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EXAR

OPERATIONAL AMPLIFIER DATA BOOK

First in Quality...First in Service • Custom, Semicustom and Standard IC's

Introduction

This Data Book contains a complete summary of technical information covering Exar's entire line of monolithic IC operational amplifier products. In addition, several design and applications articles are also included, along with a review of fundamentals of IC op-amps. To help the designer to choose the right op-amp for his application, a number of convenient cross-reference charts are also included which show the key features of each of the products discussed, in terms of different classes of applications.

EXPERIENCE AND PRODUCTS

Exar's innovativeness, product quality and responsiveness to customer needs have been the key to its success. Exar today offers a broad line of linear and interface circuits. In the field of standard linear IC products, Exar has extended its circuit technological leadership into the areas of communications and control circuits. Today Exar has one of the most complete lines of IC oscillators, timing circuits and phase-locked loops in the industry. Exar also manufactures a large family of telecommunication circuits such as tone decoders, compandors, modulators, PCM repeaters and FSK Modem Circuits. In the field of industrial control circuits, Exar manufactures a broad line of quad and dual operational amplifiers,voltage regulators, radio-control and servo driver IC's, and power control circuits.

Exar's experience and expertise in the area of bipolar IC technology extends both into custom and standard IC products. In the area of custom IC's, Exar has designed, developed, and manufactured a wide range of full-custom monolithic circuits, particularly for applications in the areas of telecommunications, consumer electronics, and industrial controls.

In addition to the full-custom capability, Exar also offers a unique semi-custom IC development capability for low to medium-volume custom circuits. This semi-custom program, is intended for those customers seeking cost-effective solutions to reduce component count and board size in order to compete more effectively in a changing marketplace. The program allows a customized monolithic IC to be developed with a turnaround time of several weeks at a small fraction of the cost of a full-custom development program.

EXCELLENCE IN ENGINEERING

Exar quality starts in Engineering where highly qualified people are backed up with the advanced instruments and facilities needed for design and manufacture of custom, semi-custom and standard integrated circuits. Exar's engineering and facilities are geared to handle all three classes of IC design: (1) semi-custom design programs using Exar's bipolar and I^2L master chips; (2) full-custom IC design; (3) development and high-volume production of standard products.

Some of the challenging and complex development programs successfully completed by Exar include analog compandors and PCM repeaters for telecommunication, electronic fuel-injection, anti-skid braking systems and voltage regulators for automotive electronics, digital voltmeter circuits, 40-MHz frequency synthesizers, high-current and high-voltage display and relay driver ICs, and many others.

NEW TECHNOLOGIES

Through company sponsored research and development activities, Exar constantly stays abreast of all technology areas related to changing customer needs and requirements. Exar has recently completed development efforts in Integrated Injection Logic (I^2L) technology, which offers unique advantages in the area of low-power, high-density logic arrays. Exar has a complete design engineering group dedicated to this new technology, and is currently supplying over twenty different custom and semi-custom I^2L products.

FIRST IN QUALITY

From incoming inspection of all materials to the final test of the finished goods, Exar performs sample testing of each lot to ensure that every product meets Exar's high quality standards. Exar's manufacturing process is inspected or tested in accordance with its own stringent Quality Assurance Program, which is in compliance with MIL-Q-9858A. Additional special screening and testing can be negotiated to meet individual customer requirements.

Throughout the wafer fab and assembly process, the latest scientific instruments, such as scanning electron microscopes, are used for inspection, and modern automated equipment is used for wafer probe, AC, DC, and functional testing. Environmental and burn-in testing of finished products is also done in-house. For special environmental or high reliability burn-in tests outside testing laboratories are used to complement Exar's own extensive in-house facilities.

FIRST IN SERVICE

Exar has the ability and flexibility to serve the customer in a variety of ways from wafer fabrication to full parametric selection of assembled units for individual customer requirements. Special marking, special packaging and military screening are only a few of the service options available from Exar. We are certain that Exar's service is flexible enough to satisfy 99% of your needs. The company has a large staff of Applications Engineers to assist the customer in the use of the product and to handle any request, large or small.

Exar reserves the right to make changes at any time in order to improve design and to supply the best product possible.

Exar cannot assume responsibility for any circuits shown or represented, as being free from patent infringement.

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Fundamentals of Operational Amplifiers

The "ideal" operational amplifier can be defined as a voltage-controlled voltage amplifier circuit which offers infinite voltage gains with an infinite input impedance, zero output impedance, and infinite bandwidth. The advantage of such an idealized block of gain is that one can perform a large number of mathematical "operations", or generate a number of circuit functions by applying passive feedback around the amplifier.

The key features of operational amplifier application can be illustrated using the simple feedback circuit of Figure 1, and assuming that the operational amplifier has infinite gain and infinite input impedance. Then, the following two conditions have to be satisfied:

a) Since the voltage gain is infinite, the net voltage across the input terminals of the operational amplifier must be zero, if the operational amplifier output voltage is to be finite. In the circuit of Figure 1, this causes the inverting input terminal of the operational amplifier to behave as a "virtual ground".

b) Since the input impedance of the ideal operational amplifier is infinite, no input current is drawn by the operational amplifier, the total current going into the circuit node connected to the inverting input of the operational amplifier (node Q in Figure 1) must be equal to the total current coming out, i.e.:

$$I_S = -I_F \quad \text{and} \quad \frac{V_{IN}}{R_S} = -\frac{V_O}{R_F} \tag{1}$$

Solving for the overall voltage gain, one obtains:

$$A_V = \frac{V_{OUT}}{V_{IN}} = -\frac{R_F}{R_S} \tag{2}$$

Because of this property, the noninverting input of an operational amplifier is often referred to as its "summing input".

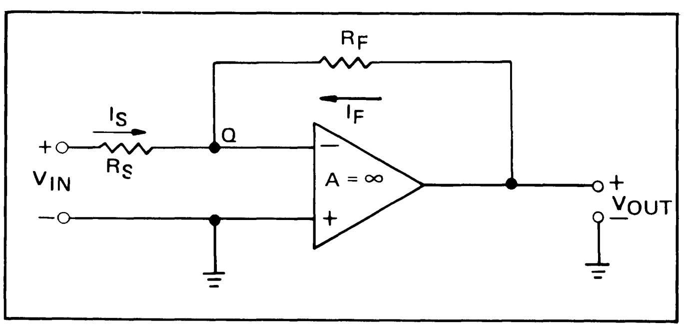


Figure 1. The "Ideal" Operational Amplifier as a Feedback Amplifier.

In the case of actual operational amplifiers, both the voltage gain and the input impedance are quite high, but still finite. Figure 2 shows the same basic feedback circuit assuming that the amplifier now has a finite input resistance, R_{IN}, and a finite voltage gain A. For simplicity, the output impedance of the operational amplifier is assumed to be negligible. The overall voltage gain of the circuit can now be expressed as:

$$A_V = V_{OUT}/V_{IN} = -\frac{R_F}{R_S}\left[\frac{1}{1 + \frac{1}{A}(1 + R_F/R_S + R_F/R_{IN})}\right] \tag{3}$$

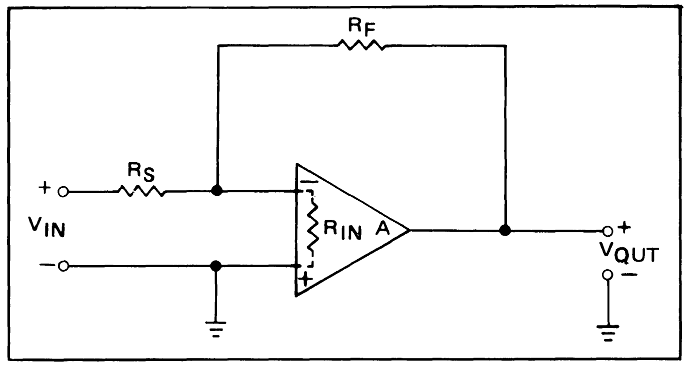


Figure 2. Basic Feedback Configuration Using an Operational Amplifier With Finite Input Impedance and Gain.

It should be noted that, for large values of R_{IN}, as the voltage gain increases (i.e. $A \to \infty$), this expression rapidly converges to that given in equation 2; and the circuit performance becomes solely determined by the external components.

In addition to having finite gain and input impedance, an actual operational amplifier circuit also has finite input bias currents as well as input offset voltage and currents. A more complete model of a practical operational amplifier is shown in Figure 3 where I_B indicates the finite input bias currents; V_{io} and I_{io} represent the voltage and current offsets associated with the circuit and R_O is the output resistance. Due to non-zero values of V_{io} and I_{io} in a practical operational amplifier circuit, $V_{OUT} \neq 0$ for $V_{IN} = 0$.

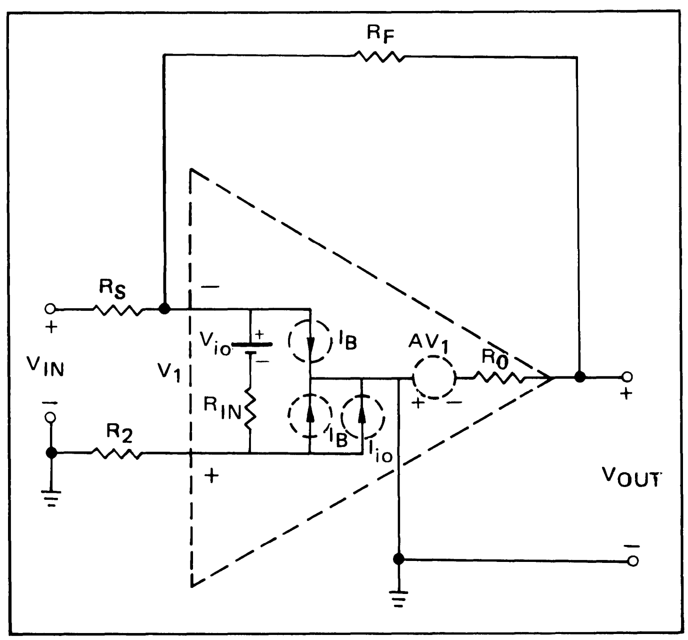


Figure 3. Equivalent Circuit of a Practical Operational Amplifier Showing the Effects of Finite Input Impedance, Current and Voltage Offsets.

Definitions of Operational Amplifier Terms

Since the operational amplifier has become a universal building block for circuit and system design, a number of widely accepted design terms have evolved which describe the comparative merits of various operational amplifiers. Some of these terms are defined below:

Input Offset Voltage: The input voltage which must be applied across the input terminals to obtain zero output voltage.

Input Offset Current: The difference of the currents into the two input terminals with the output at zero volts.

Input Bias Current: The average of the two input currents.

Input Common-Mode Range: Maximum range of input voltage that can be simultaneously applied to both inputs without causing cutoff or saturation of amplifier gain stages.

Common-Mode Rejection Ratio: Ratio of the differential open-loop gain to the common-mode open-loop gain.

Supply Voltage Rejection Ratio: Input offset voltage change per volt of supply voltage change.

Input Resistance: The ratio of the change in input voltage to the change in input current on either input with the other grounded.

Supply Current: The current required from the power supply to operate the amplifier with no load and the output at zero.

Output Voltage Swing: The peak output voltage swing, referred to zero, that can be obtained without clipping.

Large-Signal Voltage Gain: The ratio of the output voltage swing to the change in input voltage required to drive the output from zero to this voltage.

Full-Power Bandwidth: Maximum frequency over which the full output voltage swing can be obtained.

Unity-Gain Bandwidth: Frequency at which the open loop voltage gain is equal to unity.

Slew Rate: The maximum time rate of change of the output voltage, for a voltage step applied to the input. It is normally measured at the zero crossing point of the output voltage swing with the amplifier frequency compensated for unity gain.

Overload Recovery Time: Time required for the output stage to return to active region, when driven into hard saturation.

Gain Margin: The amount by which the voltage gain is below the unity (0 dB) level, at the frequency where the *excess* phase shift across the amplifier is exactly 180°. It is measured in decibels, and must be positive for unconditional stability.

Phase Margin: 180° minus the excess phase shift at the frequency where the magnitude of the open loop voltage gain is equal to unity. It is measured in degrees and must be positive for unconditional stability.

Basic Applications of Operational Amplifiers

The general usefulness of the operational amplifier stems from the fact that when used in a feedback loop, its overall performance and transfer characteristics are determined almost totally by the choice of feedback components. To be universally useful in such an application, the "ideal" operational amplifier should exhibit infinite gain, infinite input impedance and infinite bandwidth. Although these are all idealized characteristics, the practical monolithic operational amplifiers closely approximate these features, particularly for low frequency applications.

The availability and the low-cost of the integrated operational amplifier makes it an extremely versatile building block for analog system or equipment design. Therefore, it is mandatory that the circuit designer be familiar with the fundamental applications of operational amplifiers. This section of Exar's Operational Amplifier Data Book is intended to familiarize the designer with some of the simple but fundamental circuit configurations using IC operational amplifiers. The discussion is slanted toward the practical applications of operational amplifiers, as controlled by the external feedback circuitry. The particular operational amplifier parameters will be discussed as they effect the circuit performance and accuracy.

The integrated operational amplifiers shown in the figures are for the most part internally compensated, so frequency stabilization components are not shown; however, other amplifiers using external compensation may be utilized to achieve greater operating speed in many circuits.

The Inverting Amplifier

The basic operational amplifier circuit is shown in Figure 1. This circuit gives closed-loop gain of R_2/R_1 when this ratio is small compared with the amplifier open-loop gain and, as the name implies, is an inverting circuit. The input impedance is equal to R_1. The closed-loop bandwidth is equal to the unity-gain frequency divided by one plus the closed-loop gain.

The only cautions to be observed are that R_3 should be chosen to be equal to the parallel combination of R_1 and R_2 to minimize the offset voltage error due to bias current; and that there will be a DC offset voltage at the amplifier output equal to closed-loop gain times the offset voltage at the amplifier input.

Offset voltage at the input of an operational amplifier is comprised of two components, these components are identified in specifying the amplifier as input offset voltage and input bias current. The input offset voltage is fixed for a particular amplifier; however, the contribution due to input bias current is dependent on the circuit configuration used. For minimum offset voltage at the amplifier input without circuit adjustment, the source resistance for both inputs should be equal. In this case, the maximum offset voltage would be the algebraic sum of amplifier offset voltage and the voltage drop across the source resistance due to offset current. Amplifier offset voltage is the predominant error term for low source resistances, and offset current causes the main error for high source resistances.

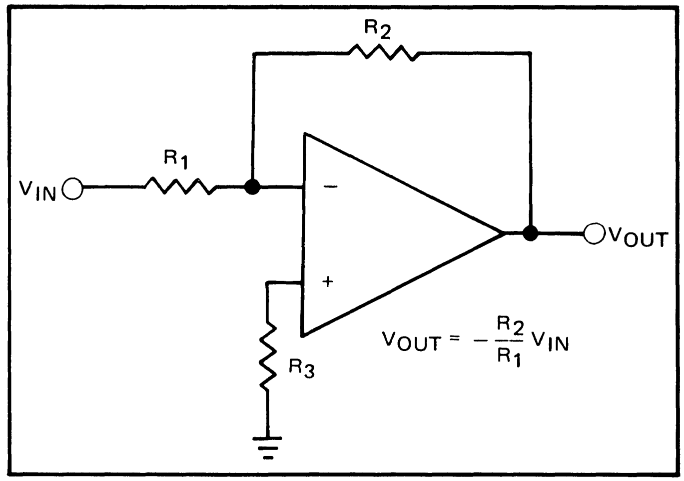


Figure 1. Inverting Amplifier

In high source resistance applications, offset voltage at the amplifier output may be adjusted by adjusting the value of R_3 and using the variation in voltage drop across it as an input offset voltage trim.

Offset voltage at the amplifier output is not as important in AC coupled applications. Here the only consideration is that any offset voltage at the output reduces the peak-to-peak linear output swing of the amplifier.

The gain-frequency characteristic of the amplifier and its feedback network must be such that oscillation does not occur. To meet this condition, the phase shift through amplifier and feedback network must never exceed 180° for any frequency where the combined gain of the amplifier and its feedback network is greater than unity. In practical applications, the phase shift should not approach 180° since this is the situation of conditional stability. Obviously, the most critical case occurs when the attenuation of the feedback network is zero.

Amplifiers which are not internally compensated may be used to achieve increased performance in circuits where feedback network attenuation is high, i.e., the amount of feedback around the amplifier is low. The compensation trade-off for a particular connection is stability versus bandwidth. Larger values of compensation capacitor yield greater stability and lower bandwidth and vice versa.

The Non-Inverting Amplifier

Figure 2 shows a high input impedance non-inverting circuit. This circuit gives a closed-loop gain equal to the ratio of $(R_1 + R_2)$ to R_1. Its closed-loop 3-dB bandwidth is equal to the amplifier unity-gain frequency divided by the closed-loop gain.

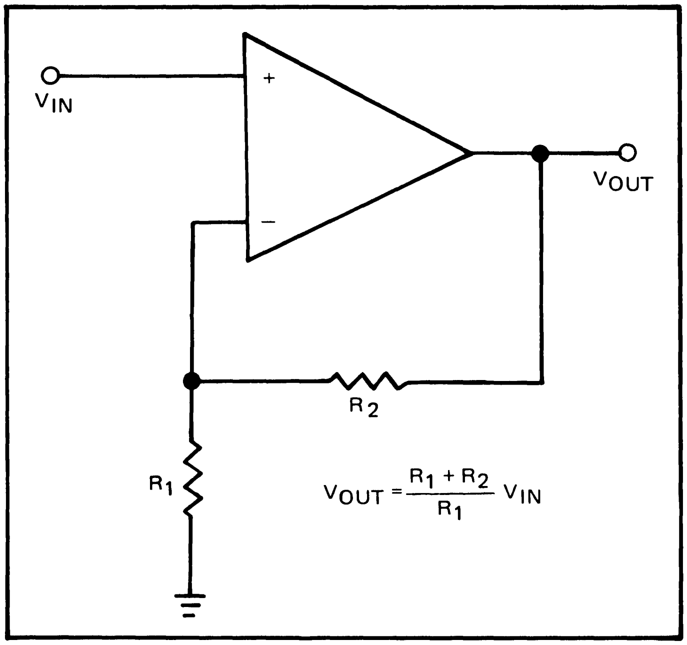


Figure 2. Non-Inverting Amplifier

The primary differences between this connection and the inverting circuit are that the output is not inverted and that the input impedance is very high and is equal to the differential input impedance multiplied by loop gain (open-loop gain/closed-loop gain). In DC coupled applications, input impedance is not as important as input current and its voltage drop across the source resistance. To minimize the output error due to the input bias current of the operational amplifier, $(R_1 + R_2)$ should be chosen equal to the source impedance of the input signal. Applications cautions are the same for this amplifier as for the inverting amplifier with one exception: the amplifier output will go into saturation if the input is allowed to float. This may be important if the amplifier must be switched from source to source. The compensation trade off discussed for the inverting amplifier is also valid for this connection.

The Unity-Gain Buffer

The unity-gain buffer is shown in Figure 3. The circuit gives the highest input impedance of any operational amplifier circuit. Input impedance is equal to the differential input impedance multiplied by the open-loop gain, in parallel with common mode input impedance. The gain error of this circuit is equal to the reciprocal of the amplifier open-loop gain or to the common-mode rejection, whichever is less. Input impedance is a misleading concept in a DC coupled unity-gain buffer. Bias current for the amplifier will be supplied by the source resistance and will cause an error at the amplifier input due to its voltage drop across the source resistance.

The cautions to be observed in applying this circuit are as follows: the amplifier must be compensated for unity-gain operation, and the output swing of the amplifier may be limited by the amplifier common-mode range. The input signal swing should not exceed the input common-mode range, since this may cause a latch-up condition.

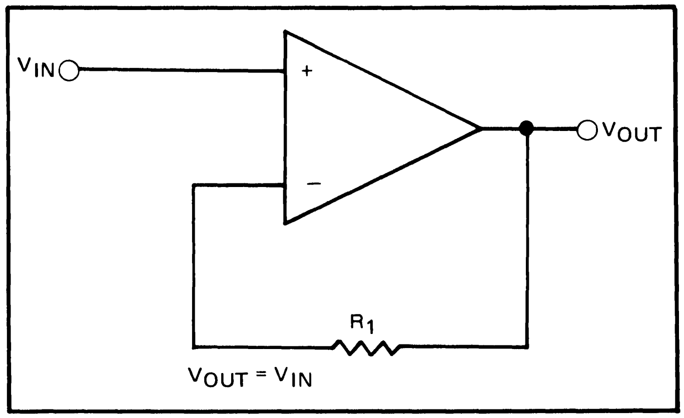


Figure 3. Unity-Gain Buffer

Summing Amplifier

The summing amplifier, a special case of the inverting amplifier, is shown in Figure 4. The circuit gives an inverted output which is equal to the weighted algebraic sum of all three inputs. The gain of any input of this circuit is equal to the inverse ratio of the appropriate input resistor to the feedback resistor, R_4. Amplifier bandwidth may be calculated as in the inverting amplifier shown in Figure 1 by assuming the input resistor to be the parallel combination of R_1, R_2, and R_3. Application cautions are the same as those for the inverting amplifier. If an uncompensated amplifier is used, compensation is calculated on the basis of this bandwidth as is discussed in the section describing the simple inverting amplifier.

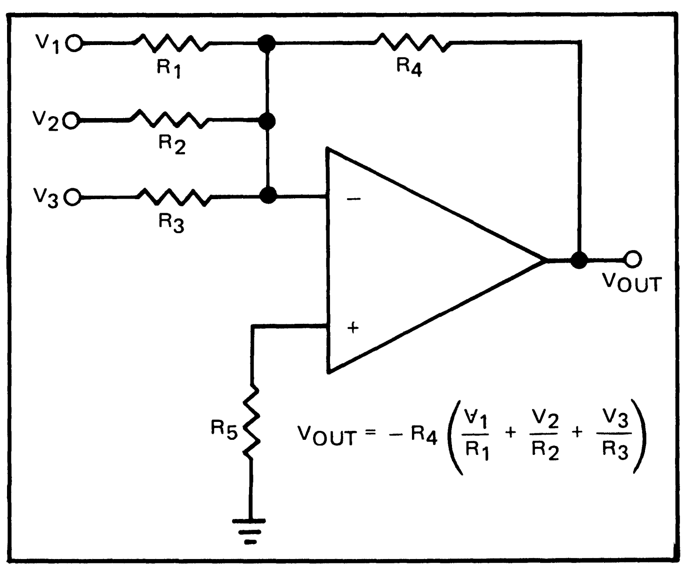


Figure 4. Summing Amplifier

The advantage of this circuit is that there is no interaction between inputs, therefore, operations such as summing and weighted-averaging are implemented very easily.

The Difference Amplifier

The difference amplifier is the complement of the summing amplifier and allows the subtraction of two voltages or, as a special case, the cancellation of a signal common to the two inputs. This circuit is shown in Figure 5 and is useful as a computational amplifier, in making a differential to single-ended conversion, or in rejecting an unwanted common-mode signal.

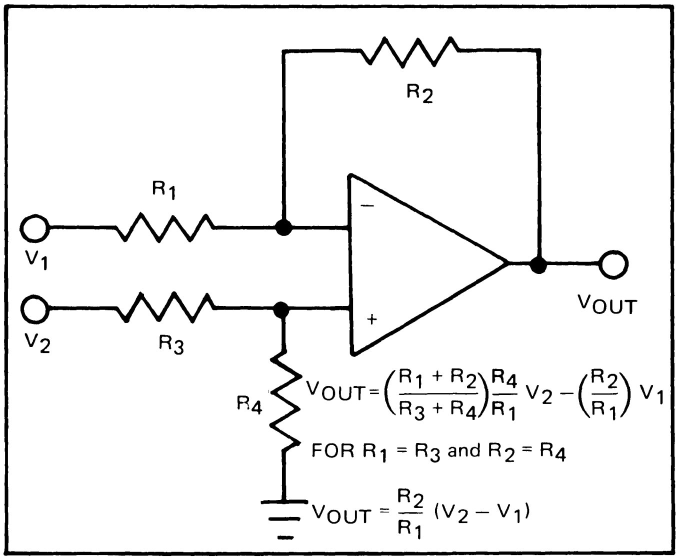


Figure 5. Difference Amplifier

Circuit bandwidth may be calculated in the same manner as for the inverting amplifier, but input impedance is somewhat more complicated. Input impedance for the two inputs is not necessarily equal: inverting input impedance is the same as for the inverting amplifier of Figure 1 and the non-inverting input impedance is the sum of R_3 and R_4. Gain for either input is the ratio of R_1 to R_2 for the special case of a differential input single-ended output where $R_1 = R_3$ and $R_2 = R_4$. The general expression for gain is given in the figure. Compensation should be chosen on the basis of amplifier bandwidth.

Care must be exercised in applying this circuit since input impedances are not equal for minimum bias current error.

Differentiator Circuit

The basic principle of a differentiator circuit is shown in the simplified connection diagram of Figure 6. However, although mathematically accurate, this particular connection is not directly useful in practice because it is extremely susceptible to high frequency noise since AC gain increases at the rate of 6 dB per octave. In addition, the feedback network of the differentiator made up of the resistor R_3 and the capacitor C_3 is an RC low pass filter which contributes 90° phase shift to the loop and may cause stability problems even with an amplifier which is compensated for unity-gain.

A practical differentiator which corrects the high frequency noise problem is shown in Figure 7. Here both the stability and noise problems are corrected by addition of two additional components, R_1 and C_2. R_2 and C_2 form a 6 dB per octave high frequency roll-off in the feedback network, and R_1C_1 form a 6 dB per octave roll-off network in the input network for a total high frequency roll-off of 12 dB per octave, to reduce the effect of high frequency input and amplifier noise. In addition R_1C_1 and R_2C_2 form lead networks in the feedback loop which, if placed below the amplifier unity-gain frequency, provide 90° phase lead to compensate the 90° phase lag of R_2C_1 and prevent loop instability.

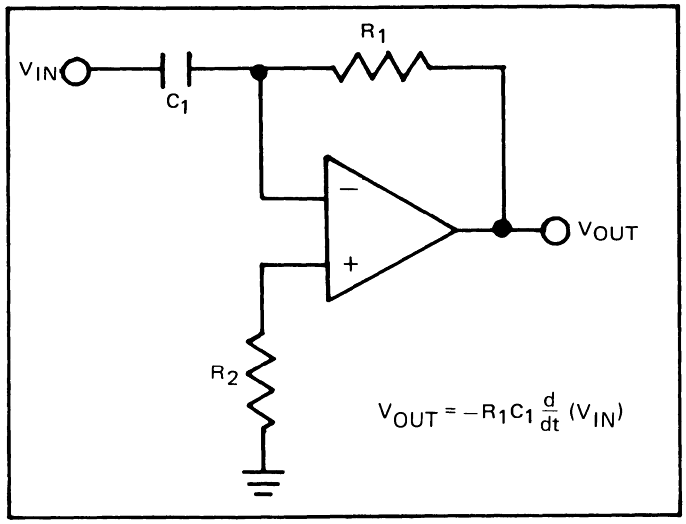


Figure 6. Basic Differentiator Connection

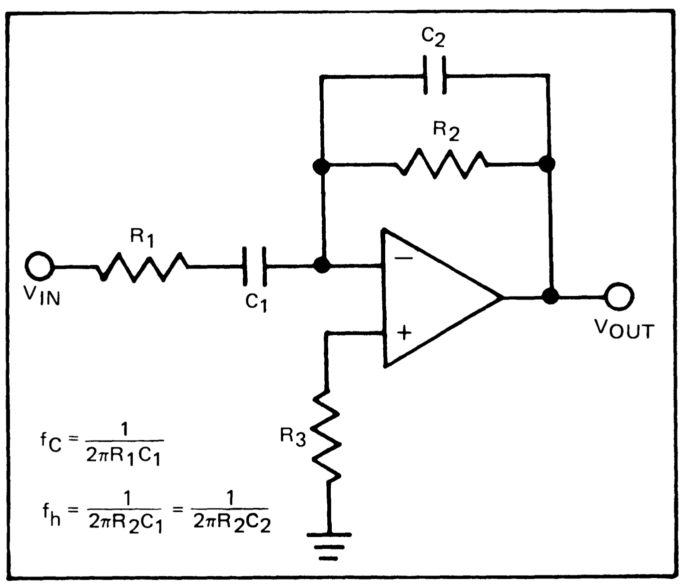


Figure 7. Practical Differentiator Circuit

Integrator Circuit

Figure 8 shows the basic circuit connection for performing the mathematical operation of integration. This circuit is essentially a low-pass filter with a constant frequency roll-off of −6 dB per octave.

The circuit must be provided with an external method of establishing initial conditions. This is shown in the figure as the double-pole, single-throw switch S_1. When S_1 is in position 1, the amplifier is connected in unity-gain configuration, and capacitor C_1 is discharged, setting an initial condition of zero volts. When S_1 is in position 2, the amplifier is connected as an integrator, and its output will be the time-integral of the input voltage.

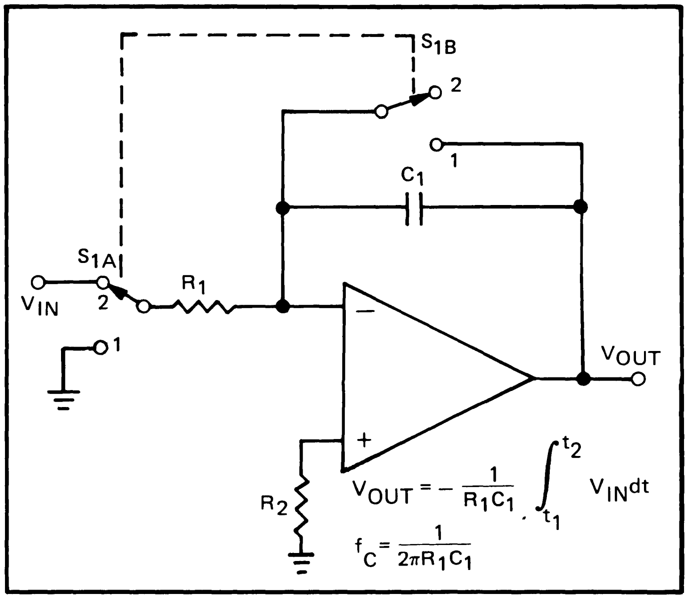


Figure 8. The Integrator Circuit

The cautions to be observed with this circuit are two: the amplifier used should generally be stabilized for unity-gain operation and R_2 must equal R_1 for minimum error due to bias current.

Simple Low-Pass Filter

The simple low-pass filter is shown in Figure 9. This circuit has a 6 dB per octave roll-off after a closed-loop 3-dB point defined by f_C. Gain below this corner frequency is defined by the ratio of R_3 to R_1. The circuit may be considered as an AC integrator at frequencies well above f_C; however, the time domain response is that of a single RC rather than an integral.

A gain vs. frequency plot of circuit response is shown in Figure 10 to illustrate the difference between this circuit and the true integrator. Note that the frequency response is flat for frequencies below f_C

$$\text{where } f_C = \frac{1}{2\pi R_3 C_1}$$

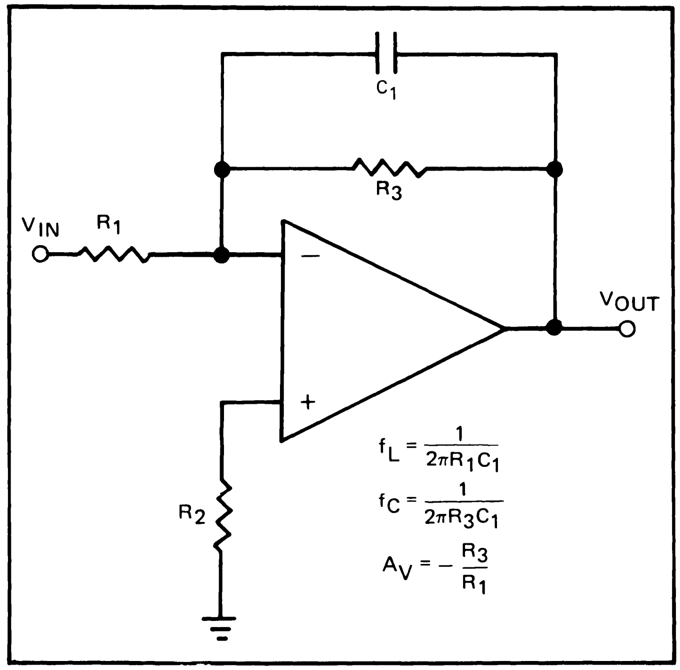


Figure 9. A Simple Low-Pass Filter Circuit

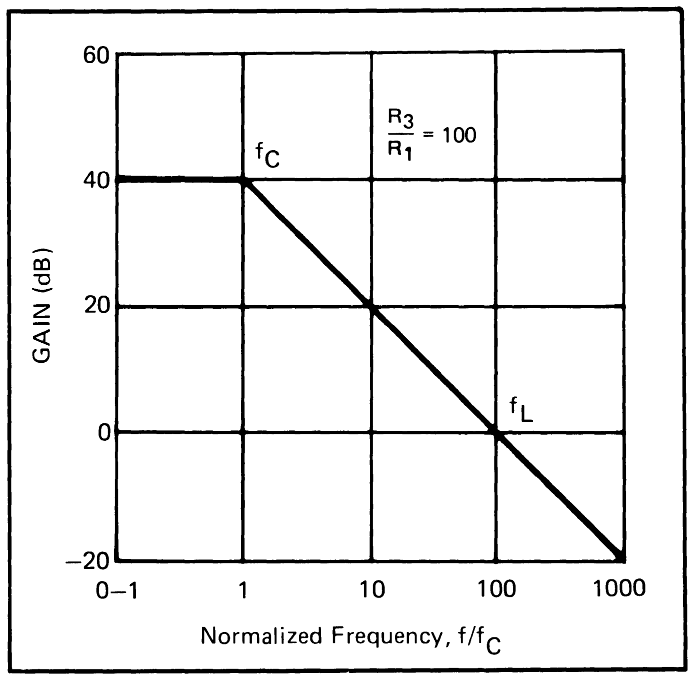


Figure 10. Frequency Response of the Simple Low-Pass Filter.

Current-to-Voltage Converter

Current may be measured in two ways with an operational amplifier: the current may be converted into a voltage with a resistor and then amplified or it may be injected directly into a summing node. Converting into voltage is undesirable for two reasons: first, an impedance is inserted into the measuring line causing an error; second, amplifier offset voltage is also amplified with a subsequent loss of accuracy. The use of a current-to-voltage converter avoids both of these problems.

The current-to-voltage converter is shown in Figure 11. The input current is fed directly into the summing note, and the amplifier output voltage changes to extract the same current from the summing node through R_1. The scale factor of this circuit is R_1 volts per ampere of current. The only conversion error in this circuit is the bias current of the operational amplifier input which is summed algebraically with the input current, I_{IN}. The main design constraints are that scale factors must be chosen to minimize errors due to bias current and since voltage gain and source impedance are often indeterminate (as with photocells) the amplifier must be compensated for unity-gain operation.

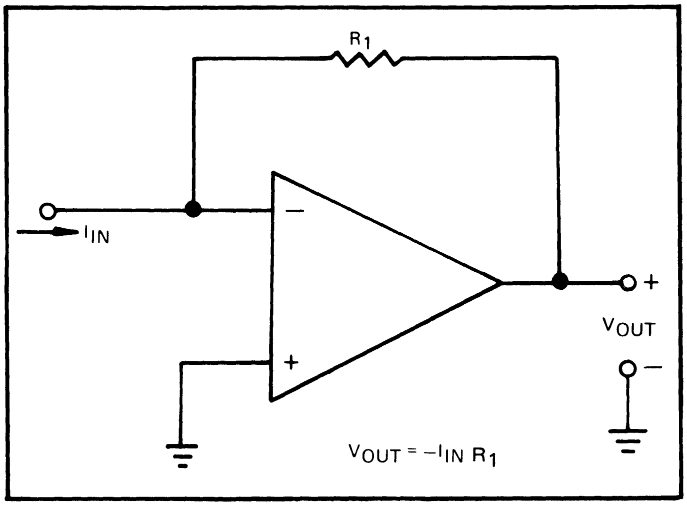


Figure 11. Operational Amplifier as a Current-to-Voltage Converter.

Voltage Controlled Current-Source

Figures 12, 13, and 14 show three simple circuit configurations for voltage-controlled constant-current stages. The circuit of Figure 12 is a basic current-sink circuit which uses a pair of Darlington connected NPN transistors external to the operational amplifier. Assuming that the base current of T_1 is negligible compared to the controlled current I_O, the current of the output transistors is equal to V_{IN}/R_1.

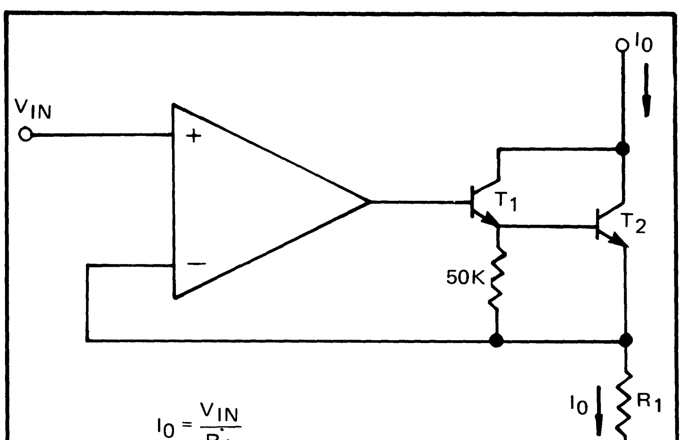


Figure 12. Voltage-Controlled Current-Sink Circuit

Figure 13 shows a current-source circuit which uses a composite connection of external PNP and NPN transistors and produces a constant output current which is proportional to the net voltage drop across the sensing resistor, R_1.

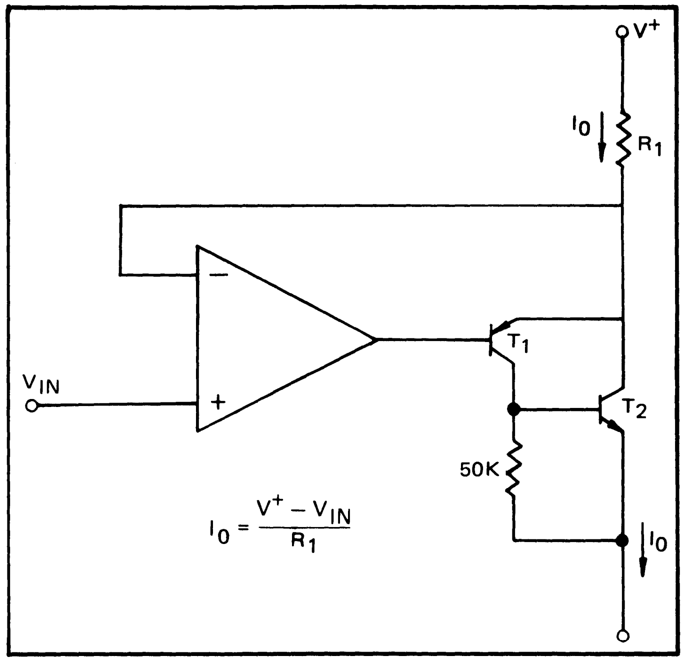


Figure 13. Voltage-Controlled Current-Source Circuit

Figure 14 shows an alternate approach to obtaining a voltage-controlled current source which does not require additional active devices. The circuit provides an output current proportional to the input voltage V_{IN}. If the resistors R_1 through R_4 are chosen to be equal and much larger than R_5, then the output current is:

$$I_{OUT} = V_{IN}/R_5$$

The above expression assumes that the current through R_3 is much smaller than I_O.

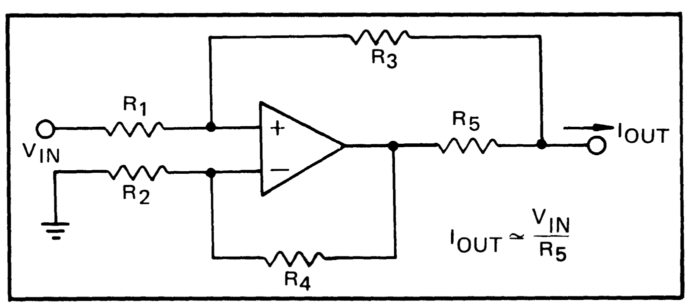


Figure 14. A Voltage-Controlled Current Source Circuit Which Does Not Require External Active Devices.

This circuit can supply an output current of either polarity, up to the maximum positive or negative output current available from the operational amplifier. The maximum voltage compliance of the output is limited by the output swing of the operational amplifier minus the voltage drop across the sensing resistor, R_5.

Triangle Wave Oscillator

A constant amplitude triangular wave generator is shown in Figure 15. This circuit provides a variable frequency triangular wave whose amplitude is independent of frequency. This entire circuit can be built inexpensively, using a dual operational amplifier IC, such as the XR-4558.

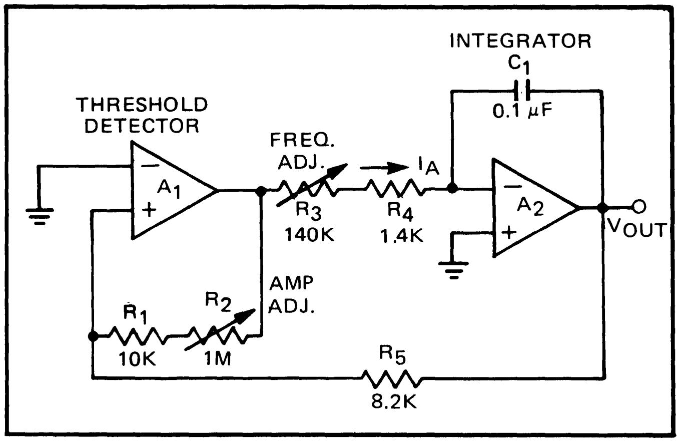


Figure 15. A Simple Triangle Wave Oscillator.

The generator embodies an integrator as a ramp generator and a threshold detector with hysterisis as a reset circuit. The integrator has been described in a previous section and requires no further explanation. The threshold detector is similar to a Schmitt trigger in that it is a latch circuit with a large dead zone. This function is implemented by using positive feedback around an operational amplifier. When the amplifier output is in either the positive or negative saturated state, the positive feedback network provides a voltage at the non-inverting input which is determined by the attenuation of the feedback loop and the saturation voltage of the amplifier. To cause the amplifier to change states, the voltage at the input of the amplifier must be caused to change polarity by an amount in excess of the amplifier input offset voltage. When this is done, the amplifier saturates in the opposite direction and remains in that state until the voltage at its input again reverses. The complete circuit operation may be understood by examining the operation with the output of the threshold detector in the positive state. The detector positive saturation voltage is applied to the integrator summing junction through the combination R_3 and R_4 causing the current I_A to flow.

The integrator then generates a negative-going ramp with a rate of I_A/C_1 volts per second until its output equals the negative trip point of the threshold detector. The threshold detector then changes to the negative output state, and supplies a negative current, I_B, at the integrator summing point. The integrator now generates a positive-going ramp with a rate of I_B/C_1 volts per second until its output equals the positive trip point of the threshold detector, where the detector again changes output state and the cycle repeats.

Triangular wave frequency is determined by R_3, R_4 and C_1 and the positive and negative saturation voltages of the amplifier A_1. Amplitude is determined by the ratio of R_5 to the combination of R_1 and R_2 and the threshold detector saturation voltages. Positive and negative ramp rates are equal and positive and negative peaks are equal if the detector has equal positive and negative saturation voltages. The output waveform may be offset with respect to ground if the inverting input of the threshold detector, A_1, is offset with respect to ground.

The generator may be made independent of temperature and supply voltage if the detector is clamped with matched zener diodes.

The integrator section should be compensated for unity-gain. The detector section may require compensation if power supply impedance causes oscillation during its transition time. The current into the integrator should be large with respect to the input bias current for maximum symmetry; and offset voltage should be small with respect to peak output voltage swing.

Active Filter Design with IC Op-Amps

INTRODUCTION

Frequency selective networks for use in the frequency range below 100 kHz have always been a problem. In this area of operation the inductors and capacitors required are large, both in value and physical size. Also, at these frequencies inductors and capacitors become quite lossy and the circuit Q's begin to suffer.

The answer to this problem is to exchange the large inductor and capacitor for a large block of gain, and use well known feedback principles to achieve selectivity with R-C active filters. Previously, to achieve a high degree of accuracy and circuit stability, a large number of active components was required in a fairly sophisticated circuit. Consequently, the design time and number of active components required made the use of active filters quite expensive.

The solution to this problem came with the advent of integrated circuits which allowed transistors to be "less expensive" than resistors. Now, excellent gain blocks can be fabricated at fairly reasonable costs. And as technology improves, the performance will continue to improve and the costs will continue to decline, making the use of active filters very economical.

The availability of low cost dual or quad operational amplifier IC's have made the operational amplifier based active filter techniques cost effective over conventional passive filters. The recent availibility of programmable quad operational amplifiers such as the XR-4202 or the XR-346 have provided the active filter designer with the flexibility to externally program gain-bandwidth product, supply current, input bias current, input offset current, input noise and the slew rate. The user, therefore, can trade off bandwidth for supply current or optimize the noise figure. Likewise, other amplifier characteristics can be programmed for a specific need.

Since the operational amplifier plays such a key role in the active filter, its characteristics are of prime importance. By using operational amplifiers as the basic gain stage of the active filter, problems previously encountered due to low input impedance, high output impedance and low gain are virtually eliminated. Operational amplifiers provide the required response for various filter types. Some of the more popular filters are multiple feedback, state variable, bi-quad and Sallen Key which can be used to obtain high pass, band pass and low pass filter functions (and which are capable of giving the designer all of the standard filter responses, i.e., Butterworth, Chebychev, Bessel, etc.)

This application article is intended to assist the designer in selecting the optimum filter for his application. It begins with a table of transfer functions and network defining equations for the high pass, low pass, band pass and the band reject filters. A guide to the three types of filter responses will be presented, also several filter realizations are illustrated with their respective merits and limitations. Finally, the entire contents are brought together to provide the designer a complete working schematic of an active filter in a modem configuration utilizing the XR-4202 Quad Programmable Operational Amplifier along with the XR-2206 Waveform Generator and the XR-2211 Precision Tone Decoder.

TRANSFER FUNCTIONS AND EQUATIONS

Table 1 is intended to give the designer a brief review of the basic transfer functions, and network defining equations. It is noted that a family of curves exists for all cases except first order low pass and high pass. This is due to the presence of α, the damping coefficient. This point will be expanded upon in the next section of filter responses.

FILTER RESPONSES

Once the transfer function has been determined, the next step in filter design is to decide upon the desired response. As previously mentioned the damping of the filter determines it's characteristics near cut off. There are three basic types of responses which are depicted in Table 2 along with their characteristics. In the case of the Butterworth and Bessel, the response has been fixed. However, for the Chebychev the α is chosen for the particular response desired. This is done by using a nomograph such as the one shown in Figure 1. To use a nomograph the information required is: A_{max} (maximum ripple in the passband), A_{min} (minimum attenuation in the stop band), and Ω_s (ratio of the A_{min} bandwidth to the A_{max} bandwidth). These terms are illustrated in Figure 2. Once these terms are known the nomograph is used by locating A_{max} and drawing a straight line through A_{min} to the left hand side of the graph. From this point a horizontal line is drawn to the intersection of Ω_s. The minimum order of the transfer function will be the number of the curve passing above this point. Once this is done the α and ωo for each stage is found by consulting the Chebychev network parameter tables for the desired passband ripple, and the number of poles. Such tables can be found in standard filter handbooks.

FILTER REALIZATIONS

There are numerous ways of realizing the transfer functions discussed. Each of these methods have their own relative merits. The configuration selected depends primarily on the specific application and the desired sensitivity parameters. Sensitivity parameters are a means of relating the resultant change in the transfer function due to an element change. Although these parameters are only directly applicable to an infinitesimal change they are easily used to evaluate performance for 1% changes, and many times are used for element changes up to 10%. Examples will be given later in this section that will help clarify this parameter.

The filter realizations presented here are to be used as a basic guide to help the designer to become more adept at designing filters. State-variable and multiple-feedback filters will be discussed and the relative merits of each will be given. It will also be shown that many of the commonly used filters are actually specific cases for the filters mentioned.

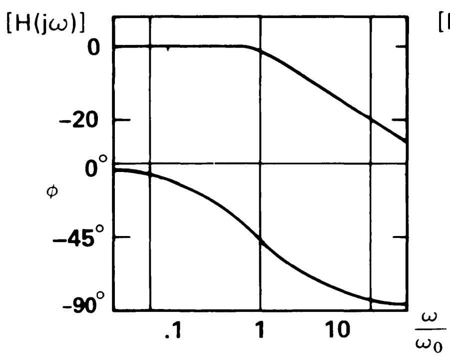
TABLE 1

Low Pass

$$H(s) = \frac{Ho\omega o}{s + \omega o}$$

$$[H(j\omega)] = \left[\frac{Ho^2 \omega o}{\omega^2 + \omega o^2}\right]^{1/2}$$

$$\phi = \text{Tan}^{-1} \frac{\omega}{\omega o}$$

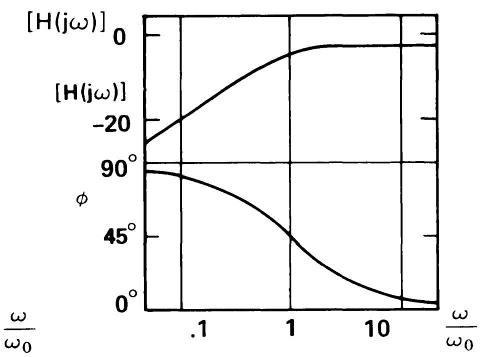


High Pass

$$H(s) = \frac{Hos}{s + \omega o}$$

$$[H(j\omega)] = \left[\frac{Ho^2 \omega o^2}{\omega^2 + \omega o^2}\right]^{1/2}$$

$$\phi = \frac{\pi}{2} - \text{Tan}^{-1} \frac{\omega}{\omega o}$$

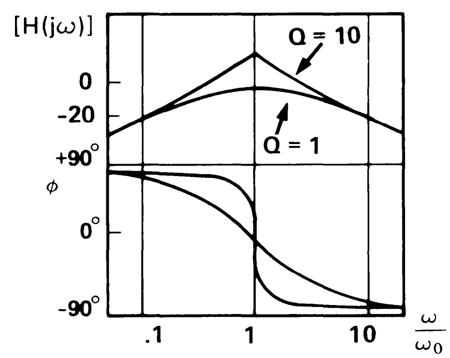


Band Pass

$$H(s) = \frac{Ho\,\alpha\,\omega os}{s^2 + \alpha\omega os + \omega o^2}$$

$$[H(j\omega)] = \left[\frac{Ho^2 \alpha^2 \omega o^2 \omega^2}{\omega^4 + \omega^2 \omega o^2(\alpha^2 - 2) + \omega o^4}\right]^{1/2}$$

$$\phi = \frac{\pi}{2} - \text{Tan}^{-1}\left(\frac{2Q\omega}{\omega o} + \sqrt{4Q^2 - 1}\right) - \text{Tan}^{-1}\left(\frac{2Q\omega}{\omega o} - \sqrt{4Q^2 - 1}\right)$$

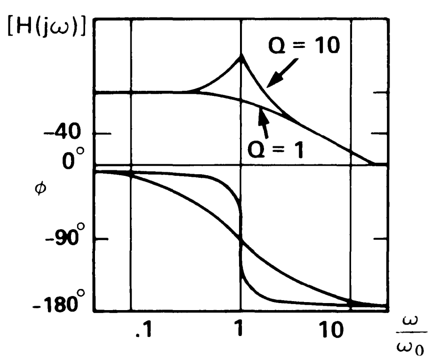


Low Pass Second Order

$$H(s) = \frac{Ho\omega o^2}{s^2 + \alpha\omega os + \omega o^2}$$

$$[H(j\omega)] = \left[\frac{Ha^2 \omega o^4}{\omega^4 + \omega^2 \omega o^2(\alpha^2 - 2) + \omega oA}\right]^{1/2}$$

$$\phi = -\text{Tan}^{-1}\left[\frac{1}{\alpha}\ 2\frac{\omega}{\omega o} + \sqrt{4 - \alpha^2}\right] - \text{Tan}^{-1}\left[\tfrac{1}{2}\ \frac{2\omega}{\omega o} - \sqrt{4 - \alpha^2}\right]$$

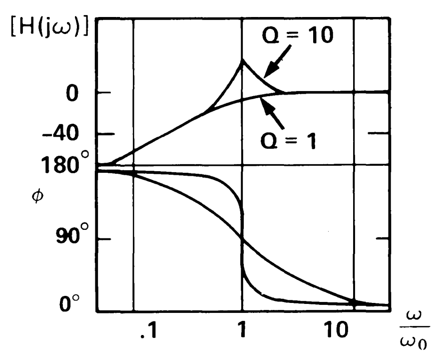


High Pass Second Order

$$H(s) = \frac{Hos^2}{s^2 + \alpha\omega os + \omega o^2}$$

$$[H(j\omega)] = \left[\frac{Ho^2 \omega^4}{\omega^4 + \omega^2 \omega o2(\alpha^2 - 2) + \omega o^4}\right]^{1/2}$$

$$\phi = \pi - \text{Tan}^{-1}\left[\frac{1}{\alpha}\left(2\frac{\omega}{\omega_0} + \sqrt{4 - \alpha^2}\right)\right] - \text{Tan}^{-1}\left[\frac{1}{\alpha}\left(2\frac{\omega}{\omega_0} - \sqrt{4 - \alpha^2}\right)\right]$$

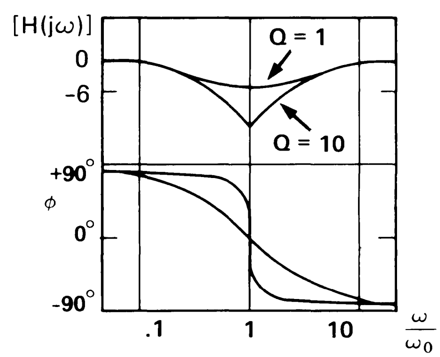


Band Reject

$$H(s) = \frac{(s^2 + \omega o^2)Ho}{s^2 + \alpha\omega os + \omega o^2}$$

$$[H(j\omega)] = \left[\frac{Ho^2 \omega^4 + \omega o^4}{\omega^4 + \omega^2 \omega o(\alpha^2 - 2) + \omega o^4}\right]^{1/2}$$

$$\phi = \frac{\pi}{2} - \text{Tan}^{-1}\left(\frac{2Q\omega}{\omega o} + \sqrt{4Q^2 - 1}\right) - \text{Tan}^{-1}\left(\frac{2Q\omega}{\omega o} - \sqrt{4Q^2 - 1}\right)$$



Definition of Terms

ωo = Cut off frequency $2\,\pi f_0$

ω_2 = Upper cut off frequency

α = Loop damping

$Q = 1/\alpha = \dfrac{\omega c}{\omega_2 - \omega_1}$

$s = \sigma + j\omega$ complex frequency

ϕ = Phase

ω_C = Center frequency

$[H(j\omega)]$ = Magniture response

ω_1 = Lower cut off frequency

$H(s)$ = Transfer function

TABLE 2

Filter Type	α	Basic Features	Amp. Response
Bessel	$\sqrt{3}$	Best time delay Smoothest phase response	
Butterworth	$\sqrt{2}$	Maximally flat amplitude response	
Chebychev	Can Vary	Passband ripple Fast cutoff slope	

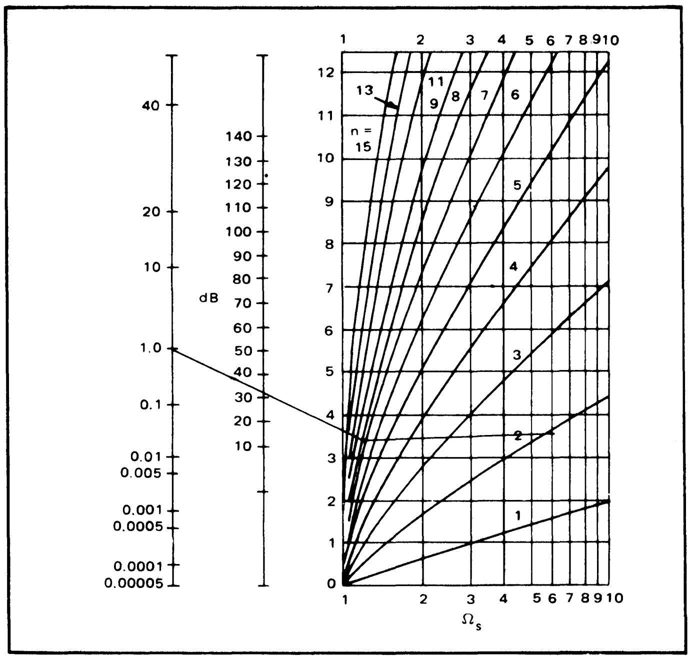


FIGURE 1

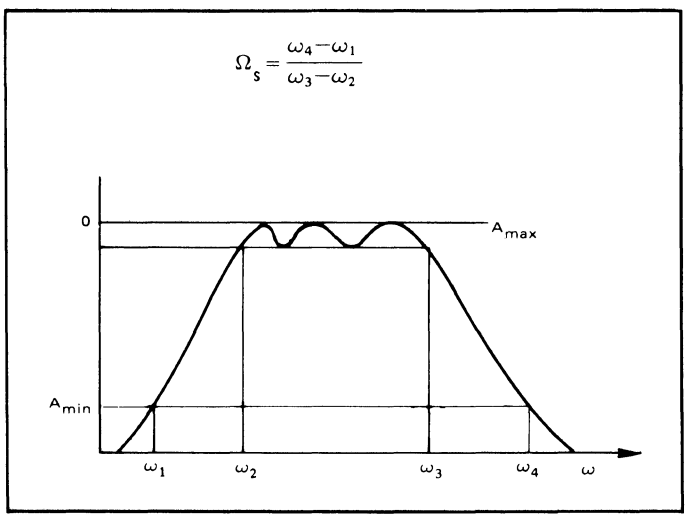


FIGURE 2

Figure 3 illustrates a typical multiple-feedback connection with the non-inverting input grounded. To minimize offset this point should be returned to ground via a resistor whose value is equal to the impedance at the inverting input. The transfer function for this circuit is given by E-1. Each element represents a single resistor or capacitor. To realize the transfer function each admittance parameter is replaced by 1/R for a resistor and sC for a capacitor. An example will help to clarify this point. If the desired response is a high pass, the

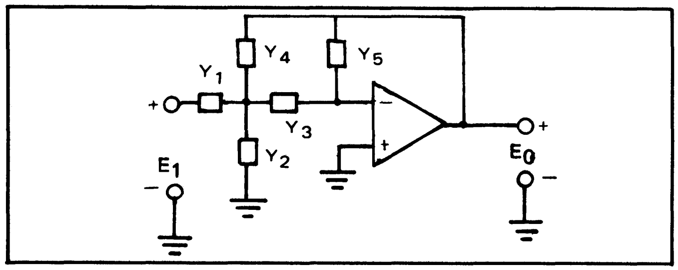


FIGURE 3

E–1

$$\frac{E_0}{E_1}(s) = \frac{-Y_1 Y_3}{Y_5(Y_1 + Y_2 + Y_3 + Y_4) + Y_3 Y_4}$$

form of the characteristic equation is given in Table 1. To transform E–1 into the high-pass characteristic, then Y_1, Y_3, and Y_4 become capacitors and Y_2 and Y_5 resistors. (It should be obvious that a low-pass function could have been fabricated by letting Y_2 and Y_5 be capacitors, and similarly a bandpass function could have been realized by making Y_3 and Y_4 capacitors.) The terms of the network function of the high-pass filter shown in Figure 4 are given in Table 3 along with their sensitivity parameters. The transfer function for Figure 4 is given by E–2.

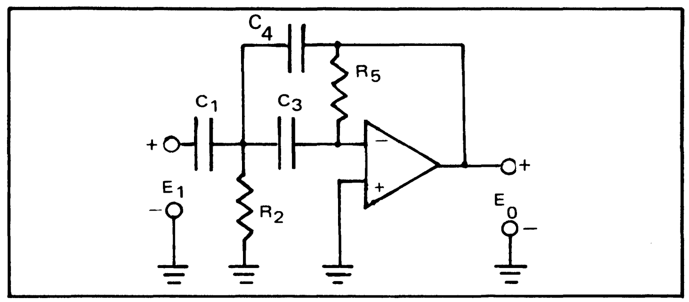


FIGURE 4

E–2

$$\frac{E_0}{E_1}(s) = \frac{-(C_1/C_4)s^2}{s^2 + s(1/R_5)(C_1/C_3C_4 + 1/C_4 + 1/C_3) + 1/R_2R_5C_3C_4}$$

As can be seen from the sensitivity parameters, there is a high degree of circuit sensitivity due to the component tolerances. Due to the interaction of components the tuning of this circuit may be rather involved. However, with tight component tolerances, these circuits give the designer very predictable results. Due to the high input impedance and low output impedance, several of these stages may easily be cascaded to achieve a higher order function. What is desired is to have a lower sensitivity to component tolerances. The most commonly used filter for this purpose is the state-variable.

The state-variable synthesis approach is used in most present day Universal Active Filters (U.A.F.). With this method the actual n^{th} order polynomial of the transfer function is simulated as it would be with an analog computer. When using the state-variable approach all three outputs (high-pass, low-pass and band-pass) are all available simultaneously. The sensitivities with respect to component tolerances are typically less than or equal to one, and the sensitivity of Q

TABLE 3

Parameter	Defining Equation	Sensitivity
H_0	$= \frac{C_1}{C_4}$	$S_{C_1}{}^{H_0} = -S_{C_4}{}^{H_0} = 1$
α	$= \sqrt{\frac{R_2}{R_5}} \left(\frac{C_1}{\sqrt{C_3C_4}} + \sqrt{\frac{C_3}{C_4}} + \sqrt{\frac{C_4}{C_3}} \right)$	$S_{C_3}{}^{\alpha} = \frac{1}{2} - \frac{1}{\alpha\omega_0 R_5 C_3} \left(\frac{C_1}{C_3} + 1 \right)$ $S_{C_4}{}^{\alpha} = \frac{1}{2} - \frac{1}{\alpha\omega_0 R_5 C_4} \left(\frac{C_1}{C_3} + 1 \right)$ $S_{C_1}{}^{\alpha} = \frac{1}{\alpha\omega_0 R_5} \cdot \frac{C_1}{C_3 C_4}$ $S_{R_2}{}^{\alpha} = -S_{R_5}{}^{\alpha} = \frac{1}{2}$
ω_0	$= \left(\frac{1}{R_2R_5C_3C_4} \right)^{1/2}$	$S_{R_2}{}^{\omega_0} = S_{R_5}{}^{\omega_0} = S_{C_3}{}^{\omega_0} = S_{C_4}{}^{\omega_0} = -\frac{1}{2}$

Note: The sensitivity of H_0 with This implies that if C_1 changes by 1% H_0 will also change by 1%. The defining equation for a sensitivity parameter is

$$S_X{}^Y = \frac{xdY}{Ydx}$$

with respect to amplifier gain is nearly zero, if the amplifier gain is high. Because of the high amplifier gain requirement these filters tend to be limited to audio range. The cost of reducing the circuit element sensitivities is the need to use (n + 2) operational amplifiers to synthesize an n^{th} order transfer function. For this reason, this type of configuration may not be cost effective in the synthesis of low Q high-pass and low-pass filters.

Figure 5 shows a typical state-variable configuration whose characteristic equations are given by E–3, E–4, and E–5. It is noted that these equations all have the same denominators; and the numerator is determined by the point at which the output is taken. This form may also be used to simulate a band-reject function by summing the high-pass and low-pass outputs. The defining equations and sensitivity parameters are given in Table 4. It is noted here that the bi-quad is actually a slight variation of a second order state-variable.

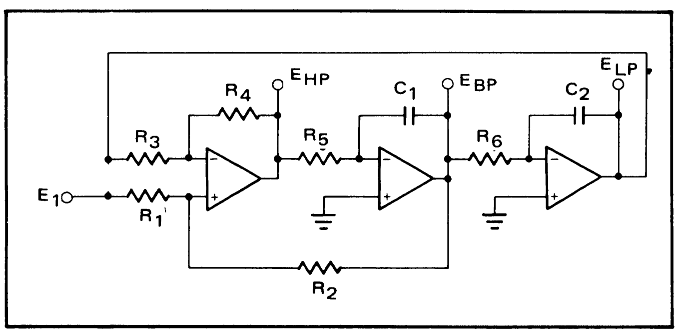


FIGURE 5

E–3 $$\frac{E_{LP}}{E_1} = \frac{\left(\frac{1}{R_5R_6C_1C_2}\right)\left(\frac{1 + R_4/R_3}{1 + R_1/R_2}\right)}{s^2 + s\left(\frac{1}{R_5C_1}\right)\left(\frac{1 + R_4/R_3}{1 + R_2/R_1}\right) + \frac{R_4}{R_3}\left(\frac{1}{R_5R_6C_1C_2}\right)}$$

E–4 $$\frac{E_{HP}}{E_1} = \frac{s^2\left(\frac{1 + R_4/R_3}{1 + R_1/R_2}\right)}{s^2 + s\left(\frac{1}{R_5C_1}\right)\left(\frac{1 + R_4/R_3}{1 + R_2/R_1}\right) + \frac{R_4}{R_3}\left(\frac{1}{R_5R_6C_1C_2}\right)}$$

E–5 $$\frac{E_{BP}}{E_1} = \frac{-s\left(\frac{1}{R_5C_1}\right)\left(\frac{1 + R_4/R_3}{1 + R_1/R_2}\right)}{s^2 + s\left(\frac{1}{R_5C_1}\right)\left(\frac{1 + R_4/R_3}{1 + R_2/R_1}\right) + \frac{R_4}{R_3}\left(\frac{1}{R_5R_6C_1C_2}\right)}$$

MODEM FILTER

A typical application for an active filter is the input stage of a frequency demodulator. Any noise or spurious signals at this point would affect the overall quality of the output. A more specific example can be cited by considering the F.S.K. system shown in Figure 6. (Frequency shift keying is a means of transmitting digital information, primarily through telecommunications links.) This type of system is thoroughly covered in Exar Application Note, AN-01 and will only be briefly discussed here.

In this system, the digital data to be transmitted is used to key the XR-2206. The frequency shift keyed output of the XR-2206 is then sent through the hybrid and out on to the line. (The hybrid is used to obtain isolation between data transmitted and data received, and also may be used to amplify the received signal.) In full duplex operation this system must be able to receive and transmit simultaneously. Due to line losses, the received signal may range from –12 dBm to –48 dBm. The output level of the transmitter is typically –6 dBm (allowing for a 6 dB loss in the hybrid), due to line mismatch, the hybrid may only provide 10 dB of isolation to the filter. (Therefore, the levels at the input of the filter, assuming a gain of 6 dB from the line through the hybrid is –6 and

TABLE 4

Output	Parameters	Defining Equation	Sensitivity
Low Pass E–3	H_0	$\frac{1 + R_3/R_4}{1 + R_1/R_2}$	$S_{R_1}^{H_0} = -S_{R_2}^{H_0} = -1/(1 + R_2/R_1)$ $S_{R_3}^{H_0} = -S_{R_4}^{H_0} = \frac{1}{H_0}\left(\frac{R_3/R_4}{1 + R_1/R_2}\right)$
	ω_0	$\left[\frac{R_4}{R_3 R_5 R_6 C_1 C_2}\right]^{1/2}$	$S_{R_3}^{\omega_0} = S_{R_5}^{\omega_0} = S_{R_6}^{\omega_0} = S_{C_1}^{\omega_0} = S_{C_2}^{\omega_0} = -S_{R_4}^{\omega_0} = -\frac{1}{2}$
	α	$\frac{1 + R_4/R_3}{1 + R_2/R_1}\left(\frac{R_3 R_6 C_2}{R_4 R_5 C_1}\right)^{1/2}$	$S_{R_4}^{\alpha} = -S_{R_3}^{\alpha} = -\frac{1}{2} + \frac{R_4/R_3}{R_5 C_1 \alpha\omega_0(1 + R_2/R_1)}$ $S_{R_1}^{\alpha} = -S_{R_2}^{\alpha} = \frac{1}{1 + R_1/R_2}$ $S_{R_6}^{\alpha} = S_{C_2}^{\alpha} = -S_{R_5}^{\alpha} = -S_{C_1}^{\alpha} = \frac{1}{2}$
High Pass E–4	H_0	$\frac{1 + R_4/R_3}{1 + R_1/R_2}$	$S_{R_1}^{H_0} = -S_{R_2}^{H_0} = -1/(1 + R_2/R_1)$ $S_{R_3}^{H_0} = -S_{R_4}^{H_0} = \frac{1}{H_0}\left(\frac{R_4/R_3}{1 + R_1/R_2}\right)$
	ω_0	SAME AS LOW PASS	
	α	$\left(\frac{1 + R_4/R_3}{1 + R_2/R_1}\right)\left(\frac{R_3 R_6 C_2}{R_4 R_5 C_1}\right)^{1/2}$	$S_{R_4}^{\alpha} = -S_{R_3}^{\alpha} = -\frac{1}{2} + \frac{R_4/R_3}{R_5 C_1 \alpha\omega_0(1 + R_2/R_1)}$ $S_{R_1}^{\alpha} = -S_{R_2}^{\alpha} = \frac{1}{1 + R_1/R_2}$ $S_{R_6}^{\alpha} = S_{C_2}^{\alpha} = -S_{R_5}^{\alpha} = -S_{C_1}^{\alpha} = 1/2.$
Band Pass E–5	H_0	$\frac{R_2}{R_1}$	$S_{R_1}^{H_0} = -S_{R_2}^{H_0} = -1$
	ω_0	SAME AS LOW PASS	
	$Q = 1/\alpha$	$\left(\frac{1 + R_2/R_1}{1 + R_4/R_3}\right)\left(\frac{R_4 R_5 C_1}{R_3 R_6 C_2}\right)^{1/2}$	$S_{R_5}^{Q} = S_{C_1}^{Q} = -S_{R_6}^{Q} = -S_{C_2}^{Q} = \frac{1}{2}$ $S_{R_4}^{Q} = S_{R_3}^{Q} = \frac{1}{2} - \frac{R_4/R_3}{R_5 C_1 \alpha\omega_0(1 + R_2/R_1)}$ $S_{R_2}^{Q} = -S_{R_1}^{Q} = \frac{1}{1 + R_1/R_2}$

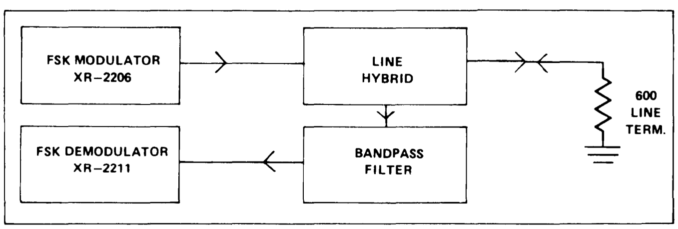


FIGURE 6

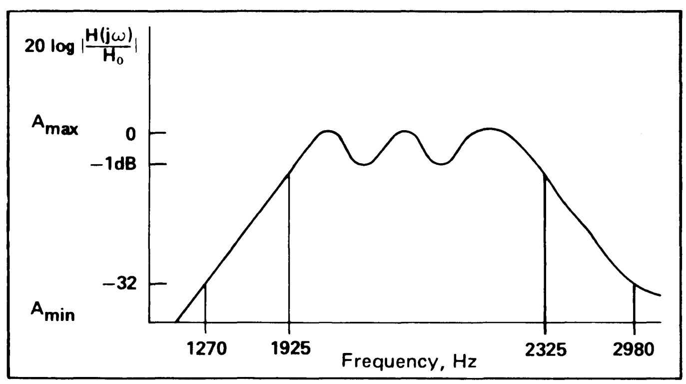


FIGURE 7

–42 dBm for the desired signal and –16 dBm from the local oscillator.) This means that in a worst case situation, the input level of the received signal is –42 dBm with the level of the local oscillator 26 dB above this. For the XR-2211 to operate with a low bit error rate, the input should be 6 dB higher than the interfering signal. This implies that the stopband A_{min} from Figure 2 is 32 dB. The XR-2211 has an internal preamplifier with a dynamic range of greater than 60 dB, and requires a minimum input level of –38 dBm to cause limiting. If we choose a filter to have a passband ripple of 1 dB and an overall gain of 5 dB, the input conditions of the XR-2211 will be satisfied. The filters introduce a phase shift that is only linear for approximately 1/2 to 1/3 of the passband, therefore, a bandwidth of 400 Hz is used for the filter. The general shape of the filter is shown in Figure 7.

Note: The values used in this filter are based on a modem using an XR-2206 as the modulator and XR-2211 as the demodulator. If digital techniques are used, the filter parameters may be different due to the harmonics generated by digital synthesis of a sine wave and higher signal to noise requirements of the demodulator.

To find the minimum number of poles required for this response the nomograph in Figure 1 is used. The point falls between a 2 and 3 pole filter. The values of $\omega_0 + \alpha$ are determined from the tables for a 3rd order chebychev response with 1 dB ripple.

From tables

$$\left.\begin{aligned} \omega_0 &= .997098 \\ \alpha &= .495609 \end{aligned}\right\} \text{ complex pole}$$

$$\omega_0 = .494171 \quad \text{– real pole.}$$

The geometric center is $\omega_0 = \sqrt{\omega_3 \omega_2}$ or $\sqrt{f_3 f_2} = f_0$

$$\text{The filter } Q_0 = Q_0 = \frac{f_0}{f_3 - f_2} = \frac{\sqrt{(1925)(2325)}}{2325 - 1925} = 5.28892$$

The Q of each section of the filter is determined by Equation 6.

$$\text{E–6} \quad QA = \left(\frac{\left(\frac{\omega_1}{Q_0}\right)^2 + 4 + \sqrt{\left[\left(\frac{\omega_1}{Q_0}\right)^2 + 4\right]^2 - 4\left(\frac{\alpha_1 \omega_1}{Q_0}\right)^2}}{2\left(\frac{\alpha_1 \omega_1}{Q_0}\right)^2} \right)^{1/2}$$

$Q_1 = 21.49 = Q_2$ Section 2 is a reflection of section one about f_0. The center frequencies are found by E–7.

$$\text{E–7} \quad M = \frac{\alpha \omega_1 Q_1}{2Q_0} + \sqrt{\left(\frac{\alpha \omega_1 Q_1}{2Q_0}\right)^2 - 1}$$

Where $M = \frac{\omega_1}{\omega_0} = \frac{\omega_0}{\omega_2} = \frac{f_1}{f_0} = \frac{f_0}{f_2}$ $\quad M = 1.0955$, $f_1 = 2317.6$, $f_2 = 1931.1$

for Section 3 the real pole is transformed into a complex pole pair.

$$Q_3 = \frac{2Q_0}{\alpha \omega_B} = 10.7$$

and $f_3 = f_0$.

The 3 filter stages are now defined:

$f_1 = 2317.6$	$Q_1 = 21.49$
$f_2 = 1931.1$	$Q_2 = 21.49$
$f_3 = 2115.56$	$Q_3 = 10.7$

In this example the multiple-feedback approach is used since 3 pole pairs can be generated with 3 op-amps, 6 capacitors and 9 resistors; an equivalent filter could have been designed with the state-variable techniques, but this would have required 9 op-amps to realize. The actual filter is shown in Figure 8. All capacitor values are chosen to be .01 μf, 5% and all resistors are 1%. The values for this filter and a low band filter are shown in Table 5.

TABLE 5

		f_0	ω_0	Q_0	R_1	R_2	R_3	C_1	C_2	H_0
Originate	A	1931.1	12.1335K	21.49	88.6K	192	354K	.01	.01	2
	B	2317.6	14.562K	21.49	74K	160	295K	.01	.01	2
	C	2115.6	13.293K	10.7	40K	355	161K	.01	.01	2
Answer	1									2
	A	1362.26	10.115K	11.827	58.5K	421	234K	.01	.01	2
	B	975.51	6129.3	11.827	96.5K	695	386K	.01	.01	2
	C	1152.78	7.243K	5.832	40.3K	1219.5	161K	.01	.01	2

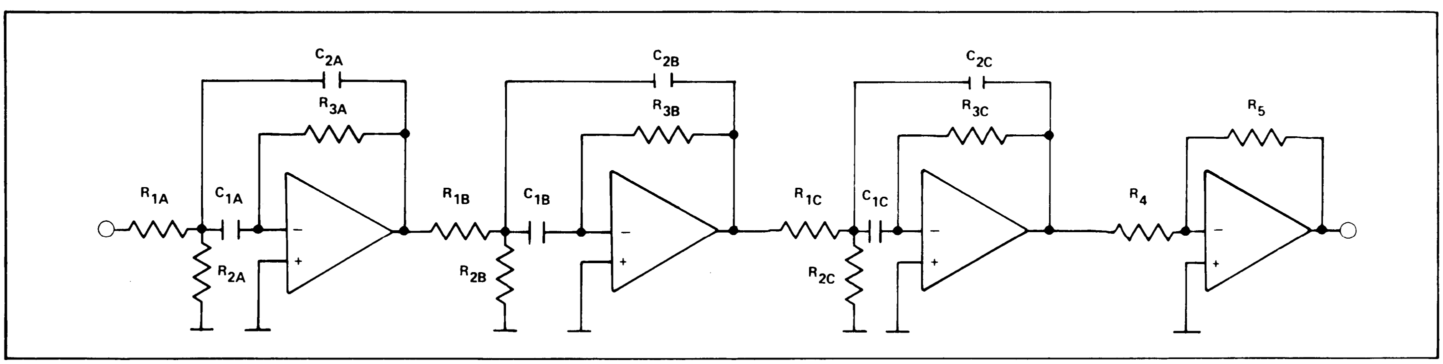


FIGURE 8

Choosing the Right Op Amp

Because of its versatility and ease of application, the op-amp is often the easiest active component to design into the circuit. However, once the initial "paper design" is accomplished, the user is faced with the key question: which op-amp is the best choice for the particular application? The availability of a very wide choice of IC op-amps of varying part numbers, types and features does not make the answer to this question an easy one. If the op-amp characteristics are not carefully considered, the total system performance may be degraded: similarly if each op-amp is overspecified with an excessive amount of "overkill" for the particular application, then the system cost will increase unnecessarily. The key selection criteria is finding the lowest cost operational amplifier which will be sufficient to meet the system performance requirements. This section provides a brief summary of various classes of IC op-amps, their features and key applications, to assist the user in choosing the most cost-effective operational amplifier for his application.

General Purpose Op-Amps

A wide variety of op-amp applications such as low-frequency amplifiers, active filters, voltage-to-current converters and voltage regulators are most economically accomplished using the low-cost general purpose IC op-amps. These op-amps are almost all variations of the basic 741-type op-amp, and offer significant cost savings over any special-purpose op-amps. They are commercially available in single, dual or quad versions. The dual and quad op-amps are particularly cost-effective for applications such as active filters which require a multiplicity of op-amps. The cost per op-amp is usually lower if one can use multiple op-amp IC's rather than single op-amps.

The single and dual general purpose op-amps are available in both internally compensated and uncompensated versions. The quad op-amps are almost invariably internally compensated, to reduce the IC package pin count. Most general purpose IC op-amps have comparable electrical characteristics, namely open loop gain of $\geqslant 20$ mV/V, small-signal unity gain bandwidth of 1 to 2 MHz and a slew rate of $\approx 1\text{V}/\mu\text{sec}$.

Exar manufactures a wide choice of dual or quad general purpose op-amps. All of these op-amps are internally compensated to make them cost-effective and reduce the external parts count. Exar's general purpose op-amps recommended for most applications are XR-1458 and XR-4558 for duals, and XR-4136, XR-4212 and XR-4741 for quad op-amps.

Ground Sensing Op-Amps

These types of op-amps have an input stage common-mode range which extends all the way to the negative supply rail. This is obtained by using Darlington-connected PNP transistors at the input stage of the op-amp. The key advantage of this class of op-amps is that they can be operated with a single positive supply, and still be able to detect or sense small signals near ground potential. The particular circuit recommended for this application is Exar's XR-3403 quad operational amplifier.

Programmable Op-Amps

Programmable op-amps allow the user to "program" or set the operating current levels within the IC op-amp by means of an external setting resistor, and thus be able to trade-off power dissipation for slew-rate or signal bandwidth. These circuits are normally available in quad form, where the power levels of all or some of the op-amps in the package can be programmed by one or two external setting resistors. The key areas of applications for programmable op-amps are active filters and telecommunication channel filters where the user is normally concerned with power dissipation. These op-amps can also be programmed to operate at micro-power levels, by the choice of external setting resistors.

The programmable quad operational amplifiers are available with either one or two separate setting controls. Those with a single setting control have all four of the operational amplifiers programmed from same current setting control. Those with two setting controls have the four op-amps on the chip programmed either in groups of two, or in groups of one and three op-amps. The advantage of partitioned programming is that some of the op-amps in the IC package can be operated at a different power or bandwidth level than the rest of the op-amps in the same chip. For example, in an active filter application, the three op-amps performing the filtering can be operated at a low-power level, yet the fourth op-amp which may be serving as an output buffer can be operated at a higher power level to provide load-drive capability.

Exar offers the broadest product line of programmable op-amps in the industry: The XR-4202, XR-146 and the XR-346-2 families of op-amps are all-bipolar programmable quad op-amp circuits. The XR-4202 offers a single current-setting control for all of the four op-amps on the chip; the XR-146 and the XR-346-2 offer partitioned programming of the four op amps. The XR-094 and XR-095 families are programmable FET-input quad op-amps which have the same pin configuration as the XR-146 and the XR-346-2 families, respectively. These programmable FET-input quad op-amps are fabricated using Exar's ion-implanted bipolar/FET or BIFET process technology which combines matched junction FETs and high-performance bipolar transistors on the same chip.

FET-Input Op-Amps

Finite input impedance or input bias currents associated with conventional bipolar op-amps can be a problem in specific applications such as sample-hold circuits or signal sensing applications from high-impedance signal sources such as transducer systems. For such applications, op-amps with junction-FET input stages offer significant performance advantages since they offer input resistances of the order of 10^{12} ohms, and input bias currents in the low pico-ampere range. Another unique feature of FET-input op-amps is their high slew-rate and wide bandwidth. For example, most FET-input op-amps offer slew-rates in excess of 10 $\text{V}/\mu\text{sec}$ and unity gain bandwidth of 3 MHz.

The FET-input op-amps offer somewhat higher offset voltages and input noise than all-bipolar op-amps; however some specially designed FET input op-amps, such as Exar's XR-072 and XR-074 series have input noise voltages comparable to conventional bipolar op-amps.

Exar offers a wide selection of FET-input dual and quad op-amps which are manufactured using Exar's ion-implanted BIFET process. The XR-082/XR-083 and the XR-072 are dual op-amps; the XR-074 and the XR-084 are quad FET-input op-amps. The XR-094 and the XR-095 are programmable quad FET-input op-amps. Because of their low power capability, the programmable BIFET op-amps are particularly suitable for low-power active filter designs.

Low Noise Op-Amps

These op-amps are particularly suited for audio amplifier and mixer applications, where low noise is of prime importance. The noise characteristics of an op-amp are determined by the noise generated at the input stage, since the noise generated at this point is amplified by the full open-loop gain of the amplifier. In most cases, input noise voltages of 10 $nV/\sqrt{Hz}$ or less is required to be suitable for high quality or professional audio signal processing applications. Such low noise characteristics are normally obtained by careful device design and manufacturing processing of the IC chips. In general, all-bipolar operational amplifiers tend to have better low noise characteristics than the FET-input op-amps.

Exar manufactures a number of low noise op-amp circuits uniquely suited to audio applications. Among Exar's family of low noise op-amps, the XR-5534 operational amplifier, and its dual versions, the XR-5532 and the XR-5533 offer the best noise performance.

Low Distortion Op-Amps

In addition to low noise characteristics, another key performance requirement for audio applications is low distortion. The distortion characteristics of op-amps are normally determined by the design of the output stage as well as the amplifier bandwidth characteristics. The total harmonic distortion (THD) is made up of three components: (a) intermodulation distortion; (b) cross-over distortion which depends on output stage design, and (c) slew-induced distortion which occurs when the output of the op-amp is forced to slew faster than its slew-rate.

The cross-over distortion can be avoided by using op-amps which have class-AB, rather than class-B type output stages. All of Exar's op-amps fall into this category.

To avoid slew-induced distortion, one should ensure that the slew rate of the amplifier is never exceeded during the excursions of the input signal. The high-speed operational amplifiers such as Exar's XR-5533 or XR-5534 op-amps which have slew rates in excess of 10 $V/\mu sec$ with a power bandwidth of 200 kHz can easily cover the entire audio frequency range without introducing slew-induced distortion.

Overview of Exar's Op Amp Products

Exar offers one of the widest selections of multiple op amps in the IC industry. These op amps vary from the general purpose 741-type quad and dual op amps to FET-input, low noise or programmable operational amplifier IC's, optimized for specific applications or performance features. Table 1 shows an overview of the wide selection of op amp products available from Exar. A summary of the key features of these op amps is given in Table 2.

TABLE 1

An Overview of Exar's Op-Amp Products

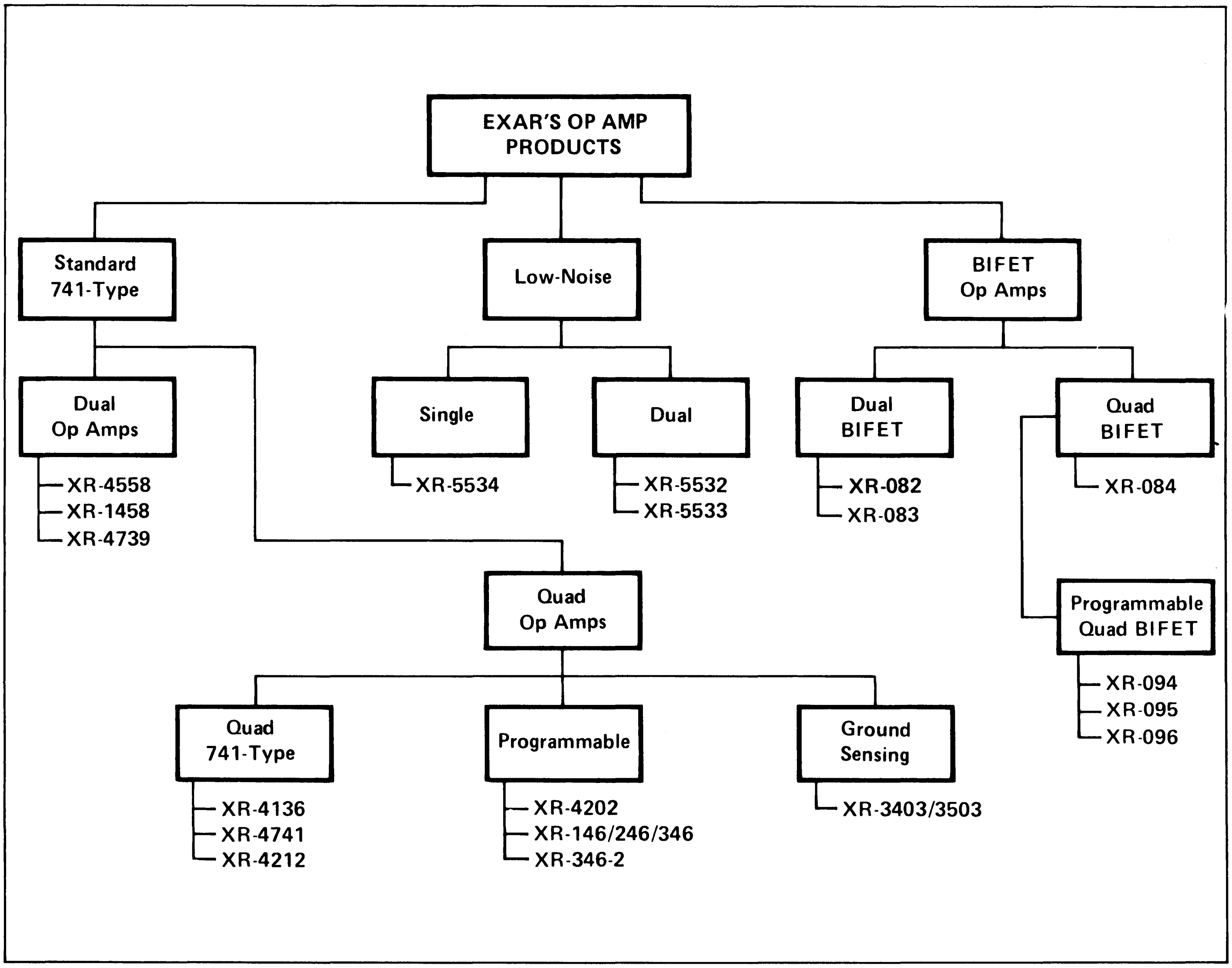


TABLE 2

Key Features of Exar's Op-Amp Products

FEATURES	XR-082/-083	XR-084	XR-094/-095	XR-096	XR-146/-246/-346	XR-1458/-4558	XR-3403/-3503	XR-4136	XR-4202	XR-4212	XR-4739	XR-4741	XR-5532	XR-5533	XR-5534
General Purpose (741-Type)					✓	✓	✓	✓	✓	✓	✓	✓			
Single Op-Amp															✓
Dual Op-Amp	✓					✓					✓		✓	✓	
Quad Op-Amp		✓	✓	✓	✓		✓	✓	✓	✓		✓			
FET Input (BIFET)	✓	✓	✓	✓											
Programmable			✓	✓	✓				✓						
Low Power (≤ 1 mA/Amp)					✓				✓	✓					
High Slew-Rate (≥ 5V/μsec)	✓	✓	✓	✓									✓	✓	✓
Wide Bandwidth (> 3 MHz)	✓	✓	✓	✓									✓	✓	✓
Low Input Current (< 10 nA)	✓	✓	✓	✓											
High Input Impedance (≥ 10 MΩ)	✓	✓	✓	✓											
Low Noise											✓		✓	✓	✓
Single Supply Operation (Ground Sensing)							✓								
High Current Drive (≥ 10 mA)													✓	✓	✓
External Offset Adj.														✓	✓
Internal Freq. Compensation	✓	✓	✓	✓	✓	✓	✓	✓	✓	✓	✓	✓	✓		

Industry-Wide Op-Amp Cross Reference

MANUFACTURER	PART NUMBER	EXAR DIRECT REPLACEMENT
Advanced Micro Devices	AM1458PC	XR-4558P
Fairchild	μA1458TC	XR-4558CP
	-3403DC	XR-3403CN
	-3403PC	XR-3403CP
	3503DM	XR-3403M
	4136DM	XR-4136M
	4136DC	XR-4136CN
	4136PC	XR-4136CP
	4558TC	XR-4558CP
Harris	HA4741-2	XR-4741M
	HA4741-5	XR-4741CP
Motorola	MC1402L	XR-4202N
	MC1402P	XR-4202P
	MC1458P	XR-4558CP
	MC3403L	XR-3403CN
	MC3403P	XR-3403CP
	MC3503L	XR-3503M
National Semiconductor	LM1458N	XR-4558CP
	LM146	XR-146M
	LM146-2	XR-146-2M
	LM246	XR-246
	LM246-2	XR-246-2
	LM346	XR-346CP
	LM346-2	XR-346-2CP
	LM13600J	XR-13600 CN
	LM13600AJ	XR-13600AN
	LM13600N	XR-13600 CP
	LM13600AN	XR-13600AP
Precision Monolithics	SSS1458	XR-4558CP
Raytheon	RC1458NB	XR-4558CP
	RC3403ADC	XR-3403CN
	RC3403ADB	XR-3403CP
	RC3503ADC	XR-3503M
	RM4136DC	XR-4136M
	RC4136DC	XR-4136CN
	RC4136DB	XR-4136CP
	RC4558NB	XR-4558CP
	RC4739DC	XR-4739CN
	RC4739DB	XR-4739CP
	HA4741-2	XR-4741M
	HA4741-5	XR-4741CP
RCA	CA1458E	XR-4558CP
Signetics	NE5558V	XR-4558CP
	NE5532FE	XR-5532N
	NE5532AFE	XR-5532AN
	NE5533F	XR-5533N
	NE5533AF	XR-5533AN
	NE5533N	XR-5533P
	NE5533AN	XR-5533AP
	SE5534F	XR-5534M
	SE5534AF	XR-5534AM
	NE5534F	XR-5534CN
	NE5534AF	XR-5534ACN
	NE5534	XR-5534CP
	NE5534A	XR-5534ACP
Texas Instruments	RM4136J	XR-4136M
	RC4136J	XR-4136CN
	RC4136N	XR-4136CP
	RC4558P	XR-4558CP
	SN72558P	XR-4558CP
	TL082CP	XR-082CP
	TL082CJG	XR-082CN
	TL082IP	XR-082P
	TL082IJG	XR-082N
	TL082MJG	XR-082M
	TL083CN	XR-083CP
	TL083CJ	XR-083CN
	TL083IN	XR-083P
	TL083IJ	XR-083N
	TL083MJ	XR-083M
	TL084CN	XR-084CP
	TL084CJ	XR-084CN
	TL084IN	XR-084P
	TL084IJ	XR-084N
	TL084MJ	XR-084M

Quality Assurance Standards

See page 76.

XR-082/083

Dual BIFET Operational Amplifiers

GENERAL DESCRIPTION – ADVANCE INFORMATION

The XR-082/XR-083 family of junction FET input dual operational amplifiers are designed to offer higher performance than conventional bipolar op-amps. Each amplifier features high slew-rate, low input bias and offset currents, and low offset voltage drift with temperature. These operational amplifier circuits are fabricated using ion-implantation technology which combines well-matched junction FETs and high-performance bipolar transistors on the same monolithic chip.

The XR-082 family of dual BIFET op-amps are packaged in 8-pin dual-in-line packages. The XR-083 family of op-amps offer independent offset adjustment for each of the individual op-amps on the same chip, and are available in 14-pin dual-in-line packages.

FEATURES

Direct Replacement for TL082/TL083 (See Chart)
Low Power Consumption
Wide Common-Mode and Differential Voltage Ranges
Low Input Bias and Offset Currents
Output Short-Circuit Protection
High Input Impedance . . . FET Input Stage
Internal Frequency Compensation
Latch-Up-Free Operation
High Slew-Rate . . . 13 V/μs, Typical

Exar Part Number	Texas Instruments Equivalent
XR-082M/XR-083M	TL-082M/TL-083M
XR-082/XR-083	TL-082AI/TL-083AI
XR-082C/XR-083C	TL-082C/TL-083C
XR-082D/DX-083D	———

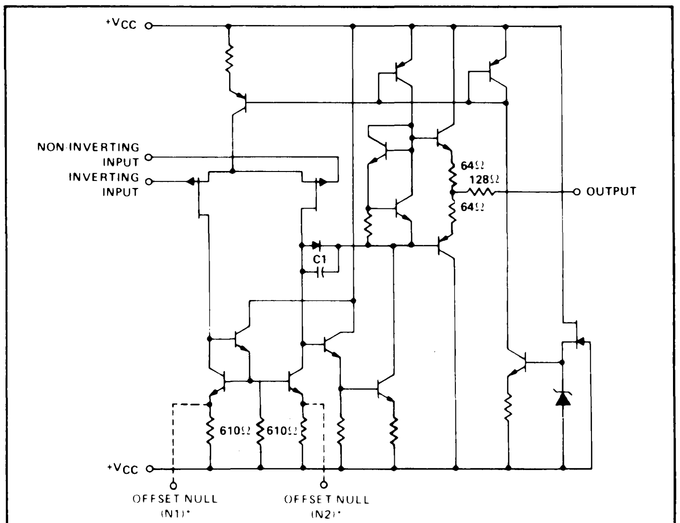
ABSOLUTE MAXIMUM RATINGS

Supply Voltage	±18V
Differential Input Voltage	±30V
Input Voltage Range (Note 1)	±15V
Output Short-Circuit Duration (Note 2)	Indefinite
Package Power Dissipation:	
Plastic Package	625 mW
Derate Above T_A = +25°C	5.0 mW°C
Ceramic Package	750 mW
Derate Above T_A = +25°C	6.0 mW/°C
Storage Temperature Range	-65°C to +150°C

AVAILABLE TYPES

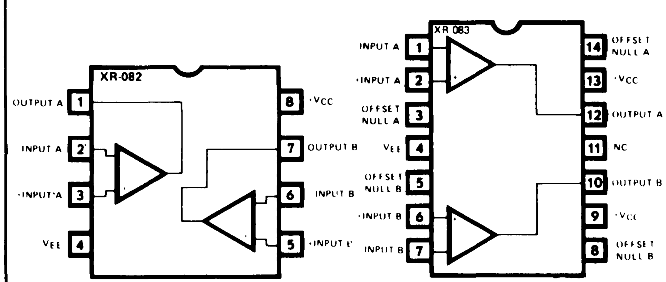
Part Number	Package	Operating Temperature
XR-082M/XR-083M	Ceramic	−55°C to +125°C
XR-082N/XR-083N	Ceramic	−25°C to +85°C
XR-082P/XR-083P	Plastic	−25°C to +85°C
XR-082CN/XR-083CN	Ceramic	0°C to +75°C
XR-082CP/XR-083CP	Plastic	0°C to +75°C
XR-082DN/XR-083DN	Ceramic	0°C to +75°C
XR-082DP/XR-083DP	Plastic	0°C to +75°C

EQUIVALENT SCHEMATIC



*Available in XR-083 Only. (One Channel Only)

FUNCTIONAL BLOCK DIAGRAM



ELECTRICAL CHARACTERISTICS T_A = 25°C, V_{CC} = ±15V, unless otherwise specified.

CHARACTERISTICS	XR-082M/XR-083M MIN.	TYP.	MAX.	XR-082/XR-083 MIN.	TYP.	MAX.	XR-082C/XR-083C XR-082D/XR-083D MIN.	TYP.	MAX.	UNITS	SYMBOL	CONDITIONS
Input Offset Voltage		3	6		3	6		5	15	mV	V_{OS}	R_S = 50Ω, T_A = 25°C
			9			9			20	mV	V_{OS}	R_S = 50Ω, T_A = Full Range
Offset Voltage Temp. Coef.		10			10			10		µV/°C	$\Delta V_{OS}/\Delta T$	R_S = 50Ω, T_A = Full Range
Input Bias Current											I_B	T_A = 25°C, Note 3
		30	200		30	200				pA		
XR-082C/XR-083C								30	400	pA		
XR-082D/XR-083D								100	800	pA		
Input Bias Current Over Temp.			50			20			20	nA	I_B	T_A = Full Range
Input Offset Current											I_{OS}	T_A = 25°C, Note 3
		5	100		5	100				pA		
XR-082C/XR-083C								5	200	pA		
XR-082D/XR-083D								20	400	pA		
Input Offset Current Over Temp.			20			10			5	nA		T_A = Full Range
Supply Current (per amplifier)		1.4	2.8		1.4	2.8		1.4	2.8	mA	I_{CC}	No Load, No Input Signal
Input Common Mode Range	±12			±12			±10			V	V_{ICM}	
Voltage Gain										V/mV	A_{VOL}	$R_L \geq 2$ KΩ, V_O = ±10V
	50	200		50	200		25	200				T_A = 25°C
	25			25			15					T_A = Full Range
Max. Output Swing (peak-to-peak)												$R_L \geq 10$ KΩ
	24	27		24	27		24	27		V	V_{OPP}	T_A = 25°C
	24			24			24					T_A = Full Range
Input Resistance		10^{12}			10^{12}			10^{12}		Ω	R_{in}	T_A = 25°C
Unity-Gain Bandwidth		3			3			3		MHz	BW	T_A = 25°C
Common-Mode Rejection	80	86		80	86		70	76		dB	CMRR	$R_S \leq 10$ KΩ
Supply-Voltage Rejection	80	86		80	86		70	76		dB	PSRR	
Channel Separation		120			120			120		dB		A_V = 100, Freq. = 1 kHz
Slew Rate		13			13			13		V/µS	dV_{out}/dt	A_V = 1, R_L = 2 KΩ, C_L = 100 pF, V_I = 10V
Rise Time		0.1			0.1			0.1		µsec	t_r	A_V = 1, R_L = 2 KΩ
Overshoot		10			10			10		%	t_o	C_L = 100 pF, V_I = 20 mV
Equivalent Input Noise Voltage		20			20			20		nV/√Hz	e_n	R_S = 100Ω, f = 1 kHz

Note 1: For Supply Voltage less than ±15V, the absolute maximum input voltage is equal to the supply voltage.

Note 2: The output may be shorted to ground or to either supply. Temperature and/or supply voltages must be limited to ensure that the dissipation rating is not exceeded.

Note 3: XR-082C/XR-083C and XR-082D/XR-083D differ only in their Input Bias Current and Input Offset Current specifications.

XR-084

Quad BIFET Operational Amplifier

GENERAL DESCRIPTION – ADVANCE INFORMATION

The XR-084 junction FET input quad operational amplifier is designed to offer higher performance than conventional bipolar quad op amps. Each of the four op-amps on the chip is closely matched in performance characteristics, and each amplifier features high slew-rate, low input bias and offset currents, and low offset voltage drift with temperature. The XR-084 FET input quad op-amp is fabricated using ion implanted bipolar/FET or "BIFET" technology which combines well-matched junction FETs and high-performance bipolar transistors on the same monolithic integrated circuit.

FEATURES

Direct Replacement for TL084 (See Chart Below)
Same Pin Configuration as XR-3403 LM324
High-Impedance Junction FET Input Stage
Internal Frequency Compensation
Low Power Consumption
Wide Common-Mode and Differential Voltage Ranges
Low Input Bias and Offset Currents
Output Short-Circuit Protection
Latch-Up-Free Operation
High Slew-Rate . . . 13 V/μS, Typical

Exar Part Type	Texas Instruments Equivalent
XR-084M	TL-084M
XR-084	TL-084AI
XR-084C	TL-084C
XR-084D	———

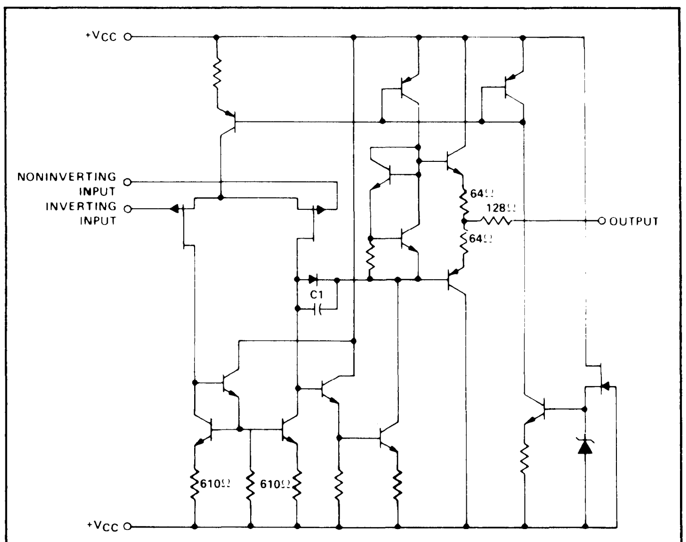
ABSOLUTE MAXIMUM RATINGS

Supply Voltage	±18V
Differential Input Voltage	±30V
Input Voltage Range (Note 1)	±15V
Output Short-Circuit Duration (Note 2)	Indefinite
Package Power Dissipation:	
Plastic Package	625 mW
Derate Above T_A = +25°C	5.0 mW/°C
Ceramic Package	750 mW
Derate Above T_A = +25°C	6.0 mW/°C
Storage Temperature Range	−65°C to +150°C

AVAILABLE TYPES

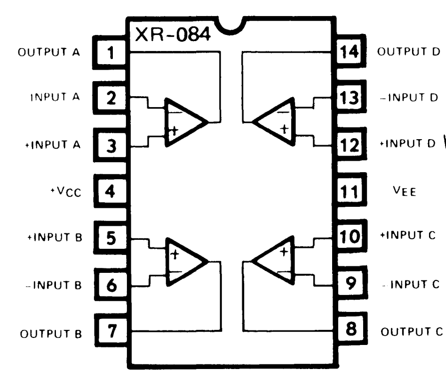
Part Number	Package	Operating Temperature
XR-084M	Ceramic	−55°C to +125°C
XR-084N	Ceramic	−25°C to +85°C
XR-084P	Plastic	−25°C to +85°C
XR-084CN	Ceramic	0°C to +75°C
XR-084CP	Plastic	0°C to +75°C
XR-084DN	Ceramic	0°C to +75°C
XR-084DP	Plastic	0°C to +75°C

EQUIVALENT SCHEMATIC



(One Channel Only)

FUNCTIONAL BLOCK DIAGRAM



ELECTRICAL CHARACTERISTICS $T_A = 25°C, V_{CC} = ±15V$, unless otherwise specified.

CHARACTERISTICS	XR-084M MIN.	XR-084M TYP.	XR-084M MAX.	XR-084 MIN.	XR-084 TYP.	XR-084 MAX.	XR-084C XR-084D MIN.	XR-084C XR-084D TYP.	XR-084C XR-084D MAX.	UNITS	SYMBOL	CONDITIONS
Input Offset Voltage		3	6 9		3	6 9		5	15 20	mV mV	V_{OS} V_{OS}	$R_S = 50\Omega, T_A = 25°C$ $R_S = 50\Omega, T_A =$ Full Range
Offset Voltage Temp. Coef.		10			10			10		μV/°C	$\Delta V_{OS}/\Delta T$	$R_S = 50\Omega, T_A =$ Full Range
Input Bias Current											I_B	$T_A = 25°C$, Note 3
XR-084M/XR-084		30	200		30	200				pA		
XR-084C								30	400	pA		
XR-084D								100	800	pA		
Input Bias Current Over Temp.			50			20			20	nA	I_B	$T_A =$ Full Range
Input Offset Current											I_{OS}	$T_A = 25°C$, Note 3
XR-084M/XR-084		5	100		5	100				pA		
XR-084C								5	200	pA		
XR-084D								20	400	pA		
Input Offset Current Over Temp.			20			10			5	nA		$T_A =$ Full Range
Supply Current (per amplifier)		1.4	2.8		1.4	2.8		1.4	2.8	mA	I_{CC}	No Load, No Input Signal
Input Common Mode Range	±12			±12			±10			V	V_{ICM}	
Voltage Gain	50 25	200		50 25	200		25 15	200		V/mV	A_{VOL}	$R_L \geq 2 K\Omega, V_O = ±10V$ $T_A = 25°C$ $T_A =$ Full Range
Max. Output Swing (peak-to-peak)	24 24	27		24 24	27		24 24	27		V	V_{OPP}	$R_L \geq 10 K\Omega$ $T_A = 25°C$ $T_A =$ Full Range
Input Resistance		10^{12}			10^{12}			10^{12}		Ω	R_{in}	$T_A = 25°C$
Unity-Gain Bandwidth		3			3			3		MHz	BW	$T_A = 25°C$
Common-Mode Rejection	80	86		80	86		70	76		dB	CMRR	$R_S \leq 10 K\Omega$
Supply-Voltage Rejection	80	86		80	86		70	76		dB	PSRR	
Channel Separation		120			120			120		dB		$A_V = 100$, Freq. = 1 kHz
Slew Rate		13			13			13		V/μS	dV_{out}/dt	$A_V = 1, R_L = 2 K\Omega$ $C_L = 100 pF, V_I = 10V$
Rise Time Overshoot		0.1 10			0.1 10			0.1 10		μsec %	t_r t_o	$A_V = 1, R_L = 2 K\Omega$ $C_L = 100 pF, V_I = 20 mV$
Equivalent Input Noise Voltage		20			20			20		$nV/\sqrt{Hz}$	e_n	$R_S = 100\Omega$ f = 1 kHz

Note 1: For Supply Voltage less than ±15V, the absolute maximum input voltage is equal to the supply voltage.
Note 2: The output may be shorted to ground or to either supply. Temperature and/or supply voltages must be limited to ensure that the dissipation rating is not exceeded.
Note 3: XR-084C and XR-084D differ only in their Input Bias Current and Input Offset Current specifications.

XR-094/095

Programmable Quad BIFET Operational Amplifier

GENERAL DESCRIPTION

The XR-094 and XR-095 junction FET input quad programmable operational amplifiers consist of four independent, high gain, internally compensated amplifiers. Two external resistors (R_{SET}) allow the user to program supply current slew-rate input noise without the usual sacrifice of gain bandwidth product. For example, the user can trade-off slew-rate for supply current or optimize the noise figure for a given source impedance. Except for the two programming pins at the end of the package, the XR-094 and XR-095 pin-out is the same as the popular 324, 3403, 124, 148 and 4741 operational amplifiers.

In the case of the XR-094, three of the op amps on the chip share a common programming pin; and the fourth op amp is programmed separately. In the case of the XR-095, each pair of op amps share a common programming pin.

FEATURES

- Same Pin Configuration as LM-346
- High-Impedance Junction FET Input Stage
- Internal Frequency Compensation
- Low Power Consumption
- Wide Common-Mode and Differential Voltage Ranges
- Low Input Bias and Offset Currents
- Output Short-Circuit Protection
- High Slew-Rate . . . 13 V/μs, Typical
- Programmable Electrical Characteristics

APPLICATIONS INFORMATION

Total Supply Current = 5.6 mA (I_{SET}/320 μA)

Slew-Rate = 13 V/μs (I_{SET}/320 μA)

I_{SET} = Current into set terminal

$$I_{SET} = \frac{V_{CC} - (V_{EE} - 0.6V)}{R_{SET}}$$

Note. I_{SET} must be $\leq$ 400 μA

ABSOLUTE MAXIMUM RATINGS

Supply Voltage	±18V
Differential Input Voltage	±30V
Input Voltage Range (Note 1)	±15V
Output Short-Circuit Duration (Note 2)	Indefinite
Package Power Dissipation:	
Plastic Package	625 mW
Derate Above T_A = +25°C	5.0 mV/°C
Ceramic Package	750 mW
Derate Above T_A = +25°C	6.0 mW/°C
Storage Temperature Range	−65°C to +150°C

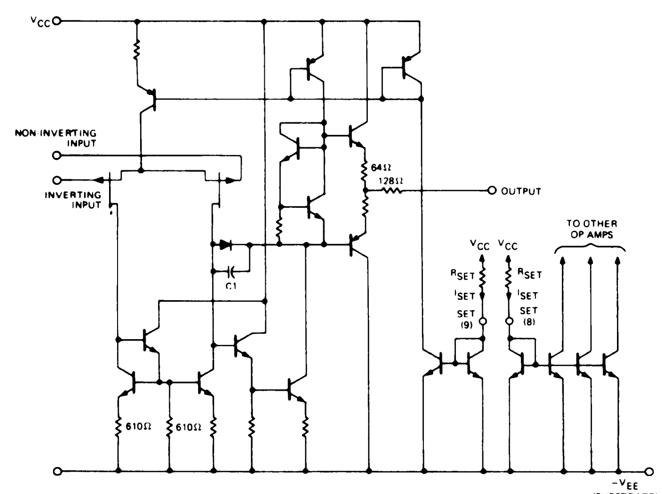
Note 1: For Supply Voltage less than ±15V, the absolute maximum input voltage is equal to the supply voltage.

Note 2: The output may be shorted to ground or to either supply. Temperature and/or supply voltages must be limited to ensure that the dissipation rating is not exceeded.

AVAILABLE TYPES

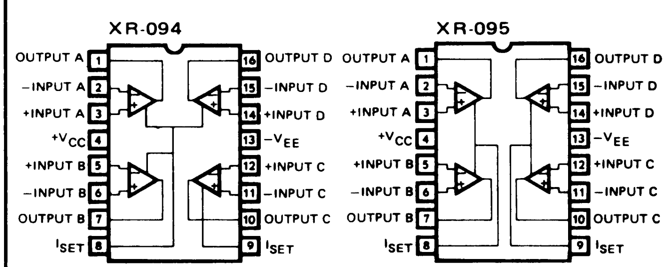
Part Number	Package	Operating Temperature
XR-094/XR-095N	Ceramic	−25°C to +85°C
XR-094/XR-095P	Plastic	−25°C to +85°C
XR-094/XR-095CN	Ceramic	0°C to +75°C
XR-094/XR-095CP	Plastic	0°C to +75°C

EQUIVALENT SCHEMATIC



(One Channel Only)

FUNCTIONAL BLOCK DIAGRAMS



ELECTRICAL CHARACTERISTICS
$T_A = 25°C$, $V_{CC} = \pm15V$, unless otherwise specified.
$I_{SET} = 320\ \mu A$.

CHARACTERISTICS	XR-094/095			XR-094C/XR-095C			UNITS	SYMBOL	CONDITIONS
	MIN.	TYP.	MAX.	MIN.	TYP.	MAX.			
Input Offset Voltage		3	6 9		5	15 20	mV mV	V_{OS} V_{OS}	$R_S = 50\Omega$, $T_A = 25°C$ $R_S = 50\Omega$, T_A = Full Range
Offset Voltage Temp. Coef.		10			10		$\mu V/°C$	$\Delta V_{OS}/\Delta T$	$R_S = 50\Omega$, T_A = Full Range
Input Bias Current		80	600 20		80	800 20	pA nA	I_B	$T_A = 25°C$ T_A = Full Range
Input Offset Current		40	300 10		40	500 5	pA nA	I_{OS}	$T_A = 25°C$ T_A = Full Range
Supply Current (per amplifier)		1.4	2.8		1.4	2.8	mA	I_{CC}	No Load, No Input Signal
Input Common Mode Range	±12			±10			V	V_{iCM}	
Voltage Gain	50 25	200		25 15	200		V/mV	A_{VOL}	$R_L \geq 2K\Omega$, $V_0 = \pm10V$ $T_A = 25°C$ T_A = Full Range
Max. Output Swing (peak-to-peak)	24 24	27		24 24	27		V	V_{OPP}	$R_L \geq 10\ K\Omega$ $T_A = 25°C$ T_A = Full Range
Input Resistance		10^{12}			10^{12}		Ω	R_{in}	$T_A = 25°C$
Unity-Gain Bandwidth		3			3		MHz	BW	$T_A = 25°C$
Common-Mode Rejection	80	86		70	76		dB	CMRR	$R_S \leq 10\ K\Omega$
Supply-Voltage Rejection	80	86		70	76		dB	PSRR	
Channel Separation		120			120		dB		$A_V = 100$, Freq. = 1 kHz
Slew Rate		13			13		$V/\mu S$	$dV_{out/dt}$	$A_V = 1$, $R_L = 2\ K\Omega$ $C_L = 100$ pF, $V_1 = 10V$
Rise Time Overshoot		0.1 10			0.1 10		μsec %	t_r t_o	$A_V = 1$, $R_L = 2\ K\Omega$ $C_L = 100$ pF, $V_1 = 20$ mV
Equivalent Input Noise Voltage		18			18		$nV/\sqrt{Hz}$	e_n	$R_S = 100\Omega$ f = 1 kHz

XR-096

Programmable Quad BIFET Operational Amplifier

GENERAL DESCRIPTION – ADVANCE INFORMATION

The XR-096 monolithic circuit contains four independently programmable BIFET operational amplifiers in a single IC package. Each of the four op amp sections on the chip has its own external bias terminal; thus its performance characteristics and power dissipation can be independently controlled, without effecting the other op amp sections on the chip. The respective bias-setting resisters, R_{SET}, connected to the programming terminals of the circuit allow one to trade-off power dissipation for slew-rate, without sacrificing the gain-bandwidth product of the circuit. These individual bias terminals can also be used to switch the op amp sections "on" and "off", and thus, multiplex between various op amp channels on the same chip.

FEATURES

Programmable Version of XR-084
Independent Programming of All Four Op Amps
Programmable for Micropower Operation
High-Impedance Junction-FET Input Stage
Internal Frequency Compensation
Low Input Bias and Offset Currents

APPLICATIONS INFORMATION

Total Supply Current = 5.6 mA (I_{SET}/320 μA)
Slew-Rate = 13 V/μs (I_{SET}/320 μA)
I_{SET} = Current into set terminal

$$I_{SET} = \frac{V_{CC} - (V_{EE} - 0.6V)}{R_{SET}}$$

Note. I_{SET} must be $\leq$ 400 μA

ABSOLUTE MAXIMUM RATINGS

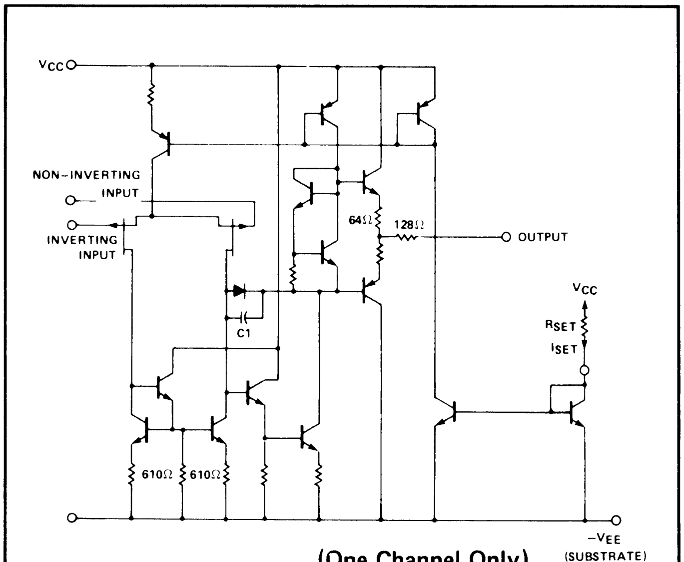
Supply Voltage	±18V
Differential Input Voltage	±30V
Input Voltage Range (Note 1)	±15V
Output Short-Circuit Duration (Note 2)	Indefinite
Package Power Dissipation:	
Plastic Package	625 mW
Derate Above T_A = +25°C	5.0 mV/°C
Ceramic Package	750 mW
Derate Above T_A = +25°C	6.0 mW/°C
Storage Temperature Range	−65°C to +150°C

Note 1: For Supply Voltage less than ±15V, the absolute maximum input voltage is equal to the supply voltage.
Note 2: The output may be shorted to ground or to either supply. Temperature and/or supply voltages must be limited to ensure that the dissipation rating is not exceeded.

AVAILABLE TYPES

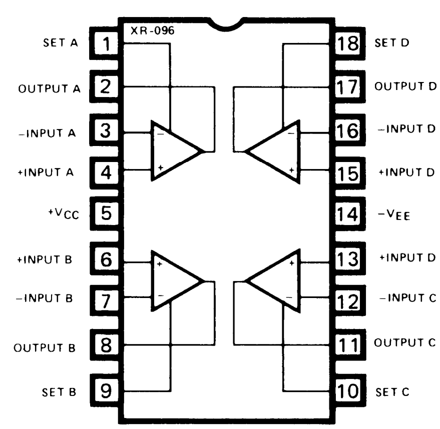
Part Number	Package	Operating Temperature
XR - 096N	Ceramic	−25°C to +85°C
XR - 096P	Plastic	−25°C to +85°C
XR - 096CN	Ceramic	0°C to +75°C
XR - 096CP	Plastic	0°C to +75°C

EQUIVALENT SCHEMATIC



(One Channel Only)

FUNCTIONAL BLOCK DIAGRAMS



ELECTRICAL CHARACTERISTICS

T_A = 25°C, V_{CC} = ±15V, unless otherwise specified.
I_{SET} = 320 μA.

CHARACTERISTICS	XR-096			XR-096C			UNITS	SYMBOL	CONDITIONS
	MIN.	TYP.	MAX.	MIN.	TYP.	MAX.			
Input Offset Voltage		3	6 9		5	15 20	mV mV	V_{OS} V_{OS}	R_S = 50Ω, T_A = 25°C R_S = 50Ω, T_A = Full Range
Offset Voltage Temp. Coef.		10			10		μV/°C	$\Delta V_{OS}/\Delta T$	R_S = 50Ω, T_A = Full Range
Input Bias Current		80	600 20		80	800 20	pA nA	I_B	T_A = 25°C T_A = Full Range
Input Offset Current		40	300 10		40	500 5	pA nA	I_{OS}	T_A = 25°C T_A = Full Range
Supply Current (per amplifier)		1.4	2.8		1.4	2.8	mA	I_{CC}	No Load, No Input Signal
Input Common Mode Range	±12			±10			V	V_{ICM}	
Voltage Gain	50 25	200		25 15	200		V/mV	A_{VOL}	$R_L \geq 2K\Omega$, V_0 = ±10V T_A = 25°C T_A = Full Range
Max. Output Swing (peak-to-peak)	24 24	27		24 24	27		V	V_{OPP}	$R_L \geq 10$ KΩ T_A = 25°C T_A = Full Range
Input Resistance		10^{12}			10^{12}		Ω	R_{in}	T_A = 25°C
Unity-Gain Bandwidth		3			3		MHz	BW	T_A = 25°C
Common-Mode Rejection	80	86		70	76		dB	CMRR	$R_S \leq 10$ KΩ
Supply-Voltage Rejection	80	86		70	76		dB	PSRR	
Channel Separation		120			120		dB		A_V = 100, Freq. = 1 kHz
Slew Rate		13			13		V/μS	$dV_{out/dt}$	A_V = 1, R_L = 2 KΩ C_L = 100 pF, V_1 = 10V
Rise Time Overshoot		0.1 10			0.1 10		μsec %	t_r t_o	A_V = 1, R_L = 2 KΩ C_L = 100 pF, V_1 = 20 mV
Equivalent Input Noise Voltage		18			18		nV/$\sqrt{Hz}$	e_n	R_S = 100Ω f = 1 kHz

XR-146/246/346

Programmable Quad Operational Amplifier

The XR-146 family of quad operational amplifiers contain four independent high-gain, low-power, programmable op-amps on a monolithic chip. The use of external bias setting resistors permit the user to program gain-bandwidth product, supply current, input bias current, input offset current, input noise and the slew rate.

The basic XR-146 family of circuits offer partitioned programming of the internal op-amps where one setting resistor is used to set the bias levels in the three op-amps, and a second bias setting is used for the remaining op-amp. Its modified version, the XR-346-2 provides a separate bias setting resistor for each of the two op-amp pairs.

FEATURES

Programmable Electrical Characteristics
Micropower Operation
Low Noise
Wide Power Supply Range
Class AB Output
Ideal Pin Out for Biquad Active Filters
Overload Protection for Input and Output
Internal Frequency Compensation

APPLICATIONS INFORMATION

Total Supply Current = 1.4 mA ($I_{SET}/10\ \mu A$)
Gain Bandwidth Product = 1 MHz ($I_{SET}/10\ \mu A$)
Slew Rate = 0.4V/μs ($I_{SET}/10\ \mu A$)
Input Bias Current ≅ 50 nA ($I_{SET}/10\ \mu A$)

I_{SET} = Current into pin 8, pin 9 (see schematic)

$$I_{SET} = \frac{V_{CC} - (V_{EE} - 0.6V)}{R_{SET}}$$

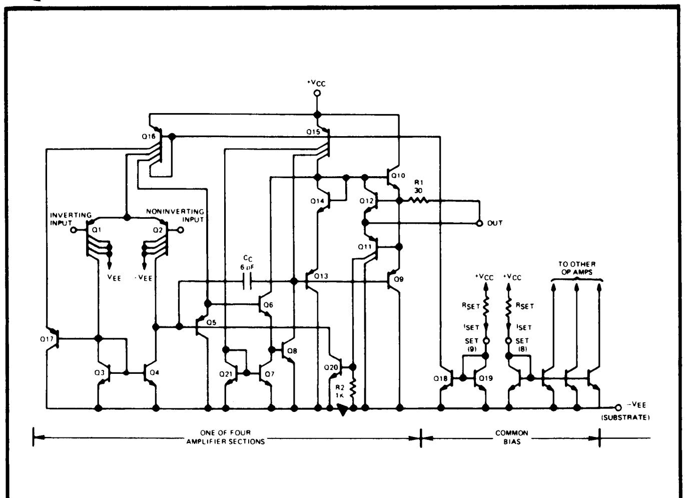
ABSOLUTE MAXIMUM RATINGS

Supply Voltage	
XR-146	±22V
XR-246/346	±18V
Differential Input Voltage (Note 1)	
XR-146/246/346	±30V
Common Mode Input Voltage (Note 1)	
XR-146/246/346	±15V
Power Dissipation (Note 2)	
XR-146	900 mW
XR-246/346	500 mW
Output Short Circuit Duration (Note 3)	
XR-146/246/346	Indefinite
Maximum Junction Temperature	
XR-146	150°C
XR-246	110°C
XR-346	100°C
Storage Temperature Range	
XR-146/246/346	−65°C to +150°C

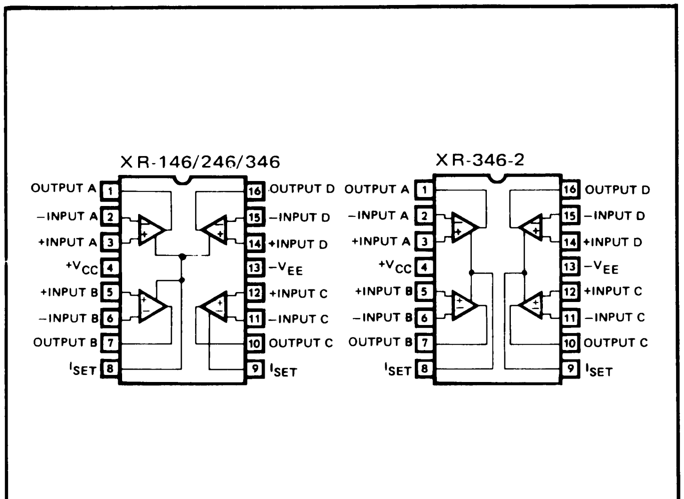
AVAILABLE TYPES

Part Number	Package	Operating Temperature
XR-146M	Ceramic	−55°C to +125°C
XR-246N	Ceramic	−25°C to +85°C
XR-246P	Plastic	−25°C to +85°C
XR-346/346-2CN	Ceramic	0°C to +70°C
XR-346/346-2CP	Plastic	0°C to +70°C

EQUIVALENT SCHEMATIC DIAGRAM



FUNCTIONAL BLOCK DIAGRAMS



ELECTRICAL CHARACTERISTICS ($T_A = +25°C$, $V_S = ±15V$, $I_{SET} = 10\ \mu A$)

PARAMETER	XR-146			XR-246/346			UNITS	CONDITIONS
	MIN.	TYP.	MAX.	MIN.	TYP.	MAX.		
Input Offset Voltage		0.5	5		0.5	6	mV	$V_{CM} = 0V$, $R_S \leq 50\Omega$
Input Offset Current		2	20		2	100	nA	$V_{CM} = 0V$
Input Bias Current		50	100		50	250	nA	$V_{CM} = 0V$
Supply Current (4 Op-Amps)		1.4	2.0		1.4	2.5	mA	
Large Signal Voltage Gain	100	1000		50	1000		V/mV	$R_L = 10\ k\Omega$, $\Delta V_{OUT} = ±10V$
Input CM Range	±13.5	±14		±13.5	±14		V	
CM Rejection Ratio	80	100		70	100		dB	$R_S \leq 10\ k\Omega$
Power Supply Rejection Ratio	80	100		74	100		dB	$R_S \leq 10\ k\Omega$
Output Voltage Swing	±12	±14		±12	±14		V	$R_L \geq 10\ k\Omega$
Short-Circuit Current	5	20	30	5	20	30	mA	
Gain Bandwidth Product	0.8	1.2		0.5	1.2		MHz	
Phase Margin		60			60		Deg	
Slew Rate		0.4			0.4		V/μs	
Input Noise Voltage		28			28		$nV/\sqrt{Hz}$	f = 1 kHz
Channel Separation		120			120		dB	$R_L = 10\ k\Omega$, $\Delta V_{OUT} = 0V$ to ±12V
Input Resistance		1.0			1.0		MΩ	
Input Capacitance		2.0			2.0		pF	

The following specifications apply over the Maximum Operating Temperature Range.

Input Offset Voltage		0.5	6		0.5	7.5	mV	$V_{CM} = 0V$, $R_S \leq 50\Omega$
Input Offset Current		2	25		2	100	nA	$V_{CM} = 0V$
Input Bias Current		50	100		50	250	nA	$V_{CM} = 0V$
Supply Current (4 Op-Amps)		1.5	2.0		1.5	2.5	mA	
Large Signal Voltage Gain	50	1000			25	1000	V/mV	$R_L = 10\ k\Omega$, $\Delta V_{OUT} = ±10V$
Input CM Range	±13.5	±14		±13.5	±14		V	
CM Rejection Ratio	70	100		70	100		dB	$R_S \leq 50\Omega$
Power Supply Rejection Ratio	76	100		74	100		dB	$R_S \leq 50\Omega$
Output Voltage Swing	±12	±14		±12	±14		V	$R_L \geq 10\ k\Omega$

ELECTRICAL CHARACTERISTICS ($T_A = 25°C$, $V_S = ±15V$, $I_{SET} = 1\ \mu A$)

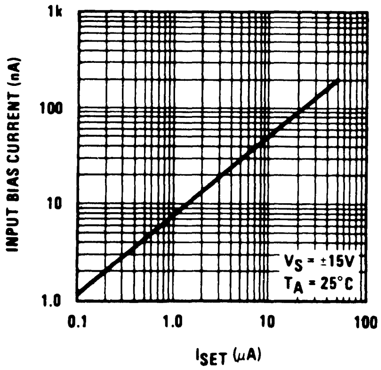
Input Offset Voltage		0.5	5		0.5	6	mV	$V_{CM} = 0V$, $R_S \leq 50\Omega$
Input Bias Current		7.5	20		7.5	100	nA	$V_{CM} = 0V$
Supply Current (4 Op-Amps)		140	250		140	300	μA	
Gain Bandwidth Product	80	100		50	100		kHz	

ELECTRICAL CHARACTERISTICS ($T_A = +25°C$, $V_S = ±1.5V$, $I_{SET} = 10\ \mu A$)

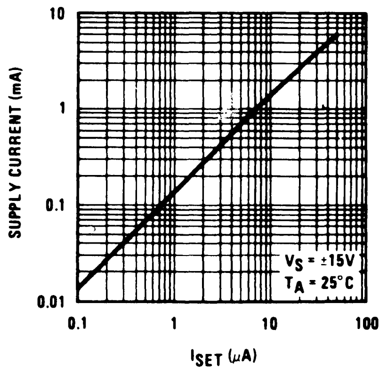
Input Offset Voltage		0.5	5		0.5	7	mV	$V_{CM} = 0V$, $R_S \leq 50\Omega$
Input CM Range	±0.7			±0.7			V	
CM Rejection Ratio		80			80		dB	$R_S \leq 50\Omega$
Output Voltage Swing	±0.6			±0.6			V	$R_L \geq 10\ k\Omega$

TYPICAL PERFORMANCE CHARACTERISTICS

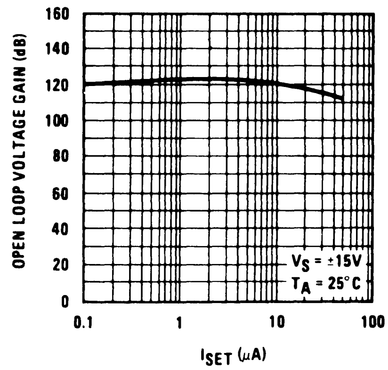
Input Bias Current vs I_{SET}



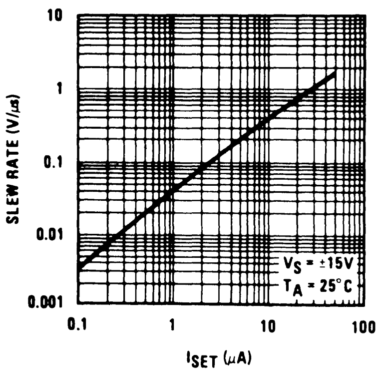
Supply Current vs I_{SET}



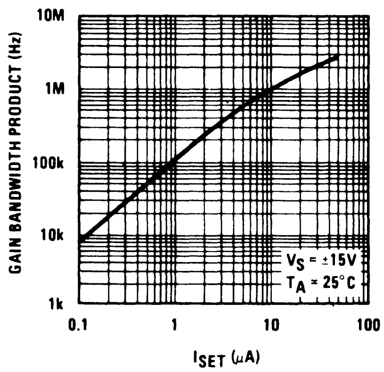
Open Loop Voltage Gain vs I_{SET}



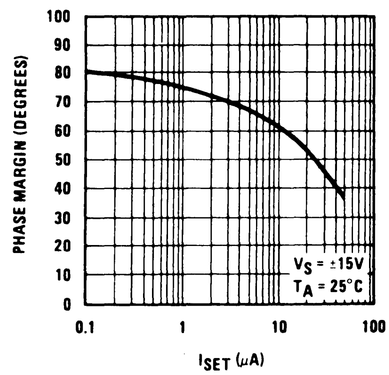
Slew Rate vs I_{SET}



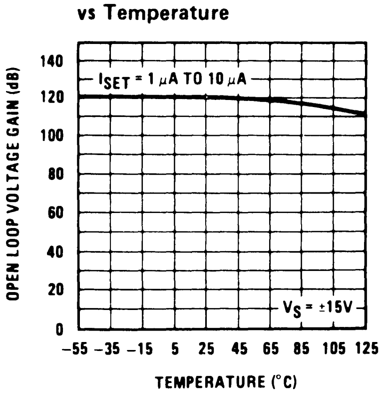
Gain Bandwidth Product vs I_{SET}



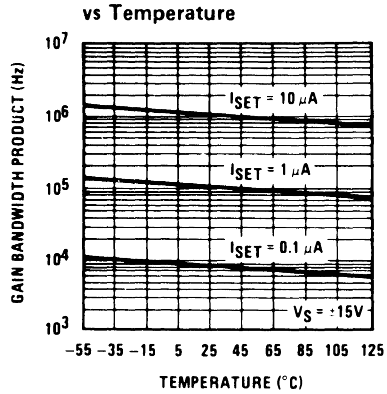
Phase Margin vs I_{SET}



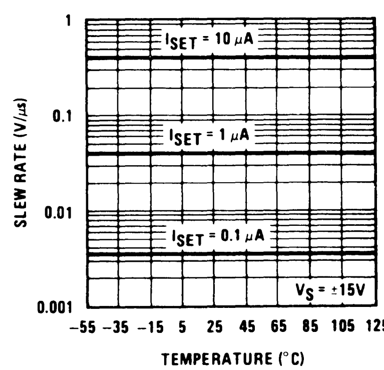
Open Loop Voltage Gain vs Temperature



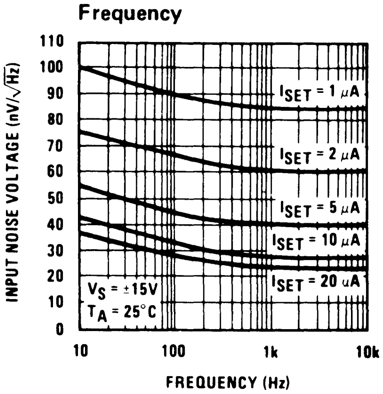
Gain Bandwidth Product vs Temperature



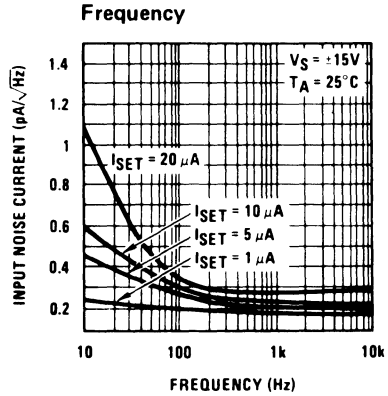
Slew Rate vs Temperature



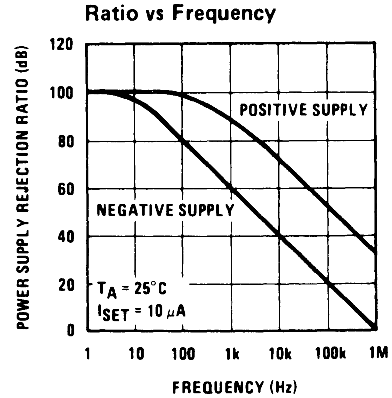
Input Noise Voltage vs Frequency



Input Noise Current vs Frequency

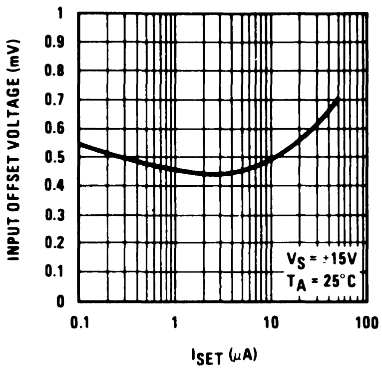


Power Supply Rejection Ratio vs Frequency

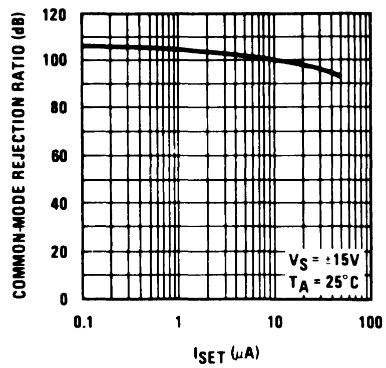


TYPICAL PERFORMANCE CHARACTERISTICS (Continued)

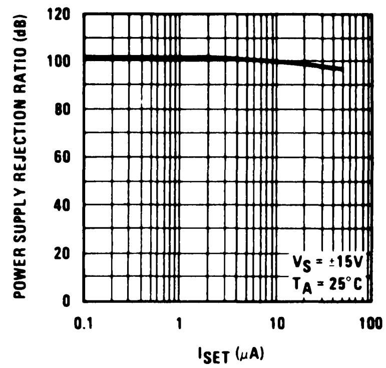
Input Offset Voltage vs I_{SET}



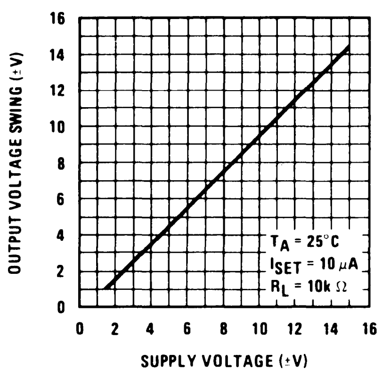
Common-Mode Rejection Ratio vs I_{SET}



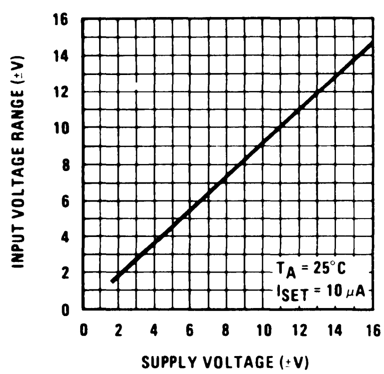
Power Supply Rejection Ratio vs I_{SET}



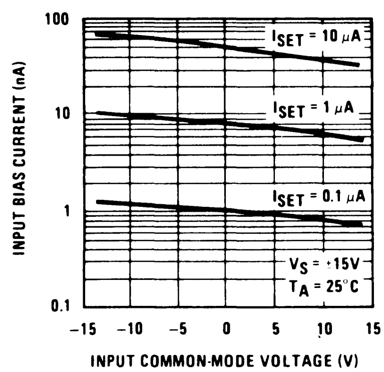
Output Voltage Swing vs Supply Voltage



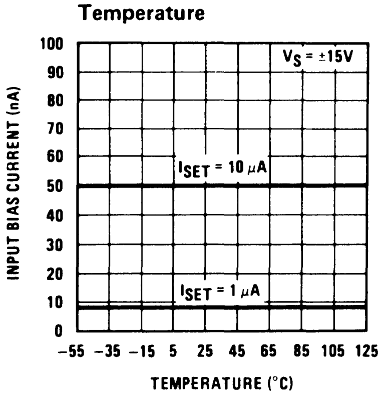
Input Voltage Range vs Supply Voltage



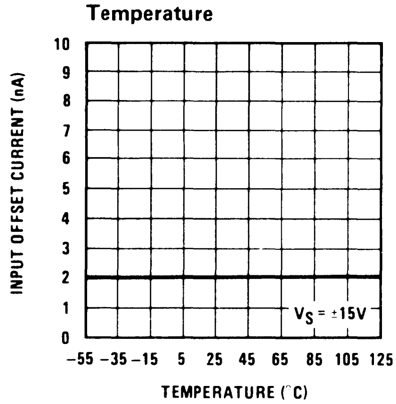
Input Bias Current vs Input Common-Mode Voltage



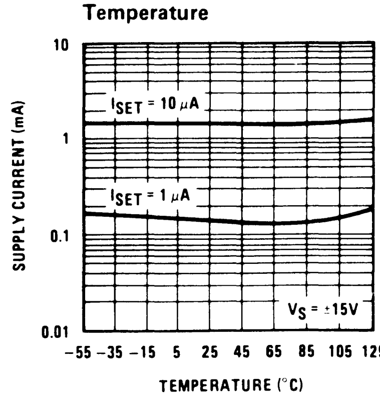
Input Bias Current vs Temperature



Input Offset Current vs Temperature



Supply Current vs Temperature



Note 1: For supply voltages less than ±15V, the absolute maximum input voltage is equal to the supply voltage.

Note 2: The maximum power dissipation for these devices must be derated at elevated temperatures and is dictated by T_{jMAX}, θjA, and the ambient temperature, T_A. The maximum available power dissipation at any temperature is $P_d = (T_{jMAX} - T_A)/\theta jA$ or the 25°C P_{dMAX}, whichever is less.

Note 3: Any of the amplifier outputs can be shorted to ground indefinitely; however, more than one should be simultaneously shorted as the maximum junction temperature will be exceeded.

XR-3403/3503

Quad Operational Amplifier

GENERAL DESCRIPTION

The XR-3403 is an array of four independent operational amplifiers, each with true differential inputs. The device has electrical characteristics similar to the popular 741. However, the XR-3403 has several distinct advantages over standard operational amplifier types in single supply applications. The XR-3403 can operate at supply voltages as low as 3.0 volts or as high as 36 volts with quiescent currents about one-fifth of those associated with the 741 (on a per amplifier basis). The common mode input range includes the negative supply, thereby eliminating the necessity for external biasing components in many applications. The output voltage range also includes the negative power supply voltage. The XR-3503 is the military-grade version of the XR-3403.

FEATURES

Short Circuit Protected Outputs
Class AB Output Stage for Minimal Crossover Distortion
True Differential Input Stage
Single Supply Operation: 3.0 to 36 Volts
Split Supply Operation: ±1.5 to ±18 Volts
Low Input Bias Currents: 500 nA Max
Four Amplifiers per Package
Internally Compensated
Similar Performance to Popular 741
Direct Pin-for-Pin Replacement for MC3403/3503, LM324 and RC4137

ABSOLUTE MAXIMUM RATINGS

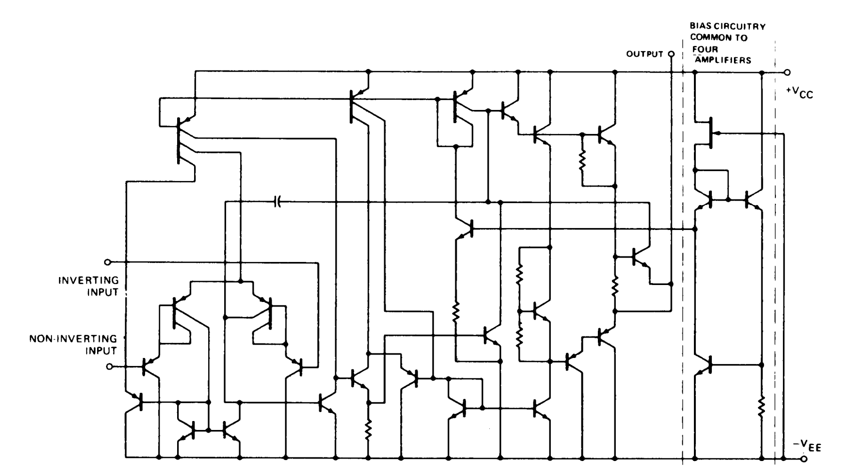
Power Supply Voltages	
Single Supply	36V
Split Supplies	±18V
Input Differential Voltage Range with Split Power Supply	±30V
Input Common Mode Voltage Range*	±15V
Package Power Dissipation:	
Plastic Package	625 mW
Derate above T_A = + 25°C	5.0 mV/°C
Ceramic Package	750 mW
Derate above T_A = +25°C	6.0 mW/°C
Storage Temperature Range	− 65°C to +150°C

*For Supply Voltage less than ±15V, the absolute maximum input voltage is equal to the supply voltage.

AVAILABLE TYPES

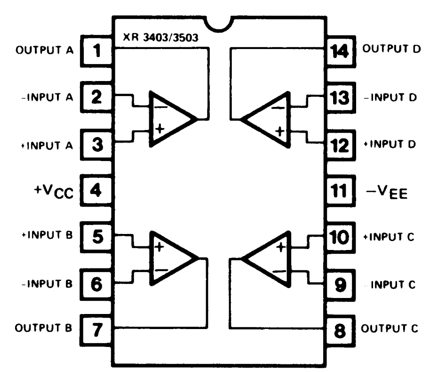
Part Number	Package	Operating Temperature
XR-3503M	Ceramic	− 55°C to +125°C
XR-3403CN	Ceramic	0°C to +75°C
XR-3403CP	Plastic	0°C to +75°C

EQUIVALENT SCHEMATIC DIAGRAM



(One Channel Only)

FUNCTIONAL BLOCK DIAGRAM



ELECTRICAL CHARACTERISTICS
(V_{CC} = +15V, V_{EE} = 15V, T_A = +25°C unless otherwise noted.)

CHARACTERISTICS	XR-3503M MIN.	XR-3503M TYP.	XR-3503M MAX.	XR-3403C MIN.	XR-3403C TYP.	XR-3403C MAX.	UNITS	CONDITIONS
Input Offset Voltage		2.0	5.0 6.0		2.0	10 12	mV	 T_A=T_{high} to T_{low}[1]
Input Offset Current		30	50 200		30	50 200	nA	 T_A=T_{high} to T_{low}
Large Signal Open-Loop Voltage Gain	 50 25	 200 300		 20 15	 200		V/mV	V_O = ±10V R_L = 2.0KΩ T_A = T_{high} to T_{low}
Input Bias Current		−200 −300	−500 −1500		−200	−500 −800	nA	 T_A = T_{high} to T_{low}
Output Impedance		75			75		Ω	f = 20 Hz
Input Impedance	0.3	1.0		0.3	1.0		MΩ	f = 20 Hz
Output Voltage Swing	±12 ±10 ±10	±13.5 ±13		±12 ±10 ±10	±13.5 ±13		V	R_L = 10 KΩ R_L = 2.0 KΩ R_L = 2.0 KΩ T_A = T_{high} to T_{low}
Input Common Mode Voltage Range	+13V-V_{EE}	+13.5V-V_{EE}		+13V-V_{EE}	+13.5V-V_{EE}		V	
Common Mode Rejection Ratio	70	90		70	90		dB	R_S<10KΩ
Power Supply Current (V_O = 0)		2.8	4.0		2.8	7.0	mA	R_L = ∞
Individual Output Short-Circuit Current[2]	±20	±30	±45	±10	±20	±45	mA	
Positive Power Supply Rejection Ratio		30	150		30	150	μV/V	
Negative Power Supply Rejection Ratio		30	150		30	150	μV/V	
Average Temperature Coefficient of Input Offset Current		50			50		pA/°C	T_A = T_{high} to T_{low}
Average Temperature Coefficient of Input Offset Voltage		10			10		μV/°C	T_A = T_{high} to T_{low}
Power Bandwidth		9.0			9.0		kHz	A_V = 1, R_L = 2.0KΩ V_o = 20V (p-p) THD = 5%
Small Signal Bandwidth		1.0			1.0		MHz	A_V = 1, R_L = 10KΩ V_o = 50 mV
Slew Rate		0.6			0.6		V/μs	A_V = 1, V_i = −10V to +10V
Rise Time		0.6			0.6		μs	A_V = 1, R_L = 10KΩ V_o = 50mV
Fall Time		0.6			0.6		μs	A_V = 1, R_L = 10KΩ V_o = 50 MV
Overshoot		20			20		%	A_V = 1, R_L = 10KΩ V_o = 50 mV
Phase Margin		60			60		Degrees	A_V = 1, R_L = 2.0KΩ C_L = 200 pF
Crossover Distortion		1.0			1.0		%	(V_{in} = 30 mV p-p V_{out} = 2.0V p-p F = 10 kHz)

[1]T_{high} = +125°C for XR-3503M, +70°C for XR-3403C
T_{low} = −55°C for XR-3503M, 0°C for XR-3403C

[2]Not to exceed maximum package power dissipation.
[3]Output will swing to ground.

ELECTRICAL CHARACTERISTICS
(V_{CC} = 5.0V, V_{EE} = Gnd, T_A = +25°C unless otherwise noted.)

CHARACTERISTICS	XR-3503M MIN.	XR-3503M TYP.	XR-3503M MAX.	XR-3403C MIN.	XR-3403C TYP.	XR-3403C MAX.	UNITS	CONDITIONS
Input Offset Voltage		2.0	5.0		2.0	10	mV	
Input Offset Current		30	50		30	50	nA	
Input Bias Current		−200	−500		−200	−500	nA	
Large Signal Open Loop Voltage Gain	20	200		20	200		V/mV	R_L = 2.0KΩ
Power Supply Rejection Ratio			150			150	μV/V	
Output Voltage Range[3]	3.5 V_{CC}−1.5V			3.5 V_{CC}−1.5V			Vp-p	R_L = 10KΩ V_{CC} = 5.0V R_L = 10 KΩ 5.0V ≤ V_{CC} ≤ 30V
Power Supply Current		2.5	4.0		2.5	7.0	mA	
Channel Separation		−120			−120		dB	f = 1.0 kHz to 20 kHz (Input Referenced)

XR-4136

Quad Operational Amplifier

GENERAL DESCRIPTION

The XR-4136 is an array of four independent internally-compensated operational amplifiers on a single silicon chip, each similar to the popular 741, but with a power consumption less than one 741. Good thermal tracking and matched gain-bandwidth products make these quad op-amps useful for active filter applications.

FEATURES

Direct Pin-for-Pin Replacement for RC4136 and RM4136
Low Power Consumption – 50 mW typ. and 120 mW max.
Short-Circuit Protection
Internal Frequency Compensation
No Latch-Up
Wide Common-Mode and Differential Voltage Ranges
Matched Gain-Bandwidth

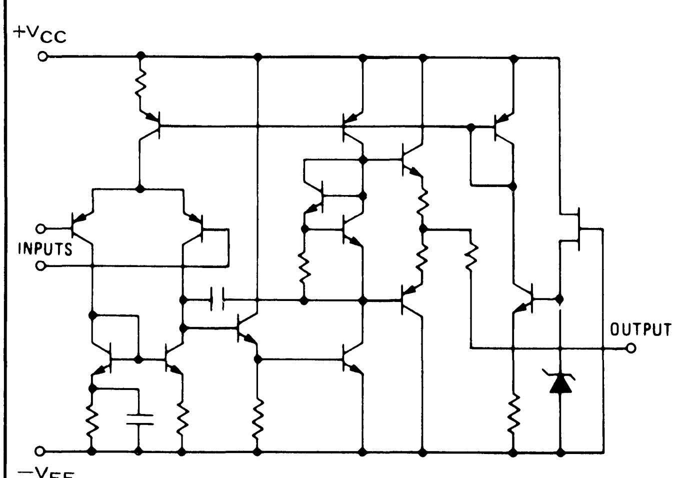
ABSOLUTE MAXIMUM RATINGS

Supply Voltage	
XR-4136M	±22V
XR-4136C	±18V
Common Mode	
Voltage Range	$-V_{EE}$ to $+V_{CC}$
Differential Input Voltage	±30V
Internal Power Dissipation	
Ceramic Package:	750 mW
Derate above T_A = +25°C	6 mW/°C
Plastic Package:	625 mW
Derate above T_A = +25°C	5 mW/°C
Storage Temperature Range:	–65°C to +150°C

AVAILABLE TYPES

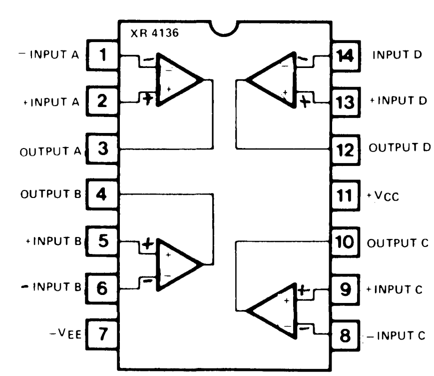
Part Number	Package	Operating Temperature
XR-4136M	Ceramic	–55°C to +125°C
XR-4136CN	Ceramic	0°C to +75°C
XR-4136CP	Plastic	0°C to +75°C

EQUIVALENT SCHEMATIC



(One Channel Only)

FUNCTIONAL BLOCK DIAGRAM



ELECTRICAL CHARACTERISTICS $T_A = +25°C$, $V_S = \pm15V$ unless otherwise specified

CHARACTERISTICS	XR4136M MIN.	XR4136M TYP.	XR4136M MAX.	XR4136C MIN.	XR4136C TYP.	XR4136C MAX.	UNITS	SYMBOLS	CONDITIONS
Input Offset Voltage		1	5.0		1	6.0	mV	$\lvert V_{io} \rvert$	$R_S \leq 10\ K\Omega$
Input Offset Current		10	200		10	200	nA	$\lvert I_{io} \rvert$	
Input Bias Current		80	500		80	500	nA	$\lvert I_b \rvert$	
Input Resistance	0.3	1.8		0.3	1.8		MΩ	R_{in}	
Large Signal Voltage Gain	50	60		20	40		V/mV	A_{VOL}	$R_L \geq 2\ K\Omega$, $V_{out} = \pm10V$
Output Voltage Swing	±12	±14		±12	±14		V	V_{out}	$R_L \geq 10\ K\Omega$
	±10	±12		±10	±12		V	V_{out}	$R_L \geq 2\ K\Omega$
Input Voltage Range	±12	±13.5		±12	±13.5		V	V_{ICM}	
Common Mode Rejection Ratio	70	105		70	105		dB	CMRR	$R_S \leq 10\ K\Omega$
Supply Voltage Rejection Ratio		10	150		10	150	μV/V	PSRR	$R_S \leq 10\ K\Omega$
Power Consumption		50	120		50	120	mW	P_i	
Transient Response (unity gain)									$V_{in} = 20\ mV$, $R_L = 2\ K\Omega$, $C_L \leq 100\ pF$
Risetime		0.07			0.07		μs	t_r	
Overshoot		20			20		%	t_o	
Unity Gain Bandwidth	2.0	3.0			3.0		MHz	BW	
Slew Rate (unity gain)		1.6			1.6		V/μs	dV_{out}/dt	$R_L \geq 2\ K\Omega$
Channel Separation (open loop)		120			120		dB		f = 10 KHz, $R_S = 1\ K\Omega$
(Gain of 100)		105			105		dB		f = 10 KHz, $R_S = 1\ K\Omega$
The following specifications apply for $-55°C \leq T_A \leq +125°C$ for XR-4136M: $0°C \leq T_A \leq +70°C$ for XR-4136C									
Input Offset Voltage			6.0			7.5	mV	$\lvert V_{io} \rvert$	$R_S \leq 10\ K\Omega$
Input Offset Current			500			300	nA	$\lvert I_{io} \rvert$	
Input Bias Current			1500			800	nA	I_b	
Large-Signal Voltage Gain	25			15			V/mV	A_{VOL}	$R_L \geq 2\ K\Omega$, $V_{out} = \pm10V$
Output Voltage Swing	±10			±10			V	V_{out}	$R_L \geq 2\ K\Omega$
Power Consumption									$V_S = \pm15V$
			150			150	mW	P_i	T_A = High
			200			200	mW	P_i	T_A = Low
Output Short-Circuit Current	5	17	35	5	17	35	mA	I_{SC}	

TYPICAL PARAMETER MATCHING:

$T_A = +25°C$, $V_S = \pm15V$ unless otherwise noted

CHARACTERISTICS	XR4136M TYP.	XR4136C TYP.	UNITS	SYMBOLS	CONDITIONS
Input Offset Voltage	±1.0	±2.0	mV	$\lvert V_{io} \rvert$	$R_S \geq 10\ K\Omega$
Input Offset Current	±7.5	±7.5	nA	$\lvert I_{io} \rvert$	
Input Bias Current	±15	±15	nA	I_b	
Voltage Gain	±0.5	±1.0	dB	A_{VOL}	$R_S \geq 2\ K\Omega$

XR-4202

Programmable Quad Operational Amplifier

GENERAL DESCRIPTION

The XR-4202 is an array of four independent operational amplifiers on a single silicon chip. The operating current of the array is externally controlled by a single resistor or current source, allowing the user to trade-off power dissipation for bandwidth.

FEATURES

- Programmable
- Micropower Operation
- Wide Input Voltage and Common Mode Range
- Internal Frequency Compensation
- No Latch-Up
- Matched Parameters
- Short-Circuit Protection

APPLICATION INFORMATION

The following approximate relations are useful for design:

Gain-Bandwidth Product	≈	50 I_{SET}	(KHz)
Power Supply Current	≈	30 I_{SET}	(μA)
Slew Rate	≈	20 I_{SET}	(V/ms)

Where: I_{SET} is in μA

$$I_{SET} = \frac{V_{EE} - V_{BE}}{R_{SET}}$$ WHERE V_{BE} DIODE VOLTAGE ≈ 0.65V

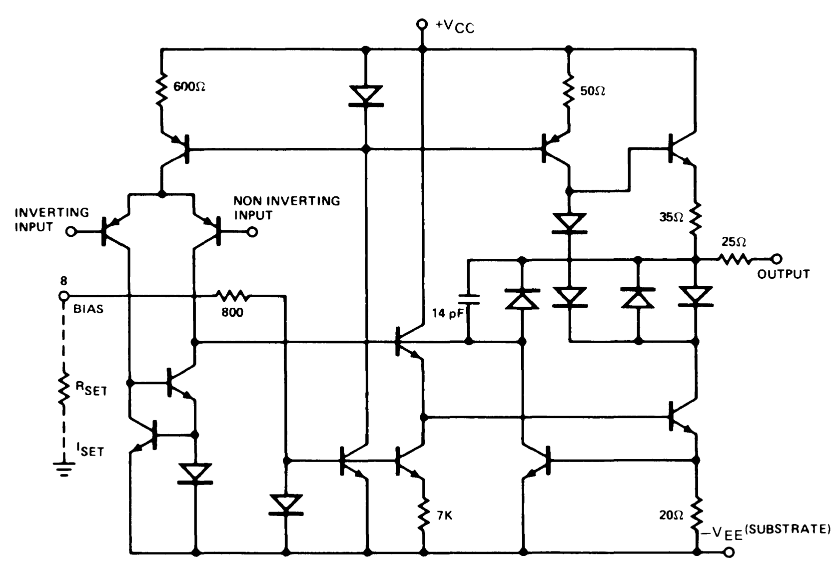
ABSOLUTE MAXIMUM RATINGS

Supply Voltage	±18V
Differential Input Voltage	±30V
Power Dissipation	
Ceramic Package:	750 mW
Derate above $T_A = +25°C$	6 mW/°C
Plastic Package:	625 mW
Derate above $T_A = +25°C$	5.0 mW/°C
Common Mode Range	V_{EE} to V_{CC}
Short Circuit Duration	Indefinite
Storage Temperature	−60°C to +150°C

AVAILABLE TYPES

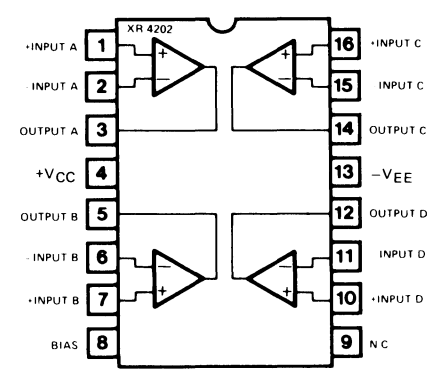
Part Number	Package	Operating Temperature
XR-4202M	Ceramic	-55°C to +125°C
XR-4202N	Ceramic	-40°C to +85°C
XR-4202P	Plastic	-40°C to +85°C

EQUIVALENT SCHEMATIC DIAGRAM



1/4 of XR-4202

FUNCTIONAL BLOCK DIAGRAM



ELECTRICAL CHARACTERISTICS HIGH POWER MODE (V_S = ±15V, I_{SET} = 75 μA and T_A = +25°C unless otherwise specified)

CHARACTERISTICS	MIN	TYP	MAX	UNITS	SYMBOL	CONDITIONS
Short Circuit Current	5	17	30	mA	I_{SC}	0°C ≤ T_A ≤ 75°C
Supply Current	0.8	1.7	6.0	mA	I_s	Note 3
Input Offset Voltage		0.8	5.0	mV	V_{io}	R_s ≤ 10 KΩ
Input Bias Current		80	500	nA	I_b	
Input Off-set Current		10	200	nA	I_{io}	
Input Resistance	0.1	0.6		MΩ	R_{in}	
Input Common Mode Voltage Range	12	±14		±V	V_{iCM}	
Common Mode Rejection Ratio	70	110		dB	CMRR	
Voltage Supply Rejection Ratio		15	150	μV/V	PSRR	
Large Signal Voltage Gain	74	88		dB	A_{VOL}	R_L = 3 KΩ; ΔV_o = ±10V
Output Voltage Swing	±10	±13.6		±V	V_{out}	R_L = 3 KΩ
Gain-Bandwidth Product		3.5		MHz	f_1	
Phase Margin		45		Deg.		
Rise Time		70		ns	t_R	ΔV_o = ±20 mV
Overshoot		20		%	t_o	ΔV_o = ±20 mV
Channel Separation		120		dB		Any amp. pair: freq. = 1 Hz, R_L = 3 KΩ
		105		dB		Any amp. pair: freq. = 10 KHz, R_L = 3 KΩ
Slew Rate		1.5		V/μs	$dV_{out/dt}$	
Input Voltage Noise		25		nV/√Hz	e_n	Bandwidth 100 Hz to 10 KHz

Note: Short circuit may be taken to either supply line or ground on only one amplifier at a time.

ELECTRICAL CHARACTERISTICS HIGH POWER MODE (V_S = ±15V, I_{SET} = 75 μA and T_A = -55°C to +125°C)

CHARACTERISTICS	MIN	TYP	MAX	UNITS	SYMBOL	CONDITIONS
Input Offset Voltage		0.8	10	mV	V_{io}	R_s ≤10 KΩ
Input Bias Current		80	1500	nA	I_b	
Input Offset Current		10	200	nA	I_{io}	
Large Signal Voltage Gain	68	88		dB	A_{vol}	R_L = 3 KΩ ΔV_o = ±10V

ELECTRICAL CHARACTERISTICS MICROPOWER MODE (I_{SET} = 1 μA, V_S = ±1.5V)

CHARACTERISTICS	MIN	TYP	MAX	UNITS	SYMBOL	CONDITIONS
Supply Current			100	μA	I_s	Note 3
Input Bias Current			200	nA	I_B	
Input Offset Current			20	nA	I_{OS}	
Input Offset Voltage		0.5	5	mV	V_{os}	R_s ≤ 10 KΩ
Input Resistance	0.5			MΩ	R_{in}	
Input Common Mode Voltage Range	0.3	±0.8		±V	V_{iCM}	
Common Mode Rejection Ratio	60	100		dB	CMRR	
Voltage Supply Rejection Ratio		20	200	μV/V	PSRR	
Large Signal Voltage Gain	66	80		dB	A_{vol}	R_L ≥ 100 KΩ
Gain-Bandwidth Product		50		KHz	f_1	
Phase Margin		75		Deg.		
Slew-Rate		20		V/ms	$dV_{out/dt}$	
Rise Time		7		μs	t_R	ΔV_o = ±20 mV
Overshoot		0		%	t_o	ΔV_o = ±20 mV
Channel Separation		120		dB		Freq. = Hz: R_L = 20 KΩ, ΔV_o = ±0.5V
		120		dB		Freq. = 1 KHz: R_L = 10 KΩ, ΔV_o = ±0.5V
Equivalent Input Voltage Noise		200		nV/√Hz	e_n	Bandwidth = 100 Hz to 10 KHz

PARAMETER MATCHING (I_{SET} = 75 μA[2])

CHARACTERISTICS	MIN	TYP	MAX	UNITS	SYMBOL	CONDITIONS
Input Offset Voltage		1		±mV	V_{os}	R_s ≤ 10 KΩ
Input Bias Current		10		±nA	I_B	
Input Offset Current		2		±nA	I_{os}	
Gain-Bandwidth Product		100		±KHz	f_1	
Slew Rate		0.2		±V/μs	$dV_{o/dt}$	

NOTES: 1. All tests refer to a single Op. amp unless otherwise specified.
2. Tests apply for parameter matching between any Op. amp pair.
3. Tests apply to four Op. amps and bias network.

XR-4212

Quad Operational Amplifier

GENERAL DESCRIPTION

The XR-4212 is an array of four independent internally compensated operational amplifiers on a single silicon chip, each similar to the popular 741, but with a power consumption less than one 741. Good thermal tracking and matched gain-bandwidth products make these Quad Op-amps useful for active filter applications.

FEATURES

Same Pinout as MC3403 and LM324
Low Power Consumption – 50 mW typ. and 120 mW max.
Short-Circuit Protection
Internal Frequency Compensation
No Latch-Up
Wide Common-Mode and Differential Voltage Ranges
Matched Gain-Bandwidth

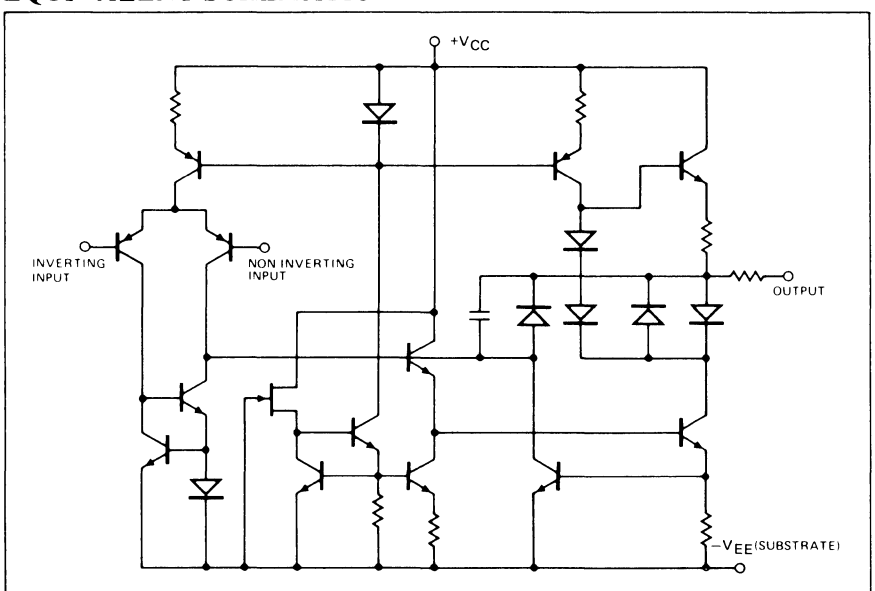
ABSOLUTE MAXIMUM RATINGS

Supply Voltage	
XR-4212M	±22V
XR-4212C	±18V
Common Mode Voltage	V_{EE} to V_{CC}
Output Short-Circuit Duration	Indefinite
Differential Input Voltage	±30V
Internal Power Dissipation	
Ceramic Package:	750 mW
Derate above $T_A = +25°C$	6 mW/°C
Plastic Package:	625 mW
Derate above $T_A = +25°C$	5 mW/°C
Storage Temperature Range:	−65°C to +150°C

AVAILABLE TYPES

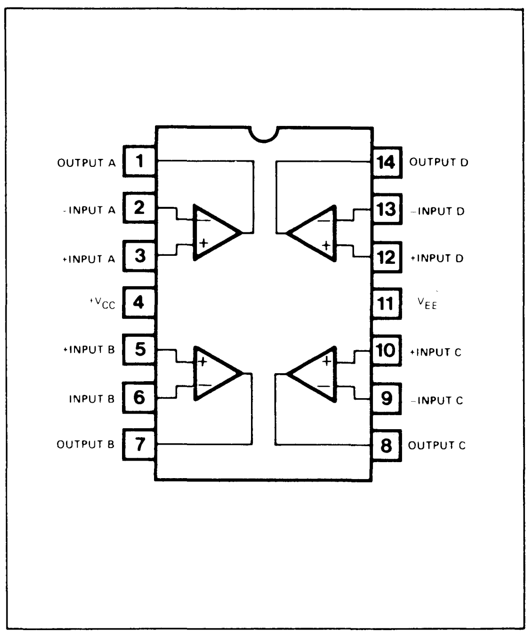
Part Number	Package	Operating Temperature
XR-4212M	Ceramic	−55°C to +125°C
XR-4212CN	Ceramic	0°C to +75°C
XR-4212CP	Plastic	0°C to +75°C

EQUIVALENT SCHEMATIC



(one channel only)

FUNCTIONAL BLOCK DIAGRAM



ELECTRICAL CHARACTERISTICS $T_A = +25°C$, $V_S = \pm 15V$ unless otherwise specified

CHARACTERISTICS	XR-4212M MIN.	XR-4212M TYP.	XR-4212M MAX.	XR-4212C MIN.	XR-4212C TYP.	XR-4212C MAX.	UNITS	SYMBOLS	CONDITIONS
Input Offset Voltage		1	5.0		1	6.0	mV	$\|V_{io}\|$	$R_s \leq 10$ KΩ
Input Offset Current		10	50		10	50	nA	$\|I_{io}\|$	
Input Bias Current		80	500		80	500	nA	$\|I_b\|$	
Input Resistance	0.3	1.8		0.3	1.8		MΩ	R_{in}	
Large Signal Voltage Gain	20	60		5	40		V/mV	AVOL	$R_L \geq 2$ KΩ $V_{out} = \pm 10V$
Output Voltage Swing	±12	±14		±12	±14		V	V_{out}	$R_L \geq 10$ KΩ
	±10	±12		±10	±12		V	V_{out}	$R_L \geq 2$ KΩ
Input Voltage Range	±12	±13.5		±12	±13.5		V	V_{iCM}	
Common Mode Rejection Ratio	70	105		70	105		dB	CMRR	$R_s \leq 10$ KΩ
Supply Voltage Rejection Ratio		10	150		10	150	μV/V	PSRR	$R_s \leq 10$ KΩ
Power Consumption		50	120		50	120	mW	P_i	
Transient Response (unity gain)									$V_{in} = 20$ mV $R_L = 2$ KΩ
Risetime		0.07			0.07		μs	t_r	$C_L \leq 100$ pF
Overshoot		20			20		%	t_o	
Unity Gain Bandwidth	2.0	3.0			3.0		MHz	BW	
Slew Rate (unity gain)		1.6			1.6		V/μs	dV_{out}/dt	$R_L \geq 2$ KΩ
Channel Separation (open loop)		120			120		dB		f = 10 KHz $R_s = 1$ KΩ
(Gain of 100)		105			105		dB		f = 10 KHz $R_s = 1$ KΩ
The following specifications apply for $-55°C \leq T_A \leq +125°C$ for XR-4212M: $0°C \leq T_A \leq +70°C$ for XR-4212C									
Input Offset Voltage			6.0			7.5	mV	$\|V_{io}\|$	$R_s \leq 10$ KΩ
Input Offset Current			200			200	nA	$\|I_{io}\|$	
Input Bias Current			1500			800	nA	I_b	
Large-Signal Voltage Gain	20			5			V/mV	AVOL	$R_L \geq 2$ KΩ $V_{out} = \pm 10V$
Output Voltage Swing	±10			±10			V	V_{out}	$R_L \geq 2$ KΩ
Power Consumption			150 200			150 200	mW mW	P_i P_i	$V_s = \pm 15V$ T_A = High T_A = Low
Output Short-Circuit Current	5	17	35	5	17	35	mA	I_{SC}	

TYPICAL PARAMETER MATCHING:

$T_A = +25°C$, $V_S = \pm 15V$ unless otherwise noted

CHARACTERISTICS	XR-4212M TYP.	XR-4212C TYP.	UNITS	SYMBOLS	CONDITIONS
Input Offset Voltage	±1.0	±2.0	mV	$\|V_{io}\|$	$R_s \geq 10$ KΩ
Input Offset Current	±7.5	±7.5	nA	$\|I_{io}\|$	
Input Bias Current	±15	±15	nA	I_b	
Voltage Gain	±0.5	±1.0	dB	AVOL	$R_s \geq 2$ KΩ

XR-4741

Quad Operational Amplifier

GENERAL DESCRIPTION

The XR-4741 is an array of four independent internally-compensated operational amplifiers on a single silicon chip, each similar to the popular 741. Each amplifier offers performance equal to or better than the 741 type in all respects. It has high slew rate, superior bandwidth, and low noise, which makes it excellent for audio amplifiers or active filter applications.

FEATURES

Short-Circuit Protection	
Internal Frequency Compensation	
No Latch-Up	
Wide Common-Mode and Differential Voltage Ranges	
Matched Gain-Bandwidth	
High Slew Rate	1.6V/μS(Typ)
Unity Gain-Bandwidth	3.5 MHz(Typ)
Low Noise Voltage	9 nV$\sqrt{Hz}$
Input Offset Current	60 nA(Typ)
Input Offset Voltage	.5 mV(Typ)
Supply Range	±2V to ±20V

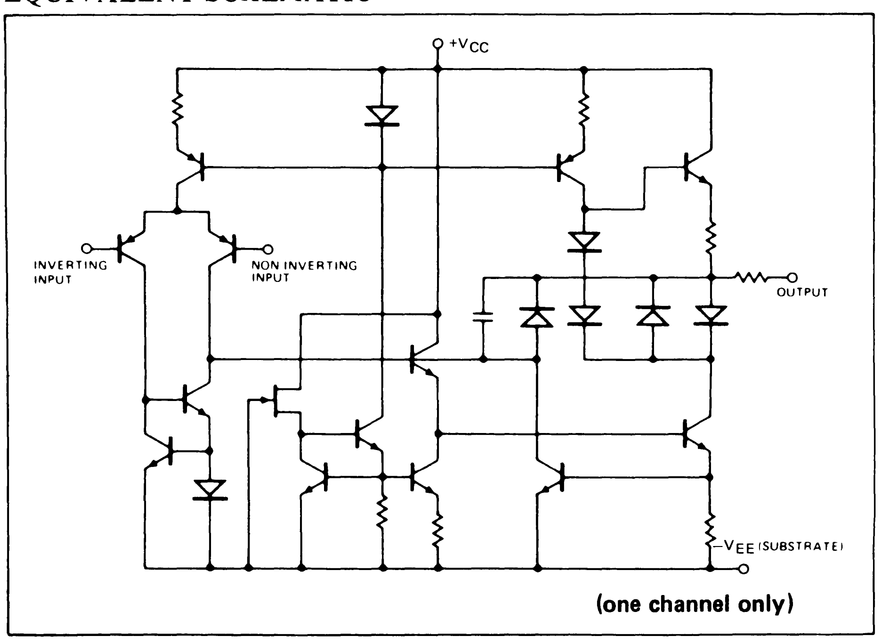
ABSOLUTE MAXIMUM RATINGS

Supply Voltage	
XR-4741	±20
Common Mode	
Voltage	V_{EE} to V_{CC}
Output Short-Circuit Duration	Indefinite
Differential Input Voltage	±30V
Internal Power Dissipation	
Ceramic Package:	880 mW
Derate above T_A = +25°C	5.8 mW/°C
Plastic Package:	625 mW
Derate above T_A = +25°C	5 mW/°C
Storage Temperature Range:	−65°C to +150°C

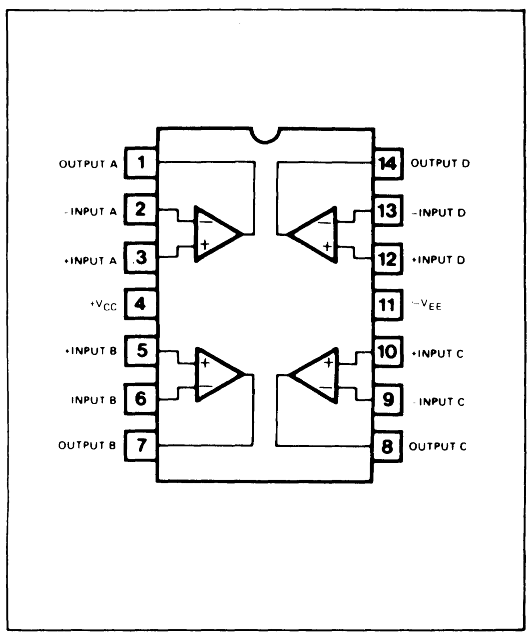
AVAILABLE TYPES

Part Number	Package	Operating Temperature
XR-4741M	Ceramic	−55°C to +125°C
XR-4741CN	Ceramic	0°C to +75°C
XR-4741CP	Plastic	0°C to +75°C

EQUIVALENT SCHEMATIC



FUNCTIONAL BLOCK DIAGRAM



ELECTRICAL CHARACTERISTICS $T_A = +25^{\circ}C$, $V_S = \pm 15V$ unless otherwise specified

CHARACTERISTICS	XR-4741M			XR-4741C			UNITS	SYMBOLS	CONDITIONS
	MIN.	TYP.	MAX.	MIN.	TYP.	MAX.			
Input Offset Voltage		0.5	3.0		1.0	5.0	mV	$\|V_{io}\|$	$R_s \leq 10$ KΩ
Input Offset Current		10	30		10	50	nA	$\|I_{io}\|$	
Input Bias Current		60	200		60	300	nA	$\|I_b\|$	
Differential Input Resistance		5			5		MΩ	R_{in}	
Input Noise Voltage (f = 1 kHz)		9			9		nV/√Hz		
Large Signal Voltage Gain	50	100		25	50		V/mV	A_{VOL}	$R_L \geq 2$ KΩ $V_{out} = \pm 10V$
Output Voltage Swing	±12	±13.7		±12	±13.7		V	V_{out}	$R_L \geq 10$ KΩ
	±10	±12.5		±10	±12.5		V	V_{out}	$R_L \geq 2$ KΩ
Full Power Bandwidth		25			25		kHz		
Output Resistance		300			300		Ω		
Input Voltage Range	±12	±13.5		±12	±13.5		V	V_{iCM}	
Common Mode Rejection Ratio	80	100		80	100		dB	CMRR	$R_s \leq 10$ KΩ
Supply Voltage Rejection Ratio		10	100		10	100	μV/V	PSRR	$R_s \leq 10$ KΩ
Power Consumption			150			210	mW	P_i	
Transient Response (unity gain)									$V_{in} = 20$ mV $R_L = 2$ KΩ
Risetime		.07			.07		μs	t_r	$C_L \leq 100$ pF
Overshoot		20			20		%	t_o	
Unit Gain Bandwidth		3.5			3.5		MHz	BW	
Slew Rate (unity gain)		1.6			1.6		V/μs	dV_{out}/dt	$R_L \geq 2$ KΩ
Channel Separation (open loop)		120			120		dB		f = 10 KHz $R_s = 1$ KΩ
(Gain of 100)		105			105		dB		f = 10 KHz $R_s = 1$ KΩ
The following specifications apply for $-55^{\circ}C \leq T_A \leq +125^{\circ}C$ for XR-4741M: $0^{\circ}C \leq T_A \leq +70^{\circ}C$ for XR-4741C									
Input Offset Voltage		4.0	5.0		5.0	6.5	mV	$\|V_{io}\|$	$R_s \leq 10$ KΩ
Input Offset Current			75			100	nA	$\|I_{io}\|$	
Input Bias Current			325			400	nA	I_b	
Input Voltage Range	±12			±12			V		
Common Mode Rejection Ratio	74			74			db		
Large-Signal Voltage Gain	25			15			V/mV	A_{VOL}	$R_L \geq 2$ KΩ $V_{out} = \pm 10V$
Output Voltage Swing	±10	±12.5		±10	±12.5		V	V_{out}	$R_L = 2$ KΩ
	±12.0	±13.7		±12	±13.7				$R_L \geq 10$ KΩ $V_s = \pm 15V$
Power Consumption			150			150	mW	P_i	T_A = High
			200			200	mW	P_i	T_A = Low
Supply Voltage Rejection Ratio		100	μV/V		100	μV/V			
Output Short-Circuit Current	±5	±15		±5	±15		mA	I_{SC}	

XR-1458/4558

Dual Operational Amplifier

GENERAL DESCRIPTION

The XR-1458/4558 is a pair of independent internally compensated operational amplifiers on a single silicon chip, each similar to the popular 741, but with a power consumption less than one 741. Good thermal tracking and matched gain-bandwidth products make these Dual Op-amps useful for active filter applications.

FEATURES

Direct Pin-for-Pin Replacement for MC1458, RC4558, N5558
Low Power Consumption – 50mW typ. and 120mW max.
Short-Circuit Protection
Internal Frequency Compensation
No Latch-Up
Wide Common-Mode and Differential Voltage Ranges
Matched Gain-Bandwidth

ABSOLUTE MAXIMUM RATINGS

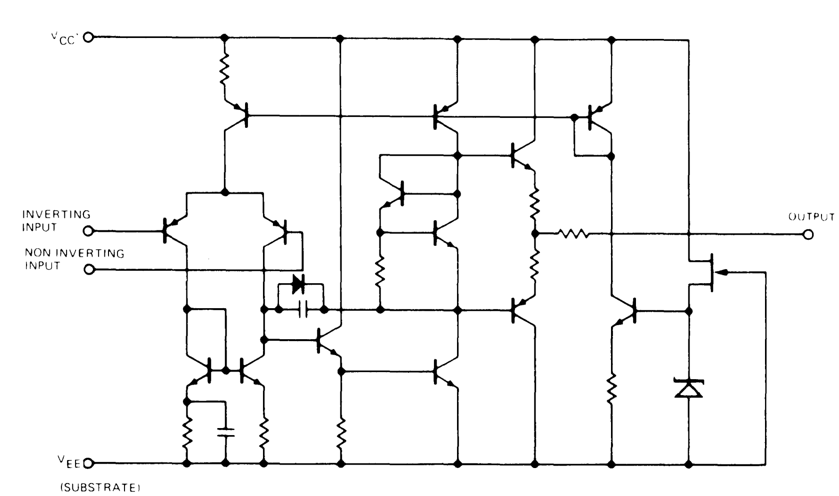
Supply Voltage	
XR-4558CP	±18V
Input Voltage (Note 1)	±15V
Common Mode Voltage Range	V_{EE} to V_{CC}
Output Short-Circuit Duration (Note 2)	Indefinite
Differential Input Voltage	±30V
Internal Power Dissipation (Note 3)	
Plastic Package:	500mW
Storage Temperature Range:	−65°C to +150°C
Operating Temperature Range:	0°C to +70°C

AVAILABLE TYPES

Part Number	Package	Operating Temperature
XR-1458CP	Plastic	0°C to +70°C
XR-4558CP	Plastic	0°C to +70°C

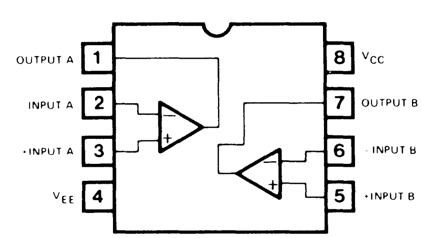
Note 1: *For supply voltages less than ±15V, the absolute maximum input voltage is equal to the supply voltage.*
Note 2: *Short circuit may be to ground or either supply. Rating applies to +125°C case temperature or +75°C ambient temperature for XR1458/4558.*
Note 3: *Rating applies for case temperatures to 125°C; derate linearly at 6.5mW/°C for ambient temperatures above +75°C for XR1458/4558.*

EQUIVALENT SCHEMATIC DIAGRAM



(One Channel Only)

FUNCTIONAL BLOCK DIAGRAM



ELECTRICAL CHARACTERISTICS

$T_A = +25^\circ C$, $V_S = \pm 15V$ unless otherwise specified

CHARACTERISTICS	XR1458/4558CP			UNITS	SYMBOLS	CONDITIONS
	MIN.	TYP.	MAX.			
Input Offset Voltage		0.5	6.0	mV	$\|V_{io}\|$	$R_S \leq 10$ KΩ
Input Offset Current		5	200	nA	$\|I_{io}\|$	
Input Bias Current		40	500	nA	$\|I_b\|$	
Input Resistance	0.3	5		MΩ	R_{in}	
Large Signal Voltage Gain	20	300		V/mV	AVOL	$R_L \geq 2$ KΩ $V_{out} = \pm 10V$
Output Voltage Swing	±12	±14		V	V_{out}	$R_L \geq 10$ KΩ
	±10	±13		V	V_{out}	$R_L \geq 2$ KΩ
Input Voltage Range	±12	±14		V	V_{iCM}	
Common Mode Rejection Ratio	70	90		dB	CMRR	$R_S \leq 10$ KΩ
Supply Voltage Rejection Ratio		30	150	μV/V	PSRR	$R_S \leq 10$ KΩ
Power Consumption		50	170	mW	P_i	
Transient Response (unity gain) Risetime Overshoot		 0.13 5		 μs %	 t_r t_o	$V_{in} = 20$ mV $R_L = 2$ KΩ $C_L \leq 100$ pF
Unity Gain Bandwidth		3.0		MHz	BW	
Slew Rate (unity gain)		1.0		V/μs	dV_{out}/dt	$R_L \geq 2$ KΩ
Channel Separation (open loop)		120		dB		f = 10 kHz $R_S = 1$ KΩ
(Gain of 100)		105		dB		f = 10 kHz $R_S = 1$ KΩ
The following specifications apply for $0^\circ C \leqslant T_A \leqslant +70^\circ C$ for XR4558CP						
Input Offset Voltage			7.5	MV	$\|V_{io}\|$	$R_S \leq 10$ KΩ
Input Offset Current			300	nA	$\|I_{io}\|$	
Input Bias Current			800	nA	I_b	
Large-Signal Voltage Gain	15			V/mV	AVOL	$R_S \geq 2$ KΩ $V_{out} = \pm 10V$
Output Voltage Swing	±10			V	V_{out}	$R_L \geq 2$ KΩ
Power Consumption		 90 120	 150 200	 mW mW	 P_i P_i	$V_S = \pm 15V$ T_A = High T_A = Low

XR-4739

Dual Low-Noise Operational Amplifier

GENERAL DESCRIPTION

The XR-4739 dual low-noise operational amplifier is fabricated on a single silicon chip using the planar epitaxial process. It was designed primarily for preamplifiers in consumer and industrial signal processing equipment. The device is pin compatible with the μA739 and MC1303, however, compensation is internal. This permits a lowered external parts count and simílified application.

The XR-4739 is available in molded dual in-line 14-pin package, and operates over the commercial temperature range from 0°C to +75°C.

FEATURES

Internally Compensated Replacement for μA739 and MC1303
Signal-to-Noise Ratio 76dB (RIAA 10mV ref.)
Channel Separation 125dB
Unity Gain Bandwidth 3MHz
Output Short-circuit Protected
0.1% Distortion at 8.5V RMS Output into 2KΩ Load

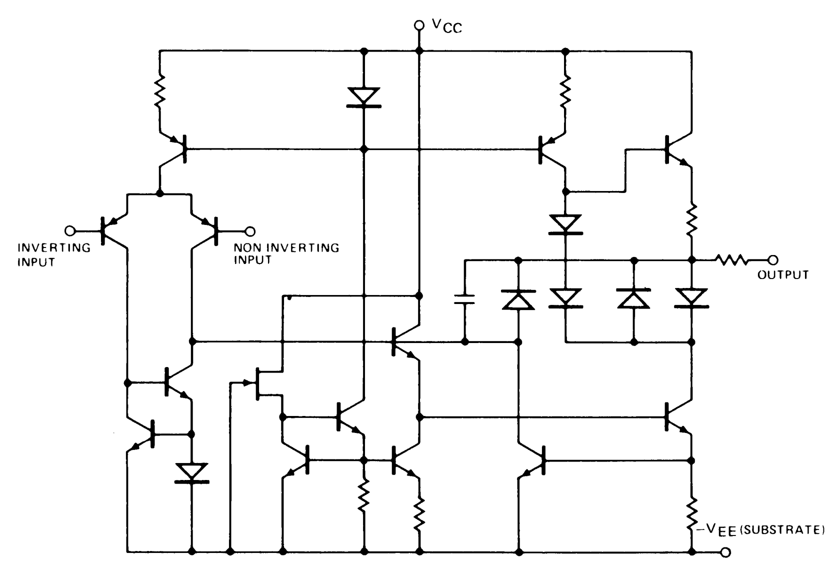
ABSOLUTE MAXIMUM RATINGS

Supply Voltage	±18V
Internal Power Dissipation (Note 1)	500 mW
Differential Input Voltage	±30V
Input Voltage (Note 2)	±15V
Storage Temperature Range	−65°C to +150°C
Lead Temperature (Soldering, 60 sec.)	300°C
Output Short-Circuit Duration (Note 3)	Indefinite

AVAILABLE TYPES

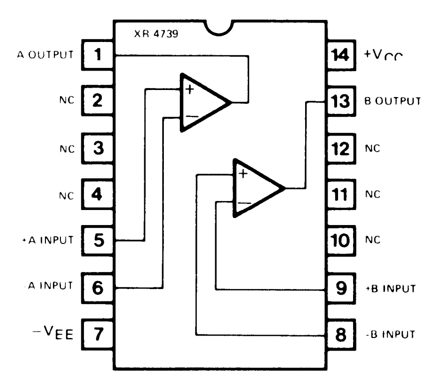
Part Number	Package Types	Operating Temperature
XR-4739CN	Ceramic	0°C to +75°C
XR-4739CP	Plastic	0°C to +75°C

SCHEMATIC DIAGRAM



(One Channel Only)

FUNCTIONAL BLOCK DIAGRAM



ELECTRICAL CHARACTERISTICS (T_A = 25°C, V_{CC} = ±15V unless otherwise specified)

PARAMETER	MIN	TYP	MAX	UNITS	CONDITIONS
Input Offset Voltage		2.0	6.0	mV	$R_S \leqslant 10$ kΩ
Input Offset Current		5.0	200	nA	
Input Bias Current		40	500	nA	
Input Resistance	0.3	5.0		MΩ	
Large-Signal Voltage Gain	20	60		K	$R_L \geqslant 2$ kΩ V_{out} = ±10V
Output Voltage Swing	±12 ±10	±14 ±13		V V	$R_L \geqslant 10$ kΩ $R_L \geqslant 2$ kΩ
Input Voltage Range	±12	±14		V	
Common Mode Rejection Ratio	70	100		dB	$R_S \leqslant 10$ kΩ
Supply Voltage Rejection Ratio		10	150	μV/V	$R_S \leqslant 10$ kΩ
Power Consumption		40	120	mW	
Transient Response (unity gain) Risetime		0.15		μs	V_{in} = 20 mV R_L = 20 kΩ $C_L \leqslant 100$ pF
Transient Response (unity gain) Overshoot		10		%	V_{in} = 20 mV R_L = 2 kΩ $C_L \leqslant 100$ pF
Slew Rate (unity gain)		1.0		V/μs	$R_L \geqslant 2$ kΩ
Broadband Noise Voltage		2.5		μV_{RMS}	B_W = 10 Hz-30 KHz R_S = 1 kΩ
Channel Separation		125		dB	f = 1.0 kHz A_V = 40 dB R_S = 1 kΩ
The following specifications apply for 0°C ≤ T_A ≤ 75°C unless otherwise specified					
Input Offset Voltage		3.0	7.5	mV	$R_S \leqslant 10$ kΩ
Input Offset Current		7.0	300	nA	
Input Bias Current		50	800	nA	
Large-Signal Voltage Gain	15,000	200,000			$R_L \geqslant 2$ kΩ V_{out} = ±10V
Output Voltage Swing	±10	±13		V	$R_L \geqslant 2$ kΩ
Power Consumption					V_S = ±15V
		100	150	mW	T_A = 70°C
		110	200	mW	T_A = 0°C

Notes:

1. Rating applies for ambient temperatures below +75°C
2. For supply voltages less than 15V, the absolute maximum input voltage is equal to the supply voltage.
3. Short-circuit may be to ground, typically 45 mA. Rating applies to +125°C ambient temperature.

XR-5532/5532A

Dual Low-Noise Operational Amplifier

– ADVANCE INFORMATION –

GENERAL DESCRIPTION

The XR-5532 dual low-noise operational amplifier is especially designed for applications in high quality professional audio equipment. The low-noise, wide bandwidth and output drive capability make it ideally suited for instrumentation and control circuits as well as active filter design.

The XR-5532A is the specially screened version of the XR-5532, with guaranteed noise characteristics.

FEATURES

Direct Replacement for Signetics NE 5532
Wide Small-Signal Bandwidth: 10 MHz
High-Current Drive Capability
(10V rms into 600Ω at VS = ±18V)
High Slew Rate: 9 V/μs
Wide Power-Bandwidth: 140 kHz
Very Low Input Noise: 5 nV/$\sqrt{Hz}$
Wide Supply Range: ±3 V to ±20V

APPLICATIONS

High Quality Audio Amplification
Telephone Channel Amplifier
Servo Control Systems
Low-Level Signal Detection
Active Filter Design

ABSOLUTE MAXIMUM RATINGS

Power Supply	±22V
Input Common-Mode Voltage	$+V_{CC}$ to $-V_{EE}$
Differential Input Voltage (Note 1)	±0.5V
Power Dissipation (Package Limitation)	
Ceramic Package 8-Pin	600mW
Derate Above $T_A = 25°C$	8mW/°C
Storage Temperature	−60°C to +150°C

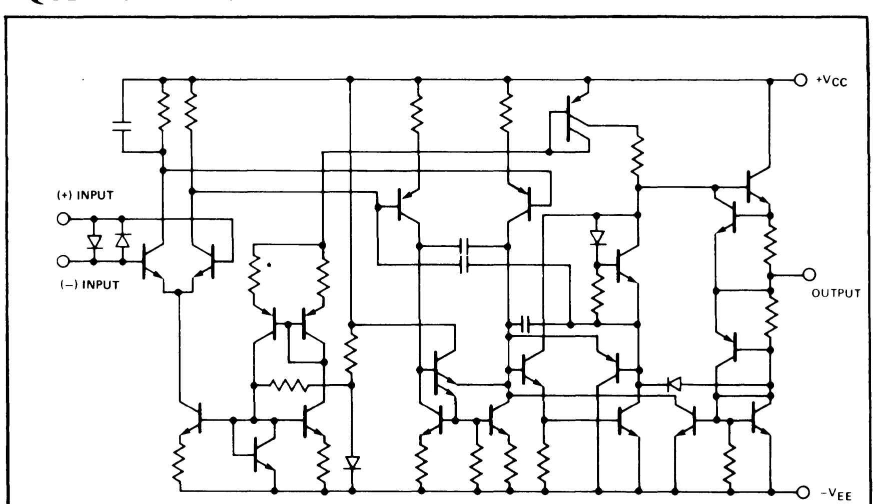
Note 1: Diodes protect the inputs against over-voltage. Therefore, unless current-limiting resistors are used, large currents will flow if the differential input voltage exceeds 0.6V. Maximum current should be limited to ±10 mA.

Note 2: Output may be shorted to ground at $V_{CC} = V_{EE} = 15$ V, $T_A = 25°C$. Temperature and/or voltages must be limited to ensure dissipation rating is not exceeded.

AVAILABLE TYPES

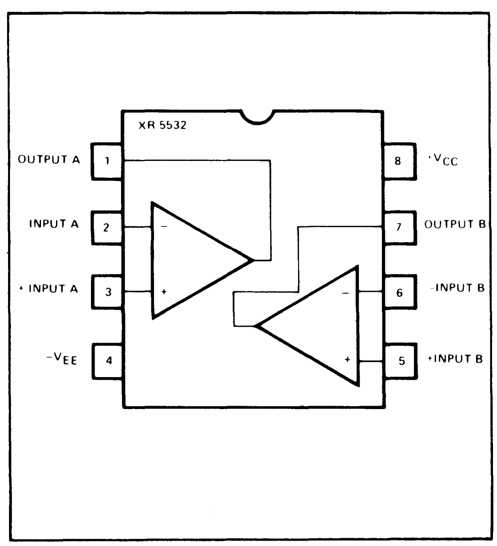
Part Number	Package	Operating Temperature
XR-5532AN	Ceramic	0°C to +75°C
XR-5532N	Ceramic	0°C to +75°C

EQUIVALENT SCHEMATIC



1/2 of XR-5532

FUNCTIONAL BLOCK DIAGRAM



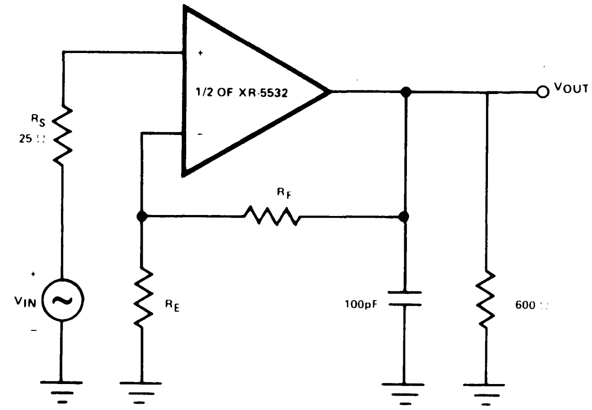
ELECTRICAL CHARACTERISTICS

$T_A = 25°C$, $V_{CC} = V_{EE} = 15V$ unless otherwise specified.

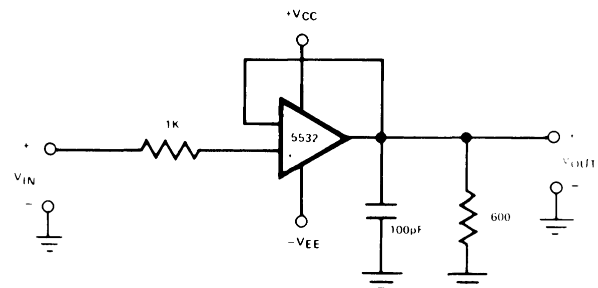
CHARACTERISTICS	XR-5532A MIN.	XR-5532A TYP.	XR-5532A MAX.	XR-5532 MIN.	XR-5532 TYP.	XR-5532 MAX.	UNITS	SYMBOL	CONDITIONS
DC CHARACTERISTICS									
Input Offset Voltage		0.5	4		0.5	4	mV	V_{OS}	$T_A = 25°C$
			5			5	mV		T_A = Full Range
Input Offset Current								I_{OS}	
		10	150		10	150	nA		$T_A = 25°C$
			200			200	nA		T_A = Full Range
Input Bias Current								I_B	
		200	800		200	800	nA		$T_A = 25°C$
			1000			1000	nA		T_A = Full Range
Large Signal Voltage Gain								A_{VOL}	$R_L \geq 600\Omega$, $V_O = \pm10V$
	25	100		25	100		V/mV		$T_A = 25°C$
	15			15			V/mV		T_A = Full Range
Supply Current		8	16		8	16	mA	I_{CC}	R_L = Open
Output Swing								V_{OUT}	$R_L \geq 600\Omega$
	±12	±13		±12	±13		V		$V_{CC} = V_{EE} = 15V$
	±15	±16		±15	±16		V		$V_{CC} = V_{EE} = 18V$
Output Short Circuit Current		38			38		mA	I_{SC}	(Note 2)
Input Resistance	30	300		30	300		kΩ	R_{IN}	
Common-Mode Range	±12	±13		±12	±13		V	V_{ICM}	
Common-Mode Rejection	70	100		70	100		dB	CMRR	
Power Supply Rejection		10	100		10	100	μV/V	PSRR	
Channel Separation		110			110			dB	f = 1 kHz, R_S = 5 KΩ
AC CHARACTERISTICS									
Transient Response									Voltage Follower
Rise Time		20			20		nsec	t_r	$R_L = 600\Omega$
Overshoot		10			10		%	t_O	V_{IN} 100 MV_{pp}, C_L = 100 pF
AC Gain									f = 10 kHz
		2.2			2.2		V/mV		
Unity-Gain Bandwidth		10			10		MHz	BW	C_L = 100 pF
Slew Rate		9			9		V/μsec		
Power Bandwidth		140			140		kHz	f_p	$V_{OUT} = \pm10V$ RL = 600Ω
Output Resistance		.3			.3		Ω	R_{OUT}	A_V = 30 dB Closed loop f = 10 kHz R_L = 600 Ω
NOISE CHARACTERISTICS									
Input Noise Voltage								e_n	
		8	10		8		nV/√Hz		f_O = 30 Hz
		5	6		5		nV/√Hz		f_O = 1 kHz
Input Noise Current								i_n	
		2.7			2.7		pA/√Hz		f_O = 30 Hz
		.7			.7		pA/√Hz		f_O = 1 kHz

TEST CIRCUITS

CLOSED LOOP FREQUENCY RESPONSE

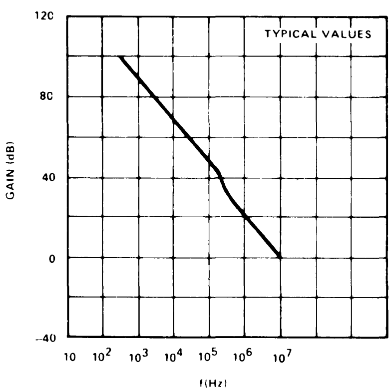


VOLTAGE FOLLOWER

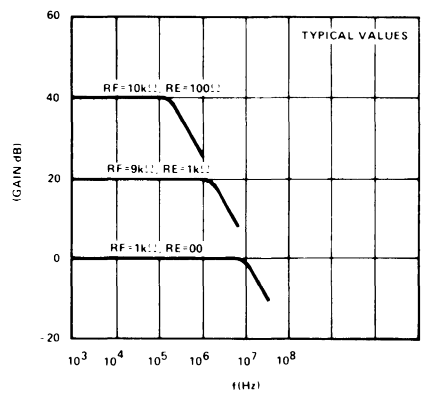


TYPICAL PERFORMANCE CHARACTERISTICS

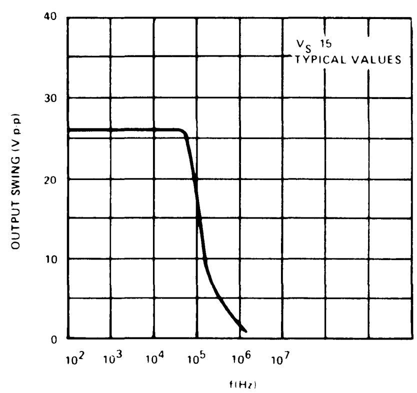
OPEN LOOP FREQUENCY RESPONSE



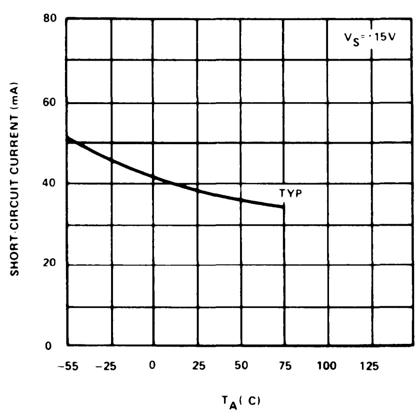
CLOSED LOOP FREQUENCY RESPONSE



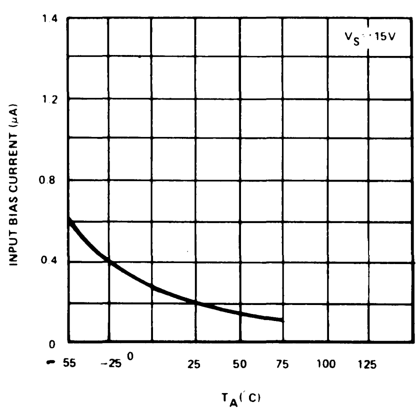
LARGE-SIGNAL FREQUENCY RESPONSE



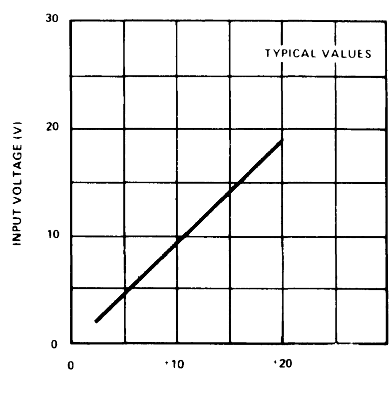
OUTPUT SHORT-CIRCUIT CURRENT



INPUT BIAS CURRENT

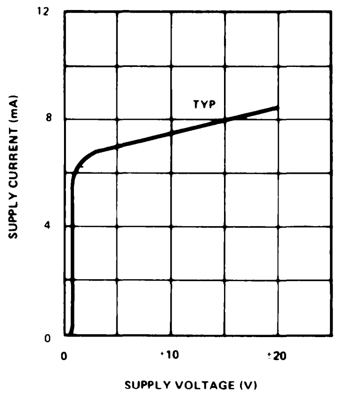


INPUT COMMON MODE VOLTAGE RANGE

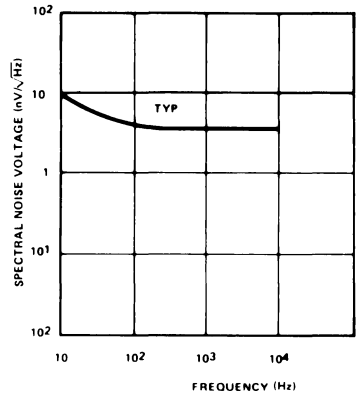


TYPICAL PERFORMANCE CHARACTERISTICS (Continued)

SUPPLY CURRENT

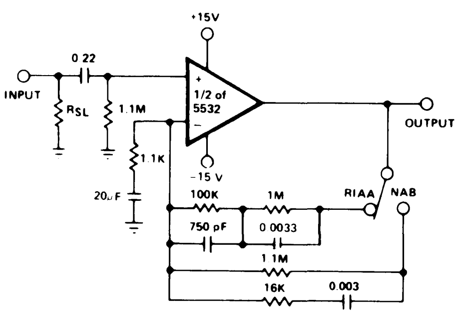


INPUT NOISE VOLTAGE DENSITY

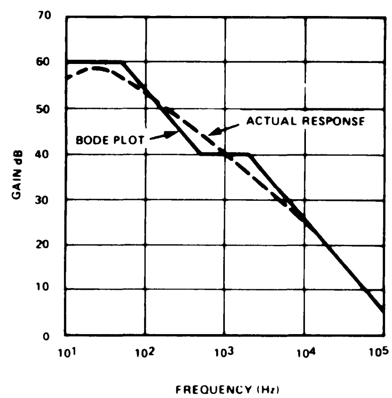


TYPICAL APPLICATION

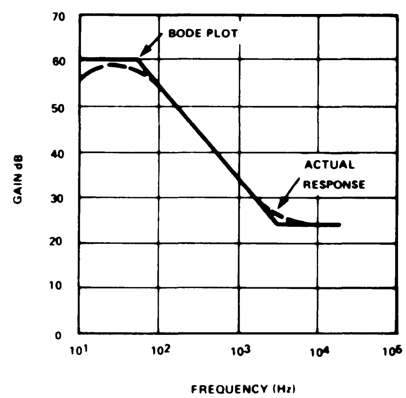
PREAMPLIFIER–RIAA/NAB COMPENSATION



*SELECT TO PROVIDE SPECIFIED TRANSDUCER LOADING
OUTPUT NOISE 0.8 mV rms (WITH INPUT SHORTED)
ALL RESISTOR VALUES ARE IN OHMS



BODE PLOT OF RIAA EQUALIZATION AND THE RESPONSE REALIZED IN AN ACTUAL CIRCUIT USING THE XR 5533



BODE PLOT OF NAB EQUALIZATION AND THE RESPONSE REALIZED IN THE ACTUAL CIRCUIT USING THE XR-5533.

XR-5533/5533A

Dual Low-Noise Operational Amplifier

– ADVANCE INFORMATION –

GENERAL DESCRIPTION

The XR-5533 dual low-noise operational amplifier is especially designed for applications in high quality professional audio equipment. The low-noise, wide bandwidth and output drive capability make it ideally suited for instrumentation and control circuits as well as active filter design.

The XR-5533A is the specially screened version of the XR-5533 with guaranteed worst-case noise specifications.

FEATURES

Direct Replacement for Signetics SE/NE 5533
Wide Small-Signal Bandwidth: 10 MHz
High-Current Drive Capability
  (10V rms into 600Ω at V_S = ±18V)
High Slew Rate: 13 V/μs
Wide Power-Bandwidth: 200 kHz
Very Low Input Noise: 4 nV/$\sqrt{Hz}$

APPLICATIONS

High Quality Audio Amplification
Telephone Channel Amplifier
Servo Control Systems
Low-Level Signal Detection
Active Filter Design

ABSOLUTE MAXIMUM RATINGS

Power Supply	±22V
Input Common-Mode Range	$-V_{EE}$ to $+V_{CC}$
Differential Input Voltage (Note 1)	±0.5V
Short Circuit Duration (Note 2)	Indefinite
Power Dissipation (Package Limitation)	
Ceramic Package 14-Pin	750 mW
Plastic Package 14-Pin	600 mW
Derate Above T_A = 25°C	5 mW/°C
Storage Temperature	–60°C to + 150°C

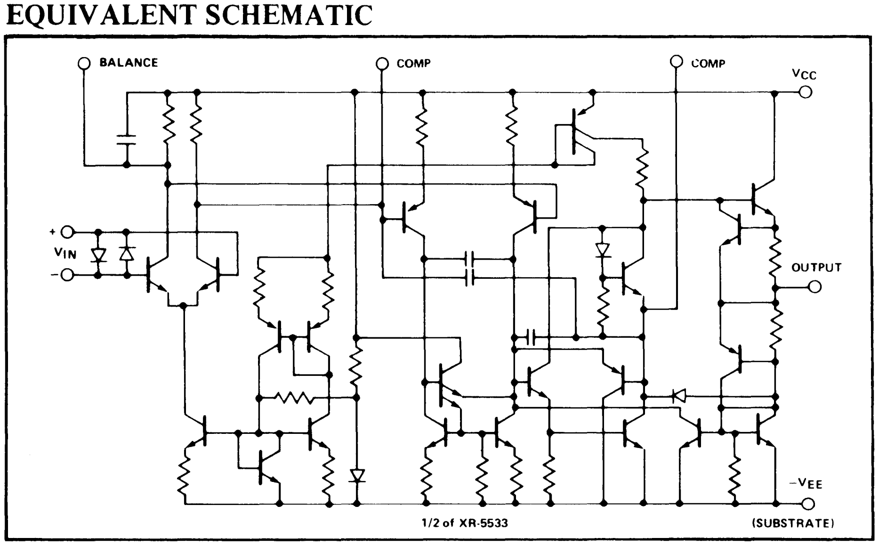
Note 1: Diodes protect the inputs against over-voltage. Therefore, unless current-limiting resistors are used, large currents will flow if the differential input voltage exceeds 0.6V. Maximum current should be limited to ±10 mA.

Note 2: Output may be shorted to ground at V_{CC} = V_{EE} = 15V, T_A = 25°C. Temperature and/or supply voltages must be limited to ensure dissipation rating is not exceeded.

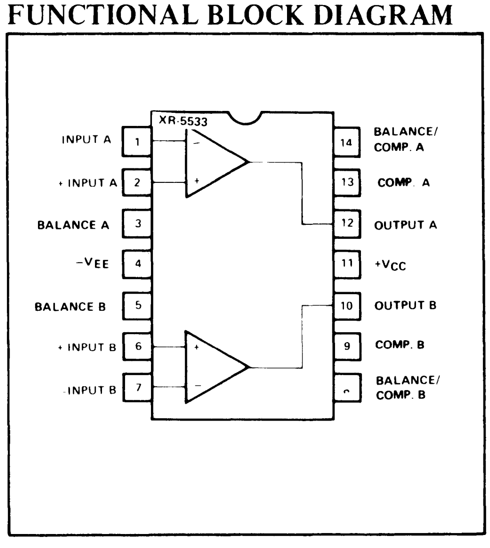
AVAILABLE TYPES

Part Number	Package	Operating Temperature
XR-5533AN	Ceramic	0°C to +75°C
XR-5533AP	Plastic	0°C to +75°C
XR-5533N	Ceramic	0°C to +75°C
XR-5533P	Plastic	0°C to +75°C

EQUIVALENT SCHEMATIC



FUNCTIONAL BLOCK DIAGRAM



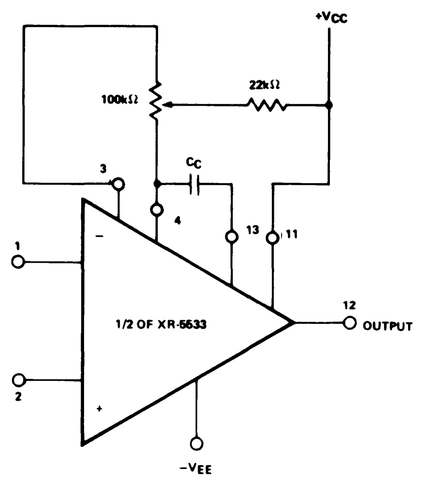
ELECTRICAL CHARACTERISTICS

$T_A = 25°C, V_{CC} = V_{EE} = 15V$ unless otherwise specified.

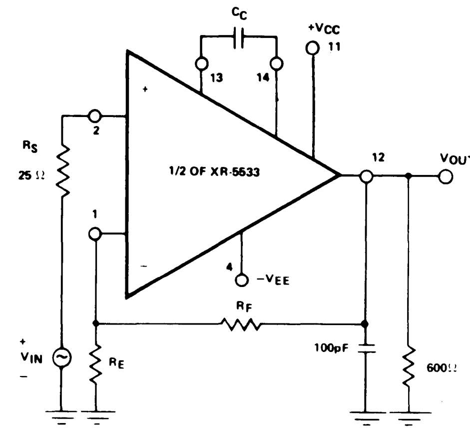
CHARACTERISTICS	XR-5533A			XR-5533			UNITS	SYMBOL	CONDITIONS
	MIN.	TYP.	MAX.	MIN.	TYP.	MAX.			
DC CHARACTERISTICS									
Input Offset Voltage								V_{OS}	
		0.5	4		0.5	4	mV		$T_A = 25°C$
			5			5	mV		T_A = Full Range
Input Offset Current								I_{OS}	
		20	300		20	300	nA		$T_A = 25°C$
			400			400	nA		T_A = Full Range
Input Bias Current								I_B	
		500	1500		500	1500	nA		$T_A = 25°C$
			2000			2000	nA		T_A = Full Range
Large Signal Voltage Gain								A_{VOL}	$R_L \geq 600\Omega, V_O = \pm 10V$
	25	100		25	100		V/mV		$T_A = 25°C$
	15			15			V/mV		T_A = Full Range
Supply Current (Each Amplifier)		4	8		4	8	mA	I_{CC}	R_L = Open
Output Swing								V_{OUT}	$R_L \geq 600\Omega$
	±12	±13		±12	±13		V		$V_{CC} = V_{EE} = 15V$
	±15	±16		±15	±16		V		$V_{CC} = V_{EE} = 18V$
Output Short Circuit Current		38			38		mA	I_{SC}	(Note 2)
Input Resistance	30	100		30	100		kΩ	R_{IN}	
Common-Mode Range	±12	±13		±12	±13		V	V_{iCM}	
Common-Mode Rejection	70	100		70	100		dB	CMRR	
Power Supply Rejection		10	100		10	100	μV/V	PSRR	
Channel Separation		110			110			dB	$f = 1$ kHz, $R_S = 5$ kΩ
AC CHARACTERISTICS									
Transient Response									Voltage Follower
Rise Time		20			20		nsec	t_r	$R_L = 600\Omega, C_C = 22$ pF
Overshoot		20			20		%	t_o	$C_L = 100$ pF $V_{IN} = 50$ mV
AC Gain									f = 10 kHz
		6			6		V/mV		$C_C = 0$
		2.2			2.2		V/mV		$C_C = 22$ pF
Unity-Gain Bandwidth		10			10		MHz	BW	$C_C = 22$ pF, $C_L = 100$ pF
Slew Rate		13			13		V/μsec		$C_C = 0$
		6			6		V/μsec		$C_C = 22$ pF
Power Bandwidth		95			95		kHz	f_p	$V_{OUT} = \pm 10V, C_C = 22$ pF
		200			200		kHz		$C_C = 0$ pF
NOISE CHARACTERISTICS									
Input Noise Voltage								e_n	
		5.5	7		7		nV/√Hz		$f_0 = 30$ Hz
		3.5	4.5		4		nV/√Hz		$f_0 = 1$ kHz
Input Noise Current								i_n	
		1.5			2.5		pA/√Hz		$f_0 = 30$ Hz
		0.4			0.6		pA/√Hz		$f_0 = 1$ kHz
Broadband Noise Figure		0.9			0.9		dB	F_N	$R_S = 5$ kΩ, f = 10 Hz to 20 kHz

TEST CIRCUITS

FREQUENCY COMPENSATION AND OFFSET VOLTAGE ADJUSTMENT CIRCUIT

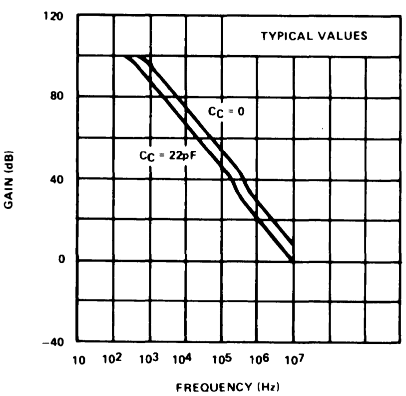


CLOSED LOOP FREQUENCY RESPONSE

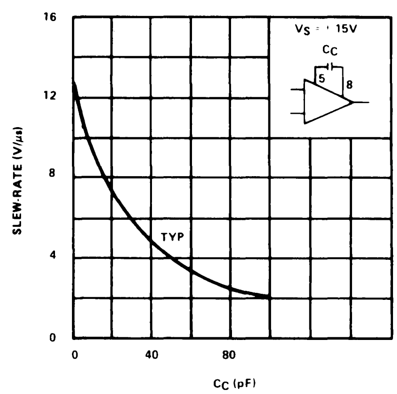


TYPICAL PERFORMANCE CHARACTERISTICS

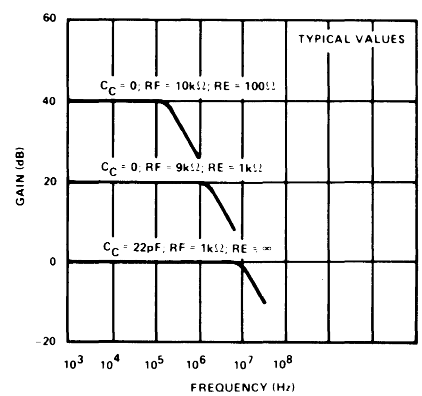
OPEN LOOP FREQUENCY RESPONSE



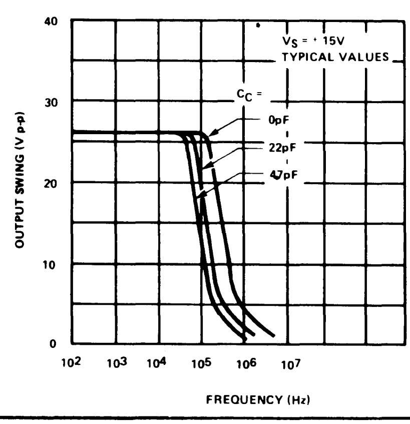
SLEW-RATE AS A FUNCTION OF COMPENSATION CAPACITANCE



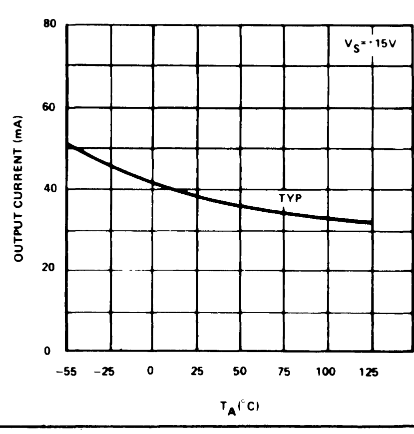
CLOSED LOOP FREQUENCY RESPONSE



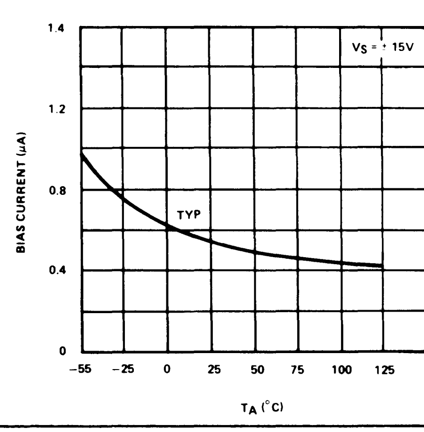
LARGE-SIGNAL FREQUENCY RESPONSE



OUTPUT SHORT-CIRCUIT CURRENT

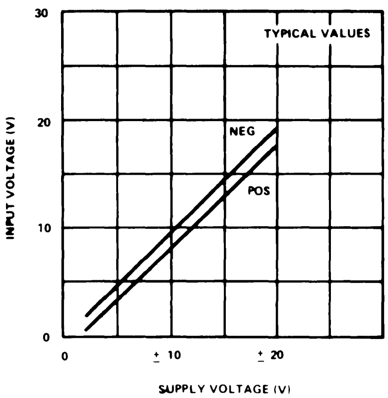


INPUT BIAS CURRENT

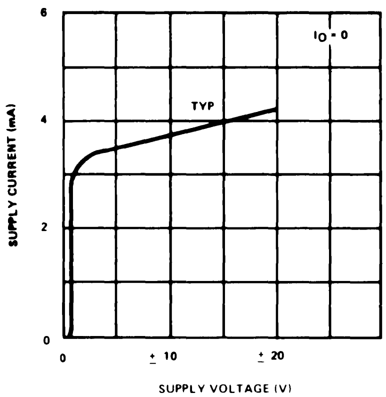


TYPICAL PERFORMANCE CHARACTERISTICS (Continued)

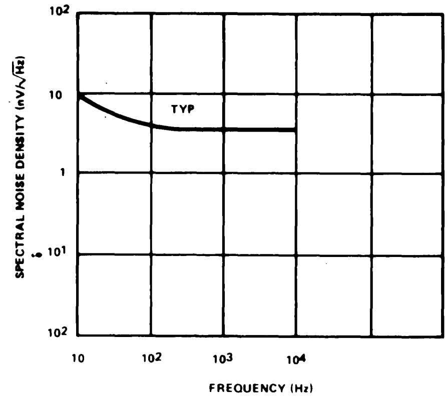
INPUT COMMON MODE VOLTAGE RANGE



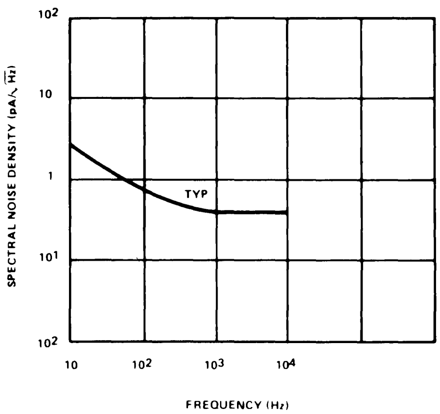
SUPPLY CURRENT PER OP AMP



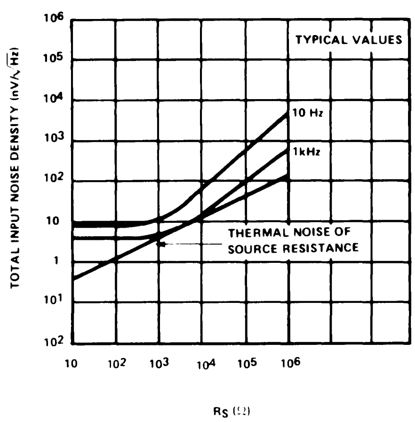
INPUT NOISE VOLTAGE DENSITY



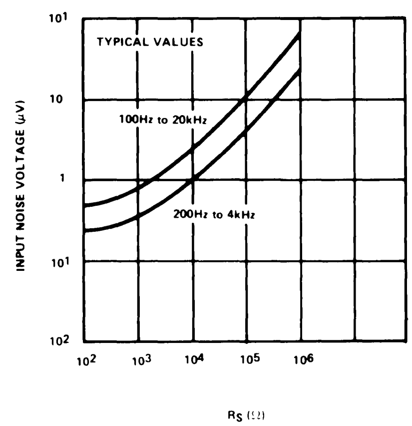
INPUT NOISE CURRENT DENSITY



TOTAL INPUT NOISE DENSITY

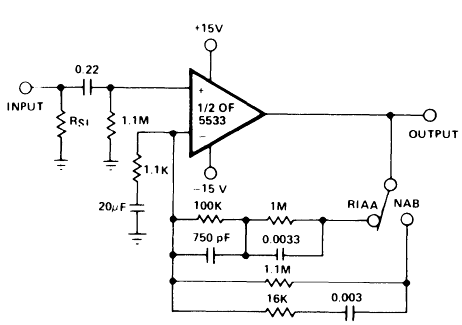


BROADBAND INPUT NOISE VOLTAGE

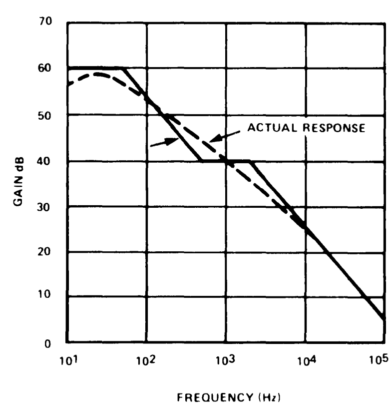


TYPICAL APPLICATION

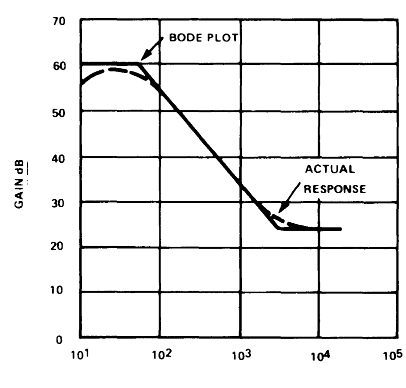
PREAMPLIFIER–RIAA/NAB COMPENSATION



*SELECT TO PROVIDE SPECIFIED TRANSDUCER LOADING
OUTPUT NOISE ≈ 0.8 mV rms (WITH INPUT SHORTED)
ALL RESISTOR VALUES ARE IN OHMS.



BODE PLOT OF RIAA EQUALIZATION AND THE RESPONSE REALIZED IN AN ACTUAL CIRCUIT USING THE XR-5533.



BODE PLOT OF NAB EQUALIZATION AND THE RESPONSE REALIZED IN THE ACTUAL CIRCUIT USING THE XR-5533

XR-5534/5534A

Low-Noise Operational Amplifier

GENERAL DESCRIPTION – ADVANCE INFORMATION –

The XR-5534 is a high performance low-noise operational amplifier especially designed for application in high quality and professional audio equipment. It offers five-fold improvement in noise characteristics, output drive capability and full-power bandwidth over conventional 741-type op-amps. The op-amp is internally compensated for gain equal to, or higher than, three. The frequency response can be optimized with an external compensation capacitor for various applications such as operating in unity-gain mode or driving capacitive loads.

The XR-5534A is a specially-screened version of the XR-5534, with guaranteed noise specifications.

FEATURES

Direct Replacement for Signetics NE/SE 5534
Wide Small-Signal Bandwidth: 10 MHz
High-Current Drive Capability
(10V rms into 600Ω at $V_S = \pm 18V$)
High Slew Rate: 13 V/μs
Wide Power-Bandwidth: 200 kHz typ.
Very Low Input Noise: 4 nV/$\sqrt{Hz}$ typ.

AVAILABLE TYPES

Part Number	Package	Operating Temperature
5534AM	Ceramic	−55°C to +125°C
5534 M	Ceramic	−55°C to +125°C
5534 ACN	Ceramic	0°C to +70°C
5534 CN	Ceramic	0°C to +70°C
5534 ACP	Plastic	0°C to +70°C
5534 CP	Plastic	0°C to +70°C

ABSOLUTE MAXIMUM RATINGS

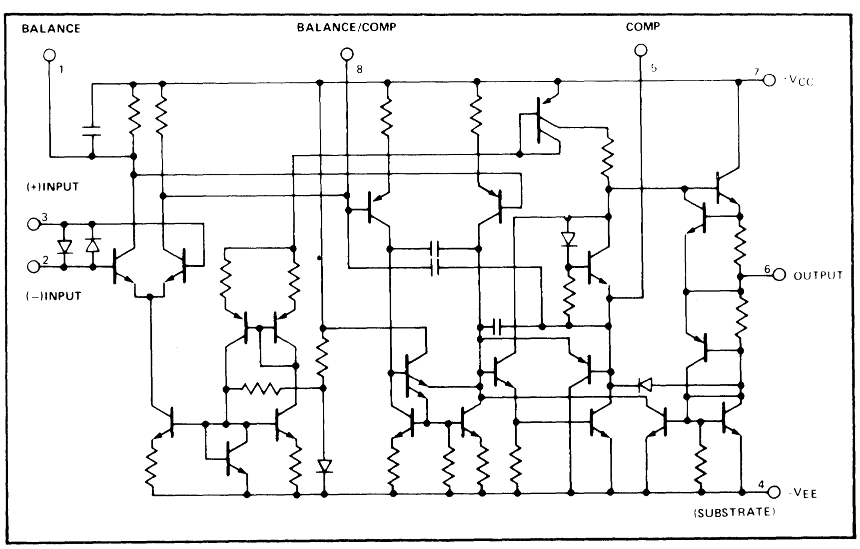
Power Supply	±22V
Input Common-Mode Voltage	$+V_{CC}$ to $-V_{EE}$
Differential Input Voltage (Note 1)	±0.5V
Power Dissipation (Package Limitation)	
Ceramic Package	385 mW
Plastic Package	300mW
Derate Above +24°C	2.5 mW/°C
Short Circuit Duration (Note 2)	Indefinite
Storage Temperature	−60°C to +150°C

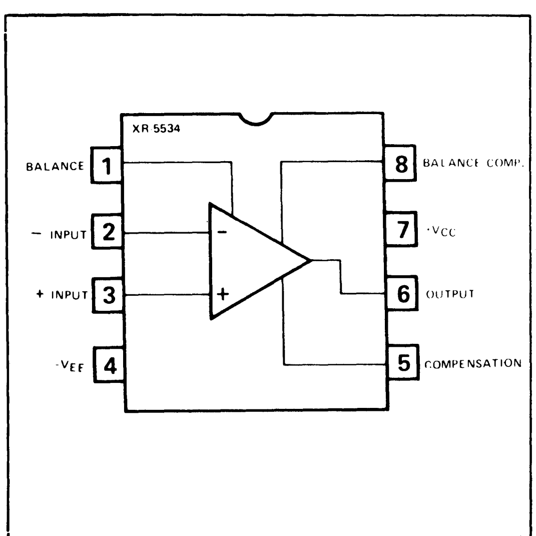
Note 1: Diodes protect the inputs against over-voltage. Therefore, unless current-limiting resistors are used, large currents will flow if the differential input voltage exceeds 0.6V. Maximum current should be limited to ±10 mA.

Note 2: Output may be shorted to ground at $V_S = \pm 15V$, $T_A = 25°C$. Temperature and/or supply voltages must be limited to ensure dissipation rating is not exceeded.

APPLICATIONS

High Quality Audio Amplification
Telephone Channel Amplifiers
Servo Control Systems
Low-Level Signal Detection
Active Filter Design





ELECTRICAL CHARACTERISTICS

$T_A = 25°C$, $V_{CC} = V_{EE} = 15V$, unless otherwise specified.

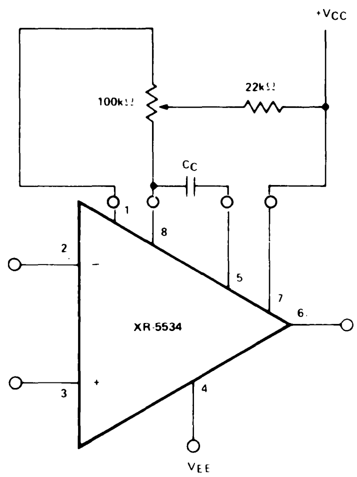
CHARACTERISTICS	XR-5534 M / 5534 AM			XR-5534AC/XR-5534C			UNITS	SYMBOL	CONDITION
	MIN	TYP	MAX	MIN	TYP	MAX			
Input Offset Voltage								V_{OS}	
		0.5	2		0.5	4	mV		$T_A = 25°C$
			3			5	mV		T_A = Full Range
Input Offset Current								I_{OS}	
		10	200		20	300	nA		$T_A = 25°C$
			500			400	nA		T_A = Full Range
Input Bias Current								I_B	
		400	800		500	1500	nA		$T_A = 25°C$
			1500			2000	nA		T_A = Full Range
Large Signal Voltage Gain								A_{VOL}	$R_L \geq 600\Omega$, $V_O = \pm 10V$
	50	100		25	100		V/mV		$T_A = 25°C$
	25			15			V/mV		T_A = Full Range
Supply Current		4	6.5		4	8	mA	I_{CC}	R_L = Open
Output Swing								V_{OUT}	$R_L \geq 600\Omega$
	±12	±13		±12	±13		V		$V_{CC} = V_{EE} = 15V$
	±15	±16		±15	±16		V		$V_{CC} = V_{EE} = 18V$
Output Short Circuit Current		38			38		mA	I_{SC}	(Note 2)
Input Resistance	50	100		30	100		KΩ	R_{in}	
Common-Mode Range	±12	±13		±12	±13		V	V_{iCM}	
Common-Mode Rejection	80	100		70	100		dB	CMRR	
Power Supply Rejection		10	50		10	100	μV/V	PSRR	
AC CHARACTERISTICS									
Transient Response									Voltage Follower
Rise Time		20			20		nSec	t_r	$R_L \geq 600\Omega$, C_C = 22 pF
Overshoot		20			20		%	t_O	C_L = 100 pF
AC Gain									f = 10 kHz
		6			6		6	V/mV	$C_C = 0$
		2.2			2.2		2.2	V/mV	C_C = 22 pF
Unity-Gain Bandwidth		10			10		MHz	BW	C_C = 22 pF, C_L = 100 pF
Slew Rate		13			13		V/μsec		$C_C = 0$
		6			6		V/μsec		C_C = 22 pF
Power Bandwidth		95			95		kHz	f_P	$V_{OUT} = \pm 10V$, C_C = 22 pF
		200			200		kHz		$C_C = 0$

NOISE CHARACTERISTICS

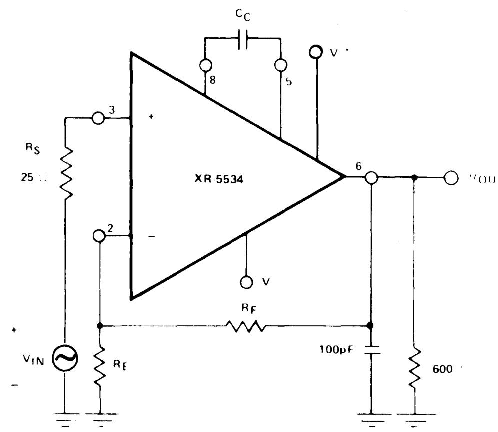
CHARACTERISTIC	XR-5534A			XR-5534			UNITS	SYMBOL	CONDITION
	MIN	TYP	MAX	MIN	TYP	MAX			
Input Noise Voltage		5.5	7		7		$nV/\sqrt{Hz}$	e_n	f_O = 30 Hz
		3.5	4.5		4		$nV/\sqrt{Hz}$		f_O = 1 kHz
Input Noise Current		1.5			2.5		$pA/\sqrt{Hz}$	i_n	f_O = 30 Hz
		0.4			0.6		$pA/\sqrt{Hz}$		f_O = 1 kHz
Broadband Noise Figure		0.9					dB	F_N	R_S = 5 kΩ
									f = 10 Hz to 20 kHz

TEST CIRCUITS

FREQUENCY COMPENSATION AND OFFSET VOLTAGE ADJUSTMENT CIRCUIT

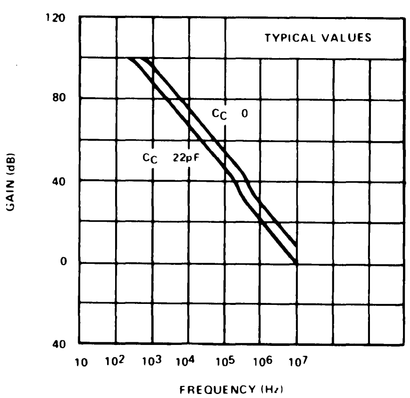


CLOSED LOOP FREQUENCY RESPONSE

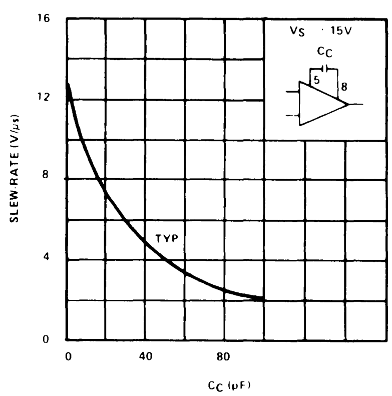


TYPICAL PERFORMANCE CHARACTERISTICS

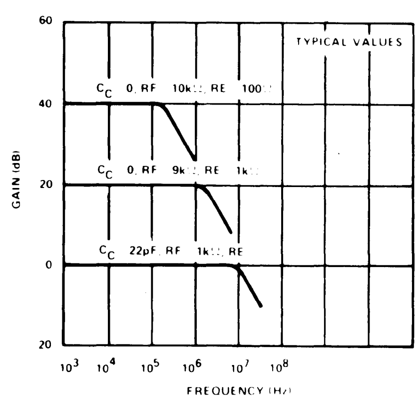
OPEN LOOP FREQUENCY RESPONSE



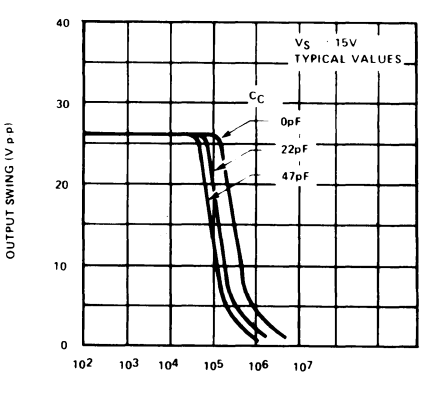
SLEW-RATE AS A FUNCTION OF COMPENSATION CAPACITANCE



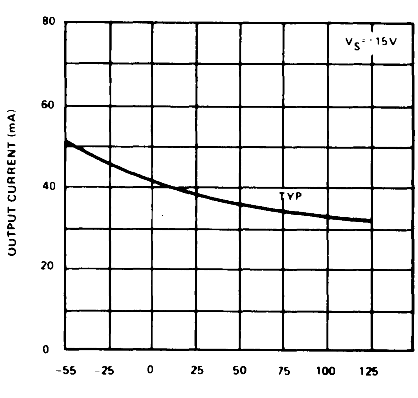
CLOSED LOOP FREQUENCY RESPONSE



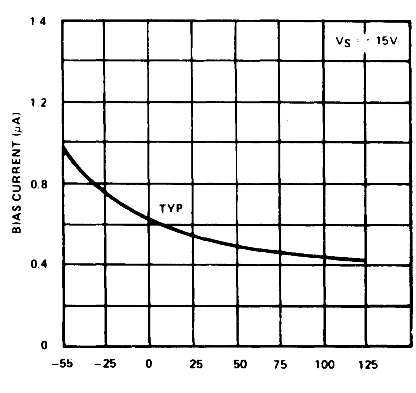
LARGE-SIGNAL FREQUENCY RESPONSE



OUTPUT SHORT-CIRCUIT CURRENT

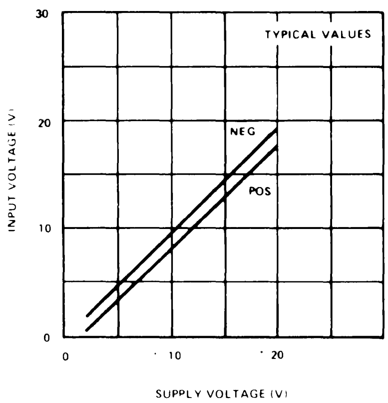


INPUT BIAS CURRENT

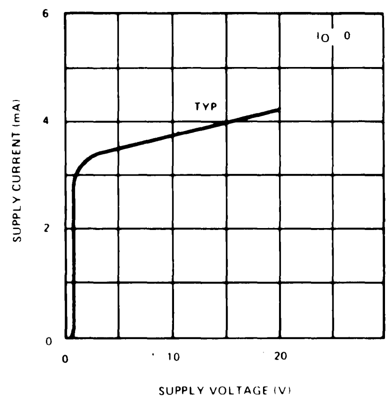


TYPICAL PERFORMANCE CHARACTERISTICS (Continued)

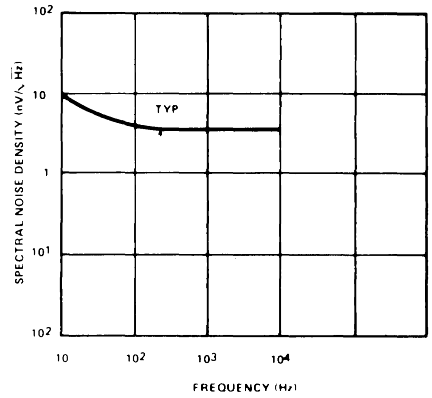
INPUT COMMON MODE VOLTAGE RANGE



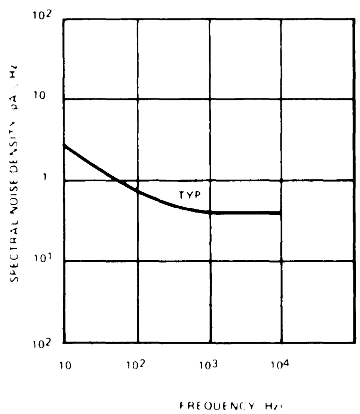
SUPPLY CURRENT



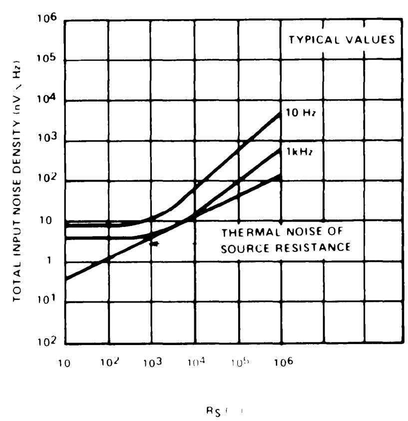
INPUT NOISE VOLTAGE DENSITY



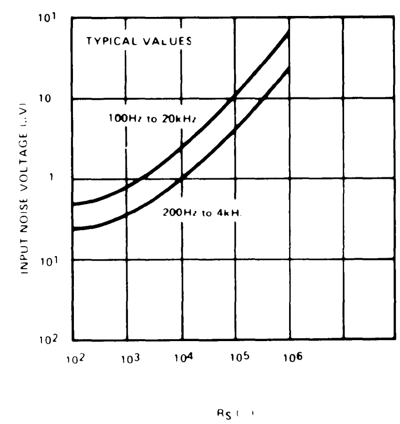
INPUT NOISE CURRENT DENSITY



TOTAL INPUT NOISE DENSITY

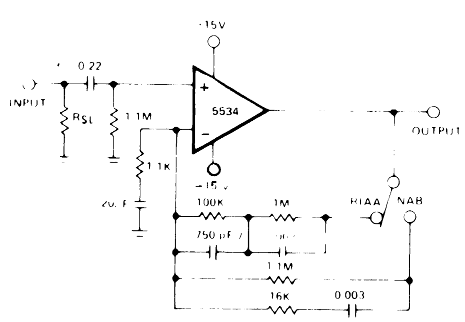


BROADBAND INPUT NOISE VOLTAGE

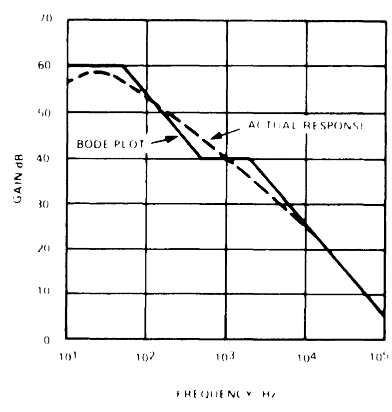


TYPICAL APPLICATION

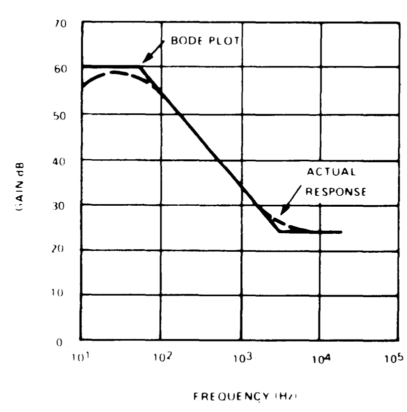
PREAMPLIFIER–RIAA/NAB COMPENSATION



*SELECT TO PROVIDE SPECIFIED TRANSDUCER LOADING
OUTPUT NOISE 0.8 mV rms WITH INPUT SHORTED
ALL RESISTOR VALUES ARE IN OHMS



BODE PLOT OF RIAA EQUALIZATION AND THE RESPONSE REALIZED IN AN ACTUAL CIRCUIT USING THE XR-5534



BODE PLOT OF NAB EQUALIZATION AND THE RESPONSE REALIZED IN THE ACTUAL CIRCUIT USING THE XR-5534

XR-13600

Dual Operational Transconductance Amplifier

GENERAL DESCRIPTION

The XR-13600 consists of 2 programmable transconductance amplifiers with high input impedance and push-pull outputs. The 2 amplifiers share common supplies but otherwise operate independently. Each amplifier's transconductance is directly proportional to its applied bias current. To improve signal-to-noise performance, predistortion diodes are included on the inputs; the use of these diodes results in a 10 dB improvement referenced to 0.5% THD. Independent Darlington emitter followers are included to buffer the outputs.

FEATURES

Direct Replacement for LM-13600 and LM-13600A
Transconductance Adjustable Over 4 Decades
Excellent Transconductance-Control Linearity
Uncommitted Darlington Output Buffers
On-Chip Predistortion Diodes
Excellent Matching Between Amplifiers
Wide Supply Range: ±2V to ±18V

APPLICATIONS

Current-Controlled Amplifiers
Current-Controlled Impedances
Current-Controlled Filters
Current-Controlled Oscillators
Multipliers/Attenuators
Sample and Hold Circuits
Electronic Music Synthesis

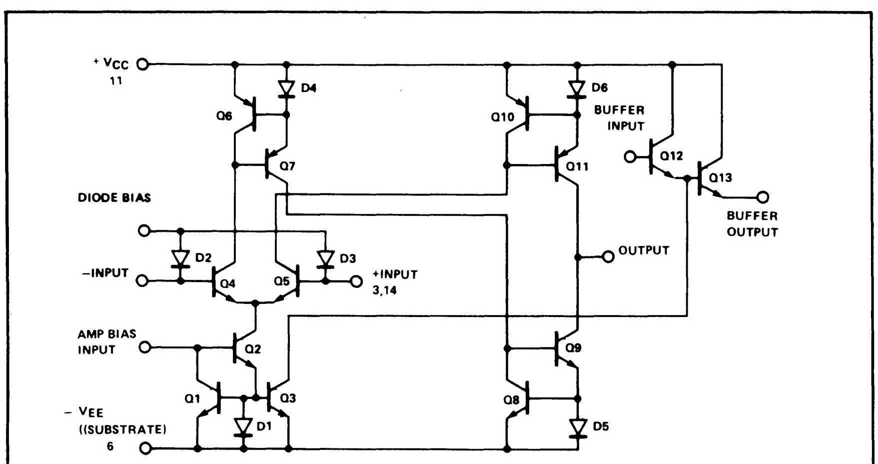
ABSOLUTE MAXIMUM RATINGS

Supply Voltage (See Note 1)	±22 V
Power Dissipation ($T_A = 25°C$, see Note 2)	625 mW
Derate Above 25°C	5 mW/°C
DC Input Voltage	$+V_{CC}$ to $-V_{EE}$
Differential Input Voltage	±5 V
Diode Bias Current (I_D)	2 mA
Amplifier Bias Current (I_B)	2 mA
Output Short Circuit Duration	Indefinite
Buffer Output Current (Note 3)	20 mA
Storage Temperature Range	-65°C to +150°C

ORDER INFORMATION

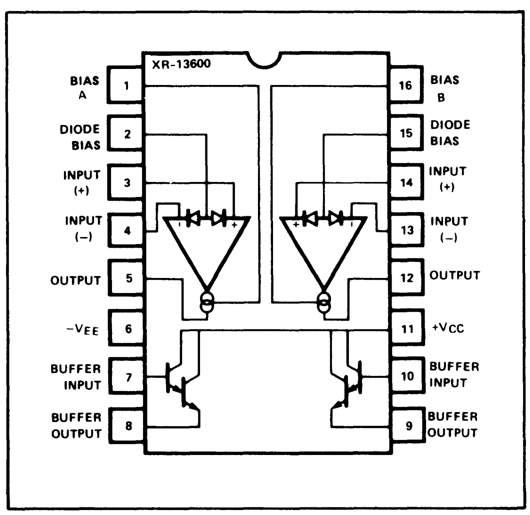
Part Number	Package	Operating Temperature
XR-13600AP	Plastic	0°C to +75°C
XR-13600CP	Plastic	0°C to +75°C

EQUIVALENT SCHEMATIC DIAGRAM



(One Channel Only)

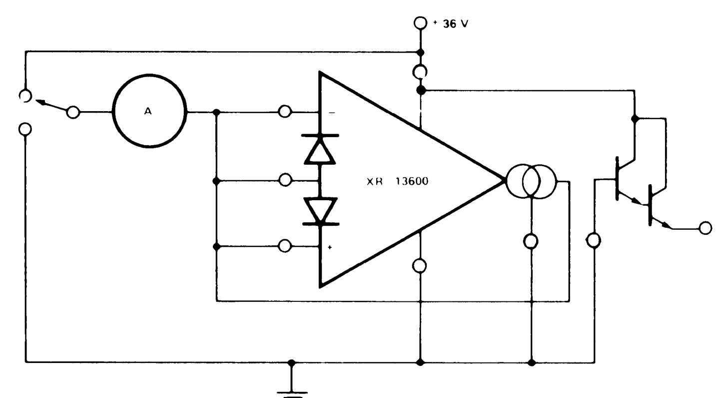
FUNCTIONAL BLOCK DIAGRAM



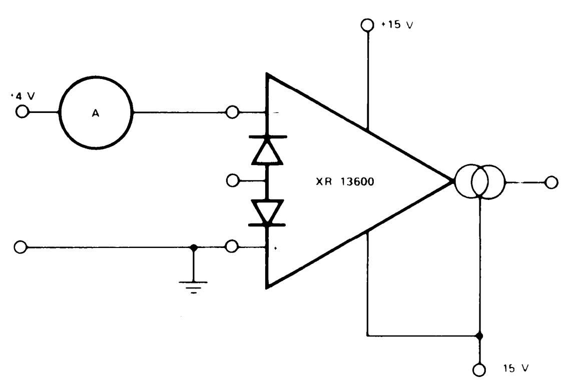
ELECTRICAL CHARACTERISTICS T_A = +25° C, Supply Voltage = ±15 V unless otherwise specified

CHARACTERISTICS	XR-13600A MIN.	XR-13600A TYP.	XR-13600A MAX.	XR-13600C MIN.	XR-13600C TYP.	XR-13600C MAX.	UNITS	CONDITIONS
Input Offset Voltage (V_{OS})		0.4	2		0.4	5	mV	
			5				mV	Over Temperature Range
		0.3	2		0.3	5	mV	I_B = 5µA
V_{OS} Including Diodes		0.5	2		0.5	5	mV	Diode Bias Current (I_D) = 500 µA
Input Offset Change		0.1	3		0.1		mV	5 µA $\leq I_B \leq$ 500 µA
Input Offset Current		0.1	0.6		0.1	0.6	µA	
Input Bias Current		0.4	5		0.4	5	µA	T_A = 25°C
		1	7		1	8	µA	Over Temperature Range
Forward Transconductance (g_m)	7700	9600	12000	6700	9600	13000	µmho	T_A = 25°C
	4000			5400			µmho	Over Temperature Range
g_m Tracking		0.3			0.3		dB	
Peak Output Current	3	5	7		5		µA	RL = 0, I_B = 5 µA
	350	500	650	350	500	650	µA	RL = 0, I_B = 500 µA
	300			300			µA	RL = 0, Over Specified Temp Range
Peak Output Voltage								
Positive	+ 12	+ 14.2		+ 12	+ 14.2		V	RL = ∞, 5 µA $\leq I_B \leq$ 500 µA
Negative	- 12	- 14.4		- 12	- 14.4		V	RL = ∞, 5 µA $\leq I_B \leq$ 500 µA
Supply Current		2.6			2.6		mA	I_B = 500 µA, Both Channels
V_{OS} Sensitivity								
Positive		20	150		20	150	µV/V	$\Delta V_{OS}/\Delta V+$
Negative		20	150		20	150	µV/V	$\Delta V_{OS}/\Delta V-$
CMRR	80	110		80	110		dB	
Common Mode Range	± 12	± 13.5		± 12	± 13.5		V	
								Referred to Input (Note 5)
Channel Separation		100			100		dB	20 Hz < f < 20 KHz
Diff. Input Current		0.02	10		0.02	100	nA	I_B = 0, Input = ± 4 V
Leakage Current		0.2	5		0.2	100	nA	I_B = 0 (Refer To Test Circuit)
Input Resistance	10	26		10	26		KΩ	
Open Loop Bandwidth		2			2		MHz	
Slew Rate		50			50		V/µSec	Unity Gain Compensated
Buff. Input Current		0.4	5		0.4	5	µA	(Note 5)
Peak Buffer Output Voltage	10			10			V	(Note 5)

TEST CIRCUITS



LEAKAGE CURRENT TEST CIRCUIT



DIFFERENTIAL INPUT CURRENT TEST CIRCUIT

Note 1. For selections to a supply voltage above ±22 V, contact factory.

Note 2. For operating at high temperatures, the device may be derated based on a 150°C maximum junction temperature and a thermal resistance of 175° C/W which applies for the device soldered in a printed circuit board, operating in still air.

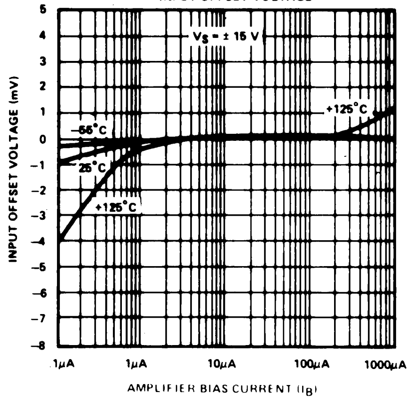
Note 3. Buffer output current should be limited so as to not exceed package dissipation.

Note 4. These specifications apply for $V_{CC} = V_{EE}$ = 15V, T_A = 25°C, amplifier bias current (I_B) = 500 µA, pins 2 and 15 open unless otherwise specified. The inputs to the buffers are grounded and outputs are open.

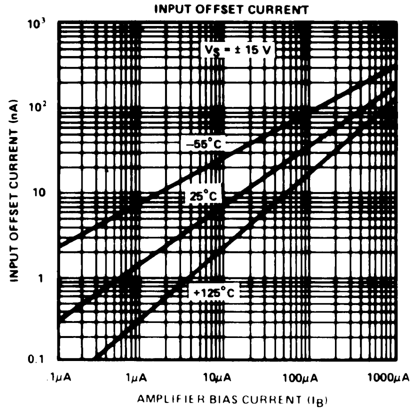
Note 5. These specifications apply for $V_{CC} = V_{EE}$ = 15 V, I_B = 500 µA, R_{OUT} = 5 kΩ connected from the buffer output to $-V_{EE}$ and the input of the buffer is connected to the transconductance amplifier output.

TYPICAL PERFORMANCE CHARACTERISTICS

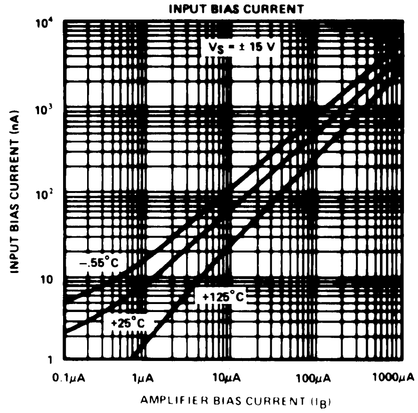
INPUT OFFSET VOLTAGE



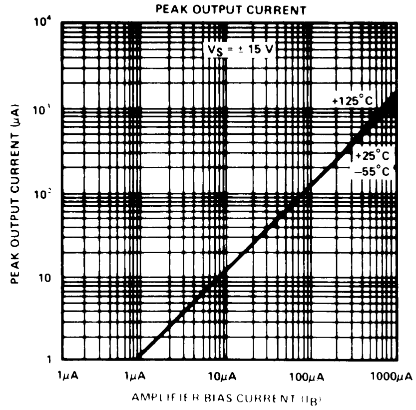
INPUT OFFSET CURRENT



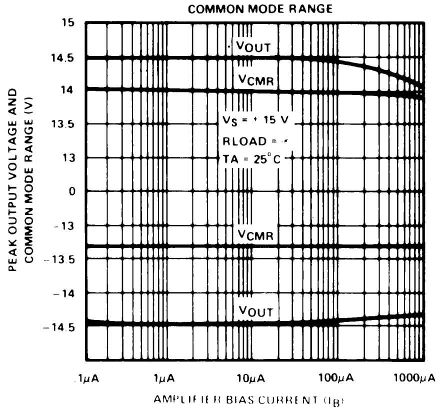
INPUT BIAS CURRENT



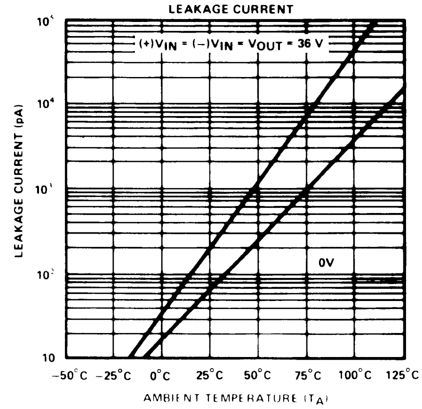
PEAK OUTPUT CURRENT



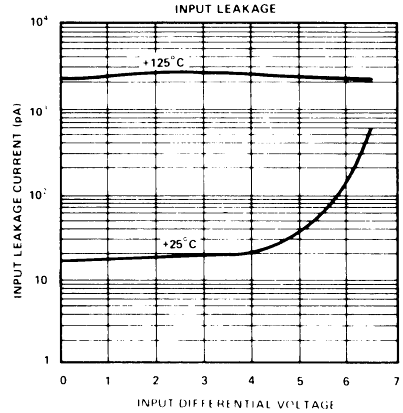
PEAK OUTPUT VOLTAGE AND COMMON MODE RANGE



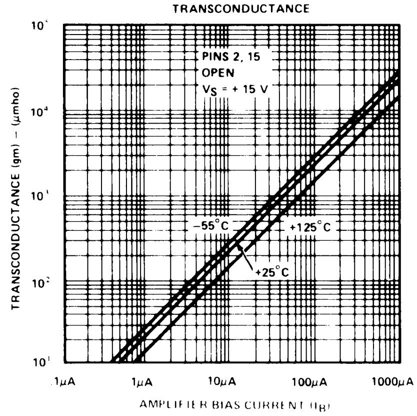
LEAKAGE CURRENT



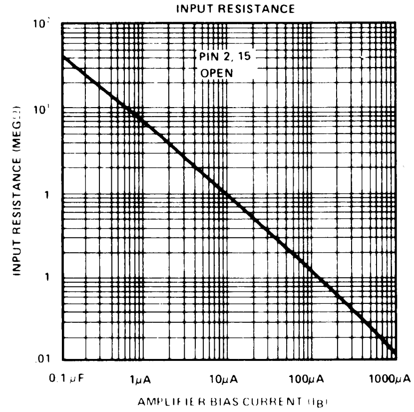
INPUT LEAKAGE



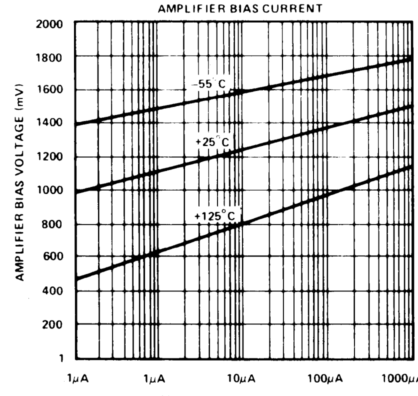
TRANSCONDUCTANCE



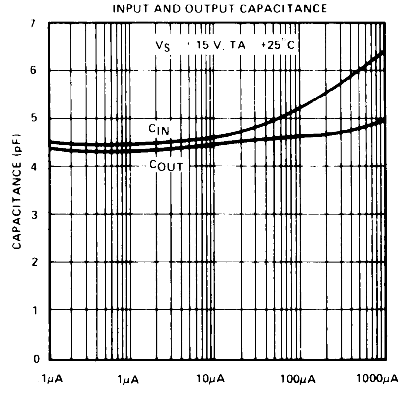
INPUT RESISTANCE



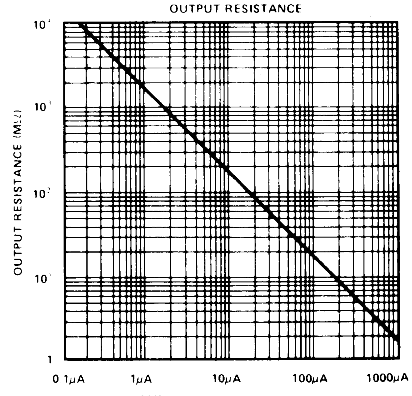
AMPLIFIER BIAS VOLTAGE VS AMPLIFIER BIAS CURRENT



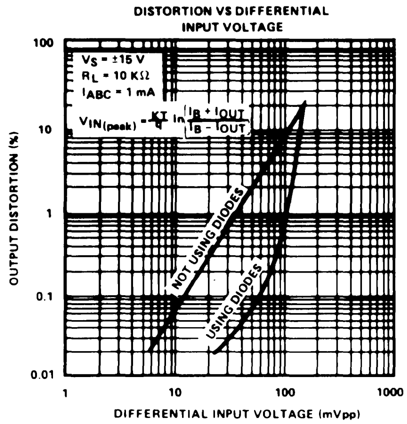
INPUT AND OUTPUT CAPACITANCE

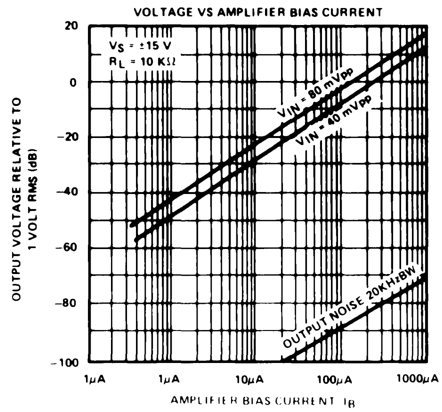


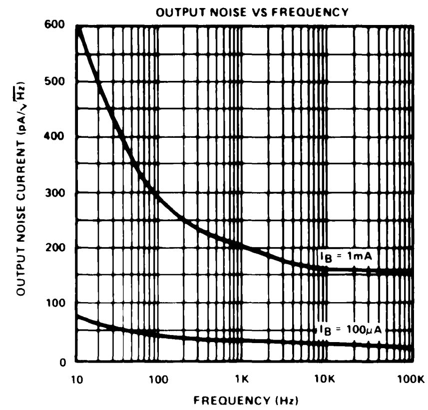
OUTPUT RESISTANCE



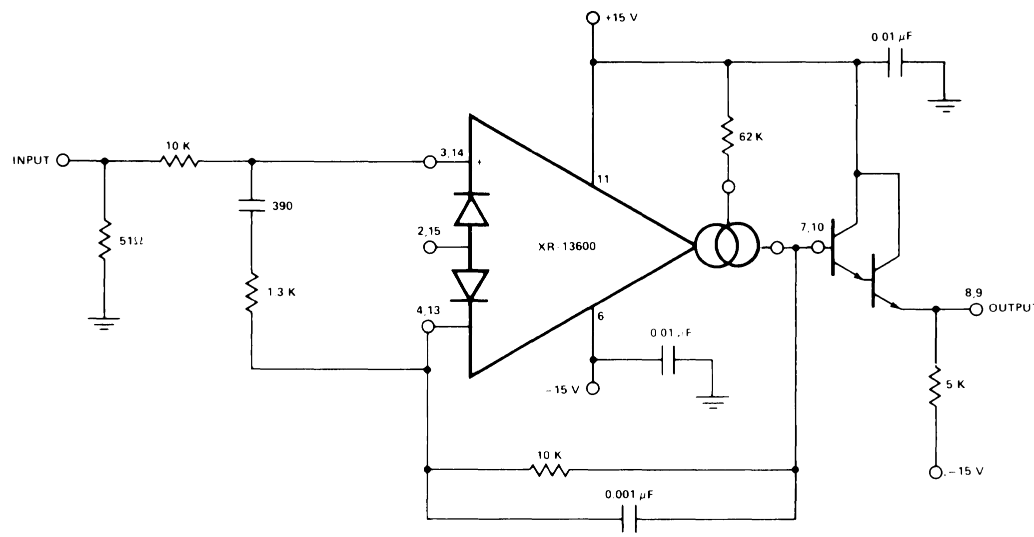
TYPICAL PERFORMANCE CHARACTERISTICS (Continued)







TYPICAL CIRCUIT CONNECTION



UNITY GAIN FOLLOWER

CIRCUIT DESCRIPTION

The differential transistor pair Q_4 and Q_5 form a transconductance stage in that the ratio of their collector currents is defined by the differential input voltage according to the transfer function:

$$V_{IN} = \frac{KT}{q} \ln \frac{I_5}{I_4} \qquad (1)$$

where V_{IN} is the differential input voltage, KT/q is approximately 26 mV at 25° C and I_5 and I_4 are the collector currents of transistors Q_5 and Q_4 respectively. With the exception of Q_3 and Q_{13}, all transistors and diodes are identical in size. Transistors Q_1 and Q_2 with Diode D_1 form a current mirror which forces the sum of currents I_4 and I_5 to equal I_B;

$$I_4 + I_5 = I_B \qquad (2)$$

where I_B is the amplifier bias current applied to the gain pin.

For small differential input voltages the ratio of I_4 and I_5 approaches unity and the Taylor series of the ln function can be approximated as:

$$\frac{KT}{q} \ln \frac{I_5}{I_4} \approx \frac{KT}{q} \frac{I_5 - I_4}{I_4} \qquad (3)$$

$$I_4 \approx I_5 \approx \frac{I_B}{2}$$

$$V_{IN}\left[\frac{(I_B)(q)}{2KT}\right] = |I_5 - I_4 \qquad (4)$$

Collector currents I_4 and I_5 are not very useful by themselves and it is necessary to subtract one current from the other. The remaining transistors and diodes form three current mirrors that produce an output current equal to I_5 minus I_4 thus:

$$V_{IN}\left[\frac{(I_B)(q)}{2KT}\right] = I_{OUT} \qquad (5)$$

The term in brackets is then the transconductance of the amplifier and is proportional to I_B.

LINEARIZING DIODES

For differential voltages greater than a few millivolts, Equation 3 is no longer accurate, and the transconductance becomes increasingly nonlinear. Figure 1 demonstrates how the internal diodes can linearize the transfer function of the amplifier. For convenience assume the diodes are biased with current sources and the input signal is the form of current I_S. Since the sum of I_4 and I_5 is I_B and the difference is I_{OUT}, currents I_4 and I_5 can be written as follows:

$$I_4 = \frac{I_B}{2} - \frac{I_{OUT}}{2} \quad , \quad I_5 = \frac{I_B}{2} + \frac{I_{OUT}}{2}$$

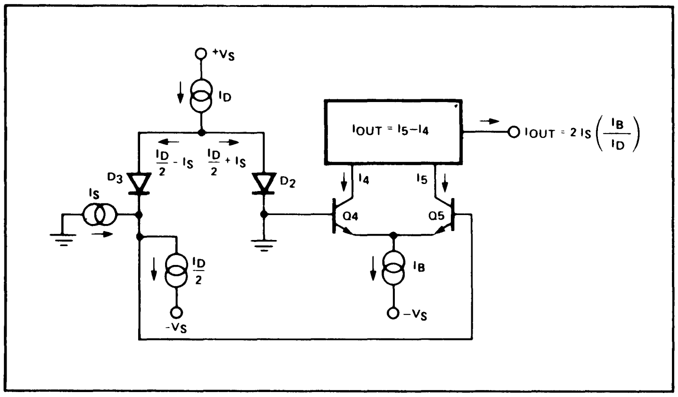


Figure 1. Linearizing Diodes

Since the diodes and the input transistors have identical geometries and are subject to similar voltages and temperatures, the following is true:

$$\frac{KT}{q} \ln \frac{\frac{I_D}{2} + I_S}{\frac{I_D}{2} - I_S} = \frac{KT}{q} \ln \frac{\frac{I_B}{2} + \frac{I_{out}}{2}}{\frac{I_B}{2} - \frac{I_{out}}{2}}$$

$$\therefore I_{out} = I_S\left(\frac{2I_B}{I_D}\right) \quad \text{for} \quad |I_S| < \frac{I_D}{2} \qquad (6)$$

Notice that in deriving Equation 6, no approximations have been made and there are no temperature dependent terms. The limitations are that the signal current not exceed $I_D/2$ and that the diodes be biased with currents. In practice, replacing the current sources with resistors will generate insignificant errors.

CONTROLLED IMPEDANCE BUFFERS

The upper limit of transconductance is defined by the maximum value of I_B (2 mA). The lowest value of I_B for which the amplifier will function therefore determines the overall dynamic range. At very low values of I_B, a buffer which has very low input bias current is desirable. A FET follower satisifies the low input current requirement, but is some what non-linear for large voltage swing. The controlled impedance buffer is a Darlington which modifies its input bias current to suit the need. For low values of I_B, the buffer's input current is minimal. At higher levels of I_B, transistor Q_3 biases up to Q_{12} with a current proportional to I_B for fast slew rate.

APPLICATIONS

VOLTAGE CONTROLLED AMPLIFIERS (VCA)

Figure 2 shows how the linearizing diodes can be used in a voltage controlled amplifier. To understand the input biasing, it is best to consider the 13 KΩ resistor as a current source and use a Therenin equivalent circuit as shown in Figure 3. This circuit is similar to Figure 1 and operates the same. The potentiometer in Figure 2 is adjusted to minimize the effects of the control signal at the output.

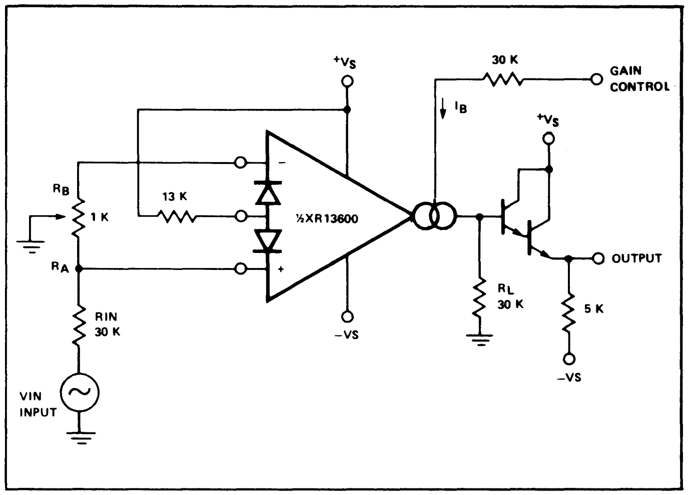


Figure 2. Voltage Controlled Amplifier (VCA) Circuit

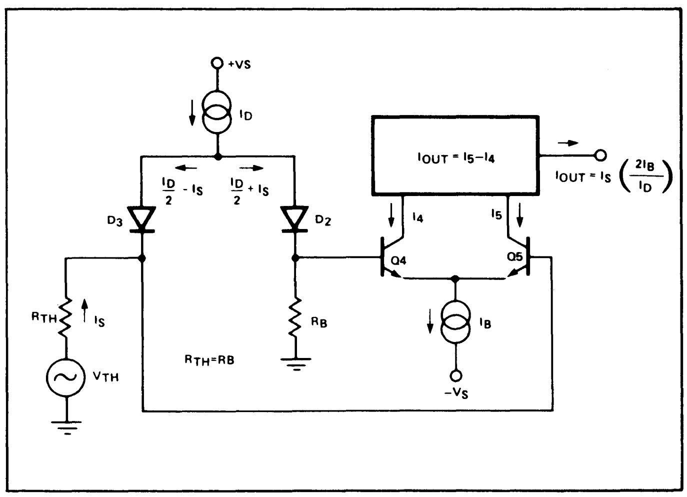


Figure 3. Equivalent VCA Input Circuit

For optimum signal-to-noise performance, I_B should be as large as possible as shown by the Output Voltage vs. Amplifier Bias Current graph. Larger amplitudes of input signal also improve the S/N ratio. The linearizing diodes help here by allowing larger input signals for the same output distortion as shown by the Distortion vs. Differential Input voltage graph. S/N may be optimized by adjusting the magnitude of the input signal via R_{IN} (Figure 2) until the output distortion is below some desired level. The output voltage swing can then be set at any level by selecting R_L.

Although the noise contribution of the linearizing diodes is negligible relative to the contribution of the amplifier's internal transistors, I_D should be as large as possible. This minimizes the dynamic junction resistance of the diodes (r_e) and maximizes their linearizing action when balanced against R_{IN}. A value of 1 mA is recommended for I_D unless the specific application demands otherwise.

STEREO VOLUME CONTROL

The circuit of Figure 4 uses the excellent matching of the two XR 13600 amplifiers to provide a Stereo Volume Control with a typical channel-to-channel gain tracking of 0.3 dB. Rp is provided to minimize the output offset voltage and may be replaced with two 510Ω resistors in AC-coupled applications. For the component values given, amplifier gain is derived from Figure 2 as being:

$$\frac{V_O}{V_{IN}} = 940 \times I_B \text{ (mA)}$$

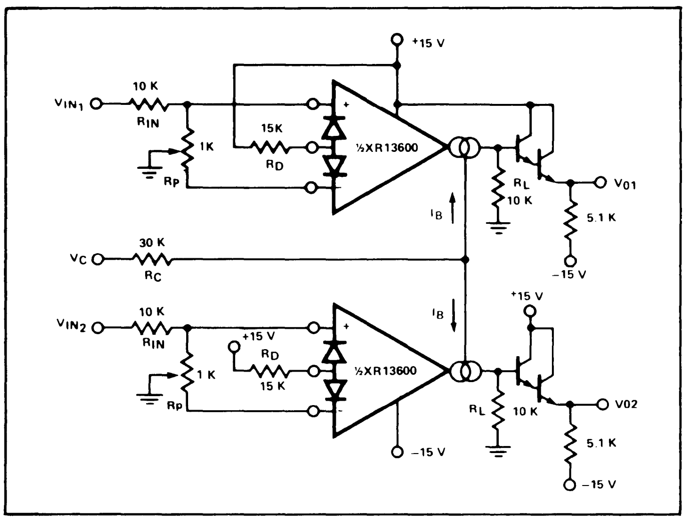


Figure 4. Stereo Volume Control

If V_C is derived from a second signal source then the circuit becomes an amplitude modulator or two-quadrant multiplier as shown in Figure 5, where:

$$I_O = \frac{-2I_S}{I_D}(I_B) = \frac{-2I_S}{I_D}\frac{V_{IN2}}{R_C} - \frac{2I_S}{I_D}\frac{(V + 1.4V)}{R_C}$$

The constant term in the above equation may be cancelled by feeding $I_S \times I_D R_C / 2(V + 1.4\text{ V})$ into I_O. The circuit of Figure 6 adds R_M to provide this current, resulting in a four-quadrant multiplier where R_C is trimmed such that $V_O = 0V$ for $V_{IN2} = 0V$. R_M also serves as the load resistor for I_O.

Noting that the gain of the XR 13600 amplifier of Figure 3 may be controlled by varying the linearizing diode current I_D as well as by varying I_B, Figure 7 shows an AGC Amplifier using this approach. As V_O reaches a high enough amplitude ($3V_{BE}$) to turn on the Darlington transistors and the linearizing diodes, the increase in I_D reduces the amplifier gain so as to hold V_O at that level.

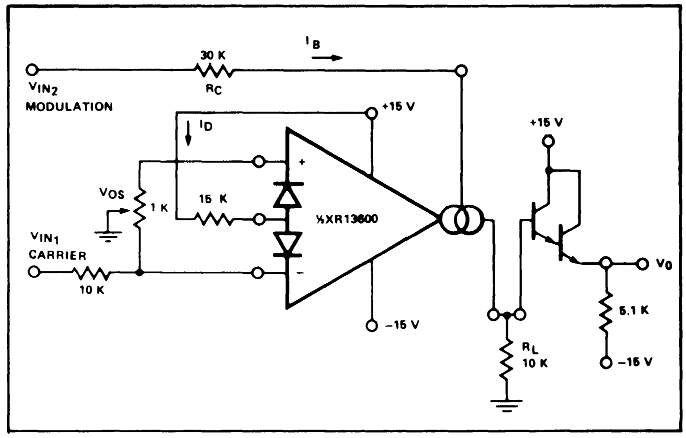


Figure 5. Amplitude Modulator

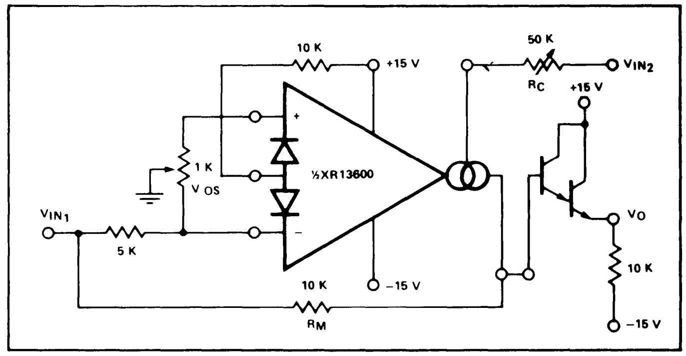


Figure 6. Four-Quadrant Multiplier

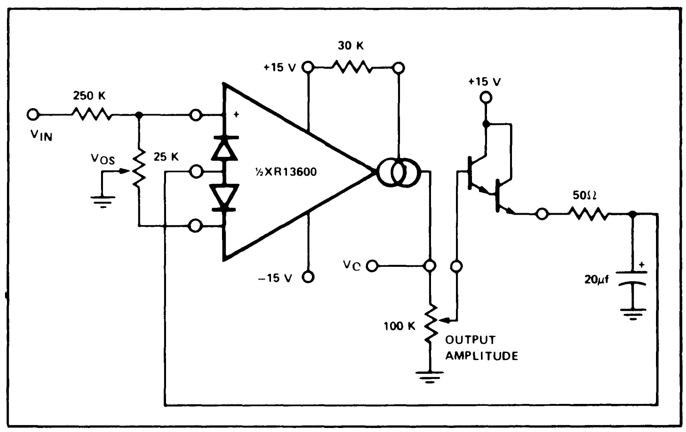


Figure 7. AGC Amplifier

VOLTAGE CONTROLLED RESISTORS (VCR)

An Operational Transconductance Amplifier (OTA) may be used to implement a Voltage Controlled Resistor as shown in Figure 8. A signal voltage applied at R_X generates a V_{IN} to the XR 13600 which is then multiplied by the gm of the amplifier to produce an output current, thus:

$$R_X = \frac{R + R_A}{g_m R_A}$$

where $g_m \approx 19.2\ I_B$ at 25°C. Note that the attenuation of V_O by R and R_A is necessary to maintain V_{IN} within the linear range of the XR 13600 input.

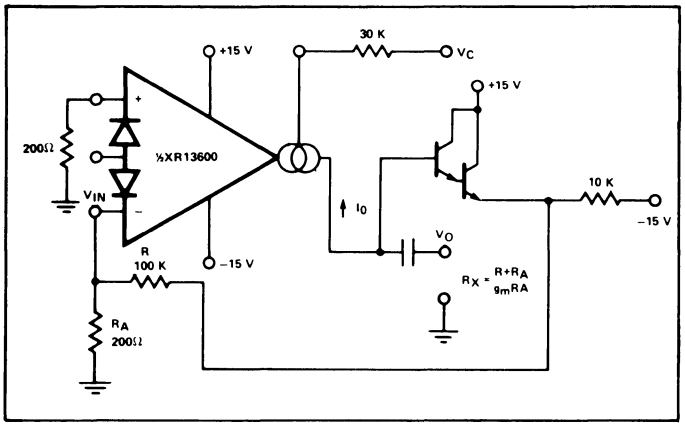


Figure 8. Voltage Controlled Resistor, Single-Ended

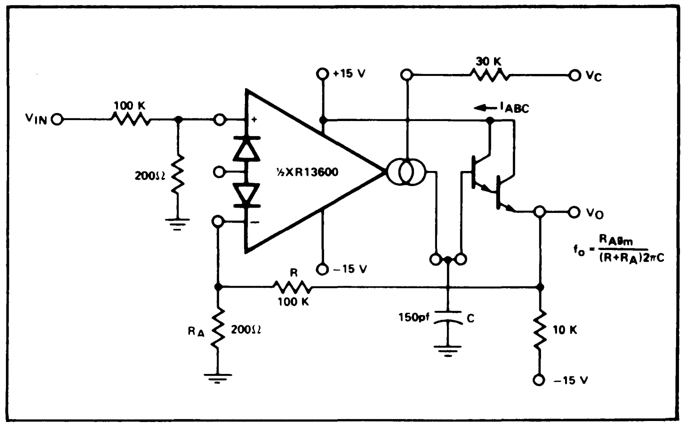


Figure 11. Voltage Controlled Low-Pass Filter

Figure 9 shows a similar VCR where the linearizing diodes are added, essentially improving the noise performance of the resistor. A floating VCR is shown in Figure 10, where each "end" of the "resistor" may be at any voltage within the output voltage range of the XR-13600.

Figure 12 shows a voltage controlled high-pass filter which operates in much the same manner, providing a single RC roll-off below the defined cut-off frequency.

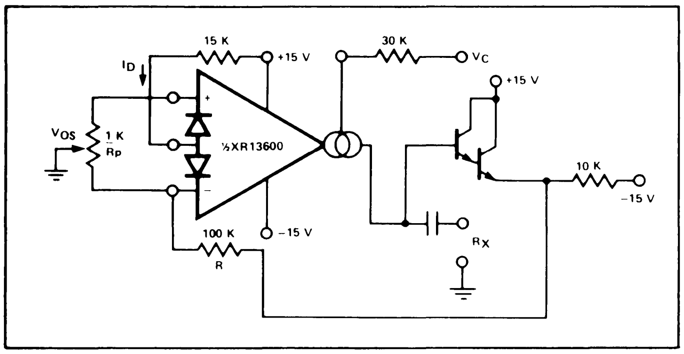


Figure 9. Voltage Controlled Resistor with Linearizing Diodes

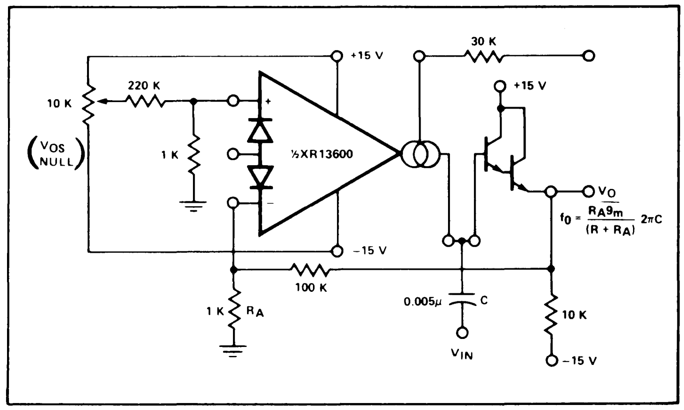


Figure 12. Voltage Controlled High-Pass Filter

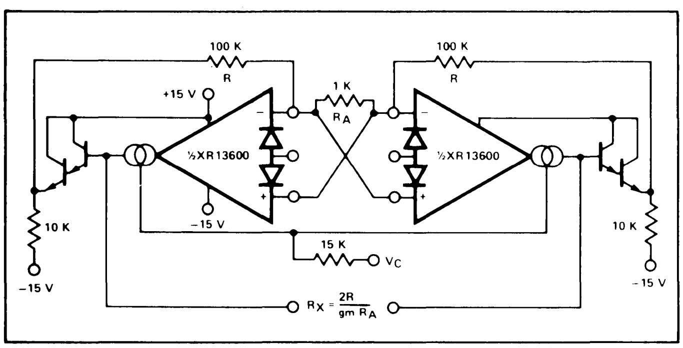


Figure 10. Floating Voltage Controlled Resistor

Additional amplifiers may be used to implement higher order filters as demonstrated by the two-pole Butterworth low-pass filter of Figure 13 and the state variable filter of Figure 14. Due to the excellent g_m tracking of the two amplifiers and the varied bias of the buffer Darlingtons, these filters perform well over several decades of frequency.

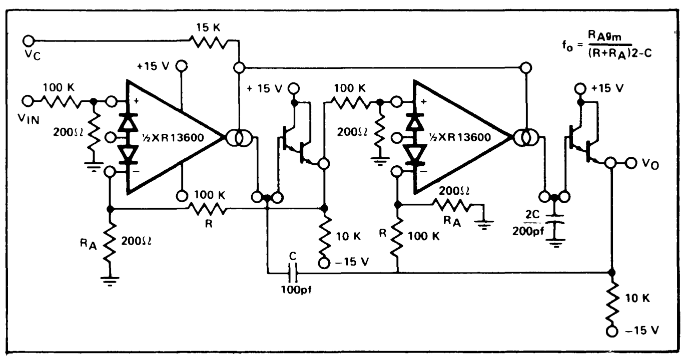


Figure 13. Voltage Controlled 2-Pole Butterworth Low-Pass Filter

VOLTAGE CONTROLLED FILTERS

OTA's are extremely useful for implementing voltage controlled filters, with the XR-13600 having the advantage that the required buffers are included on the I.C. The VC Lo-Pass Filter of Figure 11 performs as a unity-gain buffer amplifier at frequencies below cut-off, with the cut-off frequency being the point at which X_C/g_m equals the closed-loop gain of (R/R_A). At frequencies above cut-off the circuit provides a single RC roll-off (6 dB per octave) of the input signal amplitude with a -3 dB point defined by the given equation, where g_m is again $19.2 \times I_B$ at room temperature.

VOLTAGE CONTROLLED OSCILLATORS (VCO)

The classic Triangular/Square Wave VCO of Figure 15 is one of a variety of Voltage Controlled Oscillators which may be built utilizing the XR-13600. With the component values

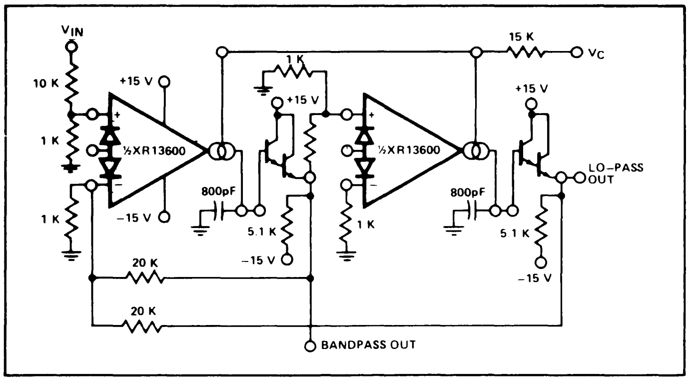


Figure 14. Voltage Controlled State Variable Filter

shown, this oscillator provides signals from 200 kHz to below 2 Hz as I_C is varied from 1mA to 10nA. The output amplitudes are set by $I_A \times R_A$. Note that the peak differential input voltage must be less than 5 volts to prevent zenering the inputs.

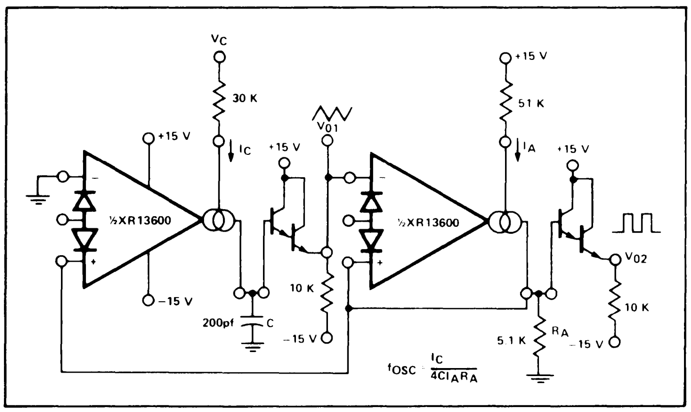


Figure 15. Triangular/Square-Wave VCO

A few modifications to this circuit produce the ramp/pulse VCO of Figure 16. When V_{O2} is high, I_F is added to I_C to increase amplifier A1's bias current and thus to increase the charging rate of capacitor C. When V_{O2} is low, I_F goes to zero and the capacitor discharge current is set by I_C.

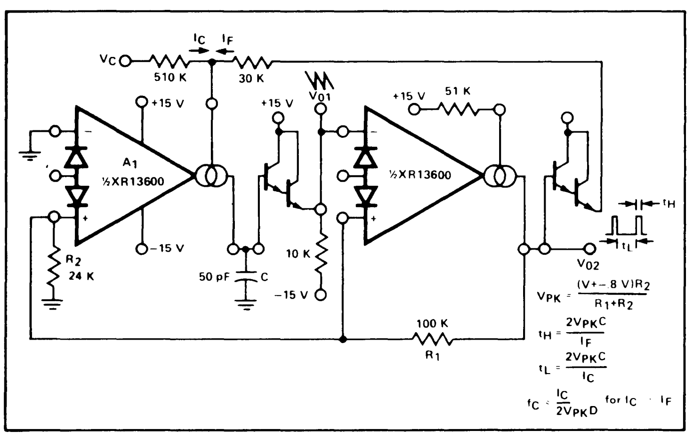


Figure 16. Ramp/Pulse VCO

The voltage-controlled low-pass filter of Figure 11 may be used to design a high-quality sinusoidal VCO. The circuit of Figure 17 employs two XR-13600 packages, with three of the amplifiers configured as low-pass filters and the fourth as a limiter/inverter. The circuit oscillates at the frequency at which the loop phase-shift is 360° or 180° for the inverter and 60° per filter stage. This VCO operates from 5 Hz to 50 kHz with less than 1% THD.

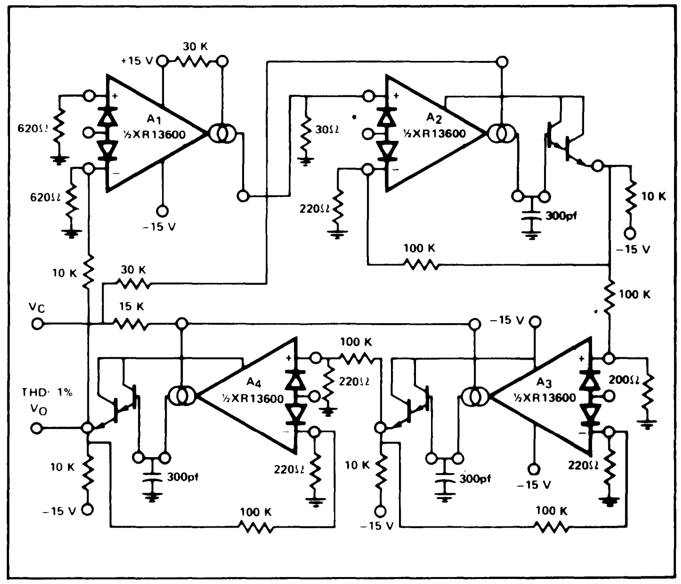


Figure 17. Sinusoidal VCO using two XR-13600 Circuits

Figure 18 shows how to build a VCO using one amplifier when the other amplifier is needed for another function.

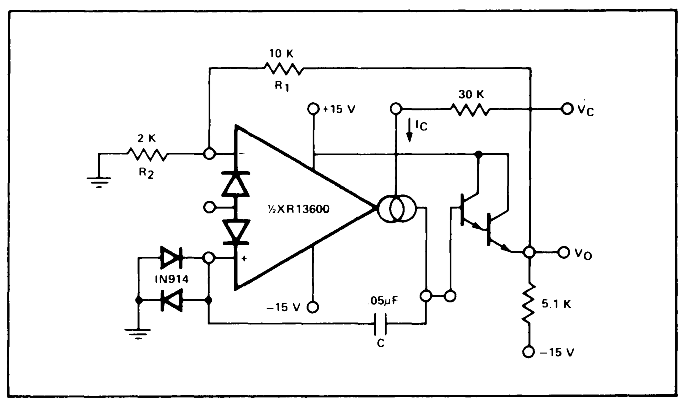


Figure 18. Single Amplifier VCO

ADDITIONAL APPLICATIONS

Figure 19 presents an interesting one-shot which draws no power supply current until it is triggered. A positive-going trigger pulse of at least 2V amplitude turns on the amplifier through R_B and pulls the non-inverting input high. The amplifier regenerates and latches its output high until capacitor C charges to the voltage level on the non-inverting input. The output then switches low, turning off the amplifier and discharging the capacitor. The capacitor discharge rate is speeded up by shorting the diode bias pin to the inverting input so that an additional discharge current flows through D_I when the amplifier output switches low. A special feature of this timer is that the other amplifier, when biased from V_O, can perform another function and draw zero stand-by power as well.

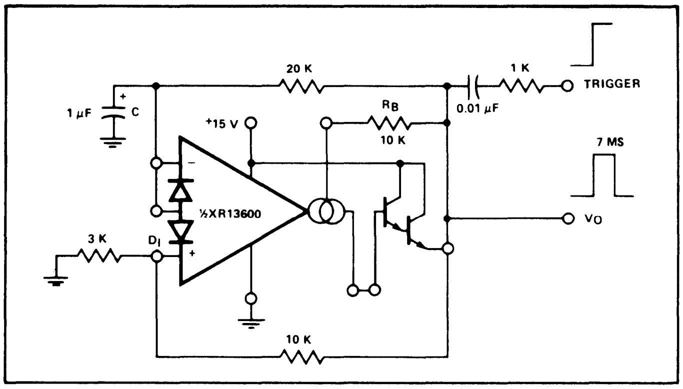


Figure 19. Timer With Zero Stand-By Power

The operation of the multiplexer of Figure 20 is very straightforward. When A_1 is turned on it holds V_O equal to V_{IN1} and when A_2 is supplied with bias current then it controls V_O. C_C and R_C serve to stabilize the unity-gain configuration of amplifiers A_1 and A_2. The maximum clock rate is limited to about 200 kHz by the XR-13600 slew rate into 150 pF when the (V_{IN1}-V_{IN2}) differential is at its maximum allowable value of 5 volts.

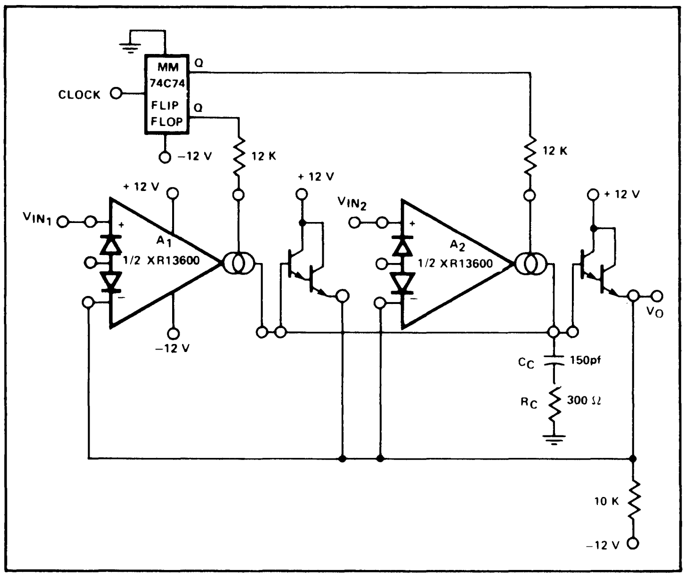


Figure 20. Multiplexer

The phase-locked loop of Figure 21 uses the four-quadrant multiplier of Figure 6 and the VCO of Figure 18 to produce a PLL with a ±5% hold-in range and an input sensitivity of about 300 mV.

The Schmitt trigger of Figure 22 uses the amplifier output current into R to set the hysteresis of the comparator; thus $V_H = 2 \times R \times I_B$. Varying I_B will produce a Schmitt trigger with variable hysteresis.

Figure 23 shows a tachometer or frequency-to-voltage converter. Whenever A1 is toggled by a positive-going input, an amount of charge equal to (V_H-V_L) C_t is sourced into C_f and R_t. This once per cycle charge is then balanced by the current of V_O/R_t. The maximum F_{IN} is limited by the amount of time required to charge C_t from V_L to V_H with a current of I_B, where V_L and V_H represent the maximum low and

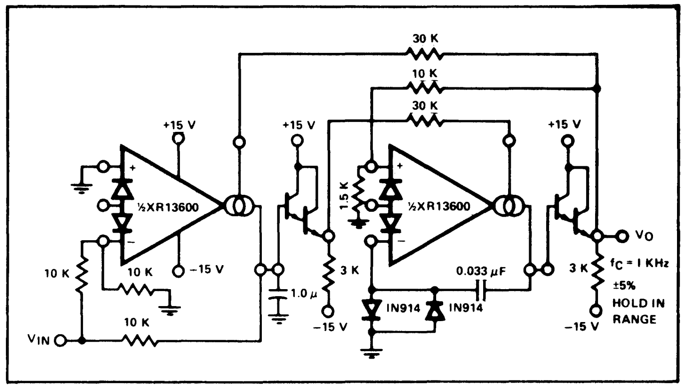


Figure 21. Phase-Locked Loop

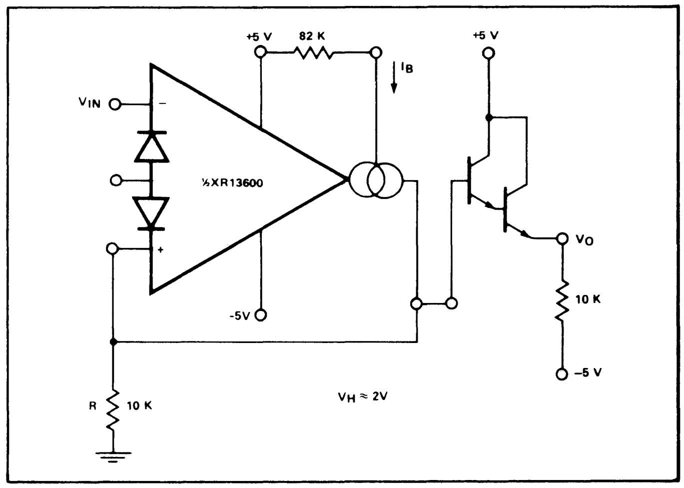


Figure 22. Schmitt Trigger

maximum high output voltage swing of the XR-13600. D1 added to provide a discharge path for C_t when A_1 switches low.

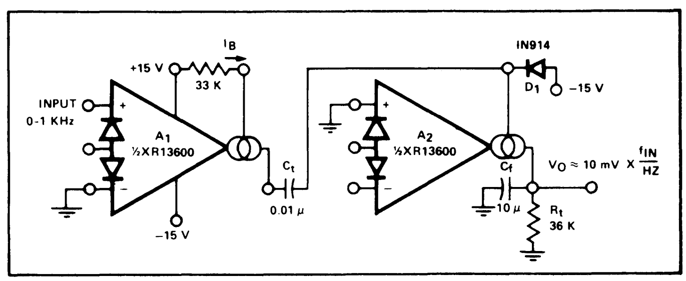


Figure 23. Tachometer

The sample-hold circuit of Figure 24 also requires that the Darlington buffer used be from the other (A_2) half of the package and that the corresponding amplifier be biased on continuously.

The peak detector of Figure 25 uses A_2 to turn on A_1 whenever V_{IN} becomes more positive than V_O. A1 then charges storage capacitor C to hold V_O equal to V_{IN} PK. One precaution to observe when using this circuit: the Darlington transistor used must be on the same side of the package as A_2 since the A_1 Darlington will be turned on and off with A_1. Pulling the output of A_2 low through D_1 serves to turn off A_1 so that V_O remains constant.

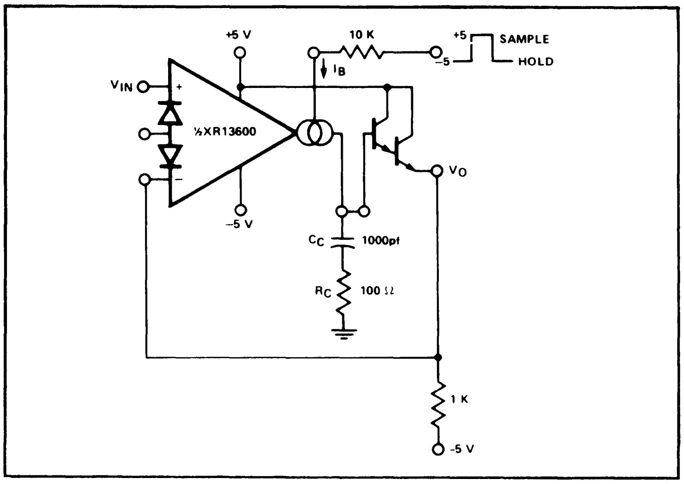


Figure 24. Sample-Hold Circuit

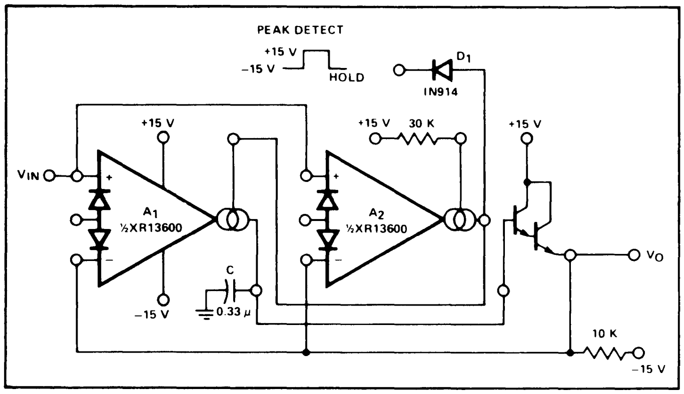


Figure 25. Peak Detector and Hold Circuit

The ramp-and-hold of Figure 26 sources I_B into capacitor C whenever the input to A_1 is brought high, giving a ramp-rate of about lV/ms for the component values shown.

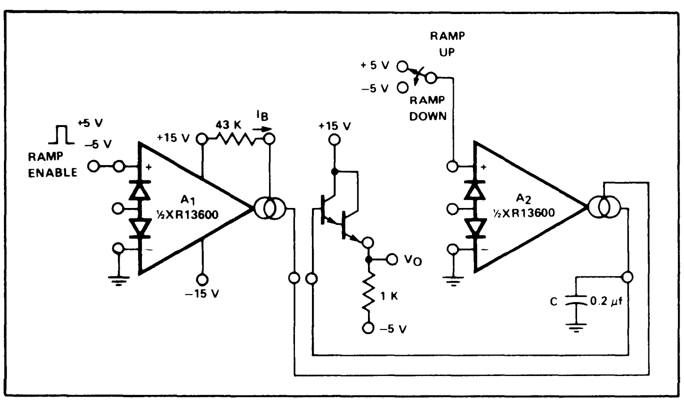


Figure 26. Ramp and Hold Circuit

The true RMS converter of Figure 27 is essentially an automatic gain control amplifier which adjusts its gain such that the AC power at the output of amplifier A_1 is constant. The output power of amplifier A_1 is monitored by squaring amplifier A_2 and the average compared to a reference voltage with amplifier A_3. The output of A_3 provides bias current to the diodes of A_1 to attenuate the input signal. Because the output power of A_1 is held constant, the RMS value is constant and the attenuation is directly proportional to the RMS value of the input voltage. The attenuation is also proportional to the diode bias current. Amplifier A_4 adjusts the ratio of currents through the diodes to be equal and therefore the voltage at the output of A_4 is proportional to the RMS value of the input voltage. The calibration potentiometer is set such that V_O reads directly in RMS volts.

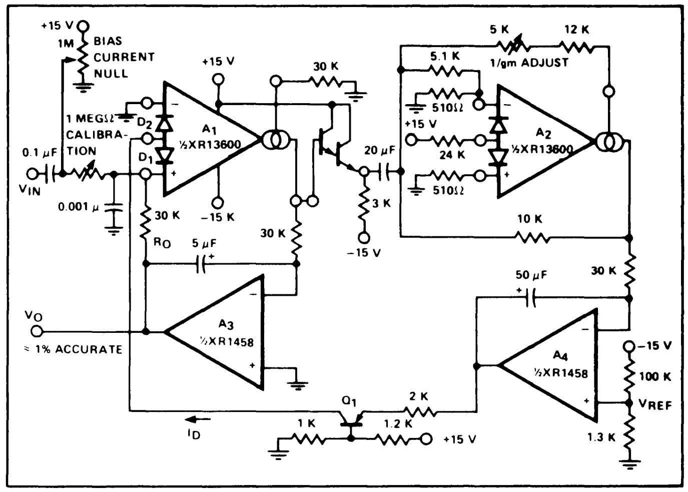


Figure 27. True RMS Converter Circuit

The circuit of Figure 28 is a voltage reference of variable Temperature Coefficient. The 100 KΩ potentiometer adjusts the output voltage which has a positive TC above 1.2 volts, zero TC at about 1.2 volts and negative TC below 1.2 volts. This is accomplished by balancing the TC of the A_2 transfer function against the complementary TC of D_1.

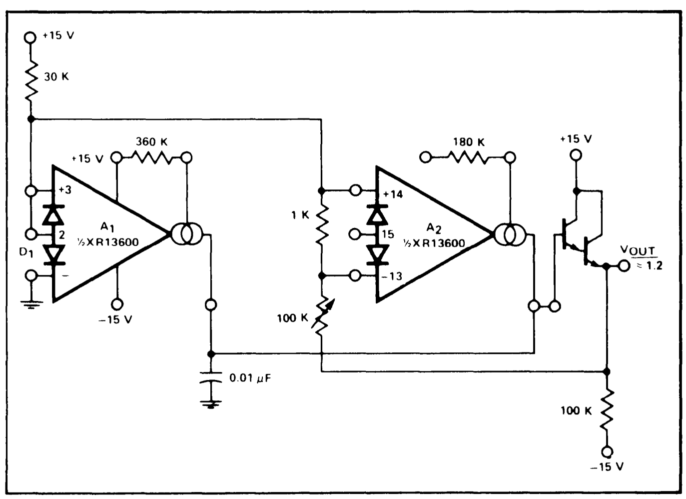


Figure 28. Delta V_{BE} Reference

The log amplifier of Figure 29 responds to the ratio of current thru buffer transistors Q_3 and Q_4. Zero temperature dependence for V_{OUT} is ensured in that the TC of the A_2 transfer function is equal and opposite to the TC of the logging transistors Q_3 and Q_4.

The wide dynamic range of the XR-13600 allows easy control of the output pulse width in the pulse-width modulator of Figure 30.

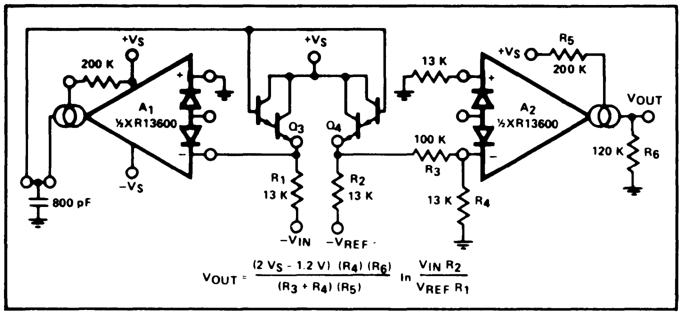


Figure 29. Log Amplifier

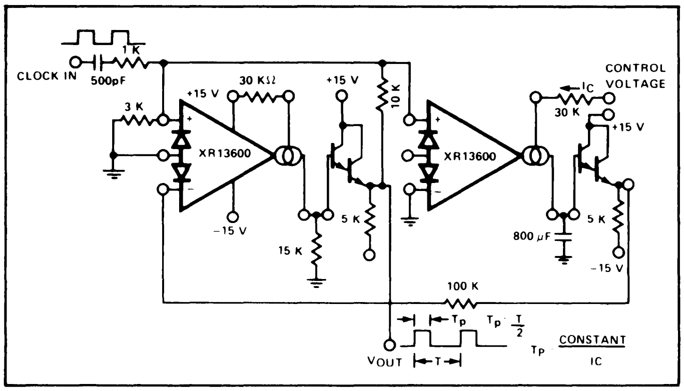


Figure 30. Pulse Width Modulator

For generating I_B over a range of 4 to 6 decades of current, the system of Figure 31 provides a logarithmic current out for a linear voltage in.

Since the closed-loop configuration ensures that the input to A_2 is held equal to 0_V, the output current of A_1 is equal to $I_3 = -V_C/R_C$.

The differential voltage between Q_1 and Q_2 is attenuated by the R_1, R_2 network so that A_1 may be assumed to be operating within its linear range. From equation (5), the input voltage to A_1 is:

$$V_{IN1} = \frac{-2KTI_3}{qI_2} = \frac{2KTV_C}{qI_2R_C}$$

The voltage on the base of Q1 is then

$$V_{B1} = \frac{(R_1+R_2)\ V_{IN1}}{R_1}$$

The ratio of the Q1 to Q2 collector currents is defined by:

$$V_{B1} = \frac{KT}{q} \ln \frac{I_{C2}}{I_{C1}} \approx \frac{KT}{q} \ln \frac{I_B}{I_1}$$

Combining and solving for I_B yields:

$$I_B = (I_1) \exp \left[\frac{2(R_1+R_2)\ V_C}{I_2R_1R_C} \right]$$

This logarithmic current can be used to bias the circuit of Figure 4 to provide temperature independent stereo attenuation characteristic.

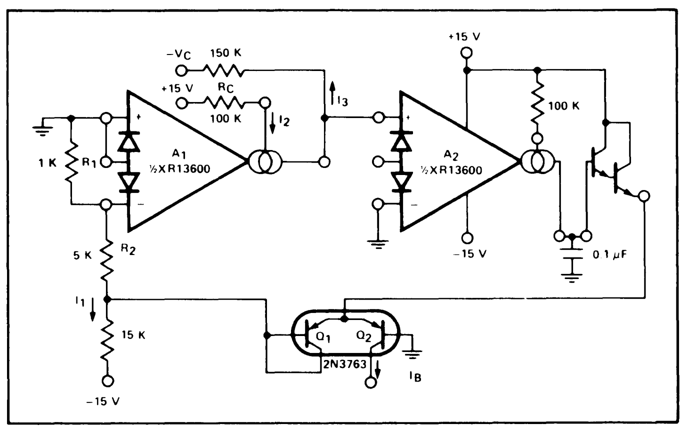


Figure 31. Logarithmic Current Source

Additional Technical Literature Available from Exar

PRODUCT GUIDE:

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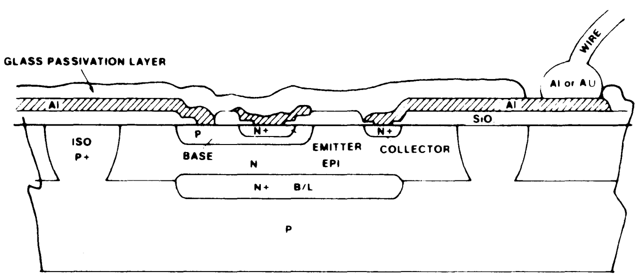
The major performance characteristics of Exar products are also available in chip form. All chips are 100% electrically tested for guaranteed DC parameters at 25°C; and 100% visually inspected at 30x to 100x magnification using Exar's standard visual inspection criteria or MIL-STD-883, Method 201, depending on the individual customer requirements. Each chip is protected with an inert glass passivation layer over the metal interconnections. The chips are packaged in waffle-pack carriers with an anti-static shield and cushioning strip plated over the active surface to assure protection during shipment. All chips are produced on the same well-proven production lines that produce Exar's standard encapsulated devices. The Quality Assurance testing of dice is provided by normal production testing of packaged devices.



Typical Bipolar Chip Cross Section

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ELECTRICAL PARAMETERS

Probing the IC chips in die form limits the electrical testing to low level DC parameters at 25°C. These DC parameters are characteristic of those parameters contained on the individual device data sheet and are guaranteed to an LTPD of 10%.

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Cavity or Waffle Pack: The dice are placed in individual compartments of the waffle pack (see figure). The plastic snap clips permit inspection and resealing.

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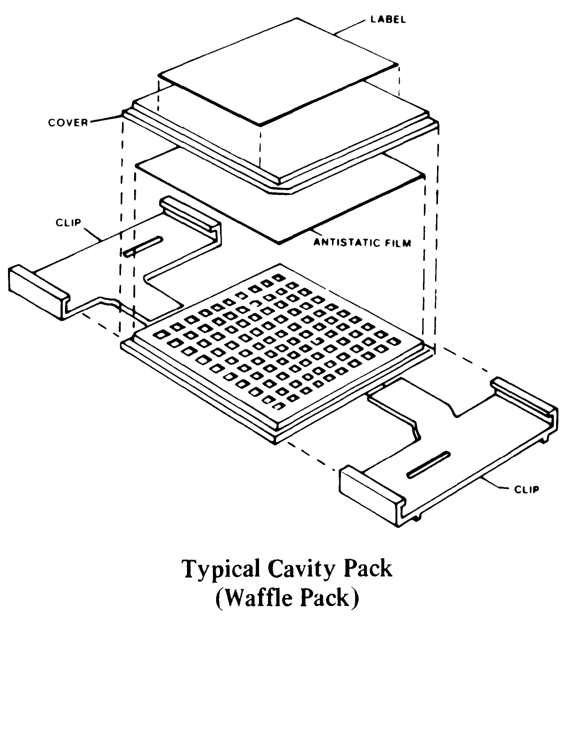
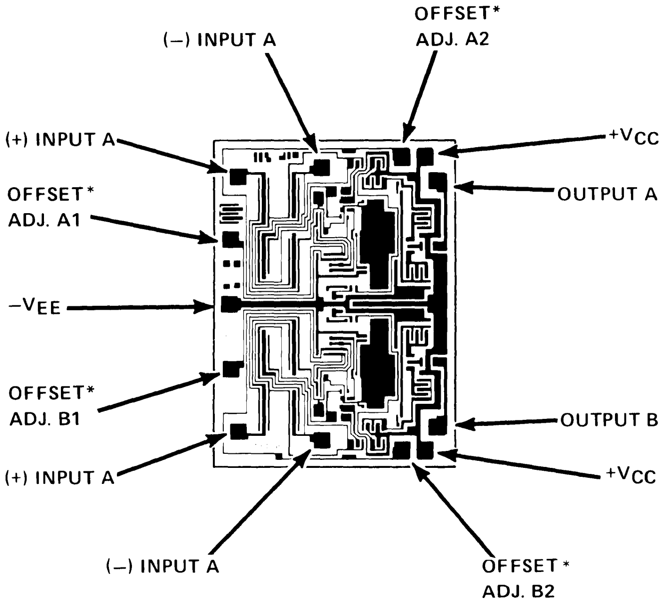
Wafer Pack: The entire wafer is sandwiched between two pieces of mylar and vacuum sealed in a plastic envelope.



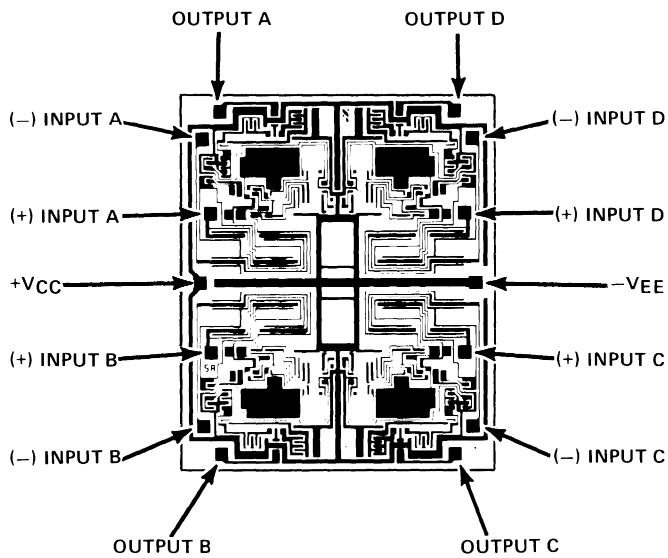
Typical Cavity Pack (Waffle Pack)

XR-082/XR-083
DUAL BIFET OPERATIONAL AMPLIFIERS



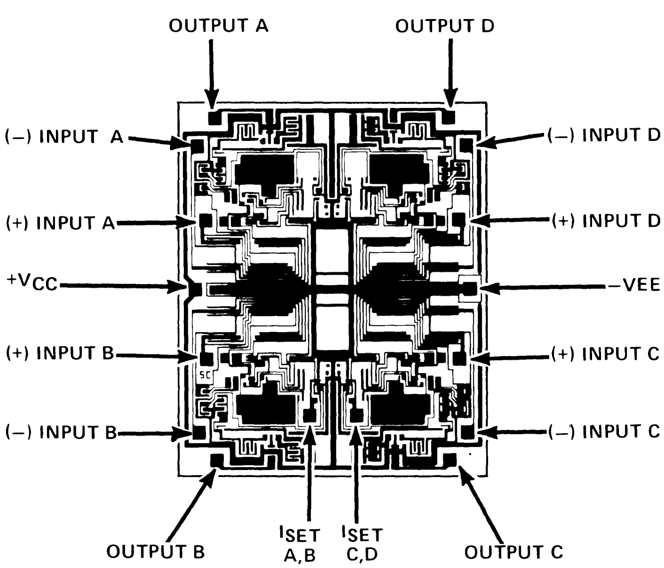
(*) Available in XR-083 Only
Chip Size: 67 mils x 90 mils
(1.70 mm x 2.28 mm)

XR-084
QUAD BIFET OPERATIONAL AMPLIFIERS



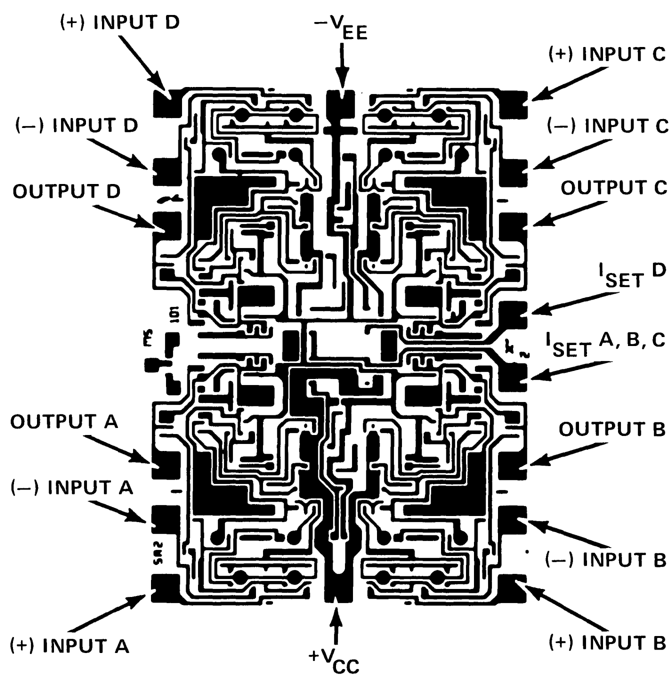
Chip Size: 100 mils x 120 mils
(2.53 mm x 3.04 mm)

XR-094/XR-095
PROGRAMMABLE QUAD BIFET OPERATIONAL AMPLIFIERS



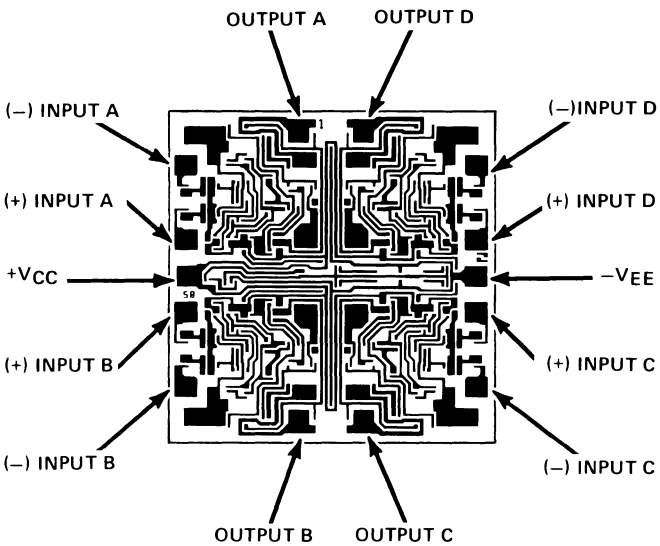
Chip Size: 100 mils x 120 mils
(2.53 mm x 3.04 mm)

XR-146/246/346 AND XR-346-2
PROGRAMMABLE QUAD OPERATIONAL AMPLIFIERS



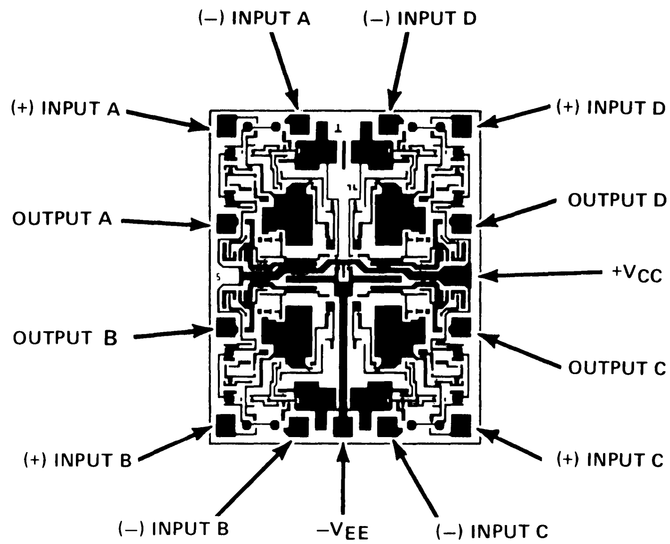
Chip Size: 67 mils x 88 mils
(2.33 mm x 1.70 mm)

XR-3403/XR-3503
QUAD OPERATIONAL AMPLIFIER



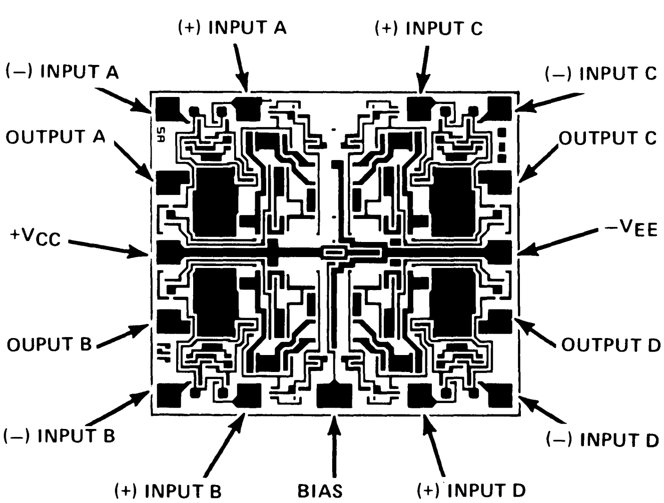
Chip Size: 73 mils x 74 mils
(1.85 mm x 1.87 mm)

XR-4136
QUAD OPERATIONAL AMPLIFIER



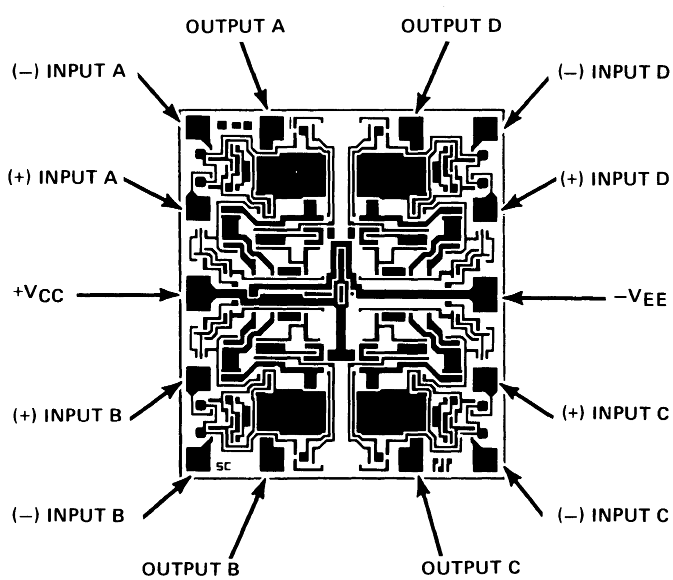
Chip Size: 68 mils x 73 mils
(1.72 mm x 1.85 mm)

XR-4202
PROGRAMMABLE QUAD OPERATIONAL AMPLIFIER



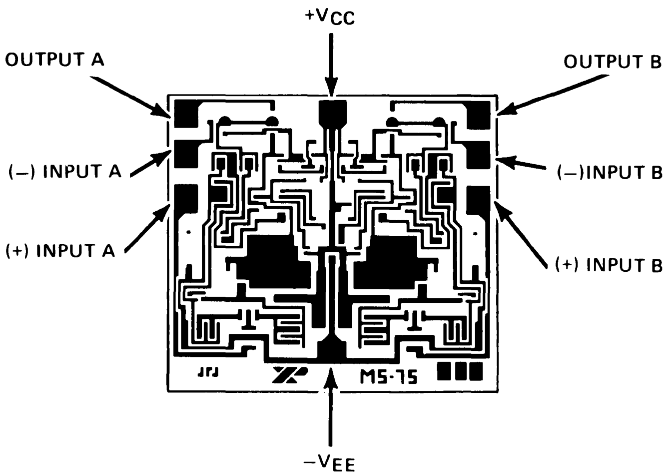
Chip Size: 63 mils x 71 mils
(1.57 mm x 1.77 mm)

XR-4212/XR-4741
QUAD OPERATIONAL AMPLIFIERS



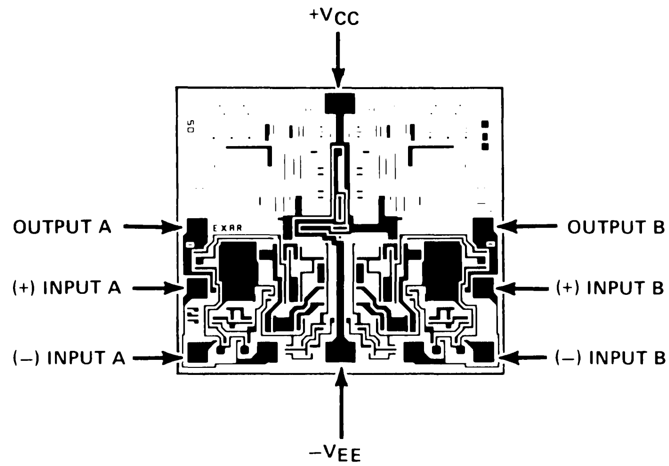
Chip Size: 63 mils x 71 mils
(1.57 mm x 1.77 mm)

XR-1458/4558
DUAL OPERATIONAL AMPLIFIER



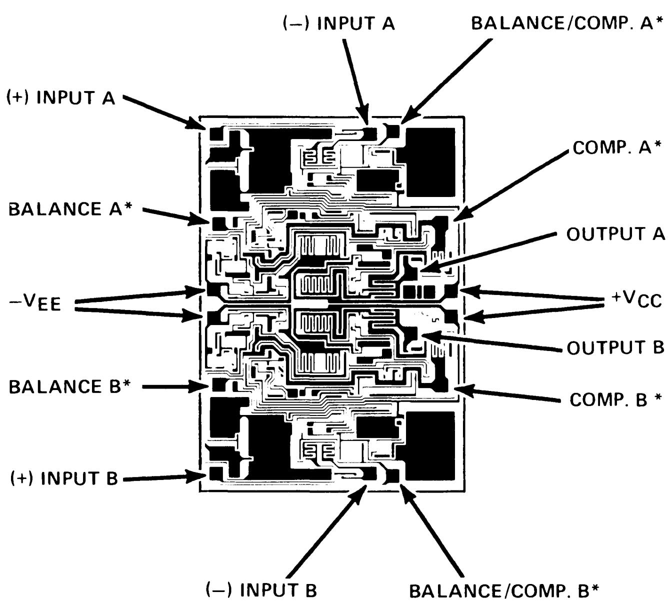
Chip Size: 61 mils x 55 mils
(1.54 mm x 1.39 mm)

XR-4739
DUAL LOW-NOISE OPERATIONAL AMPLIFIERS



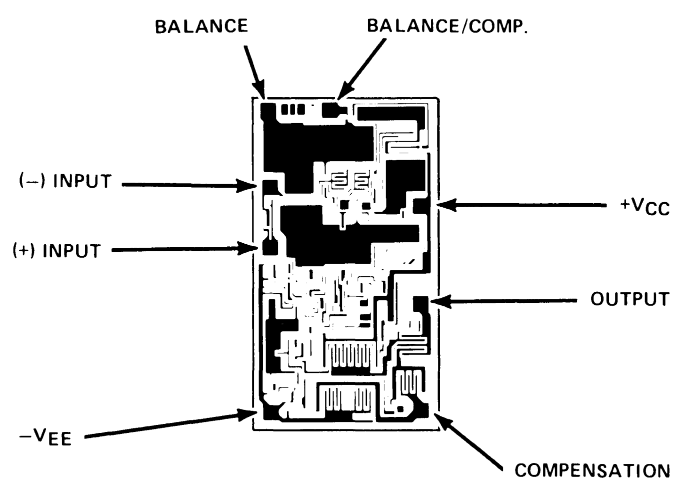
Chip Size: 63 mils x 71 mils
(1.70 mm x 1.77 mm)

XR-5532/XR-5533
DUAL LOW-NOISE OPERATIONAL AMPLIFIERS



*Available in XR-5533 Only
Chip Size: 85 mils x 118 mils
(2.15 mm x 2.99 mm)

XR-5534
LOW-NOISE OPERATIONAL AMPLIFIER



Chip Size: 52 mils x 92 mils
(1.32 mm x 2.33 mm)

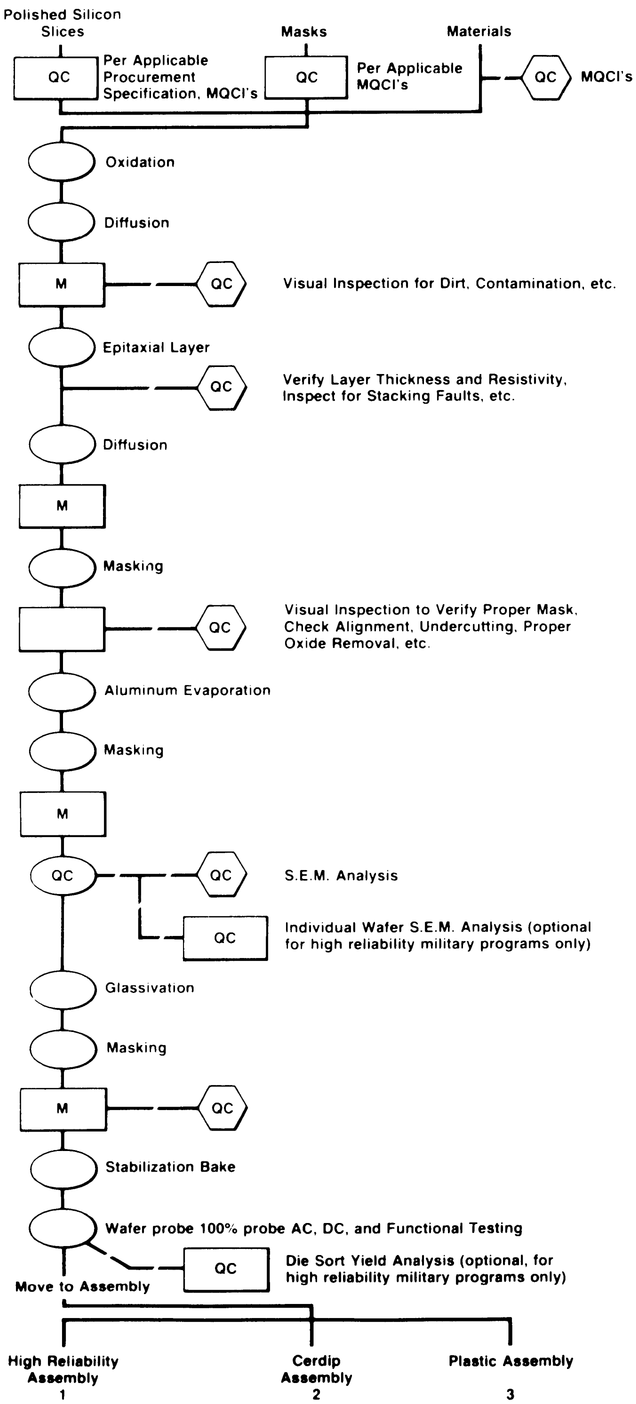
QUALITY ASSURANCE STANDARDS

The quality assurance program at Exar Integrated Systems defines and establishes standards and controls on manufacturing, and audits product quality at critical points during manufacturing. The accompanying Manufacturing/QA process flows illustrate where quality assurance audits, by inspection or test, the manufacturing process. The insertion of these quality assurance points is designed to insure the highest quality standards are maintained on Exar product during its manufacture.

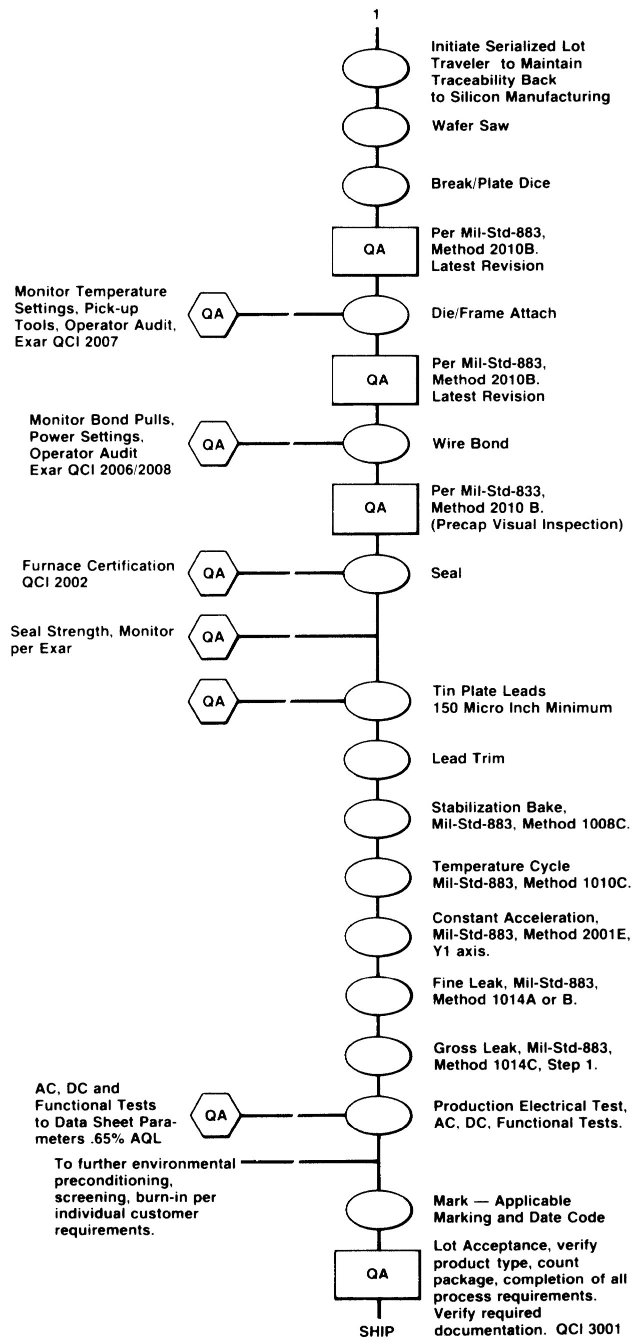
Realizing that these standard Manufacturing/QA process flows do not meet the needs of every customer's specific requirements, Exar quality assurance can negotiate and will screen product to meet any individual customer's specific requirement.

All products ending with the suffix M are fully screened to the requirements of MIL-STD-883, Method 5004, Condition C.

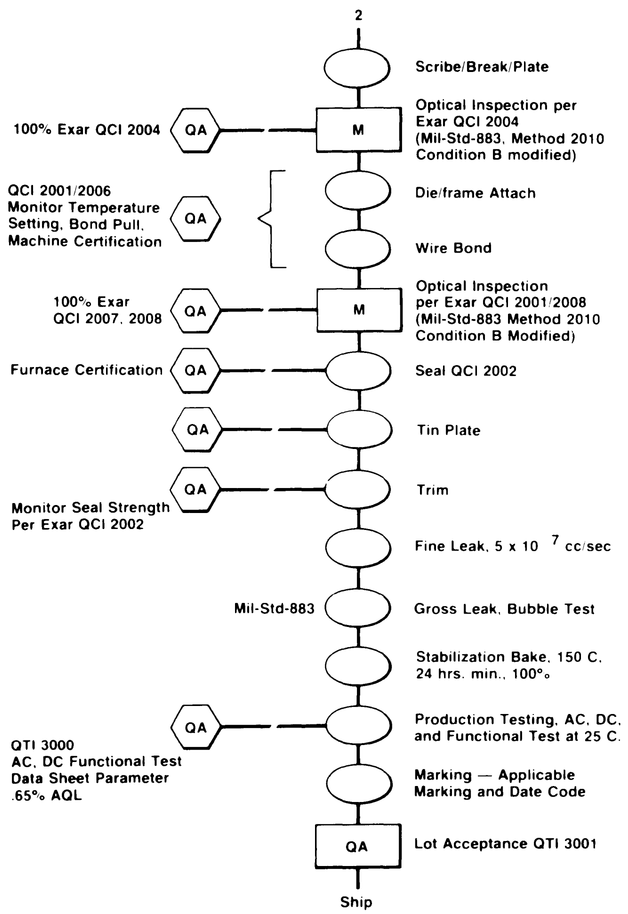
Wafer Fabrication/QA Flow



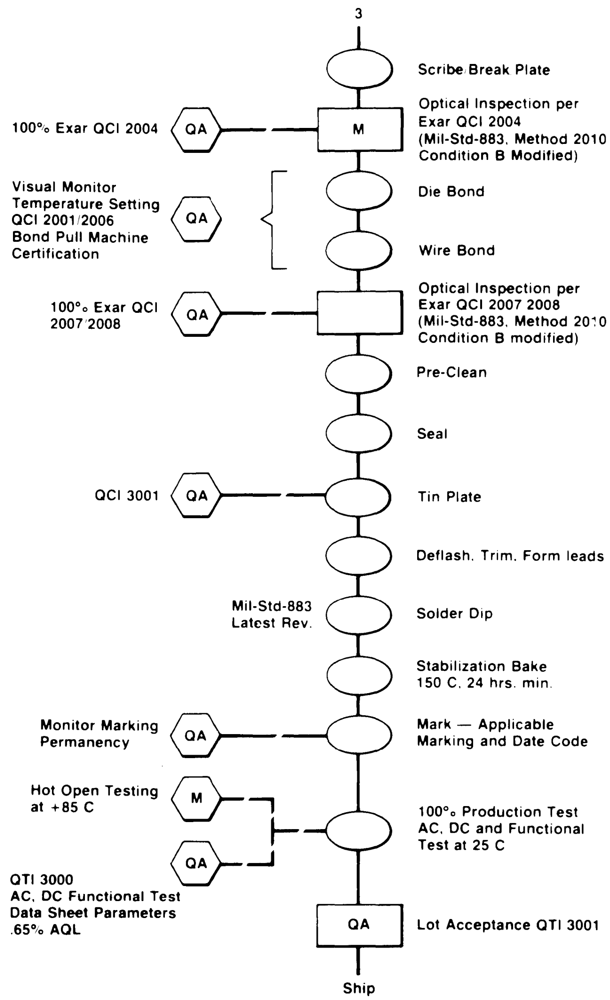
High Reliability Assembly/QA Flow



Cerdip Assembly/QA Flow



Plastic Assembly/QA Flow



PRODUCT ORDERING INFORMATION

Part Identification

XR
Manufacturer's Prefix

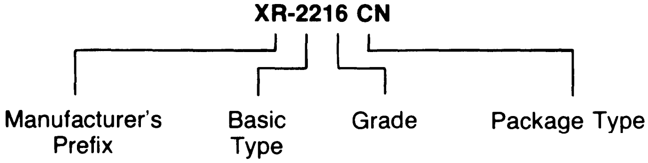
Grade
M = Military
N = Prime Electrical
P = Prime Electrical
C = Commercial
K = Kit

XXXXX
Basic Type (5 spaces)

Package Type
N = Ceramic Dual-in-line
P = Plastic Dual-in-line

Example:

XR-2216 CN



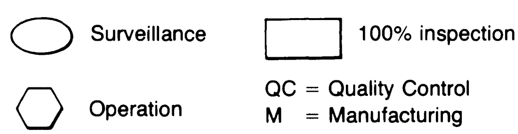
Definition of Symbols:

M = Military Grade Part, Ceramic Package Only. All Military Grades have been processed to MIL-STD-883 Level C, and are guaranteed to operate over military temperture range.
N = Prime Grade Part, Ceramic Package.
P = Prime Grade Part, Plastic Package.
CN = Commercial Grade Part, Ceramic Package.
CP = Commercial Grade Part, Plastic Package.

N, P, CN and CP parts are electrically identical and guaranteed to operate over 0°C to + 75°C range unless otherwise stated. In addition, N and P parts generally have operating parameters more tightly controlled than the CN or CP parts.

For details, consult Exar Sales Headquarters or Sales /Technical Representatives.

Legend:



QC = Quality Control
M = Manufacturing

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Pisos 6Y2
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Melbourne, 3001
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Telex (790) 31261 (TOTELEC AA)

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Telex: (855) 37360 (MEREL DK)

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Phone: 90-562 1122
Telex: 123212 (YLEOY SF)

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Rue Carle Vernet
F-92310 Sevres
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Telex: (842) 204552 (TKLEC A)

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Muehlenstrasse 70
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Telex: (841) 852330 (ROHM D)

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Ramat-Gan 52462
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20154 Milano
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Telex: (843) 332332 (ELEDRA I)

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Telex: 346369 (SUVL)

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Oslo 5
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Telex: (856) 16205 (HEFRO N)

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Telex: (960) 4-24849 (SA)

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Madrid 8,
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Telex: (831) 46786 (UTRON E)

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S-194 01 Upplands Vasby
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Ottawa, Ontario K2B 7E9
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EXAR

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March 1982