AD711—SPECIFICATIONS (V_S = ±15 V @ T_A = 25°C, unless otherwise noted.)

Parameter	J/A/S Min	J/A/S Typ	J/A/S Max	K/B/T Min	K/B/T Typ	K/B/T Max	C Min	C Typ	C Max	Unit
INPUT OFFSET VOLTAGE[1]										
Initial Offset		0.3	2/1/1		0.2	0.5		0.10	0.25	mV
T_{MIN} to T_{MAX}			3/2/2			1.0			0.45	mV
vs. Temp		7	20/20/20		5	10		2	5	µV/°C
vs. Supply	76	95		80	100		86	110		dB
T_{MIN} to T_{MAX}	76/76/76			80			86			dB
Long-Term Stability		15			15			15		µV/Month
INPUT BIAS CURRENT[2]										
V_{CM} = 0 V		15	50		15	50		15	25	pA
V_{CM} = 0 V @ T_{MAX}			1.1/3.2/51			1.1/3.2/51			1.6	nA
V_{CM} = ±10 V		20	100		20	100		20	50	pA
INPUT OFFSET CURRENT										
V_{CM} = 0 V		10	25		5	25		5	10	pA
V_{CM} = 0 V @ T_{MAX}			0.6/1.6/26			0.6/1.6/26			0.65	nA
FREQUENCY RESPONSE										
Small Signal Bandwidth	3.0	4.0		3.4	4.0		3.4	4.0		MHz
Full Power Response		200			200			200		kHz
Slew Rate	16	20		18	20		18	20		V/µs
Settling Time to 0.01%		1.0	1.2		1.0	1.2		1.0	1.2	µs
Total Harmonic Distortion		0.0003			0.0003			0.0003		%
INPUT IMPEDANCE										
Differential		$3 \times 10^{12} \| 5.5$			$3 \times 10^{12} \| 5.5$			$3 \times 10^{12} \| 5.5$		Ω‖pF
Common Mode		$3 \times 10^{12} \| 5.5$			$3 \times 10^{12} \| 5.5$			$3 \times 10^{12} \| 5.5$		Ω‖pF
INPUT VOLTAGE RANGE										
Differential[3]		±20			±20			±20		V
Common-Mode Voltage[4]		+14.5, –11.5			+14.5, –11.5			+14.5, –11.5		
T_{MIN} to T_{MAX}	$-V_S$ + 4		$+V_S$ – 2	$-V_S$ + 4		$+V_S$ – 2	$-V_S$ + 4		+V – 2	V
Common-Mode Rejection Ratio										
V_{CM} = ±10 V	76	88		80	88		86	94		dB
T_{MIN} to T_{MAX}	76/76/76	84		80	84		86	90		dB
V_{CM} = ±11 V	70	84		76	84		76	90		dB
T_{MIN} to T_{MAX}	70/70/70	80		74	80		74	84		dB
INPUT VOLTAGE NOISE		2			2			2	4	µV p-p
		45			45			45		nV/√Hz
		22			22			22		nV/√Hz
		18			18			18		nV/√Hz
		16			16			16		nV/√Hz
INPUT CURRENT NOISE		0.01			0.01			0.01		pA/√Hz
OPEN-LOOP GAIN	150	400		200	400		200	400		V/mV
	100/100/100			100			100			V/mV
OUTPUT CHARACTERISTICS										
Voltage	+13, –12.5	+13.9, –13.3		+13, –12.5	+13.9, –13.3		+13, –12.5	+13.9, –13.3		V
	±12/±12/±12	+13.8, –13.1		±12	+13.8, –13.1		±12	+13.8, –13.1		V
Current		25			25			25		mA
POWER SUPPLY										
Rated Performance		±15			±15			±15		V
Operating Range	±4.5		±18	±4.5		±18	±4.5		±18	V
Quiescent Current		2.5	3.4		2.5	3.0		2.5	2.8	mA

NOTES
[1]Input Offset Voltage specifications are guaranteed after 5 minutes of operation at T_A = 25°C.
[2]Bias Current specifications are guaranteed maximum at either input after 5 minutes of operation at T_A = 25°C. For higher temperatures, the current doubles every 10°C.
[3]Defined as voltage between inputs, such that neither exceeds ±10 V from ground.
[4]Typically exceeding –14.1 V negative common-mode voltage on either input results in an output phase reversal.

Specifications subject to change without notice.

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ABSOLUTE MAXIMUM RATINGS[1]

Supply Voltage . . . . . ±18 V
Internal Power Dissipation[2] . . . . . 500 mW
Input Voltage[3] . . . . . ±18 V
Output Short Circuit Duration . . . . . Indefinite
Differential Input Voltage . . . . . $+V_S$ and $-V_S$
Storage Temperature Range (Q, H) . . . . . –65°C to +150°C
Storage Temperature Range (N) . . . . . –65°C to +125°C
Operating Temperature Range
AD711J/K . . . . . 0°C to +70°C
AD711A/B/C . . . . . –40°C to +85°C
AD711S/T . . . . . –55°C to +125°C
Lead Temperature Range (Soldering 60 sec) . . . . . 300°C

NOTES

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

[2]Thermal Characteristics:
8-Pin Plastic Package: θ_{JC} = 33°C/Watt; θ_{JA} = 100°C/Watt
8-Pin Cerdip Package: θ_{JC} = 22°C/Watt; θ_{JA} = 110°C/Watt
8-Pin Metal Can Package: θ_{JC} = 65°C/Watt; θ_{JA} = 150°C/Watt
8-Pin SOIC Package: θ_{JC} = 43°C/Watt; θ_{JA} = 160°C/Watt

[3]For supply voltages less than ±18 V, the absolute maximum input voltage is equal to the supply voltage.

ORDERING GUIDE

Model	Temperature Range	Package Description	Package Option*
*AD711AH	–40°C to +85°C	8-Pin Metal Can	H-08A
AD711AQ	–40°C to +85°C	8-Pin Ceramic DIP	Q-8
*AD711BQ	–40°C to +85°C	8-Pin Ceramic DIP	Q-8
*AD711CH	–40°C to +85°C	8-Pin Metal Can	H-08A
AD711JN	0°C to 70°C	8-Pin Plastic DIP	N-8
AD711JR	0°C to 70°C	8-Pin Plastic SOIC	RN-8
AD711JR-REEL	0°C to 70°C	8-Pin Plastic SOIC	RN-8
AD711JR-REEL7	0°C to 70°C	8-Pin Plastic SOIC	RN-8
AD711KN	0°C to 70°C	8-Pin Plastic DIP	N-8
AD711KR	0°C to 70°C	8-Pin Plastic SOIC	RN-8
AD711KR-REEL	0°C to 70°C	8-Pin Plastic SOIC	RN-8
AD711KR-REEL7	0°C to 70°C	8-Pin Plastic SOIC	RN-8
*AD711SQ/883B	–55°C to +125°C	8-Pin Ceramic DIP	Q-8
*AD711TQ/883B	–55°C to +125°C	8-Pin Ceramic DIP	Q-8

*Not for new design, obsolete April 2002

CAUTION

ESD (electrostatic discharge) sensitive device. Electrostatic charges as high as 4000 V readily accumulate on the human body and test equipment and can discharge without detection. Although the AD711 features proprietary ESD protection circuitry, permanent damage may occur on devices subjected to high-energy electrostatic discharges. Therefore, proper ESD precautions are recommended to avoid performance degradation or loss of functionality.

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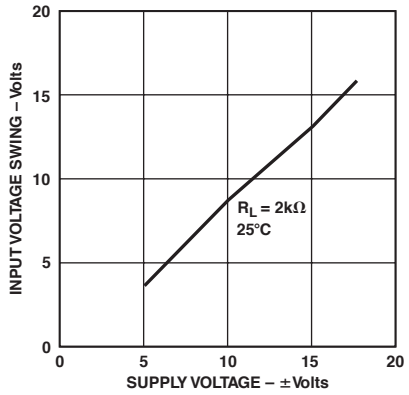
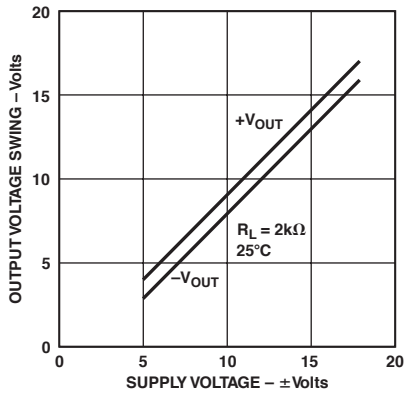
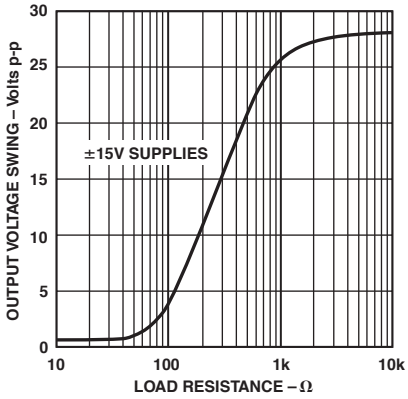
AD711–Typical Performance Characteristics



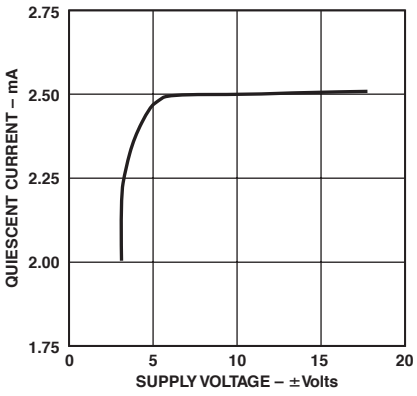
TPC 1. Input Voltage Swing vs. Supply Voltage



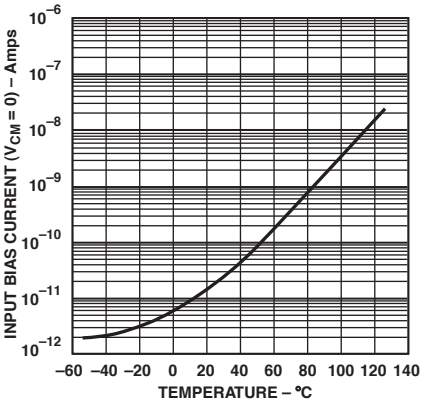
TPC 2. Output Voltage Swing vs. Supply Voltage



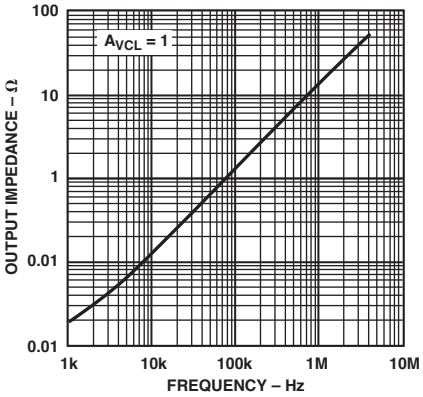
TPC 3. Output Voltage Swing vs. Load Resistance



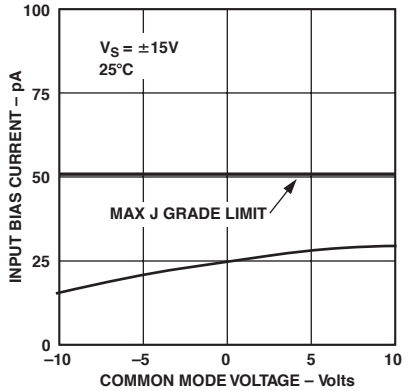
TPC 4. Quiescent Current vs. Supply Voltage



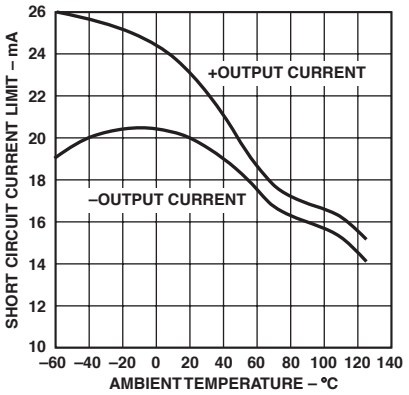
TPC 5. Input Bias Current vs. Temperature



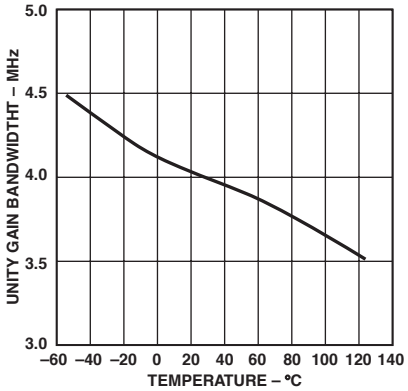
TPC 6. Output Impedance vs. Frequency



TPC 7. Input Bias Current vs. Common Mode Voltage



TPC 8. Short Circuit Current Limit vs. Temperature



TPC 9. Unity Gain Bandwidth vs. Temperature

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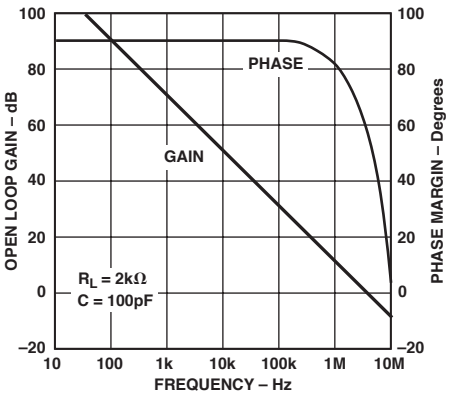
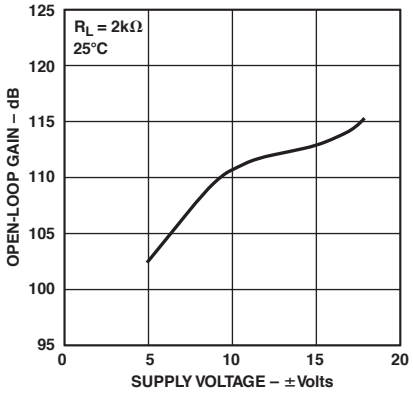
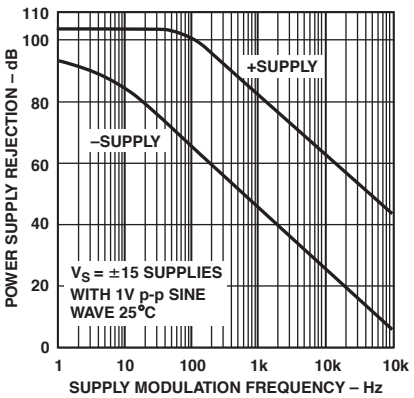


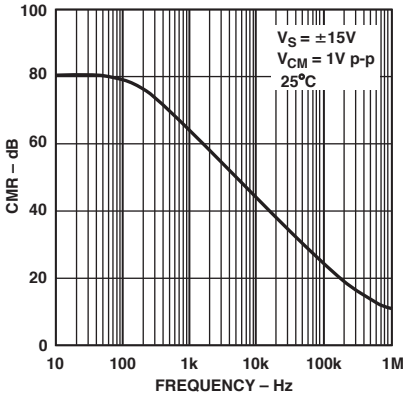
TPC 10. Open-Loop Gain and Phase Margin vs. Frequency



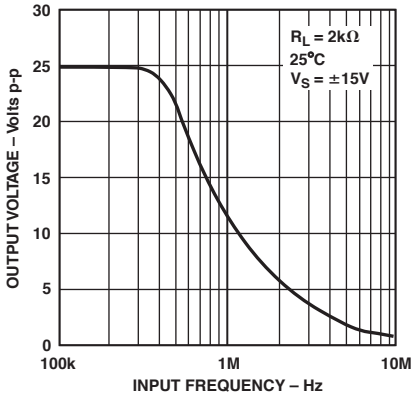
TPC 11. Open-Loop Gain vs. Supply Voltage



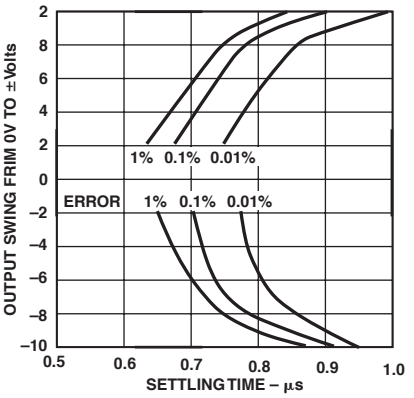
TPC 12. Power Supply Rejection vs. Frequency



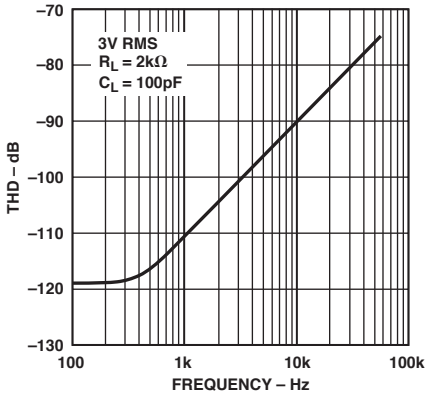
TPC 13. Common Mode Rejection vs. Frequency



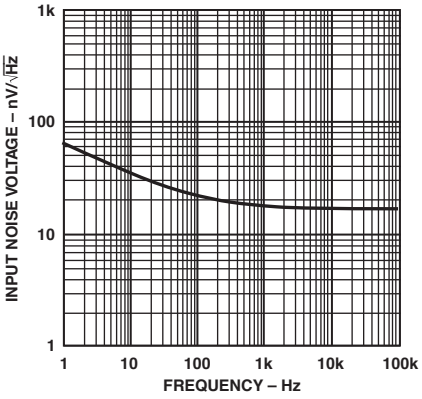
TPC 14. Large Signal Frequency Response



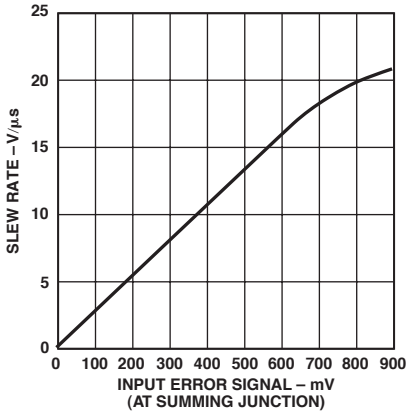
TPC 15. Output Swing and Error vs. Settling Time



TPC 16. Total Harmonic Distortion vs. Frequency



TPC 17. Input Noise Voltage Spectral Density



TPC 18. Slew Rate vs. Input Error Signal

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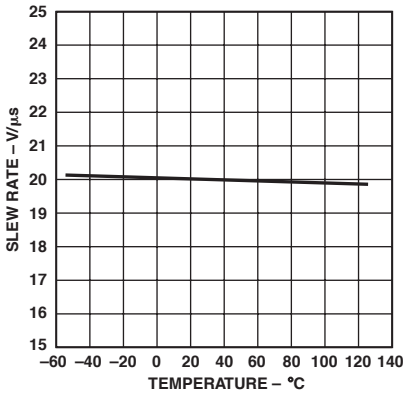
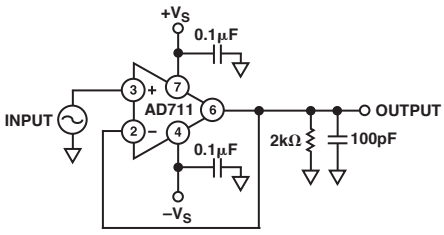
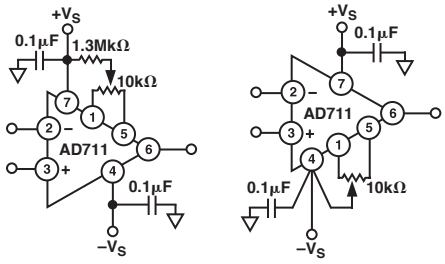


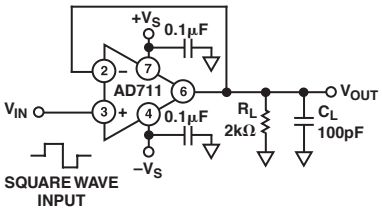
TPC 19. Slew Rate vs. Temperature



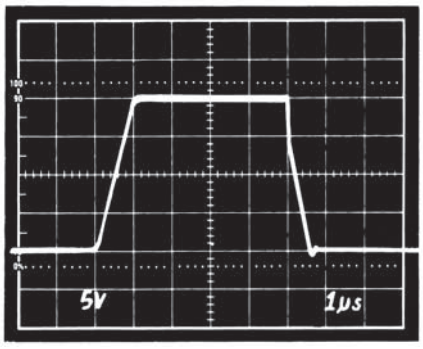
TPC 20. T.H.D. Test Circuit



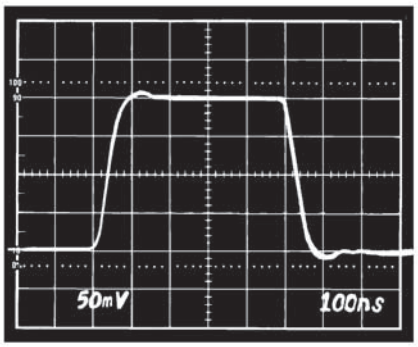
TPC 21. Offset Null Configurations



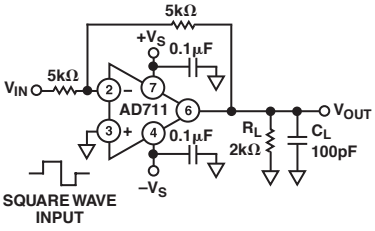
TPC 22a. Unity Gain Follower



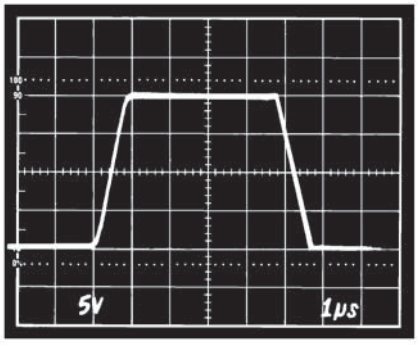
TPC 22b. Unity Gain Follower Pulse Response (Large Signal)



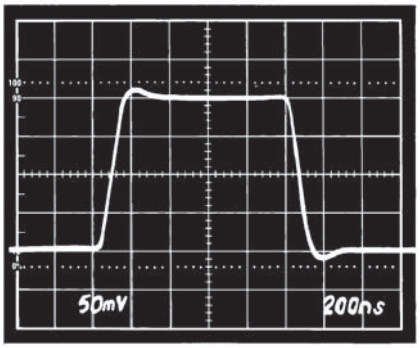
TPC 22c. Unity Gain Follower Pulse Response (Small Signal)



TPC 23a. Unity Gain Inverter



TPC 23b. Unity Gain Inverter Pulse Response (Large Signal)



TPC 23c. Unity Gain Inverter Pulse Response (Small Signal)

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OPTIMIZING SETTLING TIME

Most bipolar high-speed D/A converters have current outputs; therefore, for most applications, an external op amp is required for current-to-voltage conversion. The settling time of the converter/op amp combination depends on the settling time of the DAC and output amplifier. A good approximation is:

$$t_S\ Total = \sqrt{(t_S\ DAC)^2 + (t_S\ AMP)^2} \quad (1)$$

The settling time of an op amp DAC buffer will vary with the noise gain of the circuit, the DAC output capacitance, and with the amount of external compensation capacitance across the DAC output scaling resistor.

Settling time for a bipolar DAC is typically 100 ns to 500 ns. Previously, conventional op amps have required much longer settling times than have typical state-of-the-art DACs; therefore, the amplifier settling time has been the major limitation to a high-speed voltage-output D-to-A function. The introduction of the AD711/712 family of op amps with their 1 µs (to ±0.01% of final value) settling time now permits the full high-speed capabilities of most modern DACs to be realized.

In addition to a significant improvement in settling time, the low offset voltage, low offset voltage drift, and high open-loop gain of the AD711 family assures 12-bit accuracy over the full operating temperature range.

The excellent high-speed performance of the AD711 is shown in the oscilloscope photos of Figure 2. Measurements were taken using a low input capacitance amplifier connected directly to the summing junction of the AD711 – both photos show the worst case situation: a full-scale input transition. The DAC's 4 kΩ [10 kΩ‖8 kΩ = 4.4 kΩ] output impedance together with a 10 kΩ feedback resistor produce an op amp noise gain of 3.25. The current output from the DAC produces a 10 V step at the op amp output (0 to –10 V Figure 2a, –10 V to 0 V Figure 2b.)

Therefore, with an ideal op amp, settling to ±1/2 LSB (±0.01%) requires that 375 µV or less appears at the summing junction. This means that the error between the input and output (that voltage which appears at the AD711 summing junction) must be less than 375 µV. As shown in Figure 2, the total settling time for the AD711/AD565 combination is 1.2 microseconds.

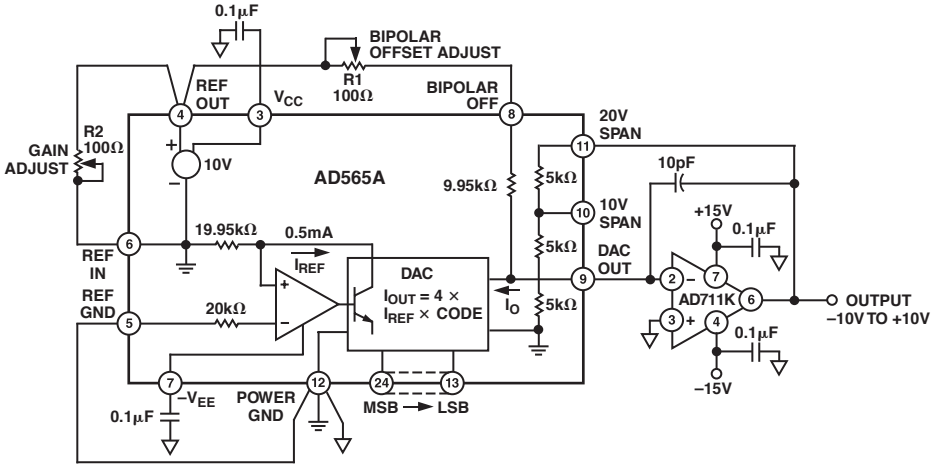
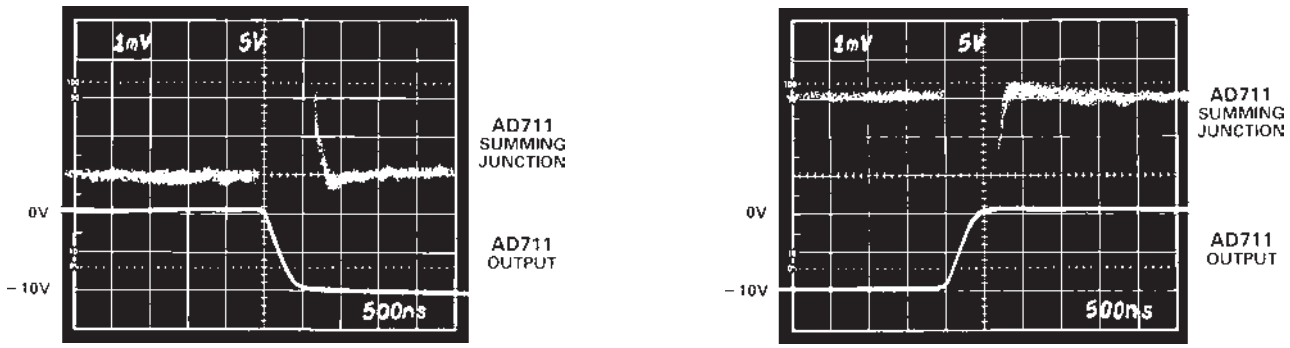


Figure 1. ±10 V Voltage Output Bipolar DAC



a. (Full-Scale Negative Transition) *b. (Full-Scale Positive Transition)*

Figure 2. Settling Characteristics for AD711 with AD565A

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OP AMP SETTLING TIME—A MATHEMATICAL MODEL

The design of the AD711 gives careful attention to optimizing individual circuit components; in addition, a careful tradeoff was made: the gain bandwidth product (4 MHz) and slew rate (20 V/μs) were chosen to be high enough to provide very fast settling time but not too high to cause a significant reduction in phase margin (and therefore stability). Thus designed, the AD711 settles to ±0.01%, with a 10 V output step, in under 1 μs, while retaining the ability to drive a 100 pF load capacitance when operating as a unity gain follower.

If an op amp is modeled as an ideal integrator with a unity gain crossover frequency of $\omega_o/2\pi$, Equation 1 will accurately describe the small signal behavior of the circuit of Figure 3a, consisting of an op amp connected as an I-to-V converter at the output of a bipolar or CMOS DAC. This equation would completely describe the output of the system if not for the op amp's finite slew rate and other nonlinear effects.

$$\frac{V_O}{I_{IN}} = \frac{-R}{\frac{R(C_f = C_X)}{\omega_o}s^2 + \left(\frac{G_N}{\omega_o} + RC_f\right)s + 1} \quad (3)$$

where:

$\frac{\omega_o}{2_\pi}$ =op amp's unity gain frequency

G_N = "noise" gain of circuit $\left(1 + \frac{R}{R_O}\right)$

This equation may then be solved for C_f:

$$C_f = \frac{2 - G_N}{R\omega_o} + \frac{2\sqrt{RC_X\omega_o + (1 - G_N)}}{R\omega_o} \quad (3)$$

In these equations, capacitor C_X is the total capacitor appearing the inverting terminal of the op amp. When modeling a DAC buffer application, the Norton equivalent circuit of Figure 3a can be used directly; capacitance C_X is the total capacitance of the output of the DAC plus the input capacitance of the op amp (since the two are in parallel).

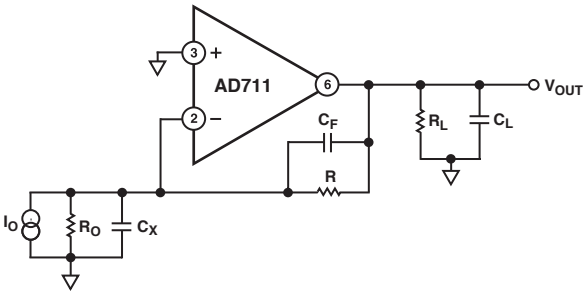

Figure 3a. Simplified Model of the AD711 Used as a Current-Out DAC Buffer

When R_O and I_O are replaced with their Thevenin V_{IN} and R_{IN} equivalents, the general purpose inverting amplifier of Figure 26b is created. Note that when using this general model, capacitance C_X is *either* the input capacitance of the op amp if a simple inverting op amp is being simulated *or* it is the combined capacitance of the DAC output and the op amp input if the DAC buffer is being modeled.

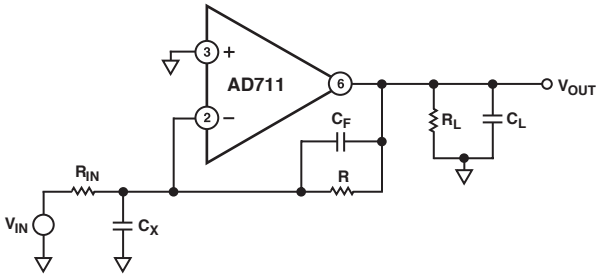

Figure 3b. Simplified Model of the AD711 Used as an Inverter

In either case, the capacitance C_X causes the system to go from a one-pole to a two-pole response; this additional pole increases settling time by introducing peaking or ringing in the op amp output. Since the value of C_X can be estimated with reasonable accuracy, Equation 2 can be used to choose a small capacitor, C_F, to cancel the input pole and optimize amplifier response. Figure 4 is a graphical solution of Equation 2 for the AD711 with R = 4 kΩ.

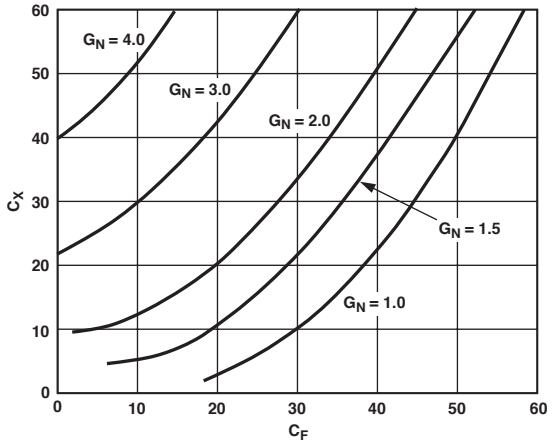

Figure 4. Value of Capacitor C_F vs. Value of C_X

The photos of Figures 5a and 5b show the dynamic response of the AD711 in the settling test circuit of Figure 6.

The input of the settling time fixture is driven by a flat-top pulse generator. The error signal output from the false summing node of A1 is clamped, amplified by A2 and then clamped again. The error signal is thus clamped twice: once to prevent overloading amplifier A2 and then a second time to avoid overloading the oscilloscope preamp. The Tektronix oscilloscope preamp type 7A26 was carefully chosen because it does not overload with these input levels. Amplifier A2 needs to be a very high speed FET-input op amp; it provides a gain of 10, amplifying the error signal output of A1.

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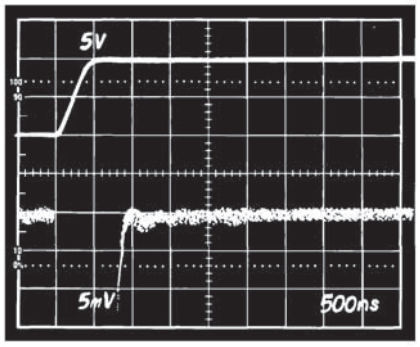

Figure 5a. Settling Characteristics 0 to +10 V Step
Upper Trace: Output of AD711 Under Test (5 V/Div)
Lower Trace: Amplified Error Voltage (0.01%/Div)

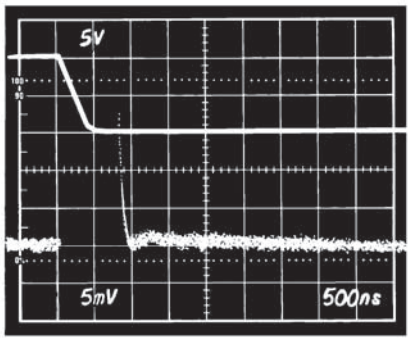

Figure 5b. Settling Characteristics 0 to –10 V Step
Upper Trace: Output of AD711 Under Test (5 V/Div)
Lower Trace: Amplified Error Voltage (0.01%/Div)

GUARDING

The low input bias current (15 pA) and low noise characteristics of the AD711 BiFET op amp make it suitable for electrometer applications such as photo diode preamplifiers and picoampere current-to-voltage converters. The use of a guarding technique such as that shown in Figure 7, in printed circuit board layout and construction is critical to minimize leakage currents. The guard ring is connected to a low impedance potential at the same level as the inputs. High impedance signal lines should not be extended for any unnecessary length on the printed circuit board.

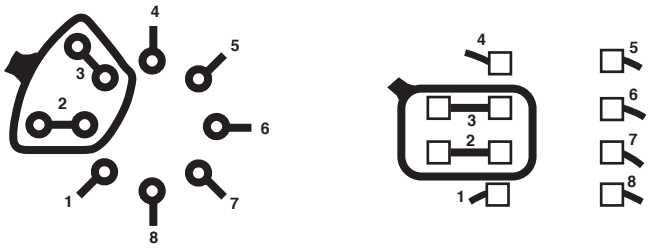

Figure 7. Board Layout for Guarding Inputs

D/A CONVERTER APPLICATIONS

The AD711 is an excellent output amplifier for CMOS DACs. It can be used to perform both 2-quadrant and 4-quadrant operation. The output impedance of a DAC using an inverted R-2R ladder approaches R for codes containing many 1s, 3R for codes containing a single 1, and for codes containing all zero, the output impedance is infinite.

For example, the output resistance of the AD7545 will modulate between 11 kΩ and 33 kΩ. Therefore, with the DAC's internal feedback resistance of 11 kΩ, the noise gain will vary from 2 to 4/3. This changing noise gain modulates the effect of the input offset voltage of the amplifier, resulting in nonlinear DAC amplifier performance.

The AD711K with guaranteed 500 µV offset voltage minimizes this effect to achieve 12-bit performance.

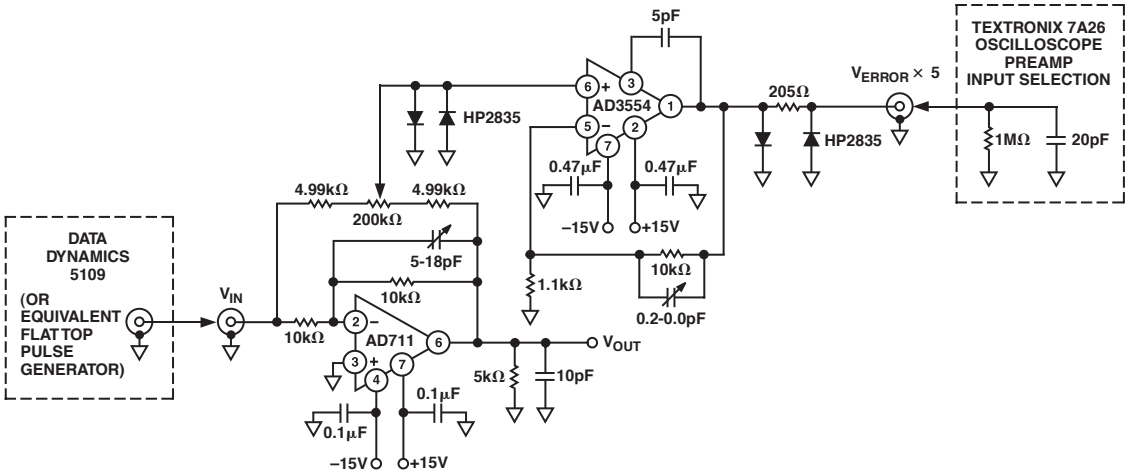

Figure 6. Settling Time Test Circuit

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Figures 8 and 9 show the AD711 and AD7545 (12-bit CMOS DAC) configured for unipolar binary (2-quadrant multiplication) or bipolar (4-quadrant multiplication) operation. Capacitor C1 provides phase compensation to reduce overshoot and ringing.

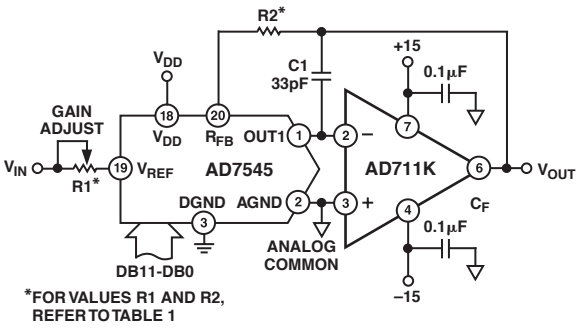


Figure 8. Unipolar Binary Operation

R1 and R2 calibrate the zero offset and gain error of the DAC. Specific values for these resistors depend upon the grade of AD7545 and are shown below.

Table I. Recommended Trim Resistor Values vs. Grades of the AD7545 for V_{DD} = 5 V

TRIM RESISTOR	JN/AQ/SD	KN/BQ/TD	LN/CQ/UD	GLN/GCQ/GUD
R1	500 Ω	200 Ω	100 Ω	20 Ω
R2	150 Ω	68 Ω	33 Ω	6.8 Ω

NOISE CHARACTERISTICS

The random nature of noise, particularly in the 1/f region, makes it difficult to specify in practical terms. At the same time, designers of precision instrumentation require certain guaranteed maximum noise levels to realize the full accuracy of their equipment.

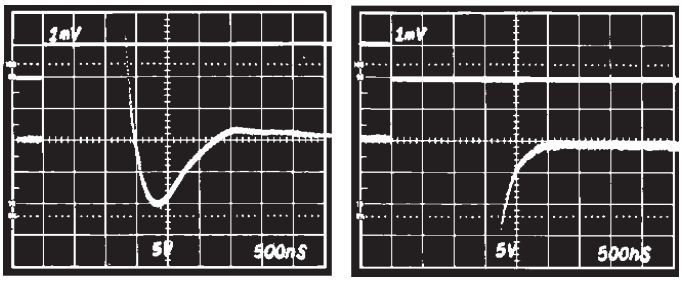
The AD711C grade is specified at a maximum level of 4.0 µV p-p, in a 0.1 Hz to 10 Hz bandwidth. Each AD711C receives a 100% noise test for two 10-second intervals; devices with any excursion in excess of 4.0 µV are rejected. The screened lot is then submitted to Quality Control for verification on an AQL basis.

All other grades of the AD711 are sample-tested on an AQL basis to a limit of 6 µV p-p, 0.1 to 10 Hz.

DRIVING THE ANALOG INPUT OF AN A/D CONVERTER

An op amp driving the analog input of an A/D converter, such as that shown in Figure 11, must be capable of maintaining a constant output voltage under dynamically changing load conditions. In successive-approximation converters, the input current is compared to a series of switched trial currents. The comparison point is diode clamped but may deviate several hundred millivolts resulting in high frequency modulation of A/D input current.

Figures 10a and 10b show the settling time characteristics of the AD711 when used as a DAC output buffer for the AD7545.



a. Full-Scale Positive Transition

b. Full-Scale Negative Transition

Figure 10. Settling Characteristics for AD711 with AD7545

compared to a series of switched trial currents. The comparison point is diode clamped but may deviate several hundred millivolts resulting in high frequency modulation of A/D input current. The output impedance of a feedback amplifier is made artificially low by the loop gain. At high frequencies, where the loop gain is low, the amplifier output impedance can approach its open loop value. Most IC amplifiers exhibit a minimum open loop output impedance of 25 Ω due to current limiting resistors. A few hundred microamps reflected from the change in converter loading can introduce errors in instantaneous input

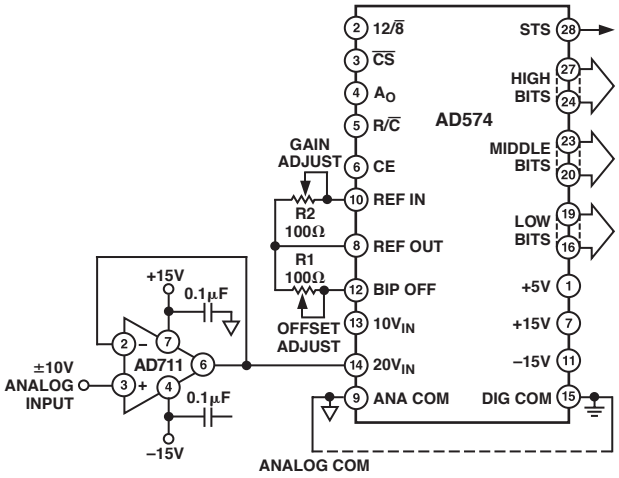


Figure 11. AD711 as ADC Unity Gain Buffer

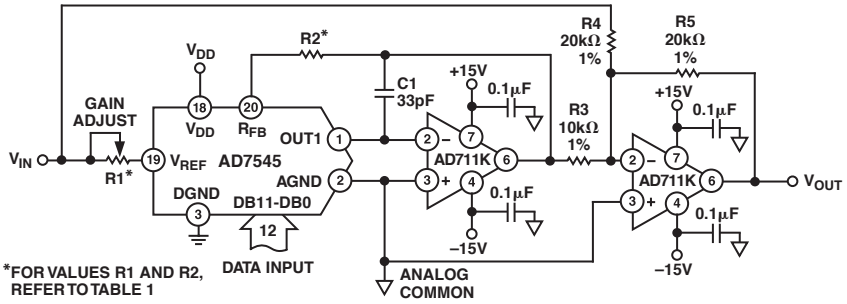


Figure 9. Bipolar Operation

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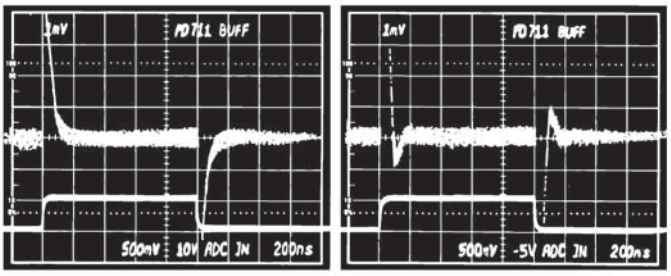
voltage. If the A/D conversion speed is not excessive and the bandwidth of the amplifier is sufficient, the amplifier's output will return to the nominal value before the converter makes its comparison. However, many amplifiers have relatively narrow bandwidth yielding slow recovery from output transients. The AD711 is ideally suited to drive high speed A/D converters since it offers both wide bandwidth and high open-loop gain.



a. Source Current = 2 mA *b. Sink Current = 1 mA*

Figure 12. ADC Input Unity Gain Buffer Recovery Times

DRIVING A LARGE CAPACITIVE LOAD

The circuit in Figure 13 employs a 100 Ω isolation resistor which enables the amplifier to drive capacitive loads exceeding 1500 pF; the resistor effectively isolates the high frequency feedback from the load and stabilizes the circuit. Low frequency feedback is returned to the amplifier summing junction via the low pass filter formed by the 100 Ω series resistor and the load capacitance, C_L. Figure 14 shows a typical transient response for this connection.

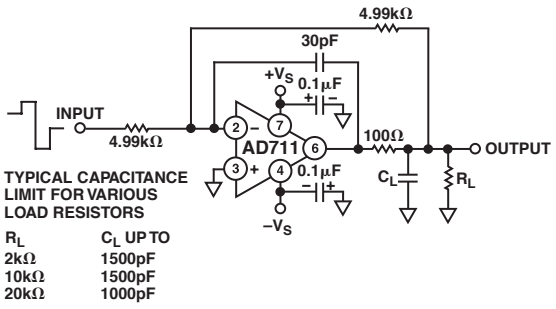


Figure 13. Circuit for Driving a Large Capacitive Load

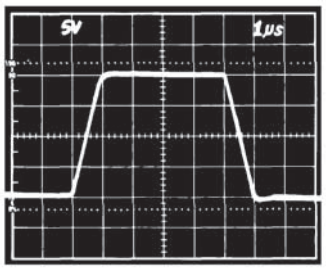


Figure 14. Transient Response R_L = 2 kΩ, C_L = 500 pF

ACTIVE FILTER APPLICATIONS

In active filter applications using op amps, the dc accuracy of the amplifier is critical to optimal filter performance. The amplifier's offset voltage and bias current contribute to output error. Offset voltage will be passed by the filter and may be amplified to produce excessive output offset. For low frequency applications requiring large value input resistors, bias currents flowing through these resistors will also generate an offset voltage.

In addition, at higher frequencies, an op amp's dynamics must be carefully considered. Here, slew rate, bandwidth, and open-loop gain play a major role in op amp selection. The slew rate must be fast as well as symmetrical to minimize distortion. The amplifier's bandwidth in conjunction with the filter's gain will dictate the frequency response of the filter.

The use of a high performance amplifier such a s the AD711 will minimize both dc and ac errors in all active filter applications.

SECOND ORDER LOW PASS FILTER

Figure 15 depicts the AD711 configured as a second order Butterworth low pass filter. With the values as shown, the corner frequency will be 20 kHz; however, the wide bandwidth of the AD711 permits a corner frequency as high as several hundred kilohertz. Equations for component selection are shown below.

$$R1 = R2 = \textit{user selected}\ (\textit{typical values}\text{: } 10\ k\Omega - 100\ k\Omega) \quad (4)$$

$$C1 = \frac{1.414}{(2\pi)(f_{cutoff})(R1)},\ C2 = \frac{0.707}{(2\pi)(f_{cutoff})(R1)} \quad (5)$$

Where:

C1 and *C2* are in farads.

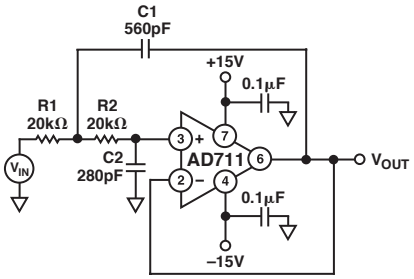


Figure 15. Second Order Low Pass Filter

An important property of filters is their out-of-band rejection. The simple 20 kHz low pass filter shown in Figure 15, might be used to condition a signal contaminated with clock pulses or sampling glitches which have considerable energy content at high frequencies.

The low output impedance and high bandwidth of the AD711 minimize high frequency feedthrough as shown in Figure 16.

The upper trace is that of another low-cost BiFET op amp showing 17 dB more feedthrough at 5 MHz.

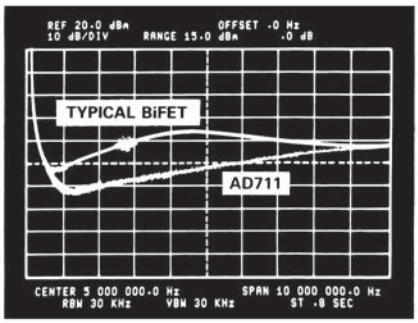


Figure 16.

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9-POLE CHEBYCHEV FILTER

Figure 17 shows the AD711 and its dual counterpart, the AD712, as a 9-pole Chebychev filter using active frequency dependent negative resistors (FDNR). With a cutoff frequency of 50 kHz and better than 90 dB rejection, it may be used as an anti-aliasing filter for a 12-bit data acquisition system with 100 kHz throughput.

As shown in Figure 17, the filter is comprised of four FDNRs (A, B, C, D) having values of 4.9395×10^{-15} and 5.9276×10^{-15} farad-seconds. Each FDNR active network provides a 2-pole response; for a total of 8 poles. The 9th pole consists of a 0.001 µF capacitor and a 124 kΩ resistor at Pin 3 of amplifier A2. Figure 18 depicts the circuits for each FDNR with the proper selection of R. To achieve optimal performance, the 0.001 µF capacitors must be selected for 1% or better matching and all resistors should have 1% or better tolerance.

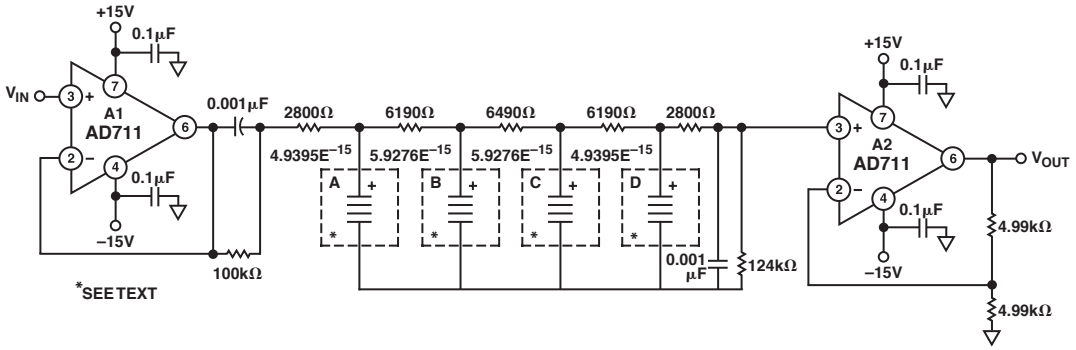

Figure 17. 9-Pole Chebychev Filter

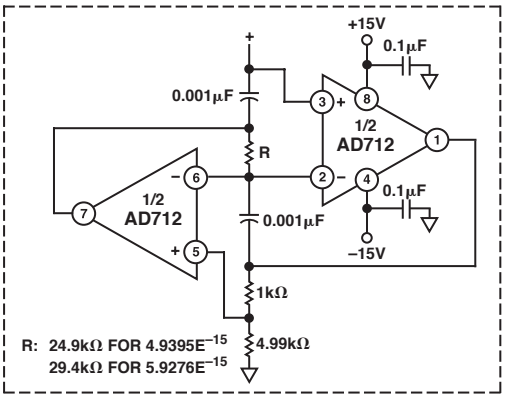

Figure 18. FDNR for 9-Pole Chebychev Filter

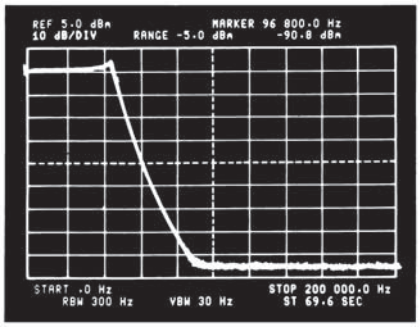

Figure 19. High Frequency Response for 9-Pole Chebychev Filter

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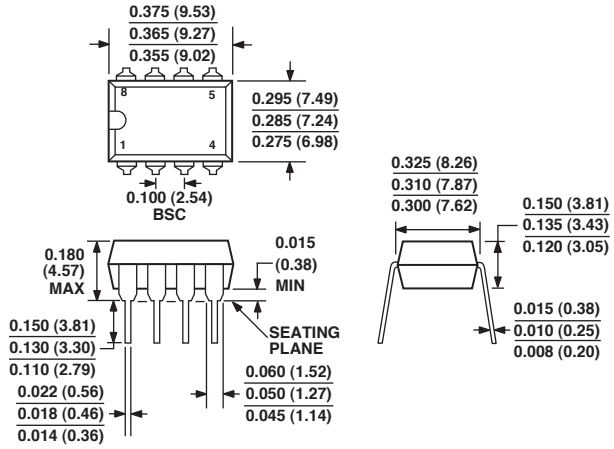
OUTLINE DIMENSIONS

8-Lead Plastic Dual-in-Line Package [PDIP]
(N-8)

Dimensions shown in inches and (millimeters)



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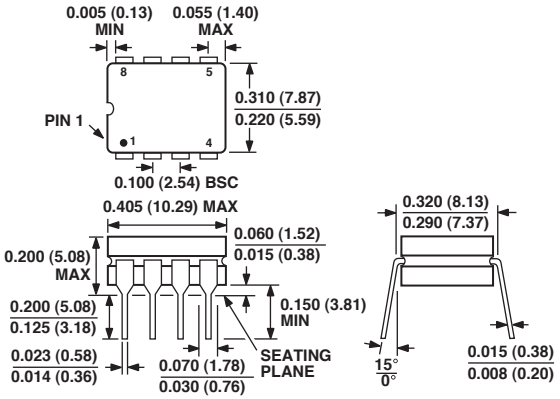
8-Lead Ceramic Dip – Glass Hermetic Seal [CERDIP]
(Q-8)

Dimensions shown in inches and (millimeters)



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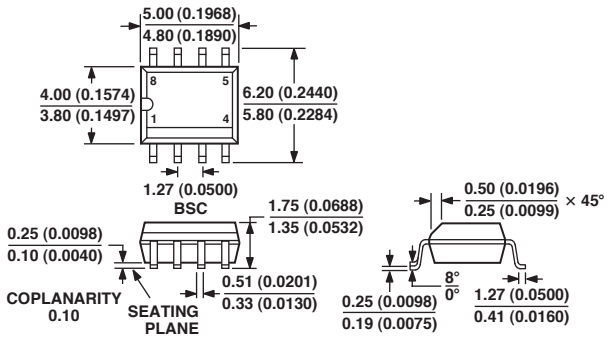
8-Lead Standard Small Outline Package [SOIC]
Narrow Body
(RN-8)

Dimensions shown in millimeters and (inches)



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8-Lead Metal Can [TO-99]
(H-8)

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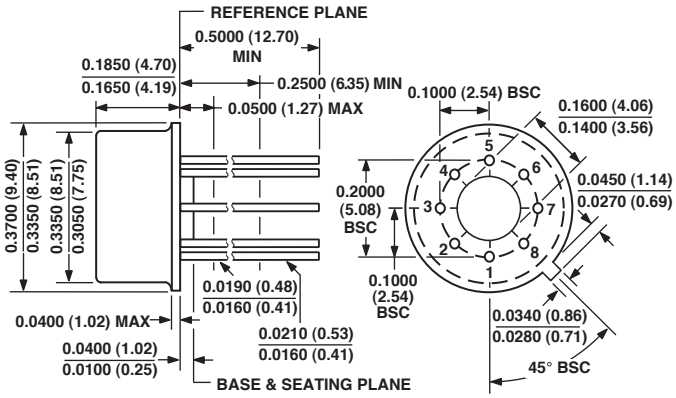

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Revision History

Location	Page
10/02—Data Sheet changed from REV. D to REV. E.	
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10/02—Data Sheet changed from REV. C to REV. D.	
Edits to CONNECTION DIAGRAMS	1
Updated OUTLINE DIMENSIONS	13
5/02—Data Sheet changed from REV. B to REV. C.	
Change from Small Outline Package (R-8)	13
Edits to ORDERING GUIDE	3
Deleted METALLIZATION PHOTOGRAPH	3

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