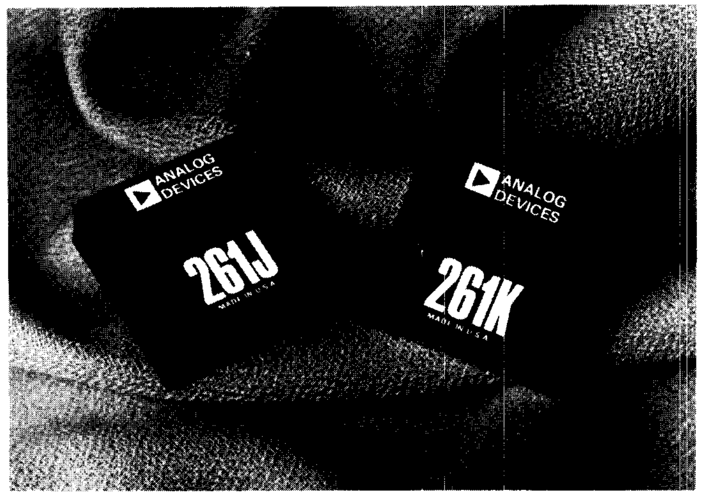
ANALOG DEVICES

Low Noise Non-Inverting Chopper Amplifier

MODEL 261

FEATURES
Non-Inverting Input
$10^9\,\Omega$ Common Mode Impedance
Protected MOSFET Chopper
Ultra Low Drift 0.1μV/°C, Max (261K)
Guaranteed Low Noise of 0.4μVp-p (0.01 to 1Hz)
Low Cost

APPLICATIONS
Microvolt & Millivolt Measurements
Meter & Recorder Preamplifier
Semiconductor Strain Gage Amplifier
Biological Sensors
Potentiometer Buffer



GENERAL DESCRIPTION
Model 261 is a low cost non-inverting chopper amplifier featuring ultra low drift of 0.1μV/°C, open loop gain of greater than 10 million V/V and guaranteed low noise performance of 0.4μV p–p max in a 0.01 to 1Hz bandwidth. It is ideally suited for low level pre-amplifier applications where high input impedance and low noise are essential.

Model 261 also offers a solution to beat frequency problems caused by a low frequency carrier mixing with harmonics of the ac line. Its carrier frequency of 3500Hz is nearly a decade higher than that of models previously available. The required harmonic of the ac line that could cause interference with a 3500Hz carrier has negligible energy content and beat frequencies are eliminated. As a further protection against interfering signals, model 261 has been completely shielded internally. This protective shield reduces interference due to RF signals, as well as carrier signals from adjacent chopper amplifiers.

Still another advantage of the 261 due to its higher chopper frequency and shielded design is an output signal that is free from both distortion and chopper spikes. The result is a design that can process low level signals while maintaining low distortion and high signal to noise ratios.

CHOPPER VS. CHOPPER-STABILIZED
Most conventional ultra-stable amplifiers are chopper-stabilized to achieve low drift. In these units, the higher frequency signal components are separated and directly amplified, while the low frequency and dc components are separately chopped, amplified, demodulated, and then summed with the high frequency components in an output stage. This method provides wide bandwidth and excellent performance at the expense of increased cost and complexity. Since many requirements for ultra-low drift amplification involve only dc and low frequency signals, the additional high frequency amplifier stage found in most chopper-stabilized amplifiers has been eliminated from the model 261. This design approach has made it possible to achieve a practical non-inverting configuration, which retains the advantages of low cost and small size. The input stage of the model 261 chops the signal at a 3500Hz rate, resulting in a maximum useful –3dB bandwidth of about 100Hz. For increased flexibility in meeting specific design requirements, terminals are provided for an external compensation capacitor, which determines the amplifier's gain-bandwidth product.

INPUT IMPEDANCE
One of the prime advantages of the non-inverting amplifier is the capability of bootstrapping the input impedance up to the level of the common mode impedance. For the model 261, this means that the 40kΩ open loop input resistance will be multiplied by the open loop gain times the feedback factor. With a typical open loop gain of 40 x 10^6, closed loop gains of up to 1600 will allow the user to realize $10^9\,\Omega$ input resistance. Even at a gain of 10,000, the effective input resistance will be over 100 megohms. (i.e.) $(40\text{k}\Omega)\dfrac{40 \times 10^6}{10^4} = 160\text{M}\Omega$

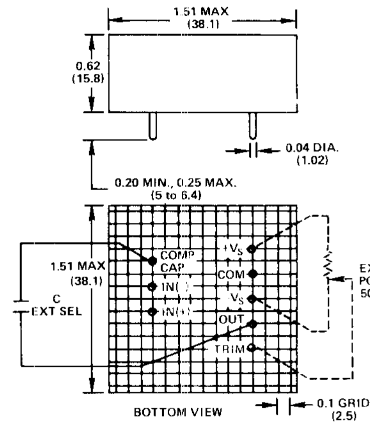
SPECIFICATIONS (typical @ +25°C and ±15V dc unless otherwise noted)

Model	261J	261K
OPEN LOOP GAIN		
DC rated load	10^7V/V min	
RATED OUTPUT		
Voltage	±10V min	*
Current	±5mA min	*
Load Capacitance Range	0 to 0.001μF	*
FREQUENCY RESPONSE[1]		
Small Signal, -3dB	100Hz	*
Full Power Response	2-50Hz min	*
Slewing Rate	100V/sec min	*
Overload Recovery	300ms	*
INPUT OFFSET VOLTAGE		
External Trim Pot[2]	50kΩ	*
Initial Offset, +25°C	±25μV max	*
Avg vs Temp (0 to +70°C)	±0.3μV/°C max	±0.1μV/°C max
Supply Voltage	±0.1μV/%	*
Time	±½μV/month	*
Warm-Up Drift	<3μV in 20 minutes	*
INPUT BIAS CURRENT		
Initial Bias, +25°C, + Input	±300pA max	*
Avg vs Temp (0 to +70°C)	±10pA/°C max	*
Initial Bias, +25°C, - Input	±10nA max	*
Avg vs Supply Voltage	±3pA/%	*
INPUT IMPEDANCE		
Differential	40kΩ‖0.01μF	*
Common Mode	10^9 Ω‖0.02μF	*
INPUT NOISE		
Voltage, 0.01 to 1Hz, p-p	0.4μV max	*
0.01 to 10Hz, p-p	1.0μV max	*
Current, 0.01 to 1Hz, p-p	8pA	*
0.01 to 10Hz, p-p	20pA	*
INPUT VOLTAGE RANGE		
Common Mode Voltage	±0.5V min	±1.0V min
Common Mode Rejection	300,000	*
Max Safe Differential Voltage	±20V	*
Max Safe Common Mode Voltage	±20V	*
POWER SUPPLY[3]		
Voltage, Rated Specification	±(14 to 16)V	*
Voltage, Operating	±(13 to 18)V	*
Current, Quiescent	±7mA	*
TEMPERATURE RANGE		
Rated Specifications	0 to +70°C	*
Operating	-25°C to +85°C	*
Storage	-55°C to +125°C	*
MECHANICAL		
Case Size	1.5" x 1.5" x 0.62"	*
Mating Socket	AC1022	*
Weight	1.75 oz. (50g)	*

[1] See applications information.
[2] Ground trim terminal if trim potentiometer is not used.
[3] Recommended power supply, ADI model 904, ±15V @ 50mA output
*Specifications same as for model 261J.
Specifications subject to change without notice.

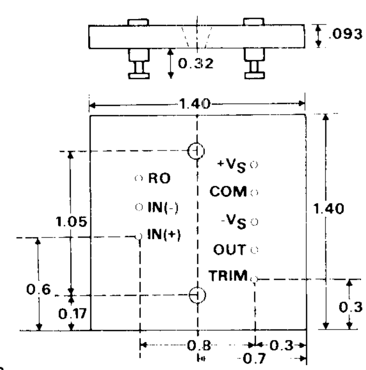
OUTLINE DIMENSIONS

Dimensions shown in inches and (mm).



Notes
1. Epoxy Case.
2. Pins 0.04 Dia. $^{+0.002}_{-0.000}$ half hard Brass, Gold plate
3. Amplifier mates with socket AC1022
4. Trim potentiometer ADI part number 79PR50K.
5. Markings do not appear on unit.
6. Trim terminal should be grounded if potentiomet is not used.
7. See Note 1 below for cap value. Use Polycarbonat Mylar, Mica, Glass, or Polystyrene capacitor for best performance.

MATING SOCKET AC1022



Notes
1. R.O. is connection for compensation capacitor.
2. Bottom View Shown.
3. Mounting holes 0.141 Dia., countersunk 82° to 0.23" Dia.
4. All in line pins spaced 0.2".
5. Dimensions in inches.
6. Markings printed on socket.

OTHER ULTRA LOW DRIFT, LOW NOISE AMPLIFIERS

Model 43K: This ultra low noise differential F amplifier has guaranteed noise performance o 2μV p-p max in a 10Hz B.W. and 3μV rms mε in a 50kHz B.W. Drift is 5μV/°C max.

Model 235: Chopper stabilized amplifier has noise of less than 1μV p-p and drift is only 0.1μV/°C (235L)

Model 184: A chopperless differential input amplifier with 0.25μV/°C drift and 1MHz bai width. Noise is 1μV p-p in a 0.01 to 1Hz B.W and 4μV rms in a 50kHz B.W.

NON-INVERTING VS. INVERTING OPERATION

The major limitation of the standard inverting type chopper stabilized amplifier is due to the practical limit on input impedance resulting from input bias current characteristics. If one attempts to obtain 10^7 ohms input impedance by using a 10^7 ohm input resistor with an inverting amplifier, this resistor will convert input current drifts of 0.5pA/°C into equivalent voltage drifts of 5μV/°C. It will also add Johnson Noise of 2.5μV p-p/$\sqrt{\text{Hz}}$ to the amplifier's input. These results negate the advantage of selecting the chopper-stabilized amplifier in the first place. Noise current will similarly increase the input uncertainty: inverting amplifier input noise currents of 10pA become 100μV noise voltages (referred to input). Furthermore, uncompensated initial bias currents of 50pA cause additional offsets of 500μV. Due to the non-inverting configuration of the model 261, these limitations are avoided. The input bias current (with its drift and noise) flows only through the signal source impedance, effectively eliminating the multiplication of drift and noise and offset caused by the input resistor in the inverting configuration. These benefits of the model 261 are shown graphically in Figure 1. When required input impedance is more than 300,000 ohms, the model 261 gives increasingly superior performance. One additional advantage is that the gain-setting precision resistors can be low-cost low value resistors instead of the more costly high resistance values.

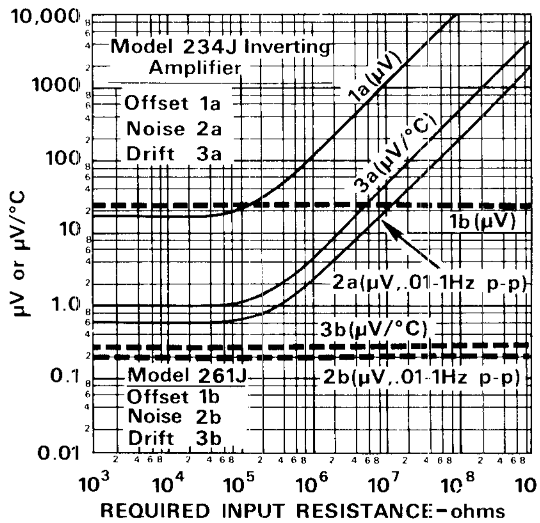


Figure 1. Offset, Drift, & Noise vs. Required Input Resistance

ION-INVERTING AMPLIFIER SELECTION CRITERIA

Model 261 vs. chopperless amplifiers)

n selecting an amplifier for low drift, one of the major conliderations is the effect of source resistance. For low values of ource resistance, the total offset (or drift) for differential mplifiers is essentially equal to the amplifier's offset (or drift) oltage. At the value of source resistance equal to the ratio of ffset voltage to difference current, or offset voltage drift to ifference current drift, the respective current is contributing n error equal to its corresponding voltage error. For values of ource resistance larger than this calculated value, the current rror's contribution will be dominant. In this section, model 261's drift and offset are compared with two low drift chopperless differential amplifiers, one with FET input, the other with bipolar input.

Fixed Source Resistance. If source resistance is fixed, bipolar chopperless amplifiers not having internal bias current drift compensation can benefit by the use of a compensating resistor in series with the (–) input. Under these conditions, Figure 2 shows a comparison of total drift /°C vs. source resistance for the model 261K, the model 184L low drift bipolar amplifier, and the model 52K low drift FET amplifier. For source resistances up to 200,000 ohms, the model 261 gives the lowest temperature drift. Total drift (R.T.I.) is equal to:

$$\Delta E_{in}/\Delta T = \Delta E_{OS}/\Delta T + R_S\,(\Delta I_{OS}/\Delta T)\ \text{(Models 184, 52)}$$

$$\Delta E_{in}/\Delta T = \Delta E_{OS}/\Delta T + R_S\,(\Delta I_b/\Delta T)\ \text{(Model 261)}$$

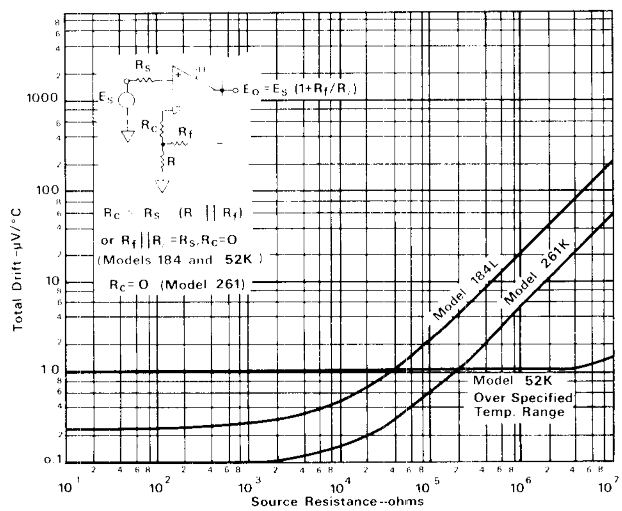


Figure 2. Offset Drift vs. Fixed Source Resistance

Variable Source Resistance. For situations where the source resistance can vary over a significant range, for instance when the amplifier's source is a multiturn potentiometer, a different set of conditions apply. The effective drift current to be considered for differential chopperless amplifiers is now the Input Bias Current/°C, rather than the Input Difference Current/°C (I_{OS}/°C). A two to one improvement in drift for the model 184 can be obtained if the bias current balancing resistor (R_C in Figure 3) is made equal to the mean value of the source resistance. Under these conditions, drift will be approximately:

$$\Delta E_{in}/\Delta T = \Delta E_{OS}/\Delta T + \tfrac{1}{2}\,(R_S\max - R_S\min)\,(\Delta I_b/\Delta T) + R_C\,(\Delta I_{OS}/\Delta T)\ \text{(Models 184, 52K)}$$

$$\Delta E_{in}/\Delta T = \Delta E_{OS}/\Delta T + (R_S\max)\,(\Delta I_b/\Delta T)\ \text{(Model 261)}$$

Figure 3 shows the total drift for a source resistance varying from zero to the chosen value, for models 52K, 184L and 261K. For values up to 200,000 ohms the model 261 again gives the lowest total temperature drift.

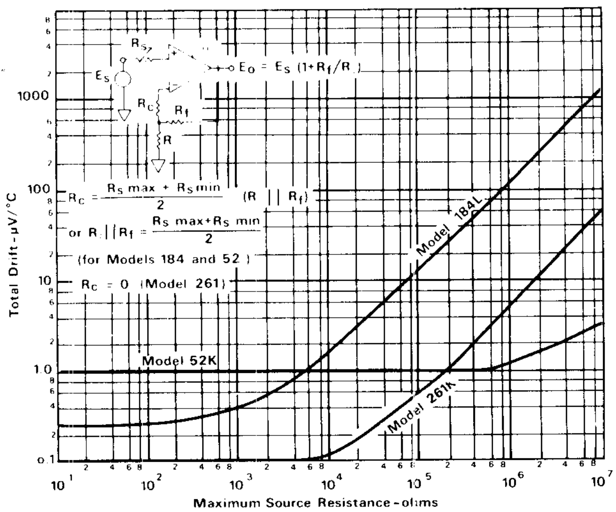


Figure 3. Offset Drift vs. Variable Source Resistance

INITIAL OFFSET

An initial offset voltage will develop due to bias current flowing through the source impedance. For fixed source impedances, this offset may be zeroed out in differential chopperless amplifiers not having internal bias current drift compensation by the use of the series compensating resistor, R_C shown in Figure 4. This offset should not be nulled out by adjusting the amplifier's offset trim because this will increase the offset voltage drift. With the model 261, however, all offsets may be zeroed out by means of the trim potentiometer. For variable source impedances, the offset should be zeroed out with the source impedance at its mean value. Figure 4 is a plot of the maximum offset which will occur with a given range of R_S variations, assuming the offsets are zeroed when operating with the mean value of R_S. Initial offset due to R_S is $I_b/2$ (R_S max - R_S min).

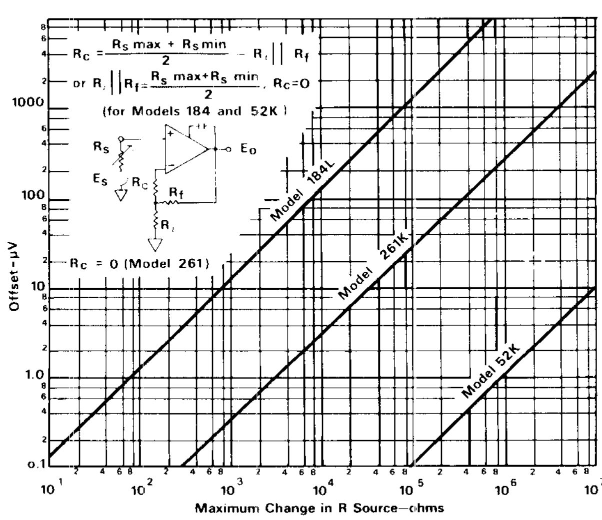


Figure 4. Offset vs. ΔR_S

LONG TERM DRIFT

Offset voltage of any amplifier will show some change with time, due to normal component aging. It is important to realize that the published drift for amplifiers does not accumulate linearly with increasing time. For example, the voltage drift of the model 261 is specified as ±½µV/month (a calculated figure believed to be quite conservative). For calculation of random long term drift, a rule of thumb is that one should multiply drift by the square root of the time factor increase. For the model 261, this yields a conservative long term drift of less than 2µV per year.

NOISE

A major criterion in the selection of an amplifier for low level signals is the amplifier input noise, since this is usually the limiting factor on system resolution. This is particularly important whenever high source impedances are encountered, since current noise through the source impedance will appear as an additional voltage noise, combining with the basic amplifier voltage noise and Johnson noise of the resistor. The sum of these noise sources will then be amplified along with the desired signal. For this reason, special care has been taken to reduce noise voltage and current to a level far below that of comparable chopper amplifiers. The one Hertz bandwidth noise voltage and current are 0.4µV p-p max and 8pA p-p respectively. For 10Hz bandwidth, corresponding values are 1µV p-p max and 20pA p-p. Figure 5A is a graph of noise vs. bandwidth for both current and voltage. Figure 5B is a spectral density plot for determining spot noise at any frequency. Figure 6 is a plot of peak to peak noise which will be encountered for these bandwidths, as a function of source resistance.

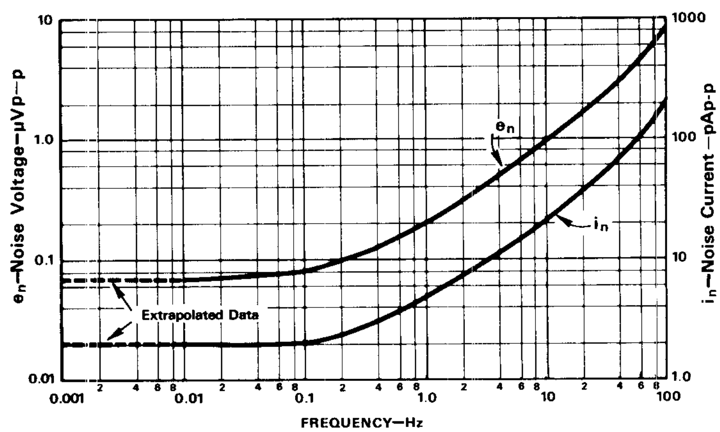


Figure 5A. Noise Current and Voltage vs. Bandwidth. Measured from 0.01Hz.

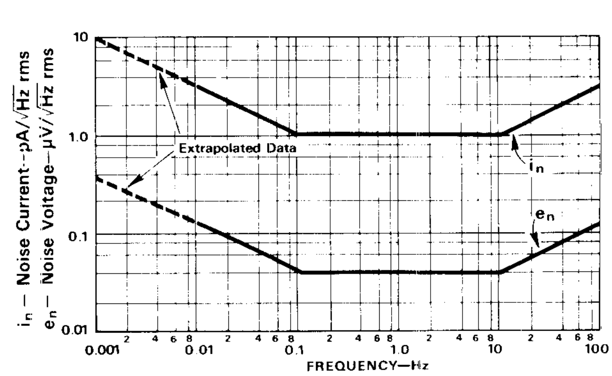


Figure 5B. Spectral Density of Current and Voltage

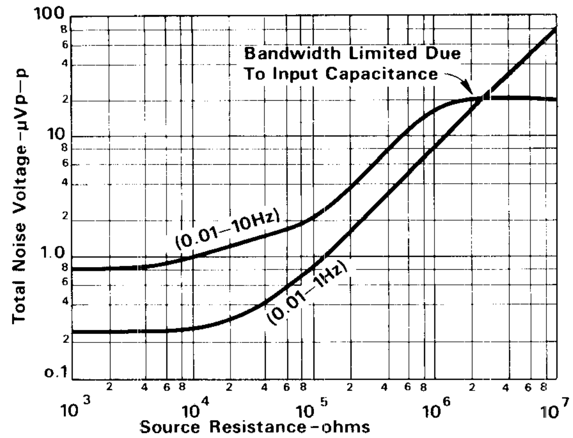


Figure 6. Total Noise vs. Source Resistance

HOW THE MODEL 261 OPERATES

As shown in Figure 7, the model 261 consists of five specific circuit functions. The input signal is fed through a resistor to the MOSFET Chopper. When the MOSFET is off (high resistance), the error signal appears at the input to the ac-coupled amplifier. When the MOSFET transistor is on (low resistance), the input to this amplifier is reduced to near zero. The difference between the on and off voltages at the amplifier is a square wave of amplitude slightly less than the error voltage. The attenuating effect of the MOSFET Chopper's "on" resistance is negligible. For example, if the attenuation were as much as 10%, the only effect would be to lower the potential open loop gain of the amplifier by the same amount.

The ac-coupled amplifier, consisting primarily of a linear integrated circuit, amplifies the resulting chopper error signal. Its output is capacitively coupled into a synchronous demodulator which reconstructs the low frequency-dc input signal, preserving polarity information. The drift of the input stage is not present in the demodulated signal since it was not chopped by the input network. The demodulated signal is filtered and further amplified by the integrator connected output dc amplifier.

Using the system just described, the reamining drift and offset, referred to the amplifier input, is equal to the output dc amplifier stage input drift and offset divided by the ac-coupled amplifier's gain. If the output stage integrated circuit amplified had a $100\mu V/^\circ C$ drift, and the ac-coupled amplifier gain is 1000, then the drift, referred to the input will be $0.1\mu V/^\circ C$ (the specification for the model 261K). The same considerations apply for offset voltage, accounting for its low value and the excellent long term stability of this amplifier.

The chopping signal is generated by a standard multivibrator. The frequency is not critical, and the multivibrator circuit is protected against latch-up.

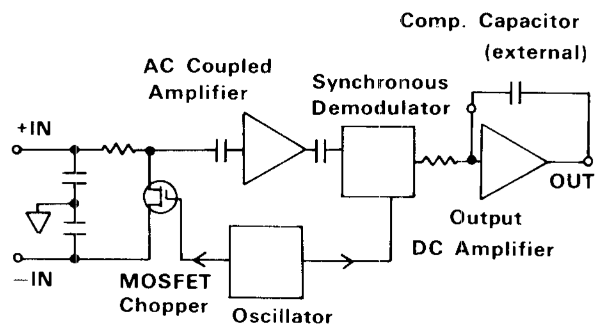


Figure 7. Model 261 Block Diagram

APPLICATION NOTES

Measurement of small signals or accurate handling of larger signals always requires care. Model 261 was specifically designed to minimize the problems raised by dc drift. To obtain best results, it is necessary to maintain good engineering practice and to observe a few requirements for optimizing performance of this precision instrument.

OFFSET VOLTAGE AND CURRENT TRIM

With the trim terminal connected to common, initial offset voltage of the model 261 is less than $25\mu V$. An additional offset voltage is developed by the flow of the input bias current through the resistance of the signal source. With a 10,000 ohm source resistance and worst case bias current of 300pA, the maximum additional offset voltage would be only $3\mu V$. For many applications, these offset voltages may be ignored, and the expense of a trim potentiometer and its adjustment is avoided. If the application requires lower offsets, an external 50,000 ohm trim potentiometer may be connected to zero the offset voltages, as shown previously. This trimming operation will not affect the drift or noise characteristics of the model 261.

INVERTING AND DIFFERENTIAL INPUT OPERATION

The input current to the amplifier's (–) terminal is less than ±10nA. Differential input operation of the amplifier is allowable, but the impedance from the inverting terminal to ground should not exceed 5000 ohms, and the common mode voltage range for best performance should not be exceeded. For purely inverting applications the user should select Analog Devices' models 234 or 235 chopper stabilized amplifiers, which are optimized for inverting operation.

SELECTABLE BANDWIDTH

For practical low-frequency applications, the model 261 uses an external compensation capacitor to determine the gain-bandwidth product. Its value may be chosen to allow the use of the maximum 100Hz -3dB bandwidth, at any given value of closed loop gain. By using a larger value of compensation capacitance, the bandwidth can be limited to any desired value below 100Hz, as required by the application. The minimum value of the required compensation capacitor, in μF, is 1000/GB, where G is the desired closed-loop dc gain, and B is the -3dB bandwidth. For example, the minimum value of recommended capaçitance (for 100Hz bandwidth to -3dB) is

10/G. Shown in Figure 8 are curves of the amplifier's response for various closed loop gains while using values of capacitance appropriate for maintaining 100Hz (−3dB) bandwidth. Figure 9 illustrates the amplifier's open-loop response with various values of the compensation capacitor. It is recommended that the capacitor be polycarbonate, mylar, mica, glass or polystyrene for best performance.

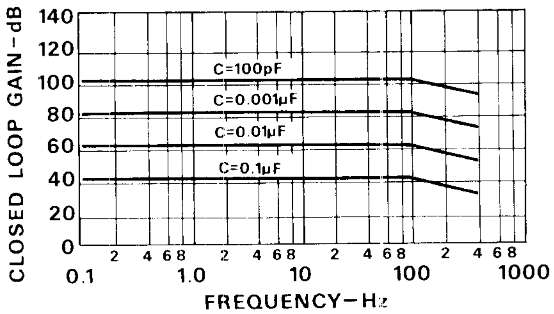


Figure 8. Compensation vs. Gain for 100Hz Bandwidth

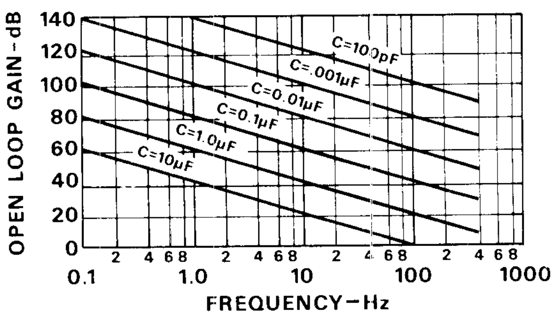


Figure 9. Open Loop Response vs. Compensation

FULL POWER RESPONSE

Full power output at any frequency can be obtained only with closed loop gains exceeding 10. This is due to the common mode voltage limitation described below.

The maximum full power output frequency is 50Hz, and will be obtained when using compensation capacitors of less than 0.013μF. For larger compensation capacitors, f_p is given by the formula:

$$f_p\ (\text{Hz}) = 0.66/C\ (\mu\text{F})$$

When using a low gain, for instance 10, the maximum f_p will be 0.66Hz due to the 1.0μF required compensation capacitor for this closed loop gain. Under such conditions the user may wish to employ the compensation circuit of Figure 10. This will increase f_p to 5Hz (for a gain of 30). For higher gains, an increase in f_p will be obtained (with the same circuit), although the rise in f_p will not be proportionately as large.

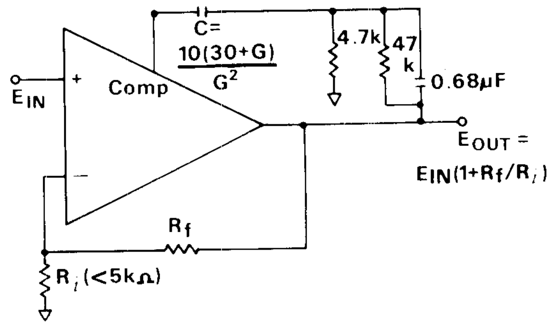


Figure 10. Compensation for Increased f_p

COMMON MODE CONSIDERATIONS

In the model 261, the maximum safe input voltage, both differential and common mode, exceeds ±20 volts. However, in order to maintain the specified Common Mode Rejection Rati of 300,000 it is necessary that common mode voltage be limited to ±1.0 volts for the model 261K and ±0.5 volts for the model 261J. These values will not be exceeded by normal input signal swings if the amplifiers closed loop gain exceeds 10 and 20, respectively. Since most applications will use this amplifier at gains of 100 or more, the specified common mod range should prove entirely adequate.

COMMON MODE REJECTION

Model 261 is designed to provide high stability, high gain and low noise in non-inverting applications where the high input impedance minimizes input signal attenuation. Although operation as a differential amplifier is possible, it is not recommen

In the non-inverting mode, there is a source of error due to the common mode voltage; however this error term can be completely ignored since the error due to open loop gain will dominate.

INVERTING INPUT TERMINAL RESISTANCE (R_i)

An attempt should be made to maintain low resistance from the inverting input terminal to ground. This will prevent the negative input's bias current from degrading the offset performance of the amplifier. This restriction in no way relates to the source resistance seen by the positive (non-inverting) input terminal.