct traces or through on-board transmission lines. For short connections, consider the trace and the input to the next device as a lumped capacitive load. Relatively wide traces (50 to 100 mils) should be used, preferably with ground and power planes opened up around them. Estimate the total capacitive load and set R_{ISO} from the plot of recommended R_{ISO} vs capacitive load. Low parasitic loads may not need an R_{ISO} since the OPA658 is nominally compensated to operate with a 2pF parasitic load.

If a long trace is required and the 6dB signal loss intrinsic to doubly terminated transmission lines 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 not necessary on board, and in fact a higher impedance environment will improve distortion as shown in the distortion vs load plot. With a characteristic impedance defined based on board material and desired trace dimensions, a matching series resistor into the trace from the output of the amplifier is used as well as a terminating shunt resistor at the input of the destination device. Remember also that the terminating impedance will be the parallel combination of the shunt resistor and the input impedance of the destination device; the total effective impedance should match the trace impedance. Multiple destination devices are best handled as separate transmission lines, each with their own series and shunt terminations.

If the 6dB attenuation loss of a doubly terminated line is unacceptable, a long trace can be series-terminated at the source end only. This will help isolate the line capacitance from the op amp output, but will not preserve signal integrity as well as a doubly terminated line. If the shunt impedance at the destination end is finite, there will be some signal attenuation due to the voltage divider formed by the series and shunt impedances.

e) Socketing a high speed part like the OPA658 is not recommended. The additional lead length and pin-to-pin capacitance introduced by the socket creates an extremely troublesome parasitic network which can make it almost impossible to achieve a smooth, stable response. Best results are obtained by soldering the part onto the board. If socketing for the DIP package is desired, high frequency flush mount pins (e.g., McKenzie Technology #710C) can give good results.

The OPA658 is nominally specified for operation using ±5V power supplies. A 10% tolerance on the supplies, or an ECL –5.2V for the negative supply, is within the maximum

specified total supply voltage of 11V. Higher supply voltages can break down internal junctions possibly leading to catastrophic failure. Single supply operation is possible as long as common mode voltage constraints are observed. The common mode input and output voltage specifications can be interpreted as a required headroom to the supply voltage. Observing this input and output headroom requirement will allow non-standard or single supply operation. Figure 3 shows one approach to single-supply operation.

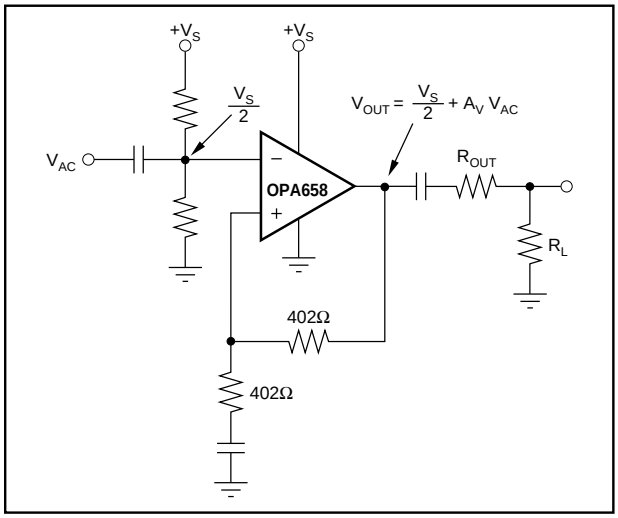


FIGURE 3. Single Supply Operation.

ESD PROTECTION

ESD static damage has been well recognized for MOSFET devices, but any semiconductor device deserves protection from this potentially damaging source. This is particularly true for very high speed, fine geometry processes.

ESD static damage can cause subtle changes in amplifier input characteristics without necessarily destroying the device. In precision operational amplifiers, this may cause a noticeable degradation of offset voltage and drift. Therefore, static protection is strongly recommended when handling the OPA658.

OUTPUT DRIVE CAPABILITY

The OPA658 has been optimized to drive 75Ω and 100Ω resistive loads. The device can drive 2Vp-p into a 75Ω load. This high-output drive capability makes the OPA658 an ideal choice for a wide range of RF, IF, and video applications. In many cases, additional buffer amplifiers are unneeded.

Many demanding high-speed applications such as ADC/DAC buffers require op amps with low wideband output impedance. For example, low output impedance is essential when driving the signal-dependent capacitances at the inputs of flash A/D converters. As shown in Figure 4, the OPA658 maintains very low closed-loop output impedance over frequency. Closed-loop output impedance increases with frequency since loop gain is decreasing with frequency.

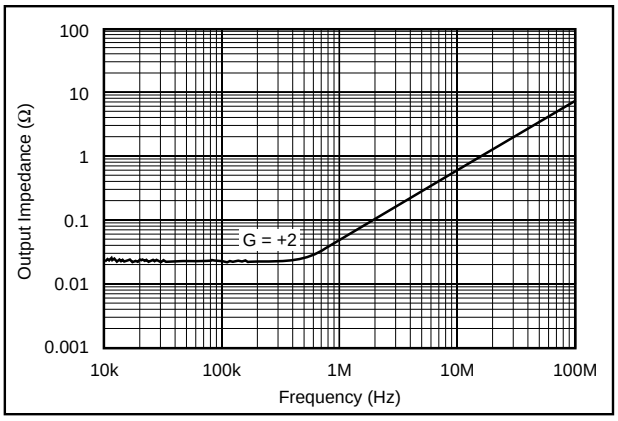


FIGURE 4. Closed-Loop Output Impedance vs Frequency.

THERMAL CONSIDERATIONS

The OPA658 will not require heatsinking under most operating conditions. Maximum desired junction temperature will set a maximum allowed internal power dissipation as described below. In no case should the maximum junction temperature be allowed to exceed 175°C.

Operating junction temperature (T_J) is given by $T_A + P_D \cdot \theta_{JA}$. The total internal power dissipation (P_D) is the sum of quiescent power (P_{DQ}) and additional power dissipated in the output stage (P_{DL}) to deliver load power. Quiescent power is simply the specified no-load supply current times the total supply voltage across the part. P_{DL} will depend on the required output signal and load but would, for a grounded resistive load, be at a maximum when the output is fixed at a voltage equal to 1/2 either supply voltage (for equal bipolar supplies). Under this condition $P_{DL} = V_S^2/(4 \cdot R_L)$ where R_L includes feedback network loading.

Note that it is the power in the output stage and not into the load that determines internal power dissipation.

As an example, compute the maximum T_J for an OPA658U at $A_V = +2$, $R_L = 100\Omega$, $R_{FB} = 402\Omega$, $\pm V_S = \pm 5V$, and the specified maximum $T_A = +85°C$. $P_D = 10V \cdot 8.5mA + 5^2/[4 \cdot (100\Omega \parallel 804\Omega)] = 155mW$. Maximum $T_J = 85°C + 0.155W \cdot 125°C/W = 104°C$.

DRIVING CAPACITIVE LOADS

The OPA658's output stage has been optimized to drive low resistive loads. Capacitive loads, however, will decrease the amplifier's phase margin which may cause high frequency peaking or oscillations. Capacitive loads greater than 5pF should be buffered by connecting a small resistance, usually 10Ω to 35Ω, in series with the output as shown in Figure 5. This is particularly important when driving high capacitance loads such as flash A/D converters.

In general, capacitive loads should be minimized for optimum high frequency performance. Coax lines can be driven if the cable is properly terminated. The capacitance of coax cable (29pF/foot for RG-58) will not load the amplifier when the coaxial cable or transmission line is terminated with its characteristic impedance.

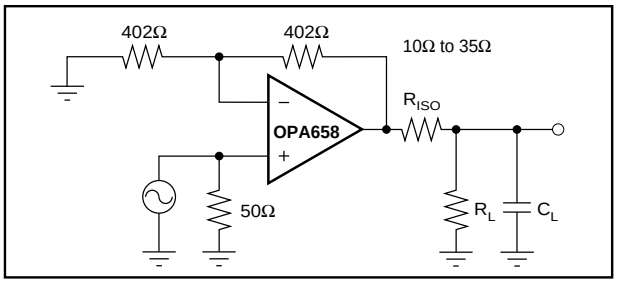


FIGURE 5. Driving Capacitive Loads.

COMPENSATION

The OPA658 is internally compensated and is stable in unity gain with a phase margin of approximately 62°, and approximately 64° in a gain of +2V/V when used with the recommended feedback resistor value. Frequency response for other gains are shown in the Typical Performance Curves.

The high-frequency response of the OPA658 in a good layout is very flat with frequency.

DISTORTION

The OPA658's Harmonic Distortion characteristics into a 100Ω load are shown versus frequency and power output in the Typical Performance Curves. Distortion can be further improved by increasing the load resistance as illustrated in Figure 6. Remember to include the contribution of the feedback resistance when calculating the effective load resistance seen by the amplifier.

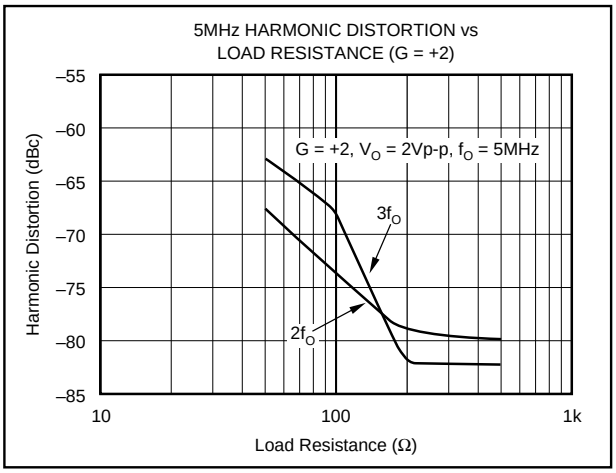


FIGURE 6. 5MHz Harmonic Distortion vs Load Resistance.

Narrowband communication channel requirements will benefit from the OPA658's wide bandwidth and low intermodulation distortion on low quiescent power. If output signal power at two closely spaced frequencies is required, third-order nonlinearities in any amplifier will cause spurious power at frequencies very near the two fundamental frequencies. If the two test frequencies, f_1 and f_2, are specified in terms of average and delta frequency, $f_O = (f_1 + f_2)/2$ and $\Delta f = |f_2 - f_1|$, the two, third-order, close-in spurious tones will appear at $f_O \pm 3 \cdot \Delta f$. The two tone, third-order spurious plot shown in Figure 7 indicates how far below these two equal power, closely spaced, tones the intermodulation spurious will be. The single tone power is at a matched 50Ω load. The unique design of the OPA658 provides much greater spurious free range than what a two-tone third-order intermodulation intercept specification would predict. This can be seen in Figure 7 as the spurious free range actually increases at the higher output power levels.

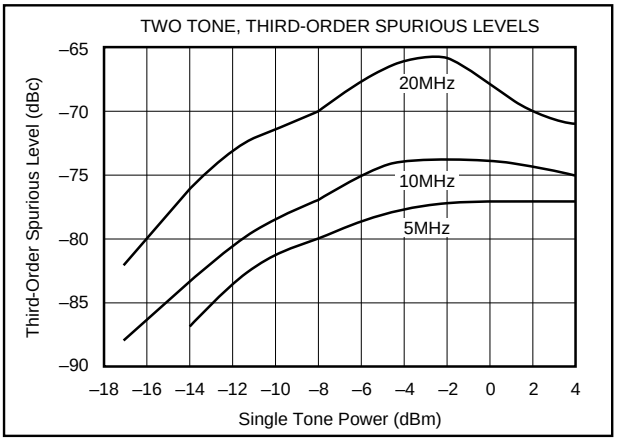


FIGURE 7. Third-Order Intercept Point vs Frequency.

DIFFERENTIAL GAIN AND PHASE

Differential Gain (dG) and Differential Phase (dP) are among the more important specifications for video applications. dG is defined as the percent change in closed-loop gain over a specified change in output voltage level. dP is defined as the change in degrees of the closed-loop phase over the same output voltage change. Both dG and dP are specified at the NTSC sub-carrier frequency of 3.58MHz and the PAL sub-carrier of 4.43MHz. All NTSC measurements were performed using a Tektronix model VM700A Video Measurement Set.

dG/dP of the OPA658 were measured with the amplifier in a gain of +2V/V with 75Ω input impedance and the output back-terminated in 75Ω. The input signal selected from the generator was a 0V to 1.4V modulated ramp with sync pulse. With these conditions the test circuit shown in Figure 8 delivered a 100IRE modulated ramp to the 75Ω input of the videoanalyzer. The signal averaging feature of the analyzer

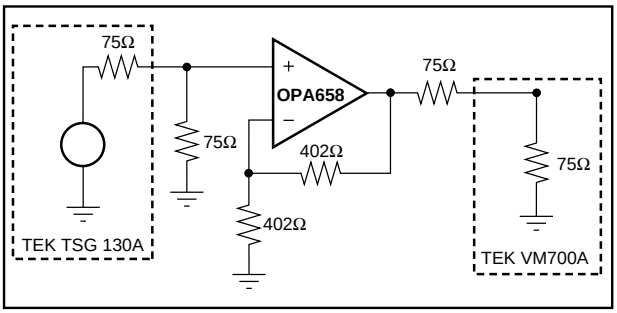


FIGURE 8. Configuration for Testing Differential Gain/Phase.

was used to establish a reference against which the performance of the amplifier was measured. Signal averaging was also used to measure the dg and dp of the test signal in order to eliminate the generator's contribution to measured amplifier performance. Typical performance of the OPA658 is 0.025% differential gain and 0.02° differential phase to both NTSC and PAL standards.

NOISE FIGURE

The OPA658's voltage and current noise spectral densities are specified in the Typical Performance Curves. For RF applications, however, Noise Figure (NF) is often the preferred noise specification since it allows system noise performance to be more easily calculated. The OPA658's Noise Figure vs Source Resistance is shown in Figure 9.

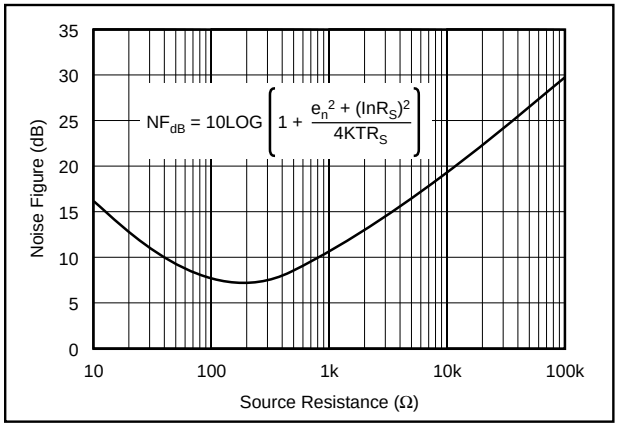


FIGURE 9. Noise Figure vs Source Resistance.

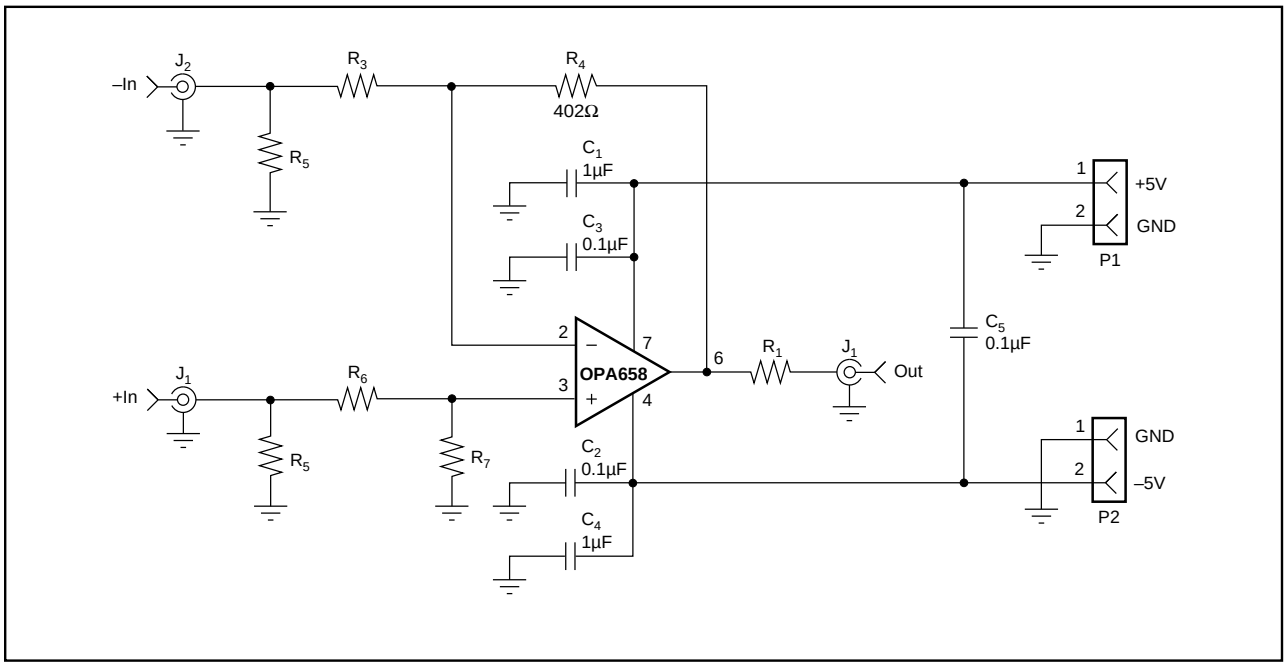


FIGURE 10. Layout Detail For Dem-OPA65X Demonstration Board.

SPICE MODELS

Computer simulation using SPICE is often useful when analyzing the performance of analog circuits and systems. This is particularly true for Video and RF amplifier circuits where parasitic capacitance and inductance can have a major effect on circuit performance. SPICE models using MicroSim Corporation's PSpice are available for the OPA658. Contract Burr-Brown applications departments to receive a SPICE Diskette.

DEM-OPA65X Demonstration Board Layout

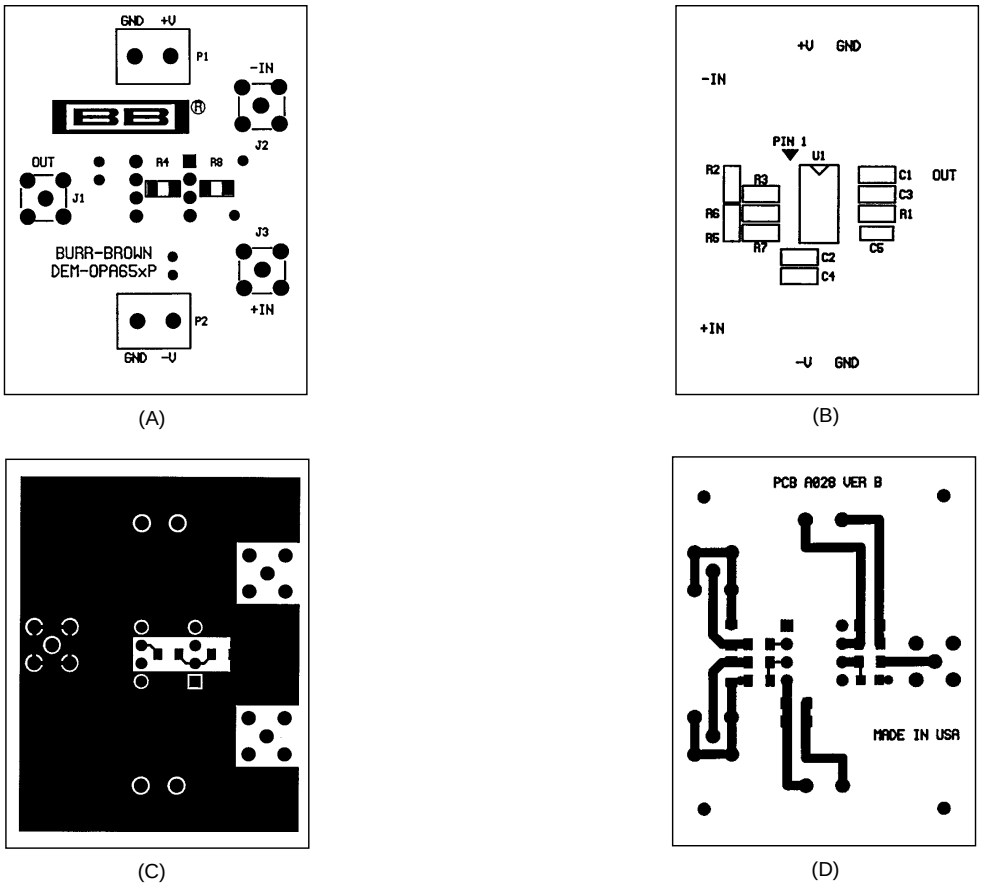


FIGURE 11a. Evaluation Board Silkscreen (Bottom). 11b. Evaluation Board Silkscreen (Top). 11c. Evaluation Board Layout (Solder Side). 11d. Evaluation Board Layout (Layout Side).

TYPICAL APPLICATION

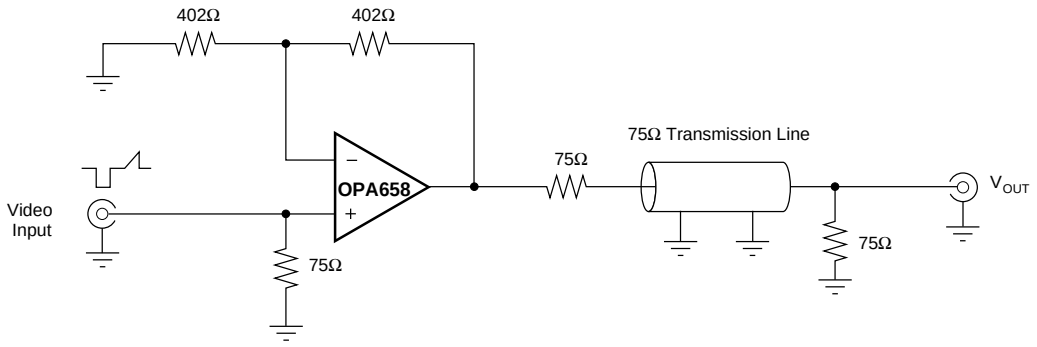


FIGURE 12. Low Distortion Video Amplifier.

PACKAGE DRAWINGS

Package Number 006 - 8-Pin Plastic, Single-Wide DIP

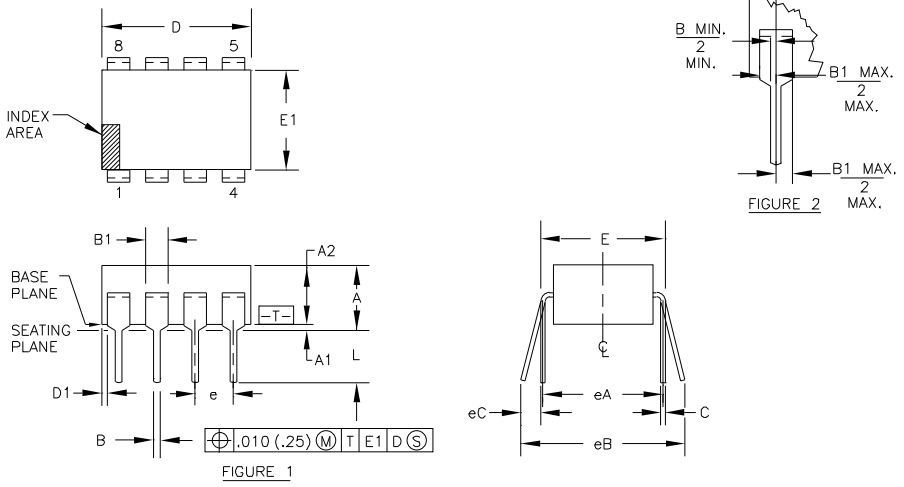


FIGURE 1

FIGURE 2

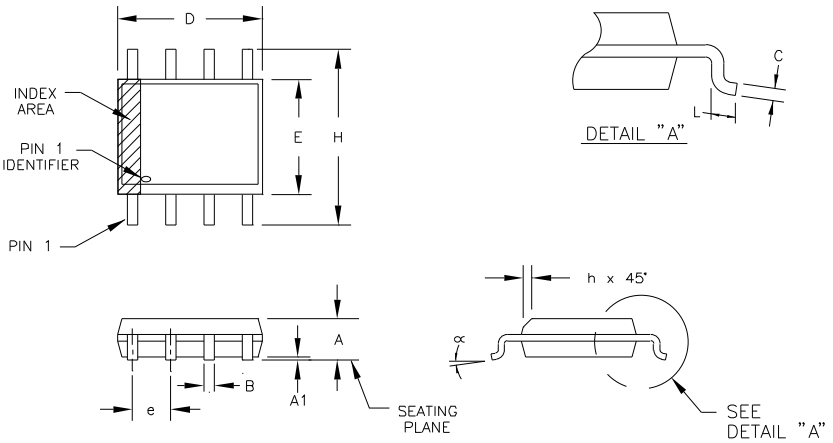
DIM	INCHES		MILLIMETERS		NOTE
	MIN.	MAX.	MIN.	MAX.	
A	--	.210	--	5.33	3
A1	.015	--	0.38	--	3
A2	.115	.195	2.92	4.95	
B	.014	.022	0.36	0.56	
B1	.045	.070	1.14	1.78	
C	.008	.015	0.20	0.38	
D	.348	.430	8.84	10.92	4
D1	.005	--	0.13	--	
E	.300	.325	7.62	8.26	5
E1	.240	.280	6.10	7.11	4
e	.100 BASIC		2.54 BASIC		
eA	.300 BASIC		7.63 BASIC		5
eB	--	.430	--	10.92	6
L	.115	.160	2.92	4.06	3

DIM	INCHES		MILLIMETERS		NOTE
	MIN.	MAX.	MIN.	MAX.	
N	8		8		7

NOTES:
1. CONTROLLING DIMENSION: INCH. IN CASE OF CONFLICT BETWEEN THE ENGLISH AND METRIC DIMENSIONS, THE INCH DIMENSIONS CONTROL.
2. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M-1982.
3. DIMENSIONS A, A1, AND L ARE MEASURED WITH THE PACKAGE SEATED IN JEDEC SEATING PLANE GAUGE GS-3.
4. D AND E1 DIMENSIONS FOR PLASTIC PACKAGES DO NOT INCLUDE MOLD FLASH OR PROTRUSIONS. MOLD FLASH OR PROTRUSIONS SHALL NOT EXCEED .010 INCH (0.25mm).
5. E AND eA MEASURED WITH THE LEADS CONSTRAINED TO BE PERPENDICULAR TO PLANE T.
6. eB AND eC ARE MEASURED AT THE LEAD TIPS WITH THE LEADS UNCONSTRAINED. eC MUST BE ZERO OR GREATER.
7. N IS THE MAXIMUM NUMBER OF TERMINAL POSITIONS.
8. CORNER LEADS (1, 4, 5, AND 8) MAY BE CONFIGURED AS SHOWN IN FIGURE 2.
9. FOR AUTOMATIC INSERTION, ANY RAISED IRREGULARITY ON THE TOP SURFACE (STEP, MESA, ETC.) SHALL BE SYMMETRICAL ABOUT THE LATERAL AND LONGITUDINAL PACKAGE CENTERLINES.

PACKAGE NUMBER: ZZ006	REV.: D
JEDEC NUMBER: MS-001	

Package Number 182 - 8-Lead SO-8 Surface Mount



DIM	INCHES		MILLIMETERS		NOTE
	MIN.	MAX.	MIN.	MAX.	
A	.054	.068	1.37	1.73	
A1	.004	.009	0.10	0.23	
B	.014	.019	0.36	0.48	
C	.008	.0098	0.20	0.25	
D	.189	.196	4.80	4.98	
E	.150	.157	3.81	3.99	
e	.050 BASIC		1.27 BASIC		
H	.229	.244	5.82	6.20	
h	.010	.019	0.25	0.48	
L	.016	.050	0.41	1.27	
N	8		8		
∝	0°	8°	0°	8°	

NOTES:
1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5-1982.
2. "D" AND "E" DO NOT INCLUDE MOLD FLASH OR PROTRUSIONS. MOLD FLASH OR PROTRUSIONS SHALL NOT EXCEED .15mm (.086 in.).
3. THE CHAMFER ON THE BODY IS OPTIONAL. IF IT IS NOT PRESENT, A VISUAL INDEX FEATURE MUST BE LOCATED WITHIN THE CROSS-HATCHED AREA.
4. "L" IS THE LENGTH OF THE TERMINAL FOR SOLDERING TO A SUBSTRATE.
5. "N" IS THE NUMBER OF TERMINAL POSITIONS.
6. LEAD TO LEAD COPLANARITY SHALL BE LESS THAN .004 INCHES FROM SEATING PLANE.

PACKAGE NUMBER: ZZ182	REV.: F
JEDEC NUMBER: MS-012	