

OPERATIONAL AMPLIFIERS

OPERATIONAL AMPLIFIERS

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- Specifications and Testing
- Amplifier Architectures
- Linearity and Distortion
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OPERATIONAL AMPLIFIERS

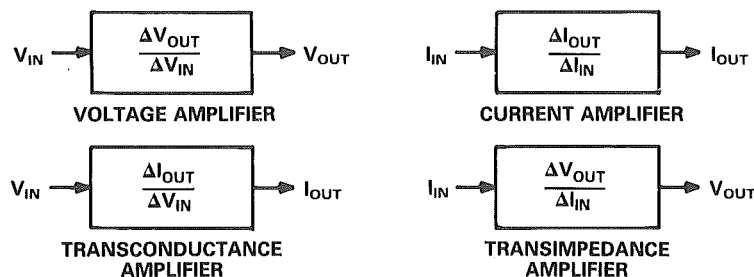
INTRODUCTION TO AMPLIFIERS

This part of our seminar discusses amplifiers, their specifications, electrical characteristics, physical realization and applications. The greatest attention is paid to the integrated circuit forms of both operational and trans-impedance amplifiers, emphasizing stability, linearity, harmonic distortion, device design, gain, bandwidth, grounding and noise. The application and selection of specific devices for use in active bias networks will also be discussed.

CLASSES OF AMPLIFIERS

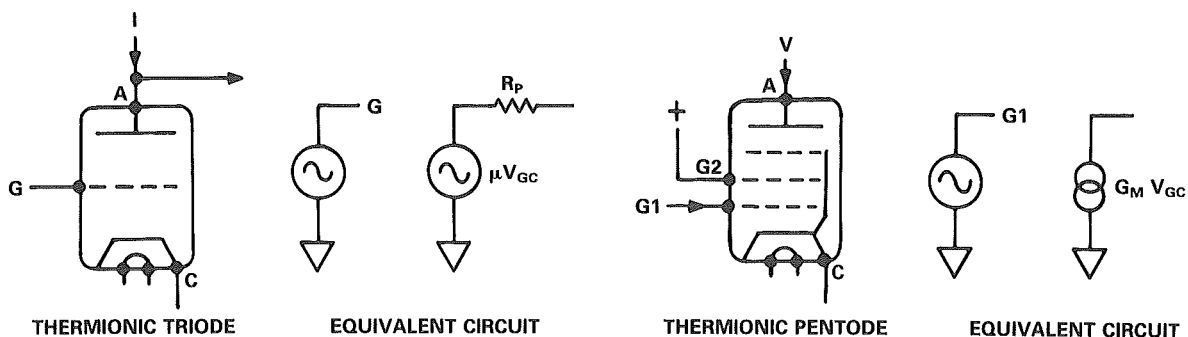
There are four general classes of amplifier, characterized by transfer characteristics expressed in volts per volts, amperes per ampere, volts per ampere, and amperes per volt.

TYPES OF AMPLIFIERS



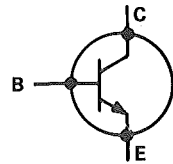
These basic amplifiers are the building blocks required to synthesize larger electronic systems. The earliest form of man-made amplifier was the triode vacuum tube, a voltage amplifier, designed by the English inventor Lee DeForest in time for World War 1. The triode is an effective amplifier but has limitations which led to the development of the pentode vacuum tube—a transconductance device. These two vacuum tube devices can be modeled as a voltage dependent voltage source and a voltage dependent current source respectively.

VACUUM TUBE AMPLIFIERS

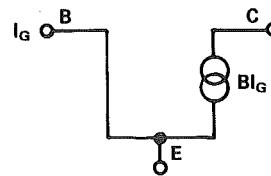


Today vacuum tubes have largely been displaced by transistors, both bipolar and field effect. These are used both as stand alone devices and incorporated into integrated circuits. Although there are simple electronic devices which are voltage amplifiers (the triode vacuum tube), current (the bipolar transistor), and transconductance amplifiers (the pentode vacuum tube and the field effect transistor), there are presently no electronic devices known which demonstrate the intrinsic characteristics of a transimpedance amplifier. Hence transimpedance amplifiers must be synthesized from voltage, current or transconductance amplifiers.

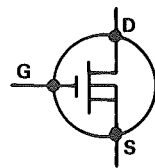
DISCRETE SEMICONDUCTOR AMPLIFIERS



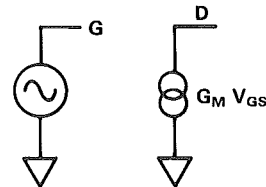
(NPN) BIPOLAR TRANSISTOR



EQUIVALENT CIRCUIT



FIELD EFFECT TRANSISTOR (FET)



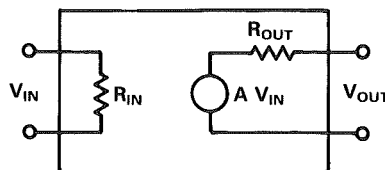
EQUIVALENT CIRCUIT

The general forms of the aforementioned amplifiers are modeled by applying Norton and Thevenin equivalent networks.

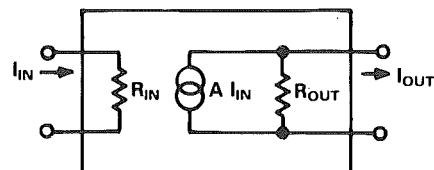
IDEAL AMPLIFIER CHARACTERISTICS

	AMPLIFIER TYPE			
	VOLTAGE	CURRENT	TRANSCONDUCTANCE	TRANSIMPEDANCE
TRANSFER FUNCTION	$V_O = A V_{IN}$	$I_O = A I_{IN}$	$I_O = G_M V_{IN}$	$V_O = R_M I_{IN}$
R_{IN} R_{OUT}	∞ 0	0 ∞	∞ ∞	0 0

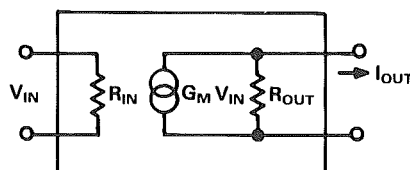
AMPLIFIER EQUIVALENT CIRCUITS



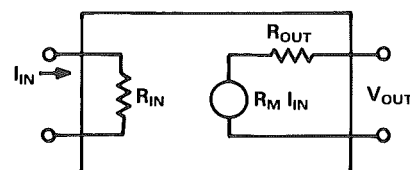
VOLTAGE AMPLIFIER



CURRENT AMPLIFIER



TRANSCONDUCTANCE AMPLIFIER



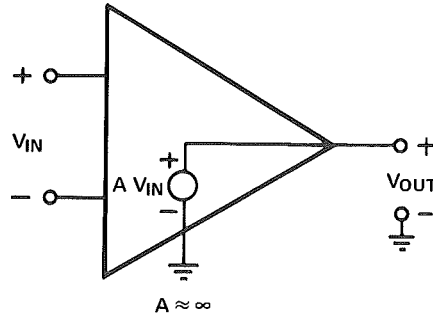
TRANSIMPEDANCE AMPLIFIER

BASIC OPERATIONAL AMPLIFIER

The operational amplifier or "op amp", is the most commonly used analog component. Op amps are available in a wide variety of technologies, specifications, prices and package styles from a multitude of vendors. When a circuit requires an operational amplifier the designer is confronted with a bewildering number of devices from which to choose. This seminar will serve to make the selection process somewhat simpler.

The ideal op amp has differential inputs, an infinite input impedance a single-ended output, and infinite gain at all frequencies. Many engineers also assume that the ideal op amp is three-terminal device and disregard Kirchoff's Law which, in essence, says that all current circuits are closed ("What goes out must come in again"). This can lead to serious circuit problems. The ideal op amp must always be considered as a four terminal device, the fourth terminal being the return path for the output current. In most designs, it is assumed that the amplifier is, in fact, an ideal, and that no current flows into the input terminals, there is no input offset voltage and no power is required for its operation.

IDEAL OP AMP



THE REAL OPERATIONAL AMPLIFIER

Input Imperfections

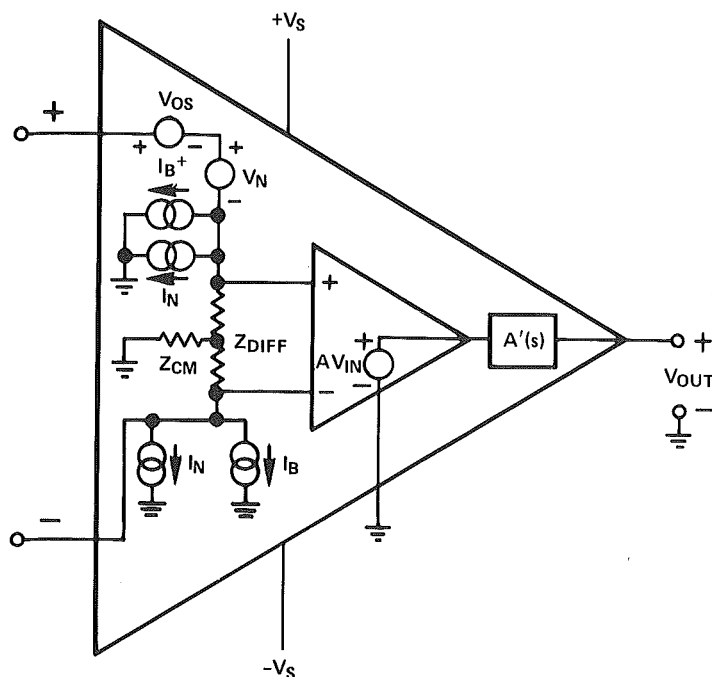
The actual characteristics of real op amps are considerably more complicated. Each input contains a dc current source (I_B , the bias current), and is a dc voltage source (V_{OS} , the offset voltage) in series with the inputs. The amplifier has differential and common-mode input impedances ($Z_{IN(DIFF)}$ and $Z_{IN(CM)}$) respectively) which are usually complex, consisting of a resistance and capacitance in parallel.

There are also three uncorrelated noise sources: two current sources (one in series with each input) and a voltage source which appears differentially. Finally the amplifier has gain w.r.t. common-mode signals, which the ideal amplifier has not, and so its common-mode rejection ratio (CMRR) must be specified.

Output Obstacles

The output side of the model is also not ideal. There is an output impedance (R_O) in series with the voltage source. The gain (A —infinite in the ideal model) is both finite and a function of frequency in a real amplifier, which also has a finite slew rate and limited output voltage and current capabilities.

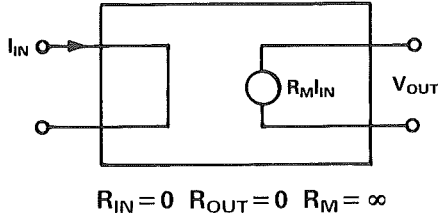
REAL OP AMP



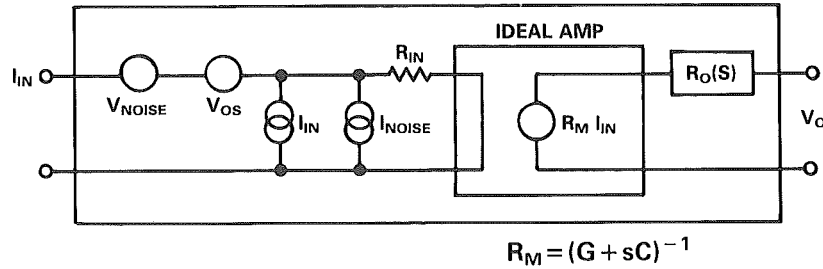
BASIC TRANSIMPEDANCE AMPLIFIER

The transfer function of a transimpedance amplifier is a scaling factor specified in volts per ampere or simply R_M . In the ideal case R_M has zero phase shift associated with it but a real transimpedance amplifier has a frequency dependent R_M .

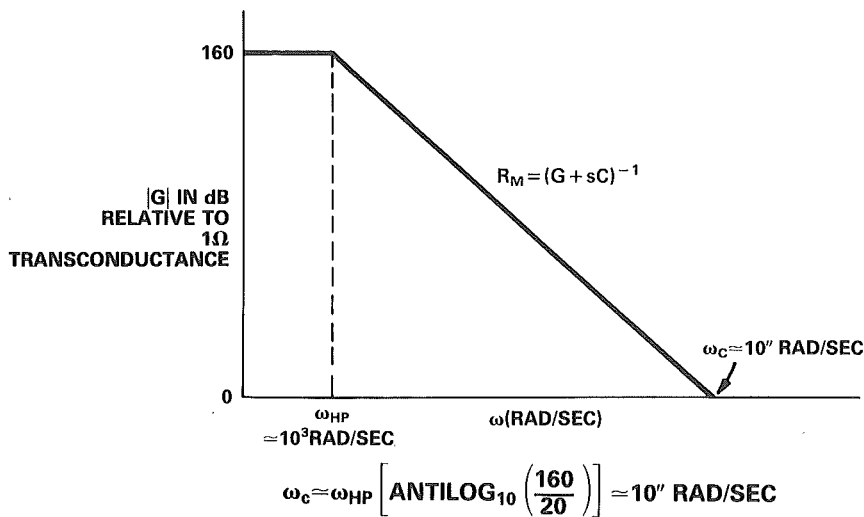
IDEAL TRANSIMPEDANCE AMPLIFIER



REAL TRANSIMPEDANCE AMPLIFIER



AD846 TRANSIMPEDANCE AMPLIFIER – THEORETICAL OPEN LOOP FREQUENCY RESPONSE

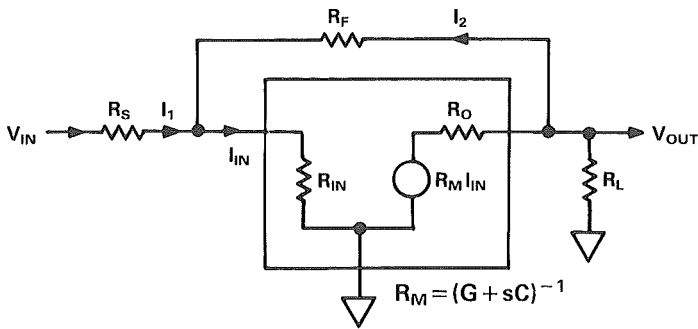


A real transimpedance amplifier and a real voltage (operational) amplifier are very similar with respect to inadequacies and deviation from the ideal model. Regardless of which type of amplifier is considered, limitations exist in fabrication techniques and manifest themselves as temperature variations of parameters, as offset voltages and currents, as nonzero bias currents, and as finite CMRR, PSRR, input and output impedances, and open loop gain.

The significant difference between the devices is that the op amp is a voltage controlled voltage source and the transimpedance amplifier is a current controlled voltage source. The computation of voltage gains, input and output impedances, in a transimpedance amplifier is very similar to those of op amps, Ohms Law and the principal of superposition are utilized.

Since the input to a transimpedance amplifier is a current and the output is a voltage the gain is expressed in ohms ($R = V/I$). Since small changes in current cause large voltage changes the gain in ohms is quite large—about 100,000,000 in the case of the AD846. In the case of the gain is expressed in dB w.r.t. 1Ω or 160dB. This is by no means the same thing as a voltage gain of 160dB. In a voltage amplifier or a current amplifier it is reasonable to talk of “unity gain” since at unity gain the input and output are equal, but in a transimpedance or transconductance amplifier the input and output signals are of different types and unity gain has little or no meaning. (Volts, amperes and ohms are, in this context, quite arbitrary quantities and to consider unity gain to be equal to a transimpedance of one volt per ampere or a transconductance of one ampere per volt has no more validity than a value of unity gain defined by any other pair of values.) It follows that the 16GHz gain-bandwidth product of the AD846 is due to the choice of unit for its gain and is not a useful parameter. It is more interesting to consider the behavior of a transimpedance amplifier in the inverting mode.

INVERTING GAIN CALCULATION FOR TRANSIMPEDANCE AMPLIFIER



$$R_{IN} \approx 67\Omega \text{ (ASSUME ZERO)}$$

$$R_{OUT} \approx 35\Omega \text{ (ASSUME ZERO)}$$

$$\therefore I_{IN} = I_1 + I_2 = \frac{V_{IN}}{R_S} + \frac{V_{OUT}}{R_F}$$

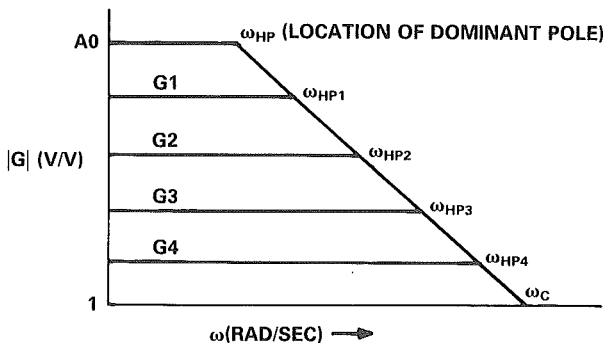
$$V_{OUT} = R_M I_{IN} \quad I_{IN} = \frac{V_{OUT}}{R_M}$$

$$\text{SUBSTITUTE: } \frac{V_{OUT}}{R_M} = \frac{V_{IN}}{R_S} + \frac{V_{OUT}}{R_F}$$

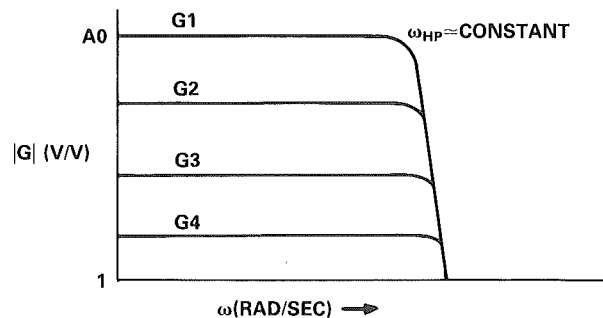
$$\text{SOLVE FOR } \frac{V_{OUT}}{V_{IN}} \quad \frac{V_{OUT}}{V_{IN}} = \frac{-\frac{R_F}{R_S}}{1 - \frac{R_F}{R_M}} = \frac{-\frac{R_F}{R_S}}{1 - R_F(G + sC)}$$

If R_F is held constant and the variation in closed loop gain is determined by altering the value of R_S , the half power point remains constant and for values of $R_F \ll R_{IN}$, $V_{OUT}/V_{IN} = -R_F/R_S$. Practical values of closed loop gains are in the order of 100 volts per volt. Similar computations, using fundamental principles, can be made to show the effects on the input and output impedances as the loop gain is varied. The transimpedance amplifier is clearly a valuable tool for the circuit designer.

COMPARISON OF OP AMP AND TRANSIMPEDANCE AMP CLOSED LOOP GAIN/FREQUENCY PERFORMANCE



DOMINANT POLE OP AMP COMPENSATION WITH
VARIOUS CLOSED LOOP RESPONSES



TRANSIMPEDANCE AMPLIFIER CONFIGURED AS AN
INVERTING VOLTAGE AMPLIFIER

AMPLIFIER SPECIFICATIONS AND TESTING

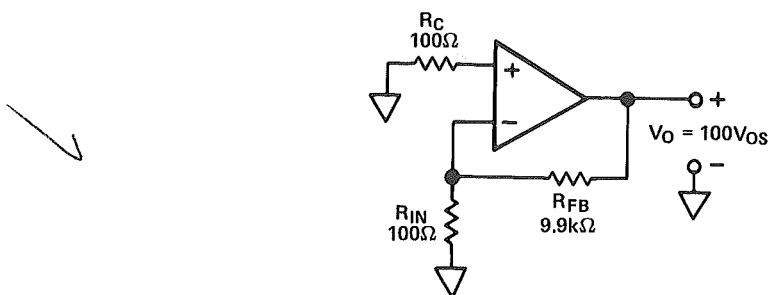
There are two other classes of amplifier, with differential inputs and a single-ended output: Instrumentation Amplifiers and Isolation Amplifiers. They are both voltage amplifiers and differ from the op amp in having gains which are lower (generally in the range 1–1,000), very stable, and accurately defined without negative feedback connections to their inputs. The isolation amplifier also has a common-mode range of many hundreds or thousands of volts. These amplifiers are described in detail in another part of this seminar but since they have a similar architecture they are tested in the same way as op amps. All the op amp test circuitry described in this section may also be used to test instrumentation and isolation amplifiers.

Although op amps and transimpedance amps have similar error terms the application of each part warrants discussing errors in separate contexts. Some specifications will be examined separately, but similarities will be pointed out when appropriate.

In general, dc errors result from lack of symmetry and inadequate matching of components. Consider the op amp. Its offset voltage (V_{OS}) is defined as the voltage which must be applied to the input to cause the output to be zero. Offset voltage is the result of mismatch in the base-emitter voltages of the differential input transistors (gate-source voltage mismatch in FET-input amplifiers) and is indistinguishable from a dc input signal. This offset can be trimmed to zero with a potentiometer, which adjusts the balance of the operating currents in the input stage until V_{BE1} and V_{BE2} (or V_{GS1} and V_{GS2}) are equal. Even if the offset voltage is trimmed to zero at one particular temperature it is a temperature variable phenomenon. When a bipolar transistor op amp is trimmed for minimum offset it is thereby trimmed for minimum temperature drift but this is not the case for FET-input op amps. The physics of this difference is described in a later section of the seminar.

Many circuits exist for measuring offset voltage. The simplest and most commonly illustrated in textbooks uses only three resistors in addition to the op amp under test. The gain of the amplifier is set by the two resistors R_{IN} and R_{FB} and the noninverting input is grounded via a bias compensation resistor (unnecessary if the op amp is a low bias current type). The offset voltage is then amplified by the closed-loop gain of the system. The output voltage is therefore a multiple of the offset voltage and it may be measured and the offset voltage calculated. Of course the closed-loop gain must be less than 1% of the open loop gain if the measurement is to be very accurate, and the circuit is THEORETICALLY unsatisfactory, although in practice perfectly adequate, because the offset voltage is defined as that voltage producing zero output and this circuit actually produces a nonzero output.

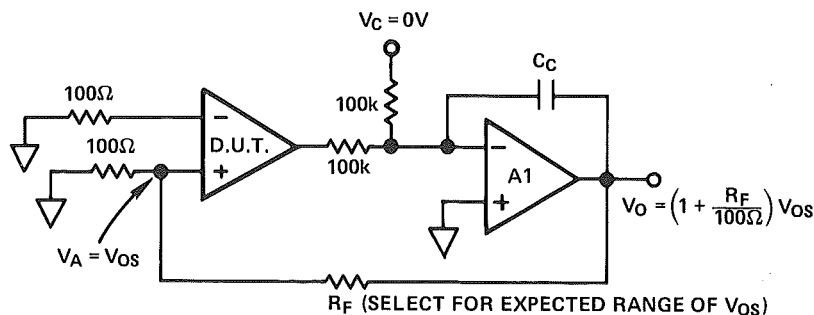
SIMPLE V_{OS} TEST CIRCUIT



A circuit which is not only theoretically satisfactory but may also be used to measure a number of other op amp parameters is illustrated in the next diagram. It consists of a servo loop built around the device under test (DUT) to determine its output voltage. A second amplifier (with a large feedback capacitor to ensure that it has negligible HF response) is used to provide feedback to force the output of the DUT to be equal in magnitude and opposite in sign to the voltage (V_C) applied to the control input. This feedback amplifier must have high gain, low offset voltage and low bias current but does NOT need to have very good performance, and certainly does not need to have specifications which are better than those of the DUT.

When measuring offset voltage the control voltage, V_C , is set to zero. This forces the output of the DUT to zero as well. Since the output of the DUT will go to zero only when a voltage equal to its input offset voltage is applied to its input, V_A must equal V_{OS} . Thus the output of the feedback amplifier is equal to $V_{OS} (1 + R_F/100)$ and the offset voltage may be calculated from the feedback amplifier output voltage.

OP AMP OFFSET VOLTAGE TEST CIRCUIT

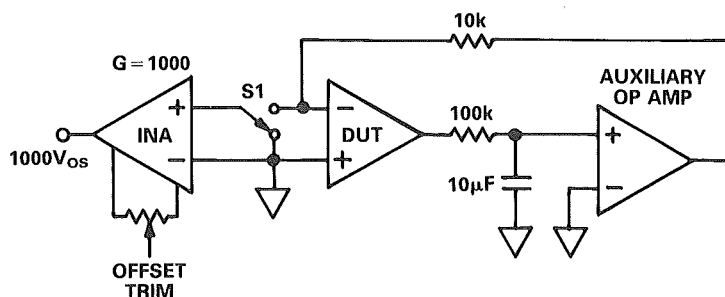


Transimpedance Amp Offset Voltage

The transimpedance amp, like the op amp, has a dc error term known as offset voltage which was shown in the transimpedance amplifier model as a voltage source in series with the inverting input terminal. This offset voltage is indistinguishable from an external voltage and introduces errors since it causes additional current to flow in the input of the transimpedance amplifier resulting in an error in the output voltage.

Since the transimpedance amplifier has a very low input impedance it is not possible to use exactly the same measurement system as is used for op amps but a similar arrangement will work. The transimpedance amplifier under test is connected to an auxiliary amplifier as before but feedback is taken to the inverting input of the DUT and therefore the auxiliary amplifier must be used in the noninverting mode. Since the input impedance of the DUT is very low a measurement at the output of the auxiliary amplifier does not tell us anything about the offset voltage at the DUT. That is measured with an instrumentation amplifier set to a gain of 1000. Any offset voltage in the instrumentation amplifier will affect the accuracy of the measurement. The in-amp offset must be set to zero by either automatic or normal procedures.

TRANSIMPEDANCE AMPLIFIER V_{OS} TEST CIRCUIT



Op Amp Input Bias Current

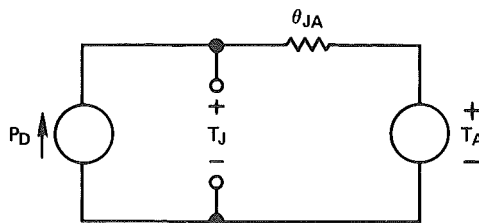
Another dc parameter of op amps is input bias current (I_B). If an op amp uses bipolar transistors in its input stage base current must be supplied from somewhere to bias them into their active operating region. Since Kirchoff's Law tells us that current must flow in circles this current must also return to its origin through a dc path. Thus operational amplifiers cannot be used with input signal sources which are not referred to the same power source as the amplifier itself. Although FETs do not require base current they nevertheless have a leakage current from their gate junction diode which results in an input bias current.

In many applications, the errors due to bias currents are actually less than the errors caused by the mismatch of the bias currents on the two inputs. This difference between the bias currents is called the input offset current, and is usually specified along with the bias current.

Like the input offset voltage, bias currents also vary with temperature. In an amplifier with a bipolar input stage the bias current decreases with increasing temperature because the transistors' β increases, and since their emitter current remains constant, the base current decreases. In the case of FET input amplifiers where the bias current is the gate leakage current of an FET (which is the leakage current of a reverse-biased junction diode). Such leakage currents double for every 10°C rise in junction temperature.

It is important to consider the test conditions under which bias current is specified, particularly in the case of a FET input op amp. Some manufacturers specify bias current at a junction temperature of 25°C . This corresponds roughly to the bias current immediately after power is applied to the amplifier. Unfortunately most circuits are not operated in a pulsed mode, and the effects of amplifier self-heating must be considered. This effect can be far from trivial. For example, an amplifier which draws 5 milliamps of supply current from ± 15 volt supplies dissipates 150 milliwatts. The thermal resistance from junction to ambient for 8-lead IC packages is typically $150^\circ\text{C}/\text{W}$ which means that after warm-up in an ambient of 25°C the junction temperature of the amplifier in question will be 22.5°C higher or 47.5°C . If the bias current was specified as (say) 50pA at 25°C junction temperature it will be over 200pA after warm-up. For this reason Analog Devices makes a practice of specifying the bias current of FET input op amps for a given AMBIENT temperature rather than a particular junction temperature. During production testing the device temperature is allowed to stabilize for about 10 seconds after the application of power before the bias current is measured.

THERMAL CIRCUIT MODEL FOR IC OP AMP



$$\begin{aligned}\theta_{JA} &= 150^\circ\text{C}/\text{W} \text{ FOR TO-99 TYPE PACKAGES} \\ &= 155^\circ\text{C}/\text{W} \text{ FOR EPOXY MINI-DIP} \\ &= 100^\circ\text{C}/\text{W} \text{ FOR 14/16 PIN CERAMIC DIP}\end{aligned}$$

CONSIDER AN OP AMP SPECIFIED FOR 50pA I_B AT $T_J = 25^\circ\text{C}$

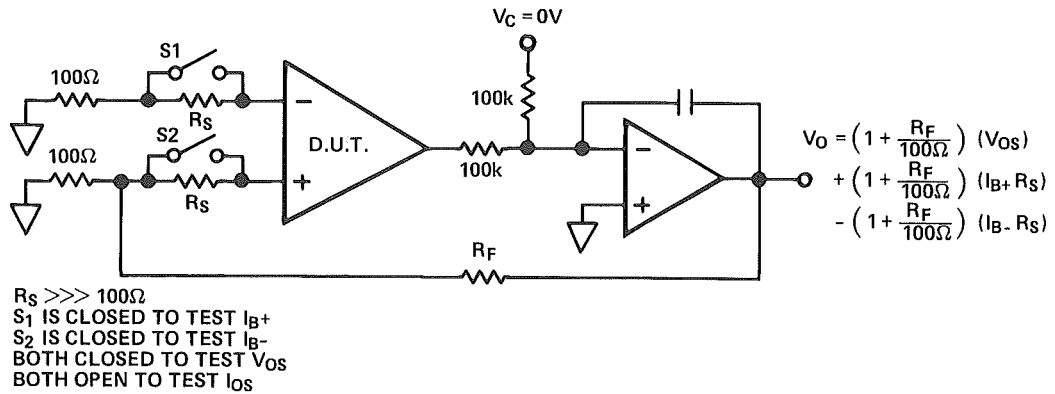
If the Amplifier Draws 5mA from $\pm 15\text{V}$,

$$\begin{aligned}P_D &= 30 \times 5 = 150\text{mW} \\ T_J &= T_A + (150\text{mW} \times 150^\circ\text{C}/\text{W}) \\ &= 25^\circ\text{C} + 22.5^\circ\text{C}\end{aligned}$$

Therefore, I_B will be Four Times Higher than Spec.

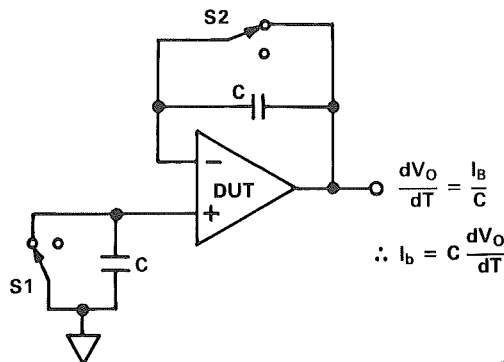
Bias current may be measured by essentially the same method as is used to measure offset voltage. A large resistance is inserted in series with the input under test, creating an apparent additional offset voltage equal to $I_B \cdot R_S$. If the actual V_{OS} has been measured and recorded the change in apparent V_{OS} due to the change in R_S can be determined and I_B is easily computed. The offset current is obtained by taking the difference between the bias current on the inverting input and the bias current on the noninverting input. Typical R_S values vary from $100k\Omega$ for bipolar amplifiers to $10^{12}\Omega$ for some FET input types.

BIAS/OFFSET CURRENT TEST CIRCUIT



Very low bias currents may also be measured by integration techniques. The current is used to charge a capacitor and the rate of charge is measured. If the capacitor and general circuit leakage is negligible (and this is very difficult to ensure when currents of under $10fA$ are to be measured) the current may be calculated directly from the rate of change of output of the test circuit when the switch which short-circuits the capacitor is opened.

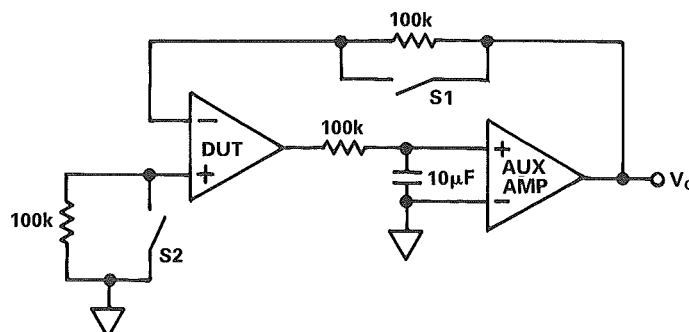
MEASUREMENT OF VERY LOW BIAS CURRENT



Transimpedance Amp Input Current

Unlike the op amp, the transimpedance amplifier does not have an input bias or offset current. Instead an input current is specified. This input current is a measure of how well the input stages and associated circuitry are matched and balanced. This input current is not really a bias current, since an ideal transimpedance amplifier, unlike an op amp, does not require an external current to function correctly.

TRANSIMPEDANCE AMPLIFIER INPUT CURRENT TEST CIRCUIT



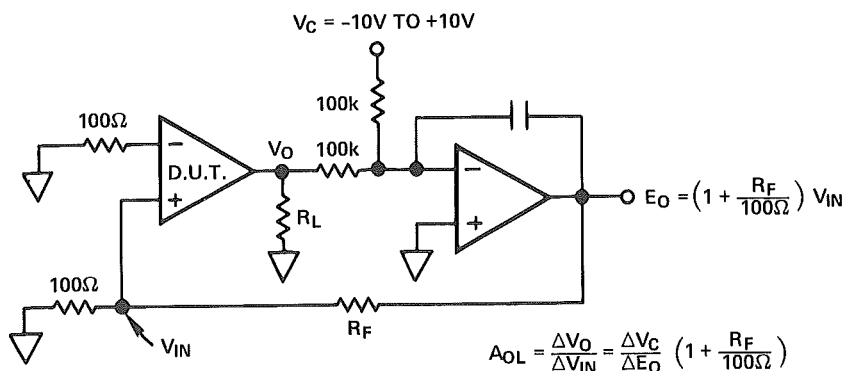
Nevertheless the circuitry used to measure this non-bias current is very similar to the circuitry used to measure op amp bias currents except that, as in the offset measurement circuitry, the feedback is taken to the inverting input of the transimpedance amplifier and the auxiliary amplifier is used in the noninverting mode. The input currents are allowed to flow into resistors (R_3) which are short-circuited by switches. By measuring the output of the auxiliary amplifier with the switches open and closed the voltage drop in the resistors may be measured and the input currents calculated.

Op Amp Open Loop Voltage Gain

Another op amp parameter which distinguishes a real amplifier from an ideal amplifier is open-loop gain. The open loop gain of an ideal op amp is assumed to be infinite. The same assumption is occasionally made of real amplifiers, with unfortunate results, as will be demonstrated later in this chapter. (Instrumentation and isolation amplifiers, on the other hand, have gains which are usually between 1 and 1000 and are very accurately defined.)

The open loop gain of an operational amplifier is an interesting parameter to measure. It is not generally practical to measure open loop gain directly by applying a signal at the input and observing the output change. Op amps generally have around 20 volts of output swing and gains of over one million—the input would therefore need to be of the order of one microvolt and it is very hard to handle such signals without unacceptable errors due to thermoelectric potentials). However, by using the same feedback system that we used to measure V_{OS} and I_B it is possible to measure the change in input voltage required to produce a known change in output voltage.

OPEN LOOP GAIN TEST CIRCUIT

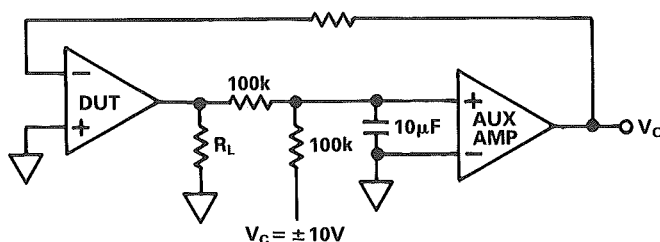


In this circuit, the control voltage, V_C is varied from $-10V$ to $+10V$, causing the DUT output, V_O , to vary from $+10V$ to $-10V$. The DUT output is varied by the change in V_{IN} produced by negative feedback through the second amplifier. V_{IN} is attenuated from E_O by the $R_F/100$ voltage divider. E_O is easily measured, and hence the open loop gain can readily be computed. This circuit works perfectly well for instrumentation and isolation amplifiers but it is not necessary to use it with them since their lower gain does allow the use of the simpler technique of measuring the input and output voltages.

Open Loop Transimpedance

The open loop transimpedance of an ideal transimpedance amplifier is infinite. Where such an amplifier has a finite transimpedance this is specified in ohms (volts per ampere). Practical monolithic transimpedance amplifiers have dc transimpedance of the order of 300 million volts per ampere (300 megohms). The sheer magnitude of this transimpedance imposes restrictions on the techniques used to measure it. In principle all that is required is to inject a known current into the input, measure the resulting output voltage and compute the transimpedance. In practice the smallness of the current which must be injected, combined with the value of the input current, makes such a method less than ideal and an electronic servo approach yields more satisfactory results.

OPEN LOOP TRANSIMPEDANCE TEST CIRCUIT



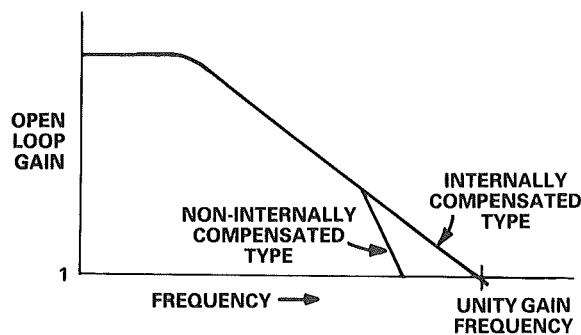
Again the circuitry is similar to the op-amp circuitry but uses the inverting input of the DUT and the auxiliary amplifier in a noninverting mode. Otherwise the procedure is the same as is used for op amps.

As with most amplifier parameters the transimpedance varies with temperature. Under normal operation of the transimpedance amplifier open loop gain variation does not significantly affect circuit performance, because the required system gains are controlled by precision resistors used as feedback elements.

Frequency Response

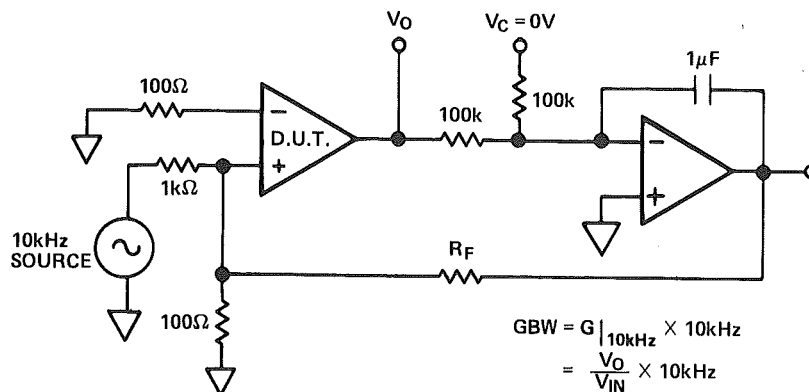
Most operational amplifiers have a very simple frequency response. The gain is constant at dc and very low frequencies and then has a single-pole rolloff, falling at 6dB/octave (-20dB/decade). It is obvious that throughout the region where this single pole frequency response applies the product of gain and frequency is a constant—this constant is known as the “gain-bandwidth product” of the amplifier and is a measure of its high frequency performance. In the majority of op amps this single pole response continues past the point where the gain has dropped to unity (such amplifiers are known as “internally compensated” or “unity gain stable” amplifiers). Some amplifiers have a more complex response and at a gain of something less than 10 a second pole appears. These amplifiers are not stable at low closed loop gains but generally have better high frequency performance than internally compensated types.

OPERATIONAL AMPLIFIER GAIN/ FREQUENCY PLOT



Since all op amps have a single pole frequency response at gains of more than 10 it is not usual to measure gain at a number of frequencies. It is evident that if the gain is measured at dc and one frequency the response of the amplifier throughout the region where it has a single pole response may be reduced from the gain-bandwidth product.

GAIN-BANDWIDTH PRODUCT TEST CIRCUIT



In the circuit shown the dc output of the DUT is held at 0 volts by V_C and the integrator amplifier, and a low amplitude ac signal is applied to the input of the DUT. The frequency of this ac signal is chosen so that the open loop gain is somewhere between 20 and 1000, since lower gain might result in complications from any second pole in the DUT's response and higher gain would necessitate too small an input signal. For most op amps 10kHz is a very suitable frequency. Since the integrator has very low gain at 10kHz there is no ac feedback and the DUT is effectively running open-loop for the ac signal. The ac output from the DUT can be measured and the gain at 10kHz can be calculated. [For example, an AD741 amplifier has an open loop gain of approximately 100 at 10kHz and an easily generated 100mV input will produce an easily measured 10V output. This corresponds to a 1MHz gain-bandwidth product.]

Slew Rate

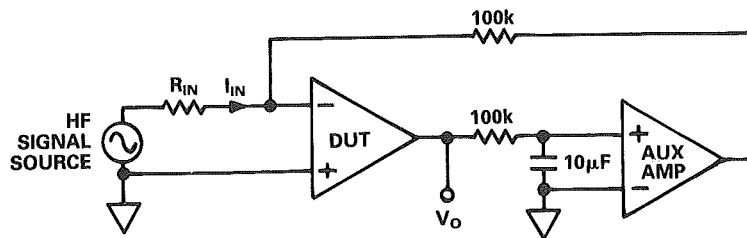
The same test circuit may be used to measure the slew rate of the DUT. The slew rate of an amplifier is defined as the rate at which the output voltage can change when high drive is applied to the input. When the circuit is used to measure slew rate the input signal is a fast square wave of sufficient amplitude to drive the output of the DUT to saturation. As oscilloscope is used to observe the slew rate of the DUT output.

Open Loop Transimpedance Amplifier Frequency Response

Like the operational amplifier, the transimpedance amplifier generally has a single pole frequency response and the product of its gain and frequency is a constant "gain bandwidth product". The gain need be measured at dc and only one frequency in order to calculate the overall response. The transimpedance amplifier does not have infinite bandwidth but, as will be shown later, it can be configured as a voltage amplifier whose bandwidth is not dependent on closed loop gain as is the case with the dominant pole compensated operational amplifier.

Once again the test circuitry is similar to that used for the op amp except that, again, the feedback is taken to the inverting input of the transimpedance amplifier and the auxiliary amplifier is used in the noninverting mode. The RC filter between the DUT and the auxiliary amplifier prevents any feedback at the test frequency. The signal (which is a current, not a voltage) is applied to the inverting input of the DUT. The amplified signal is measured at the DUT output and the transimpedance calculated.

TRANSIMPEDANCE AMPLIFIER H.F. OPEN LOOP GAIN MEASUREMENT

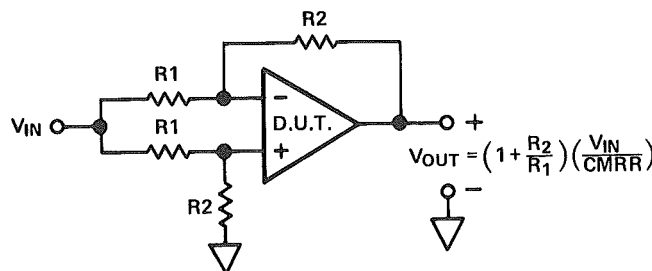


Common-Mode Rejection Ratio (CMRR)

The ideal operational amplifier has only differential gain and is insensitive to the absolute voltage on the inputs. A real amplifier has several nonideal characteristics associated with input levels. First of all the range of input voltage is limited. Few IC op amps will operate when the voltages on the input terminals are outside the supply voltages. The second, and perhaps more subtle, characteristic is the common-mode rejection ratio (CMRR). The CMRR is defined as the ratio of a change in common-mode voltage to the change in differential input voltage which would produce the same change in output. It is often convenient to specify this parameter in dB.

The common-mode rejection ratio can be measured in several ways. One method uses four precision resistors to configure the op amp as a subtractor amplifier, a signal is applied to both inputs and the change in output is measured—an amplifier with infinite CMRR would have not change in output. The disadvantage inherent in this circuit is that the ratio match of the resistors is as important as the CMRR of the op amp in determining the CMRR of the whole system. A mismatch of 0.1% between resistor pairs will result a CMRR of only 60dB, no matter how good the op amp. Since most op amps have CMRRs of between 80 and 120dB it is clear that this circuit is only marginally useful for measuring CMRR (although it does an excellent job in measuring its own resistance match!).

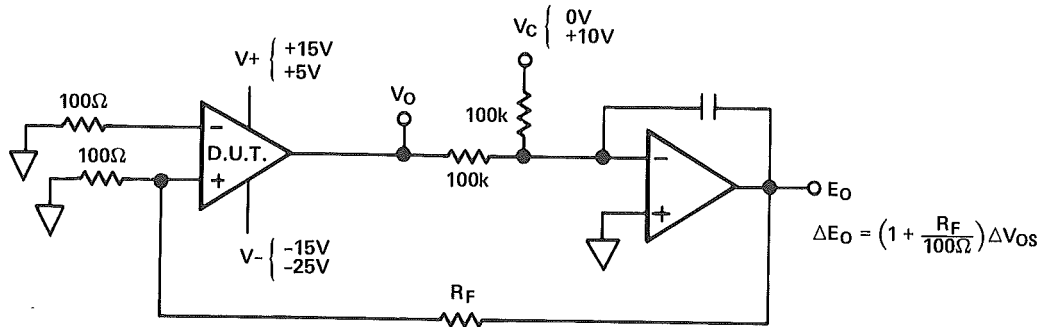
SIMPLE CMR TEST CIRCUIT



RESISTORS MUST MATCH WITHIN 1ppm (0.0001%)
IN ORDER TO MEASURE CMR > 100dB

The circuit used for the measurement of offset, bias and gain will, with slight modification, measure CMRR without requiring accurately matched resistors. The circuit is used to measure the offset voltage of the DUT when it is operating with $\pm 15\text{V}$ supplies (referred to the overall system). Its output, is centered between those supplies by setting V_C to 0V . The measurement is repeated with the DUT operating with $+5\text{V}$ and -25V supplies and its output still centered between them (at -10V) by setting V_C to $+10\text{V}$. This change is equivalent to applying $+10\text{V}$ common-mode signal at the input. The CMRR is the ratio of the 10V common-mode signal to the change in offset voltage.

COMMON-MODE REJECTION TEST CIRCUIT



For example if this 10V change in common-mode voltage creates a 1mV change in V_{Os} , the CMRR is $10\text{V}/1\text{mV}$ ($= 10000$ or 80dB)

Transimpedance Amplifier Common-Mode Rejection

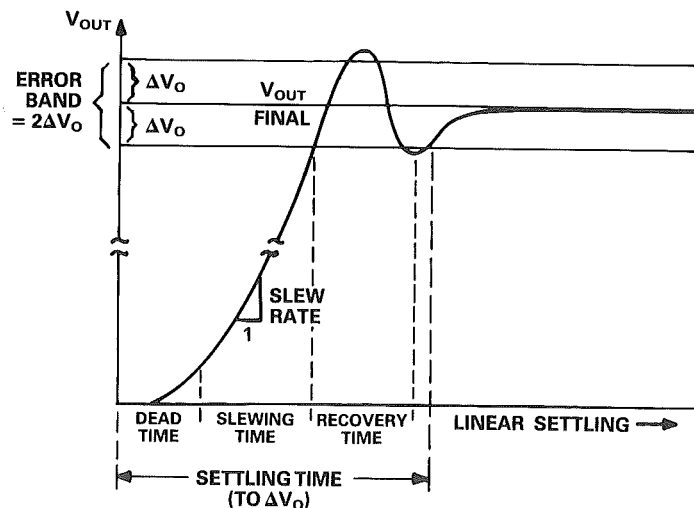
The transimpedance amplifier is designed to be used in the inverting mode and as such the CMRR term associated with op amps is not applicable. The noninverting terminal is grounded and should not be considered as a signal input.

Settling Time

Settling time is an important op amp specification especially when the amplifier must handle rapidly changing signals. In applications such as multiplexers, sample/hold amplifiers (SHAs) and amplifiers used with A/D and D/A converters it is the amplifier settling time which often determines the maximum data rate for a specified accuracy.

Settling time is defined as the time which elapses between the application of a step input to the time at which the amplifier output enters, and remains within, a specified error band symmetrical about the final output value.

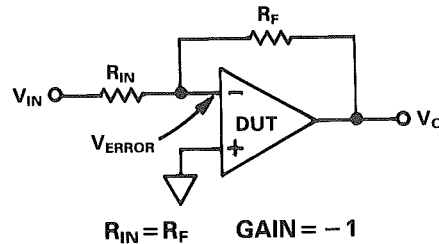
AMPLIFIER SETTLING TIME



Settling time is determined by both the linear and the nonlinear characteristics of the amplifier. It varies with the input signal level and is greatly affected by impedances external to the amplifier. For these reasons extrapolation of settling times from one set of operating conditions to another becomes virtually impossible. Settling time cannot be predicted from open loop specifications, such as slew rate or small signal bandwidth, as it is a closed loop parameter. The best way to know how fast an amplifier will settle in a particular application is to measure it.

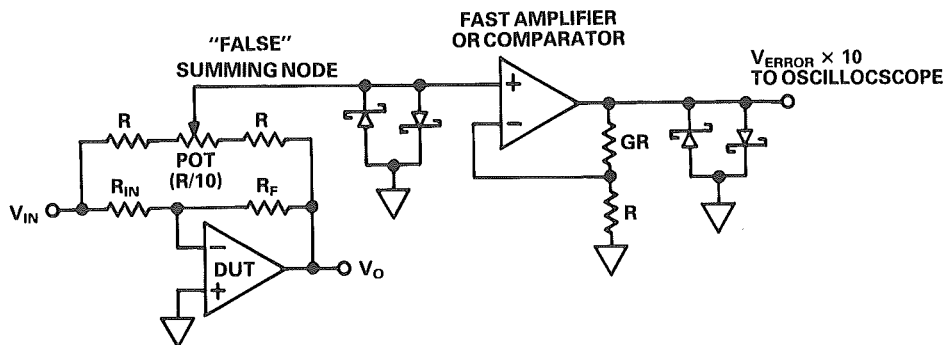
In many test configurations the amplifier is measured in the unity gain inverting mode. An error voltage is present at the inverting input which is generated by the input voltage being subtracted from the output voltage by means of two resistors.

SIMPLE SETTLING TIME TEST CIRCUIT



A transition in the input test voltage will cause large spikes in the error voltage. However a measurement cannot be made directly at this summing junction without influencing the result of the measurement but a second "false" summing junction may be formed by the use of two more resistors.

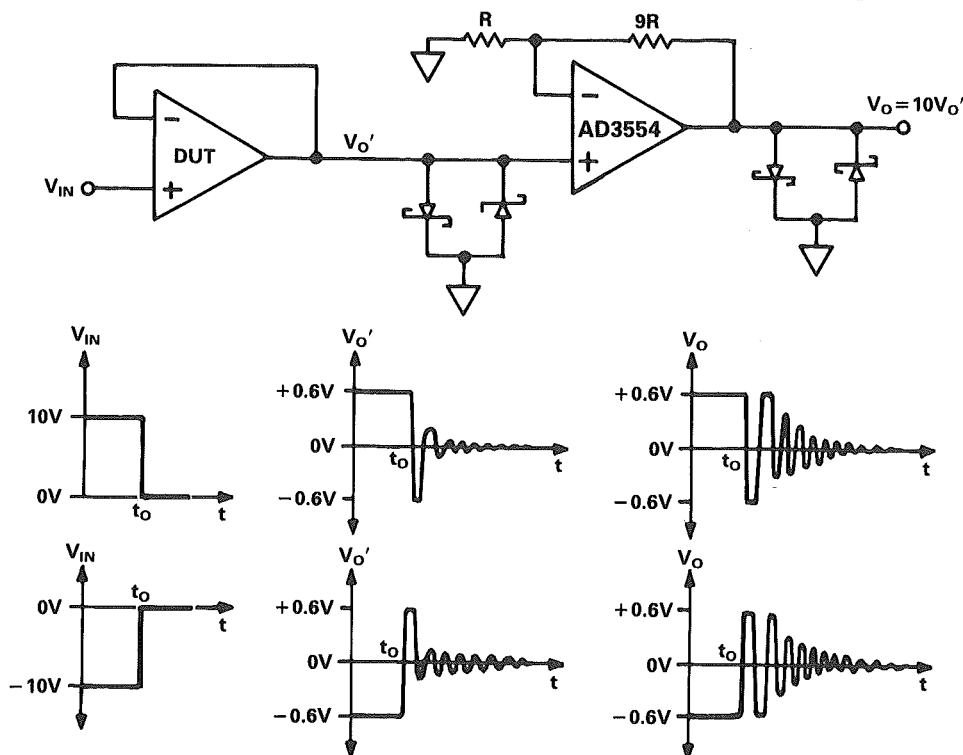
PRACTICAL SETTLING TIME TEST CIRCUIT WITH "FALSE" SUMMING POINT



These resistors form a Wheatstone bridge with resistors R_{IN} and R_F which may be balanced with a potentiometer. The voltage at the artificial junction follows the voltage at the summing node. A comparator is used to observe this error voltage. It usually has gain, and it may also settle more slowly than the amplifier under test, since comparator settling time introduces only second-order errors into the measurement. Clamping minimizes overdrive of both the comparator and the monitoring oscilloscope.

Measuring the settling time of amplifiers in the noninverting mode must be approached differently. As there is no error junction it is unreasonable to attempt to form a false one. The most common noninverting configuration to be tested is the unity gain follower which requires direct comparison of the input step and the output signal as they swing through the entire common-mode range. In this case the comparator measures the difference directly rather than observing an error point. It is subjected to wider voltage swings and must settle at least 2 or 3 times faster than the DUT.

ALTERNATIVE SETTling TIME TEST CIRCUIT



An alternative approach is to limit the input of the amplifier to a ten volt transition with a final value of zero volts. This allows direct analysis of the amplifier's output without overdriving the scope.

NOISE

In addition to noise present on the input signal there are two types of noise in circuits using op amps: external interference and the inherent noise of the circuit itself. Interference is noise which originates from sources not related to the actual circuit. Such noise sources include ground and power-supply noise created by other circuitry in a system, stray electromagnetic pickup of line frequency energy (and the harmonics thereof) and radio and radar transmissions, contact arcing in mechanical switches and relays, and transients due to switching in reactive circuits. Even mechanical vibration can create noise in high impedance amplifier circuits, either by piezo-electric pickup due to the use of piezo-electric plastic materials in cables or circuit boards, or by capacitance variation as the circuit vibrates. External interference can often be eliminated once the interfering source is identified and appropriate action taken.

The second type of noise is the inherent noise of the circuit itself. Unlike interference, it cannot be totally eliminated since it is caused by components in the actual circuit such as resistors and sources within the amplifier. The best that can be accomplished is to minimize the noise in a specific bandwidth of interest. To do this effectively it is important to construct reasonably accurate models of the noise sources. In op amps there are a current noise source in each input and a voltage noise source which appears differentially. These three noise sources are uncorrelated (although the two current noise sources have similar rms amplitudes), which is to say that there is no connection between them and their instantaneous outputs are not related to each other in any way. When two or more uncorrelated noise sources are to be added together it is done by taking the square root of the sum of their squares. It follows that, for a small number of noise sources, if the effect of any noise source is more than three or four times larger than that of any other only the largest need be considered when analyzing the noise of the system.

Voltage and Current Noise Model

A considerable amount of research at Analog Devices and elsewhere over more than fifteen years has demonstrated that the noise sources in op amps really are uncorrelated and are either true Gaussian or white noise sources or differ from true white noise sources by such a small amount that the difference is insignificant for all practical purposes. There are also sources with a "pink" noise spectrum i.e., the noise power density is inversely proportional to the square root of the frequency. The same research has demonstrated that they are true pink noise sources within the present limits of experimental accuracy. We can therefore consider the noise sources in op amps in terms of their statistical properties: rms value, peak value, and frequency content. Measurement of the rms

value of a random noise waveform requires an averaging interval which is long compared to the period of the lowest frequency of interest. The rms noise value is a useful and meaningful way to characterize the amount of noise generated by an amplifier.

$$E_n^2 = \int_0^T e_n^2(t) dt$$

Where E_n (rms) is the rms Noise Voltage Value
 T is the Interval of Observation
 e_n is the Instantaneous Noise Voltage

Low frequency noise is difficult to measure, since very long observation intervals are needed. In some cases low frequency noise is measured by direct observation on a storage oscilloscope screen using very slow sweep periods.

It is common to specify the peak-to-peak noise of an op amp. Although this makes system noise analysis much simpler (the argument goes: if the noise never exceeds X value the system perturbation never exceeds KX and all will be well) it is dangerous unless the statistical nature of noise is clearly understood. For Gaussian noise and a given rms noise value statistics tell us that the chance of a particular peak-to-peak value being exceeded decreases sharply as that value increases—BUT THIS PROBABILITY NEVER BECOMES ZERO. Thus for a given rms noise it is possible to predict the percentage of time that a given peak-to-peak value will be exceeded but it is NOT possible to give a peak-to-peak value which will never be exceeded.

RMS TO PEAK-TO-PEAK RATIOS

Nominal Peak-to-Peak	% of Time Noise will Exceed Nominal Peak-to-Peak Value
$2 \times \text{rms}$	32%
$3 \times \text{rms}$	13%
$4 \times \text{rms}$	4.6%
$5 \times \text{rms}$	1.2%
$6 \times \text{rms}$	0.27%
$6.6 \times \text{rms}$	0.10%
$7 \times \text{rms}$	0.046%
$8 \times \text{rms}$	0.006%

Peak-to-peak noise specifications, therefore, must always be written with a time limit. A suitable one is 6.6 times the rms value which is exceeded only 0.1% of the time. Since a manufacturer presented with such a specification will always assume Gaussian noise and measure the rms value, the op amp user is well advised to master the relationship between rms and peak values and then specify rms.

Types of Noise

Four types of noise are commonly encountered in operational amplifiers: popcorn noise, flicker noise, shot noise and Johnson noise.

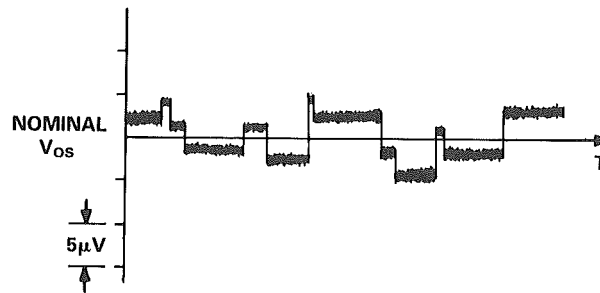
TYPES OF NOISE ON OP AMPS

- Popcorn Noise
- Flicker (1/f) Noise
- Schottky (Shot) Noise
- Johnson Noise

Popcorn Noise

When monolithic op amps were first introduced in the late 1960s popcorn noise was the dominant noise source. Whole advertising campaigns were fought to persuade users that Company X had lower popcorn noise than Company A, Company B and Company C. However, all of them A, B, C and X, were pretty bad. Today the causes of popcorn noise are quite well understood and production tests for it are reliable so that no reputable amplifier manufacturer has any difficulty in shipping products that are substantially free from popcorn noise. (Occasionally a digital circuit manufacturer will attempt to manufacture op amps on a TTL diffusion schedule in order to mop up spare plant capacity. These op amps will invariably suffer badly from popcorn noise since the very process steps which improve TTL speed and yield also promote popcorn noise.)

"POPCORN NOISE" CONSISTS OF RANDOM STEP CHANGES IN V_{OS} IN THE 10ms + TIMEFRAME



Popcorn noise is so-called because when played through an audio system it sounds like cooking popcorn. It consists of random step changes of offset voltage which take place at random intervals in the 10+ ms timeframe. Such noise results from high levels of contamination and crystal lattice dislocation at the surface of the silicon chip, which in turn results from inappropriate processing techniques or poor quality raw materials. It is quite easy to screen for popcorn noise during the manufacture of op amps so even if, for reasons beyond a manufacturer's control, devices should be produced with high levels of popcorn noise they should be detected and rejected at an early stage.

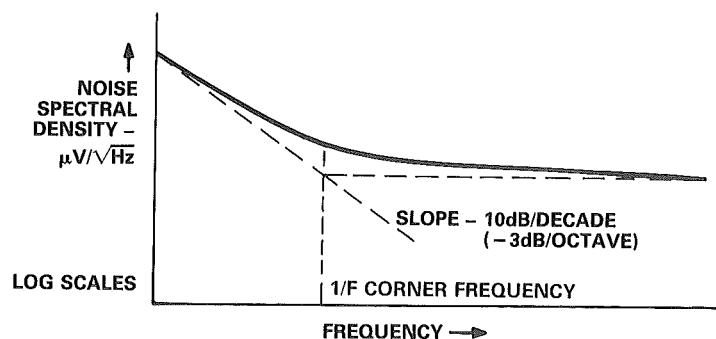
"Flicker" or "1/f" Noise

Flicker noise is the dominant noise at low frequencies. It has a power spectral density which is inversely proportional to frequency (hence the term "1/f Noise"). The noise voltage spectral density is therefore inversely proportional to the square root of the frequency.

The noise spectral density drops at 10dB/decade (-3dB/octave) with rising frequency. Noise with such a spectrum is called "pink" noise.

In early op amps (once popcorn noise had been conquered) flicker noise was the dominant noise source below frequencies as high as 1kHz. Today it is rarely significant above 50Hz and in devices such as the AD OP-27 it is insignificant above one or two Hz. The frequency at which the flicker noise characteristic of an amplifier intersects the white noise floor is known as the "1/f corner frequency" and is a figure of merit for the amplifier—the lower the better.

"FLICKER" OR "1/f" NOISE



Johnson Noise

Thermal excitation of the electrons in conductors causes random movement of charge. In a resistance this random current causes a noise voltage, known as Johnson Noise, whose amplitude is given by the formula:

JOHNSON NOISE VOLTAGE

$$E_N \text{ (rms)} = \sqrt{4kTRB}$$

Where: k = Boltzmann's Constant ($1.38 \times 10^{-23} \text{ J/}^\circ\text{K}$)

T = Temperature $^\circ\text{K}$

R = Resistance in Ohms

B = Bandwidth in Hertz

At Room Temperature (25°C) This May Be Simplified To

$$E_N \text{ (rms)} \approx \frac{1}{8} \sqrt{RB} \quad \text{or} \quad e_n \approx 4\sqrt{R}$$

E_N = Total Noise in $\mu\text{V rms}$

R = Resistance in Kilohms

B = Bandwidth in Kilohertz

e_n = Spectral Density in $\text{nV}/\sqrt{\text{Hz}}$

R = Resistance in Kilohms

Johnson noise is a fundamental property of resistances and must always be considered when designing low noise circuitry. There are only three ways in which it may be reduced: by reducing the temperature, the resistance itself or the working bandwidth.

Reducing the temperature is generally impractical. Since the function is a square root cooling a resistor from room temperature ($25^\circ\text{C}/298\text{K}$) to liquid nitrogen temperature ($-196^\circ\text{C}/77\text{K}$) will only reduce the noise by 42%. Reducing the resistance itself or the working bandwidth are generally more useful. As a reference point it is useful to remember that at room temperature a 1K resistor has $4\text{nV}/\sqrt{\text{Hz}}$ white noise. This is equivalent to 128nV rms noise in a 1kHz bandwidth.

Although Johnson noise is one of the causes of internal noise in an amplifier it is generally less important inside the amplifier than it is in the external resistors of the circuit. In many cases the noise from the external resistors is much higher than amplifier generated noise.

“Shot” (or Schottky) Noise

Current in a conductor consists of a flow of electrons each of which has a discrete charge. Current in a semiconductor consists of a flow of electrons or “holes” depending on whether it is an N-type or P-type semiconductor but the quantized nature of the flow is the same in either case. There are statistical variations in the rate of electron flow and these manifest themselves as a noise current.

SHOT NOISE

$$I_N = 5.7 \times 10^{-4} \sqrt{I_J B}$$

Where I_N is Noise Current in Picoamps rms

I_J is Junction Current in Picoamps

B is Bandwidth of Interest (in Hertz)

Since the electronic charge is extremely small the noise current is very small as well and is only significant when the bandwidth is very large or when the noise current is an appreciable fraction of the total current since noise current is proportional to the square root of the current the second case occurs only at very low currents. Thus shot noise is important at high frequencies and in amplifiers with very low bias currents but rarely elsewhere.

Noise Specifications

The data sheet of a high performance operational amplifier should contain some noise specifications. Noise testing is slow and time consuming and therefore expensive, so it is more normal for manufacturers to quote typical specifications rather than absolute maxima. Specifications may be quoted however, which are “guaranteed but not tested”. Amplifiers intended for low noise applications may be tested, and priced accordingly, and many manufacturers are prepared to test amplifiers for noise as a special at additional cost.

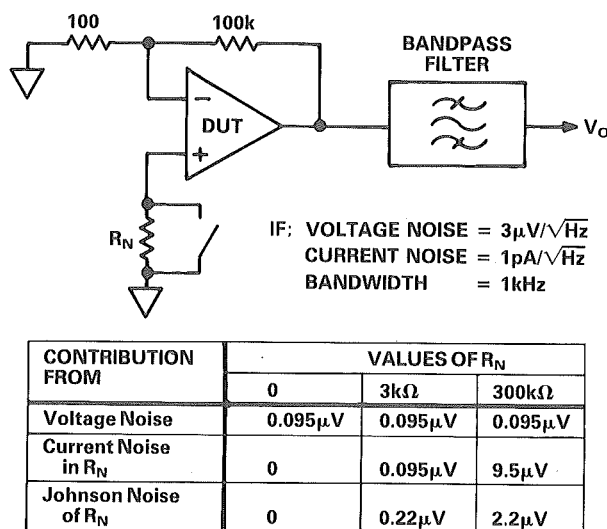
Noise specifications may be presented in a wide variety of ways. In order to understand and compare the noise specifications of devices from different manufacturers, or even those of different devices from the same manufacturer, it is necessary to understand the principles behind noise specifications so that the data available may be manipulated to its most useful form.

Early op amp specifications often referred to a “noise figure” expressed in dB when the op amp was used with a stated source impedance. This noise figure is the ratio of the actual noise in the circuit to the theoretical minimum noise of a perfect amplifier used in the same circuit. Although noise figure is still widely specified for RF amplifiers and discrete devices it is no longer commonly used for op amps.

Since there are three noise sources in an op amp (as mentioned above, the two current noise sources are of similar magnitude—though uncorrelated) an ideal data sheet would specify the magnitude and spectral composition of each. This is almost never done, simply because the amount of work involved in characterization, and in monitoring production devices, would be prohibitively expensive.

Although the voltage noise is always present and unavoidable, the current noise of an op amp is only amplified if it flows in a resistor and develops a noise voltage. Thus the voltage noise is dominant when an amplifier is driven from a low source resistance and the current noise is dominant when it is driven from a high source resistance. Depending on the actual values of voltage and current noise in the amplifier the Johnson noise of the source resistance itself may dominate both amplifier noise sources. This forms the basis of a simple noise measuring circuit. The DUT is configured as a voltage amplifier, its output passes through a bandpass filter and the output noise is measured both when the input is grounded directly and when it is grounded through a resistor. The resistor is chosen to be large enough that when the current noise of the amplifier flows in it the noise voltage developed is more than three times greater than either the amplifier voltage noise or its own Johnson noise. The first measurement gives the voltage noise of the amplifier, the second its current noise.

OPERATIONAL AMPLIFIER NOISE MEASUREMENT



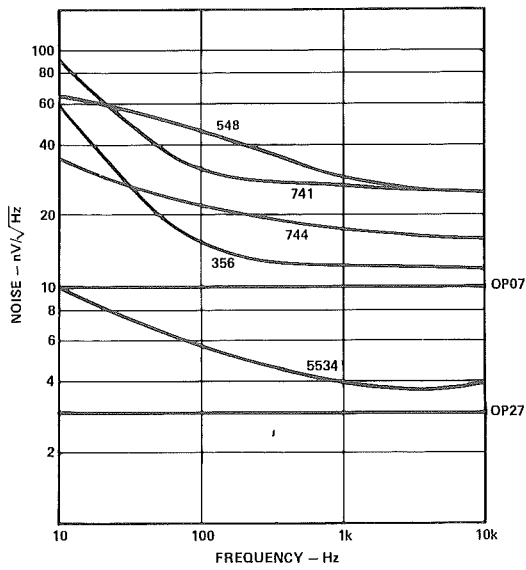
Where a single value is given for the noise of a device (e.g., Noise = 1 μV rms) it should be associated with a band of frequencies and is the total voltage noise of the device within that band. If no frequency information is given the figure should be for the noise in the band from dc to the unity gain frequency. This noise may be specified at a particular source resistance—the specified noise is then the rms sum of the voltage noise and the current noise flowing in that resistance.

A more useful noise specification consists of noise spectral density figure for mid frequencies and the 1/f corner frequency mentioned above. Although such a specification may not contain as much information as might be desired on the relative magnitudes of voltage and current noise it does provide a firm basis for comparisons between devices. Noise is sometimes specified in terms of a graph of total noise up to a given frequency. Such curves may be useful when the noise gain versus frequency characteristics of a circuit are known but it is important to consider both the voltage and current noise curves (and the gain to each with respect to frequency) in a computation of total noise.

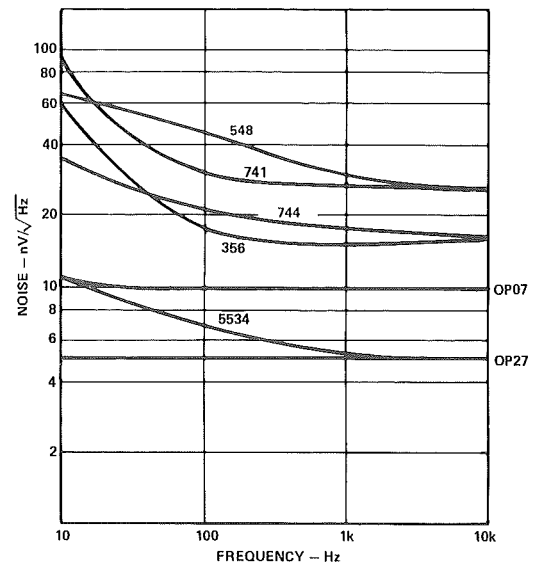
Yet another useful specification is a set of noise spectral density figures (at particular frequencies or plotted as a graph against frequency). The noise spectral density, e_n , at a given frequency is defined as the rms noise voltage in a 1Hz band surrounding a specified value of frequency. It is usually expressed in nanovolts per root hertz (nV/ $\sqrt{\text{Hz}}$). The awkward units arise from the fact that it is noise power which is actually being characterized and voltage is proportional to the square root of power.

The following family of curves demonstrates the noise performance of several common low noise amplifiers at several different values of input resistance. These curves take into account the amplifier voltage noise, the Johnson noise of the source resistance and the voltage noise generated by the input noise current flowing through the source resistance. The relative noise performance of each amplifier is critically dependent upon the source resistance.

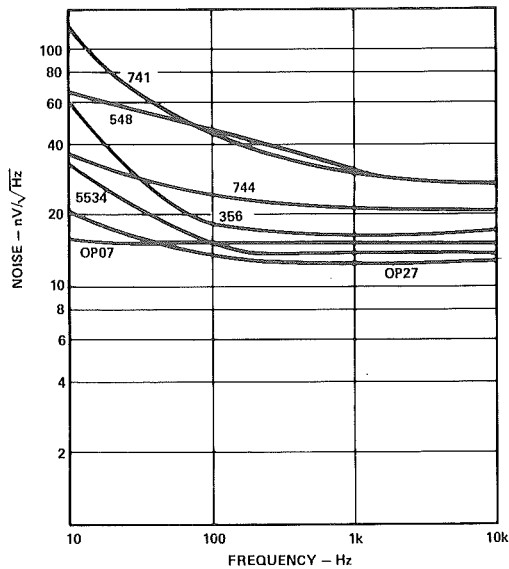
$R_S = 100\Omega$



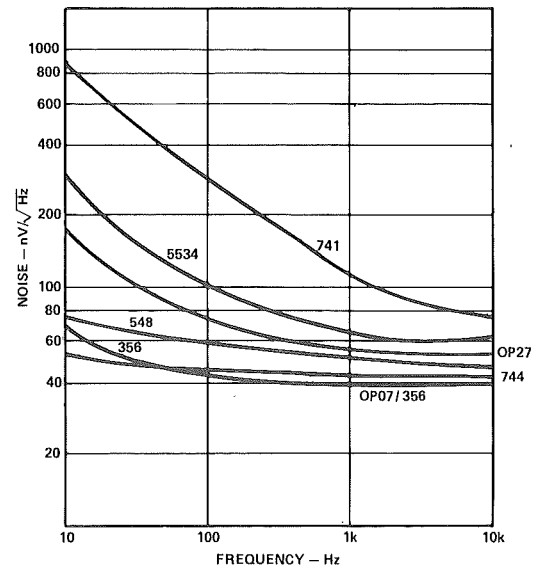
$R_S = 1k\Omega$



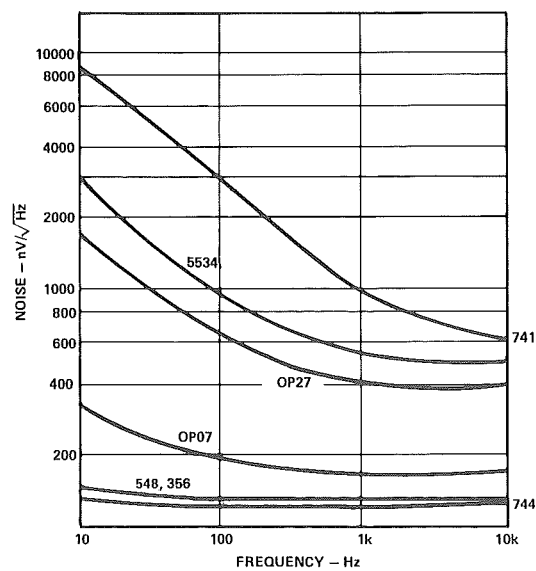
$R_S = 10k\Omega$



$R_S = 100k\Omega$



$R_S = 1M\Omega$



A series of spot noise figures at particular frequencies (10Hz, 100Hz, 1kHz, 10kHz) may be given instead of plots. These numbers are useful for direct comparison of noise characteristics of several amplifiers. Computation of total rms noise in a particular bandwidth requires that the shape of the spectral density curve be known. In the low frequency region, where flicker noise is dominant, the noise at a particular frequency can be computed from the formula.

$$E_n = k/f$$

where k is the value of E_n at $f = 1\text{Hz}$.

The total 1/f noise in a given bandwidth is given by

$$E_n = k \sqrt{\int_{f_1}^{f_2} \frac{df}{f}} = k \sqrt{\ln \frac{f_2}{f_1}}$$

The total flicker noise in a given band is a function of the ratio of the low and high band edge frequencies since the actual frequency cancels out. It is necessary, however, that the upper band edge is still in the 1/f region. The formula implies that a decade contributes 1.52K (or an octave 0.83K) where K is the value of E_n at 1Hz. For example consider an amplifier with $E_n = 100\text{nV}/\sqrt{\text{Hz}}$ at 1Hz. The total 1/f noise from 0.1 to 10Hz is therefore:

$$\begin{aligned} E_n &= 0.1\mu\text{V}/\sqrt{\text{Hz}} \times 1.52 \sqrt{\log \frac{10}{0.1}} \\ &= 0.1 \times 1.52 \times \sqrt{2} \\ &= 0.21\mu\text{V} \end{aligned}$$

For white noise the computation of total noise is also straightforward. The total rms noise in the band between f_1 and f_2 is given by:

$$E_n = \sqrt{\int_{f_1}^{f_2} e_n^2 df} = e_n \sqrt{f_2 - f_1}$$

If f_1 is less than 10% of f_2 , then ignoring f_1 (i.e., treating it as zero) will introduce less than 5% error. Thus the noise below 10kHz of an amplifier with $10\text{nV}/\sqrt{\text{Hz}}$ will be

$$E_n = 10\text{nV}/\sqrt{\text{Hz}} \times \sqrt{10^4 \text{ Hz}}$$

$$E_n = 1\mu\text{V rms}$$

In practice it is virtually impossible to measure noise within specific frequency limits with no contribution from outside those limits since practical filters have finite rolloff characteristics. Fortunately the measurement error induced by a single pole low pass filter is readily computed. The noise in the spectrum above the filter cutoff frequency extends the effective corner frequency to 1.57 times the original f_c and so if a single pole filter is used in a noise measurement circuit its corner frequency should be set to 0.66 of f_2 . Similarly, a 2 pole filter has an apparent corner frequency of $1.2f_c$.

The key points to remember about noise in op amps are:

SUMMARY OF OP AMP NOISE SPECS

1. Low Frequency Noise Generally Follows a "1/f" (Actually $1/\sqrt{f}$) Characteristic. Total rms Noise in a 1/f Region can be Computed from $E_n = 1.52k \sqrt{\log \frac{f_2}{f_1}}$ Where $k = e_n$ at 1Hz.
2. High Frequency Noise Generally Follows a White Noise Characteristic. Total rms White Noise is Computed from $E_n = e_n \sqrt{f_2 - f_1}$ (or Simply $e_n \sqrt{f_2}$, if $f_2 \geq 10f_1$).
3. The "1/f Corner Frequency" can Serve as a Figure of Merit for Comparing Amplifiers.
4. Broadband Noise Test Circuits Should Use Filter Cutoff Frequencies Set for 2/3 of the Desired Bandwidth, to Allow for the Slope of the Filter's Roll-Off.
5. Uncorrelated Noise Sources Add in a Root-Sum-of-Squares Fashion.

The specifications defined in this section are thus far only component specifications. In practical applications it is necessary to examine the importance of each specification relative to the actual application and determine the possible tradeoffs.

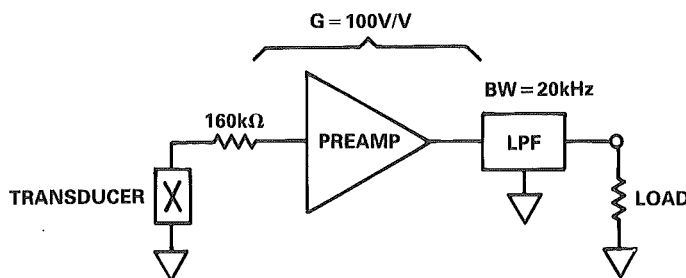
MAKING SENSE OUT OF NOISE

The sources of noise have been discussed at length, and although we have experienced the limitations imposed by noise (the proverbial background hiss) we have not yet shown quantitatively how the noise renders the input signal less useable. This however will be remedied through the use of the following example.

A well known Beethoven symphony is "Wellington's Victory". This symphony portrays in music Great Britain's famous general, Lord Wellington, commanding an army of 15,000 infantry and 3,000 cavalry, defeating Napoleon and his army at the Battle of Waterloo. The opening movement to the symphony begins with the distant sound of troops marching forward to engage the enemy in an impending battle. Steadily the music increases in intensity, as the troops move within battle range. Suddenly the sound erupts to enormous proportion as cannons boom and rifle shots crack. This absolutely beautiful symphony has to be a favorite, and the significance of this movement lies in the wide dynamic range of sound intensity and easily varies 80dB with peak sound pressure levels reaching 110dB. In the case of this symphony not only is the loudness of the cannons important, but of equal importance is the silence of death as this stillness permeates the battle field which was once filled by living soldiers and charging dragoons.

When this symphony is played from a recording, a limiting factor determining how quiet the music can sound is often the quality of the preamplifier used in the front end of the playback system. In any event this preamplifier should be designed to achieve the lowest possible noise reflected to the input. Generally, the preamplifier has a voltage gain of 40dB, an output signal level of 0.5V rms, a half power frequency (bandwidth) of 20kHz, and an output signal to noise ratio of 80dB. If the transducer driving the input to the preamplifier has an impedance equal to 160k Ω , what is the minimum signal level that must be available from the transducer to achieve the required preamplifier signal to noise output if the device selected is the AD OP-27?

SCHEMATIC OF TRANSDUCER AND AMPLIFIER



First compute the noise contributed by the source resistor:

RESISTOR NOISE VOLTAGE COMPUTATION

$$V_{\text{NOISE}} = \text{RESISTOR} = \sqrt{4 \times K \times T \times B \times R}$$

Where K is Boltzman's Constant = 1.38049×10^{-23} Joules/Kelvin

T is the Temperature in Kelvin

B is the Bandwidth in Hertz

R is the Resistor Value in Ohms

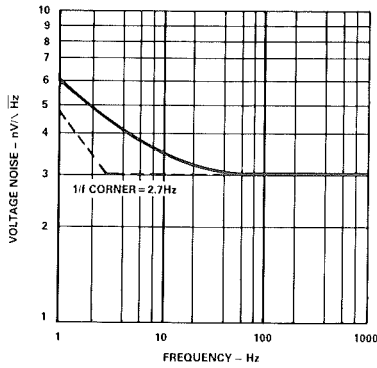
$$V_{\text{NOISE}} = \sqrt{4 \times 1.38049 \times 10^{-23} \times 298 \times 20 \times 10^3 \times 160 \times 10^3} = 7.256 \times 10^{-4} \text{ Volts rms}$$

$$0.162 \mu V$$

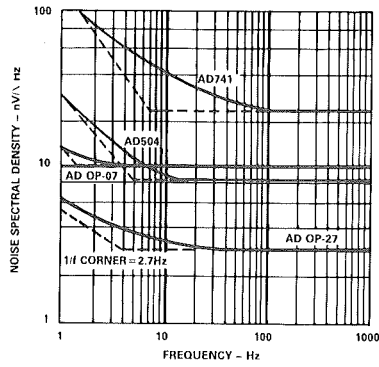
$$\sqrt{7.256 \times 10^{-4}}$$

Second compute the noise contributed by the AD OP-27:

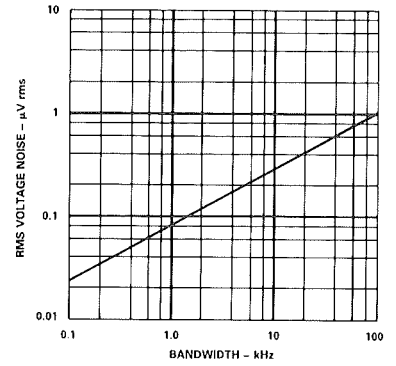
AD OP-27 NOISE CHARACTERISTICS



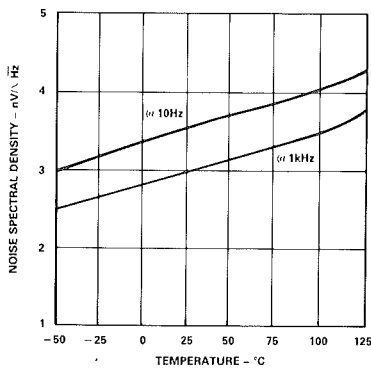
INPUT VOLTAGE NOISE SPECTRAL DENSITY



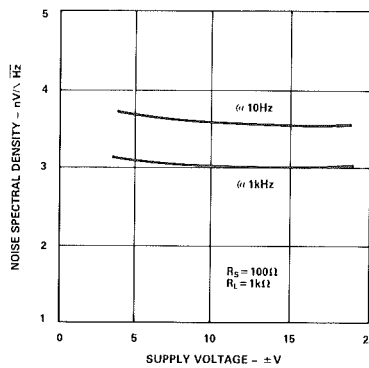
COMPARISON OF OP AMP INPUT VOLTAGE NOISE SPECTRUMS



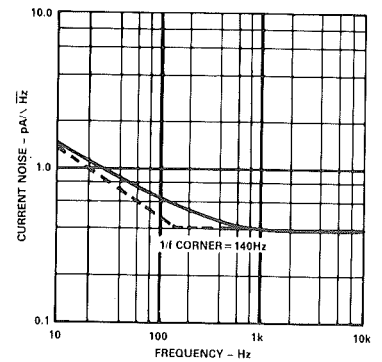
INPUT WIDEBAND NOISE VS. BANDWIDTH (0.1Hz TO FREQUENCY INDICATED)



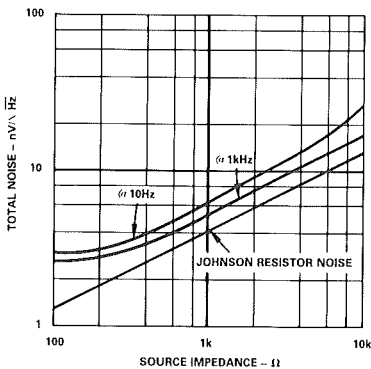
INPUT VOLTAGE NOISE VS. TEMPERATURE



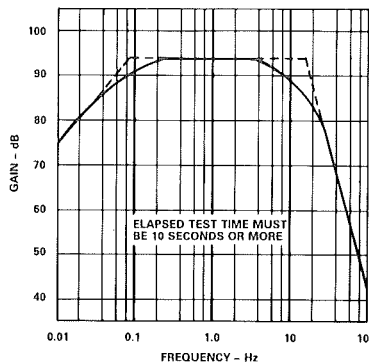
INPUT VOLTAGE NOISE VS. SUPPLY VOLTAGE



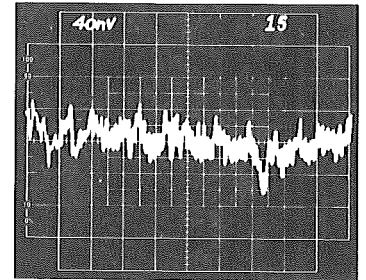
INPUT CURRENT NOISE SPECTRAL DENSITY



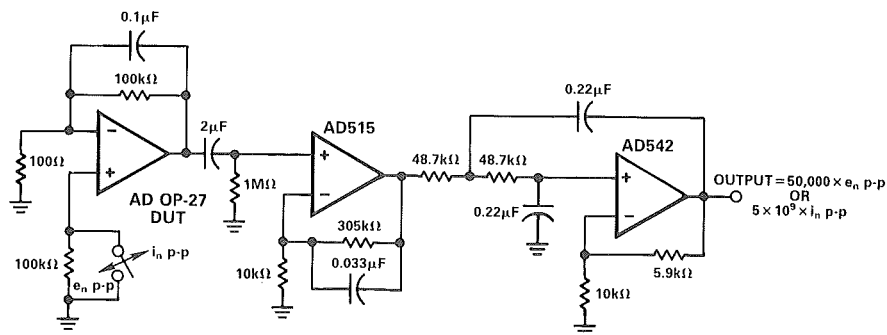
TOTAL NOISE VS. SOURCE IMPEDANCE



0.1Hz TO 10Hz NOISE TEST FREQUENCY RESPONSE



0.1Hz TO 10Hz P-P VOLTAGE NOISE

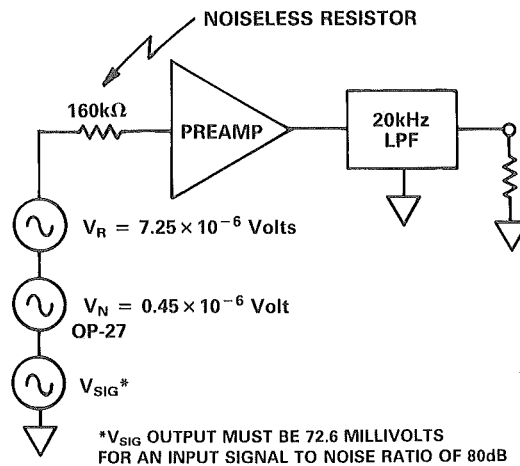


NOTE: ALL CAPACITORS MUST BE NONPOLARIZED

0.1Hz TO 10Hz NOISE TEST BANDPASS FILTER (VOLTAGE GAIN = 50,000)

The AD OP-27 data sheet reveals that the wideband noise of the AD OP-27 reflected to the input is 0.45 microvolts rms in a 20kHz bandwidth. The method in which the wideband noise is presented for the AD OP-27 makes the noise comparison easier, for this method gives a sum total regardless of the device specific contributing factor. Clearly in this example the AD OP-27 noise contribution is significantly less than the noise voltage generated by the transducer.

TRANSDUCER AND AMPLIFIER NOISE CONTRIBUTIONS



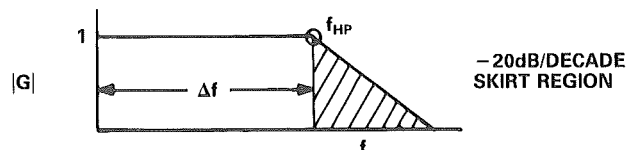
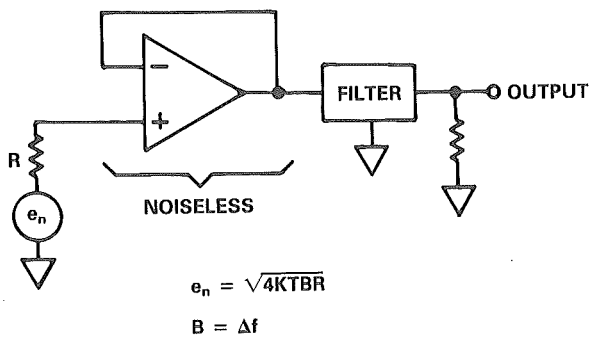
The two sources of noise are the transducer source resistance, and the AD OP-27. It is important to remember that these noise sources are not correlated and the sum contribution is equal to the root sum square of the quantities. This computation is simply the SQUARE ROOT of the sum of the squares.

ROOT SUM SQUARE CALCULATION

$$\begin{aligned}
 V_{NOISE} &= \sqrt{V_R^2 + V_N^2} \\
 &= \sqrt{(0.45 \times 10^{-6})^2 + (7.25 \times 10^{-6})^2} = \\
 &= 7.26 \times 10^{-6} \text{ Volts rms}
 \end{aligned}$$

For the preamp output to have an 80dB signal to noise ratio the peak input signal must be 80dB greater than the amplifier noise reflected to the input. Hence the transducer must be capable of producing a peak unloaded signal of 72.6 millivolts.

NOISE BANDWIDTH CORRECTION



$$G(f) = \frac{1}{1 + jf/f_{HP}}$$

$$f_{NOISE \text{ BANDWIDTH}} = f_n = \int_0^{\infty} |G(f)|^2 df = \int_0^{\infty} \left| \frac{1}{1 - j(f/f_{HP})} \right|^2 df$$

$$f_n = \int_0^{\infty} \frac{1}{(1 + f/f_{HP})^2} df = \frac{\pi}{2} f_{HP}$$

The 20kHz bandwidth used in the computation of the noise contribution assumed the low pass filter had an infinitely steep skirt response. However, for the mathematicians and purists the bandwidth is corrected by multiplying by $\pi/2$, or 1.57. The factor accounts for the noise contained in the skirt region of the filter whose asymptotic response is -20dB per decade.

Noise figure is another means of describing the noise characteristics of an amplifier, and can be represented mathematically as follows:

NOISE FIGURE EQUATIONS

Noise Factor = F

$$F = \frac{\text{Total available output noise power}}{\text{Noise power at output due to source resistance}}$$

OR

$$F = 1 + \frac{\text{Noise power output due to FET}}{\text{Noise power output due to source resistance}}$$

$$F = 1 + \frac{V_n^2 + I_n^2 R_s^2}{4KTR_s B}$$

R_s = Source Resistance

V_n = AD OP-27 Noise Voltage

I_n = AD OP-27 Noise Current

K = Boltzman's Constant

T = Temperature in Kelvin

B = Bandwidth

$$\text{Noise Figure} = 10 \log \text{Noise Factor} = 10 \log F$$

However, the noise figure can be misleading when applied to small signal amplifiers in the audio range. The misleading aspect arises from the fact that the noise figure of the amplifier in many cases is dependent upon the source resistance. To achieve the optimum noise figures in FETs, for example, it may be necessary to increase the source impedance to tens or megohms. The noise figure may have been improved at the expense of adding considerably more Johnson noise. This is also true to some extent in the design of microwave amplifiers, as there are optimum source impedances to produce the lowest noise figure. The optimum noise source impedance results in a lower gain design, but the minimum discernable signal is lower.

Therefore, it is suggested that whenever considering signals in the presence of noise, *reflect all noises sources to the input* to permit a signal to noise comparison on a one-on-one basis.

AMPLIFIER TECHNOLOGIES

It is useful for the design engineer to know not only the types of operational amplifier products available, but also the technologies used to produce them. He can then select an amplifier for a particular application from a suitable category of amplifiers and have some familiarity with its probable idiosyncrasies.

Standard Bipolar Process

The standard junction-isolated bipolar process used in the majority of op amps produces three basic transistors: a high quality vertical NPN transistor, a high quality vertical PNP transistor, and a somewhat lower quality lateral PNP transistor. The vertical PNP is of limited utility since its collector is always connected to the negative power supply. Thus the two transistors which can be used elsewhere in the amplifier circuit are the vertical NPN and the lateral PNP. The lateral PNP is very low performance (low β low f_t) and is used primarily in biasing circuitry. It is the NPN, therefore, which is used in the actual signal path as much as possible. An amplifier which uses standard bipolar transistors has bias currents generally in the 100nA to 1 μ A range, reasonably low offset voltage and drift, and low voltage noise. Examples of these amplifiers include the 741 and 301 types.

Super-Beta

Super-Beta processing is an addition to the standard bipolar process. With one additional diffusion step, an NPN transistor with a β of several thousand can be produced, reducing input bias current by an order of magnitude, down to 10nA or less. The increased gain of the input stage reduces input bias current and improves common-mode rejection, which are two of the more important specifications of precision amplifiers. Typical open loop gain of a super-beta op amp is several million, common-mode rejection is over 100dB, and input offset voltage characteristics are similar or superior to standard bipolar types. Examples of super-beta amplifiers include the 308, AD510 and AD517 types.

Dielectric-Isolated (DI) Bipolar

In conventional bipolar and super beta ICs, the individual transistors are isolated from each other by the use of reverse-biased p-n junctions. It is the parasitic capacitance of these junctions which limits the bandwidth of the lateral PNP transistors (and ultimately, the amplifier). The dielectric isolation (DI) process uses a thin oxide layer to provide the isolation between transistors. It then becomes possible to fabricate a high speed PNP transistor and therefore, a high speed amplifier.

The DI process is not without its limitations, however as the oxide layer is easily punctured by electrostatic discharge, resulting in device destruction. Another drawback is the fact the DI circuits require larger geometries than junction-isolated equivalent circuits, resulting in somewhat larger chip sizes.

Complementary Bipolar

The complementary bipolar process was developed so that both wideband NPN and PNP transistors could be fabricated on the same chip without the use of dielectric isolation. CB is capable of F_T 's of about 1GHz yielding amplifiers with gain bandwidth's of over 1GHz.

BIFET

The BIFET process uses ion-implantation to produce a JFET with high breakdown voltage on a chip also containing standard bipolar devices. A pair of these JFETs can be used as input devices for an op amp. The added performance, however, is gained at the expense of generally poorer offset voltage, drift, CMR, and noise specifications. Newer designs allow for factory trimming of BIFET op amp offset voltage and drift.

BI-MOS

Since JFETs can be used for high-impedance input stages, it is tempting to consider MOSFETs for the same application. Some manufacturers have developed processes which permit MOSFETs to be included on a bipolar IC. Rather than junction leakages inherent in JFETs, MOSFETs ideally have only oxide leakages, which are much lower, potentially reducing input bias current. MOSFETs, however, are ESD sensitive and require protection diodes on the inputs. Very often, these diodes exhibit leakages which are at least as high as the input bias current of a JFET-input amplifier. MOSFETs further tend to be much noisier than JFETs in the audio frequency spectrum, and dc offsets are difficult to control. When MOSFETs are used in the output stage of an op amp, it becomes possible to swing the output very close to the power supply rails. In a conventional bipolar output stage, output swing is limited by saturation voltages and other effects. It is important to note, however, that a MOSFET output stage must be lightly loaded to minimize the effects of R_{ON} .

CMOS

Amplifiers constructed entirely of MOSFETs are also available. These amplifiers exhibit poor performance if built along classical op amp designs. Newer designs use CMOS switches and external capacitors to provide offset voltage cancellation similar to the methods used in chopper-stabilized amplifiers. These designs suffer from high noise, poor output drive characteristics, and limited supply voltage ranges.

MONOLITHIC IC PROCESSES

STANDARD BIPOLAR

Advantages

- Many Suppliers
- Low Noise
- Low V_{OS} , V_{OS} Drift

Disadvantages

- High I_B
- Lateral PNP Limits Bandwidth

DIELECTRIC ISOLATED BIPOLAR

Advantages

- Allows Wide Bandwidth PNP
- Otherwise Similar to Standard Bipolar

Disadvantages

- Limited Number of Suppliers
- More Expensive
- ESD-Sensitive

BIFET

Advantages

- Allows High Breakdown JFET to be Fabricated on Bipolar IC
- High Input Impedance
- Low I_B

Disadvantages

- Additional Process Step
- Higher V_{OS} , V_{OS} Drift
- Poorer CMRR
- I_B Doubles Every 10°C Rise in Temperature

BI-MOS

Advantages

- Uses MOSFET for Potentially Very High Input Impedance
- Inputs can Swing to Supply Rail

Disadvantages

- Protection Diode Leakage Defeats Low MOS I_B
- High Noise

CMOS

Advantages

- Low Power
- Logic Circuitry and Switches can be Added
- Allows Fabrication of Monolithic Chopper-Stabilized Amplifier

Disadvantages

- High Noise
- Limited Output Drive Capability
- Limited Supply Voltage Range

COMPLEMENTARY BIPOLAR

Advantages

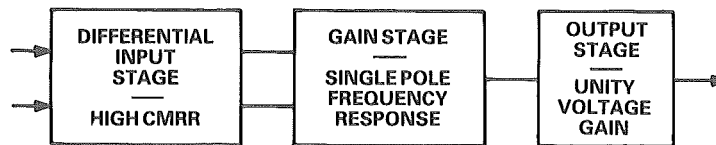
- Allows Wide Bandwidth PNP & NPN
- Very Fast Settling
- High Current Output
- Low Noise

Disadvantages

- Limited Supply Voltage Rating

There is an aspect of circuit technology which has an immediate impact on the choice of an op amp—the structure of the input stage. Almost all operational amplifiers have three stages: a differential input stage with good common-mode rejection, a high gain stage with a single pole frequency response which converts the differential output of the first stage to a single-ended signal to drive the output stage, and an output stage.

TYPICAL OPERATIONAL AMPLIFIER ARCHITECTURE



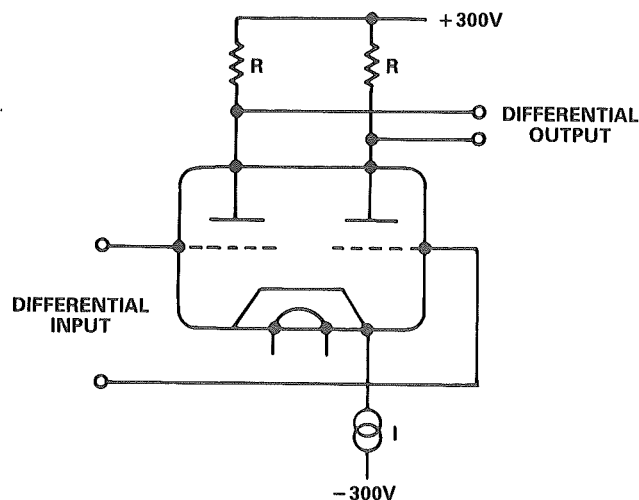
There are differences in the gain and output stages which are possible with different technologies but they are relatively unimportant. The choice of input stage technology is critical. There are three basic input stage technologies: simple bipolar, bias compensated bipolar, and field-effect transistor (FET).

CHARACTERISTICS OF OP AMP INPUT STAGES

	SIMPLE BIPOLAR	BIAS-COMPENSATED BIPOLAR	FET
OFFSET VOLTAGE	LOW	LOW	MEDIUM
OFFSET DRIFT	LOW	LOW	MEDIUM
BIAS CURRENT	HIGH	MEDIUM	LOW – VERY LOW
BIAS MATCH	EXCELLENT	POOR (CURRENT CAN BE IN OPPOSITE DIRECTIONS)	FAIR
BIAS/TEMP VARIATION	LOW	LOW	BIAS DOUBLES FOR EVERY 10°C RISE
NOISE	LOW	LOW	FAIR

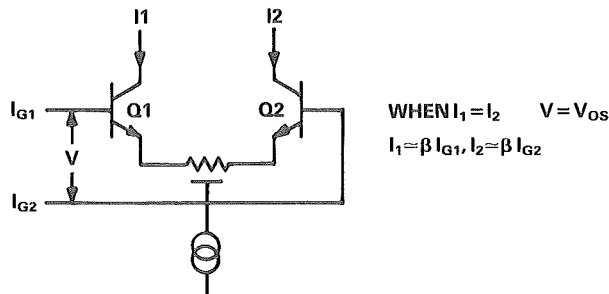
All three differential input stages use the basic “long-tailed pair” of amplifying devices. These design was originally developed with thermionic tubes but may be constructed with any amplifying device: tubes, transistors or FETs. Two similar devices have a common cathode/emitter/source connection, fed from a current source, the input is applied differentially to the grids/bases/gates, and the output is taken differentially from two resistive loads in the anodes/collectors/drains.

BASIC “LONG-TAILED PAIR” DIFFERENTIAL AMPLIFIER



Let us consider op amps with a bipolar transistor input stage. By diffusing them close together on a single chip and paying close attention to geometry and photo-engraving accuracy it is possible to manufacture a pair of bipolar transistors whose characteristic are very closely matched. When such a pair of transistors is used in a differential amplifier they will have virtually identical values of V_{BE} when their emitter currents are equal and so the offset voltage of the amplifier will be very low. Moreover, by using a chip-trimming technique such as zener zap or laser trim it is possible to reduce this offset still further.

SIMPLE BIPOLAR TRANSISTOR DIFFERENTIAL AMPLIFIER

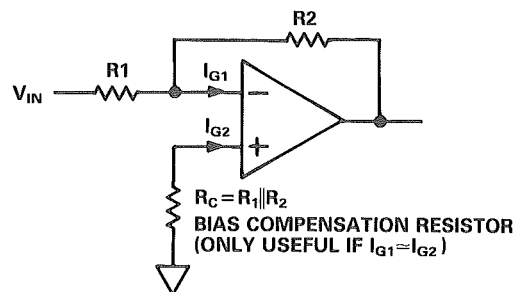


Such a simple stage has excellent characteristics. It has very good offset and low offset drift with temperature (it is a useful feature of bipolar differential amplifiers that when they are trimmed for minimum offset they are thereby also trimmed for minimum drift). In addition they have low noise, wide common-mode range and good frequency response provided that the current in the input stage is not too low.

Their major disadvantage is their bias current. Normal bipolar transistors have current gain (β) in the range 100 to 500 and there is a type (known as the super- β transistor) with values as high as 5000. This means that the bias current in a simple bipolar op amp will be somewhere between 0.02% and 1% of the input stage emitter current—which means that it will probably lie between 50nA and 2 μ A. This is a major disadvantage in many op amp applications and is only partially offset by the excellent matching of these bias currents and their limited variation with temperature (β increases with temperature so bias currents decrease—typically I_B at 125°C is roughly 50% of I_B at -55°C).

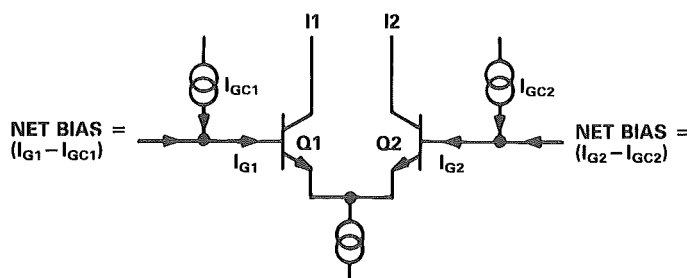
Because the currents are so well matched it is sometimes possible to compensate for their magnitude by the use of a bias compensation resistor. If the resistance seen by each input of an op amp is the same, the voltage drop due to bias currents will cancel out IF THE BIAS CURRENTS ARE EQUAL. Bias compensation is thus useful only with op amps having roughly equal bias currents in the inverting and noninverting inputs and in applications where the signal source impedance is known and constant.

A BIAS COMPENSATION RESISTOR CAN MINIMIZE BIAS CURRENT ERRORS



The second type of op amp input stage is known as the “bias compensated bipolar input stage”. Despite the similar names it has no relation to the circuit technique described above. In a bias compensated input stage we have a standard bipolar input stage which is equipped, on chip, with current sources to provide its input bias currents.

BIAS COMPENSATED BIPOLAR INPUT STAGE (Net Bias Current May Flow In Either Direction)

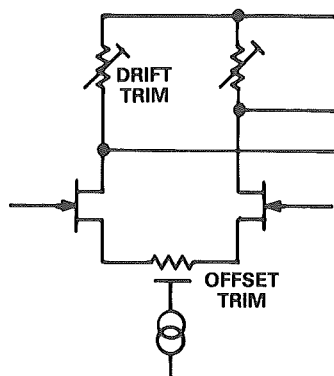


Such an architecture has most of the advantages of normal bipolar input stage: low offset, low drift, low noise and the possibility of high frequency operation. It also has much lower bias currents since the bias current flowing in the external circuitry is only the residual mismatch current between the bias current source and the transistor base. Typical values lie between 1 and 20nA and do not vary greatly with temperature.

However the bias currents in the inverting and noninverting inputs in op amps of this type are not well matched and may even flow in opposite directions (if the bias source current is greater than the base current there will be a net current flow OFF the chip—if it is less than the base current the flow will be ONTO the chip). Bias compensating resistors are therefore inappropriate for amplifiers of this type.

Even bias currents of a few nanoamps can be inconvenient in many applications. Where this is the case FET input op amps should be used. Originally it was necessary to use hybrid techniques in order to combine an FET input stage with bipolar op amp circuitry but the introduction of the “BiFET” process allows the fabrication of both types of device on the same chip. Recent developments in the BiFET process allow the manufacture of monolithic op amps with bias currents as low as 60fA (the AD549).

FET OP AMP INPUT STAGE Offset and Drift Must Be Trimmed Separately



The gate current of a junction FET is not related to its source current but is the leakage current of the reverse biased diode forming its gate. The bias current of FET input op amps is therefore independent of the current in the input stage and so high speed amplifiers may still have very low bias current.

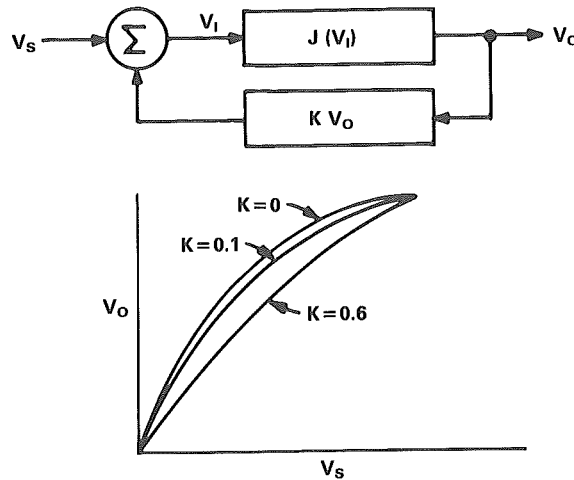
However, there are disadvantages to the architecture. Diode leakage current is not temperature stable, doubling for every 10°C increase in chip temperature, so that an amplifier with a bias current of 50pA at room temperature (25°C) has a bias current of 50nA at the military maximum of 125°C. The bias currents of an FET op amp are not well matched so bias compensation resistors may not be used with it.

Like bipolar amplifiers, FET amplifiers may have their offset voltage trimmed on the chip by laser or zener zap. Unlike bipolar amplifiers there is no intrinsic drift trim in the process—if an FET amplifier is trimmed for minimum offset its temperature coefficient is not minimized, but it may be separately laser trimmed to a minimum (zener zap is unsuitable for this), although it is necessary to make measurements with the silicon wafer at two different temperatures in order to do this. The result of this added complexity of trim is that the offset voltage and drift of an FET amplifier are larger than those of bipolar amplifiers, even though the bias currents are far lower. Which type of amplifier is chosen will depend on the application.

LINEARITY AND DISTORTION

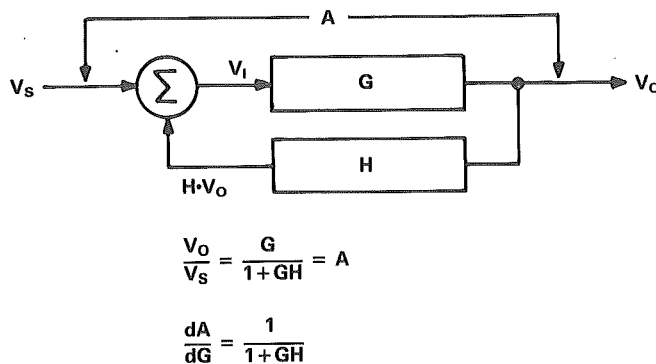
An ideal amplifier produces an exact scaled replica of its input signal at its output. To do this the slope of its transfer characteristic must be constant. Altering the shape of the input signal between the input and the output is referred to as “distorting” it. Distortion is a result of a processing a signal in a nonlinear system. A remarkable property of feedback amplifiers is their ability to improve linearity through the use of feedback. The actual mechanism by which this is accomplished is not obvious but may be approached by using a combination of analytical and graphical techniques.

FEEDBACK IMPROVES LINEARITY



The effect of adding feedback to improve linearity may be treated mathematically. These methods of analysis become increasingly more important as the total harmonic distortion (THD) in an amplifier is used as a criterion for selecting it.

NONLINEARITY GAIN SENSITIVITY

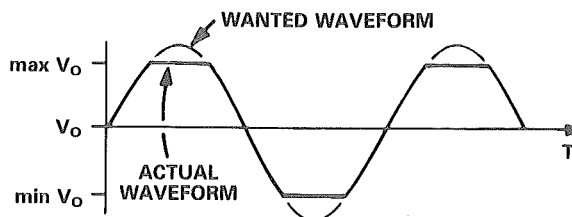


INCREASING THE PRODUCT GH REDUCES THE SENSITIVITY OF THE OVERALL AMPLIFIER TO VARIATION OF G AND HENCE REDUCES NONLINEARITY.

It is difficult to discern the presence of distortion on a sinusoidal waveform visually by using an oscilloscope—our eyeballs are just not calibrated to detect slightly distorted sinusoids (it is sometimes less difficult to perceive distortion on non-curved waveforms such as pulses and square, triangular and sawtooth waves). The most effective means of measuring distortion is to use a spectrum analyzer and measure the harmonically related components resulting from the nonlinear characteristics of an amplifier. A truly linear amplifier does not change the shape of an input signal—it merely scales it, likewise an amplifier with nonlinearity does change the waveshape and so causes distortion.

The most common form of distortion is "limiting" or "clipping" and occurs when the required output voltage from an amplifier is larger than the maximum voltage that the amplifier can actually provide (the output is said to exceed the amplifier's "headroom" or to "limit"). Symmetrical clipping introduces high levels of odd-harmonics into a waveform, asymmetrical clipping introduces even harmonics as well. The cures are either to reduce the gain of the amplifier or to increase its supply voltages.

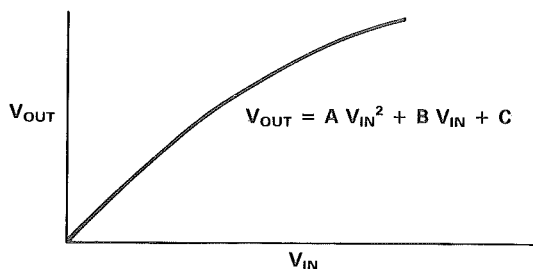
CLIPPING OCCURS WHEN THE REQUIRED OUTPUT AMPLITUDE EXCEEDS THE POSSIBLE OUTPUT AMPLITUDE



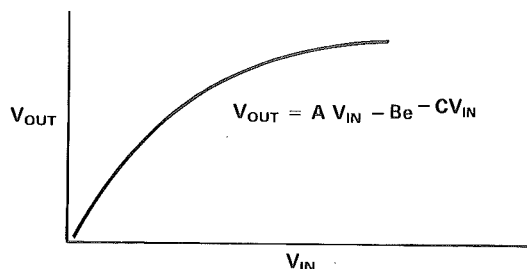
Although clipping may be considered a gross form of nonlinearity the term is normally reserved for phenomena which cause the central part of the amplifier transfer characteristic to deviate from a straight line. Consider a 12-bit system with a full scale range of 10 volts. If the system is specified to be linear to 12-bits it implies that the maximum deviation from the ideal transfer curve is 1LSB (2.44mV).

The two most common type of nonlinearity are square law and logarithmic. Square law nonlinearity occurs when there is a quadratic term in the transfer characteristic of the amplifier and can be produced by a field effect transistor with a resistive load. Logarithmic nonlinearity is produced by a logarithmic (inverse exponential) term in the transfer characteristic and can be caused by bipolar transistors and diodes.

BASIC TYPES OF NONLINEARITY (In Reality Many Forms of Distortion Will Contain Both)



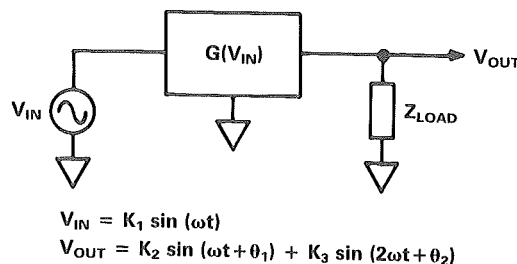
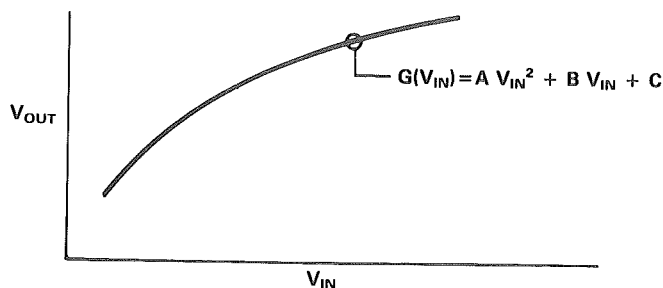
SQUARE LAW NONLINEARITY



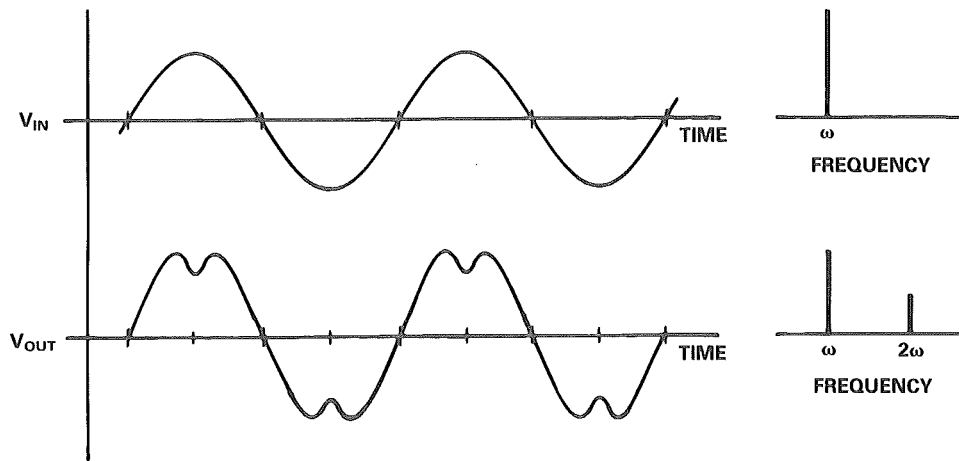
LOGARITHMIC NONLINEARITY

These types of distortion may be modelled by adding either a parabola or a rising exponential (or both) to the device transfer characteristic but it must be appreciated that distortion mechanisms are rarely simple and generally have several distinct causes. Models are thus rarely very accurate. Square law approximation of a nonlinear transfer function can be accomplished using curve fitting techniques, but the important thing to remember is square law transfer characteristics produce second harmonic terms only.

QUADRATIC DISTORTION

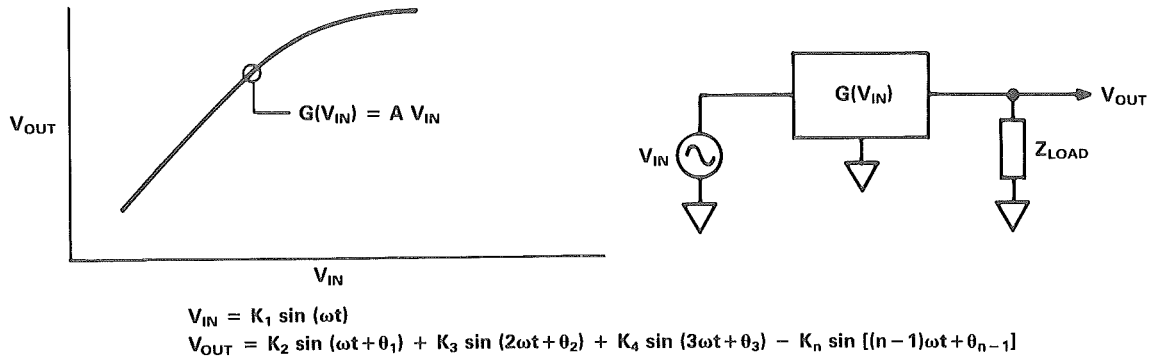


QUADRATIC DISTORTION

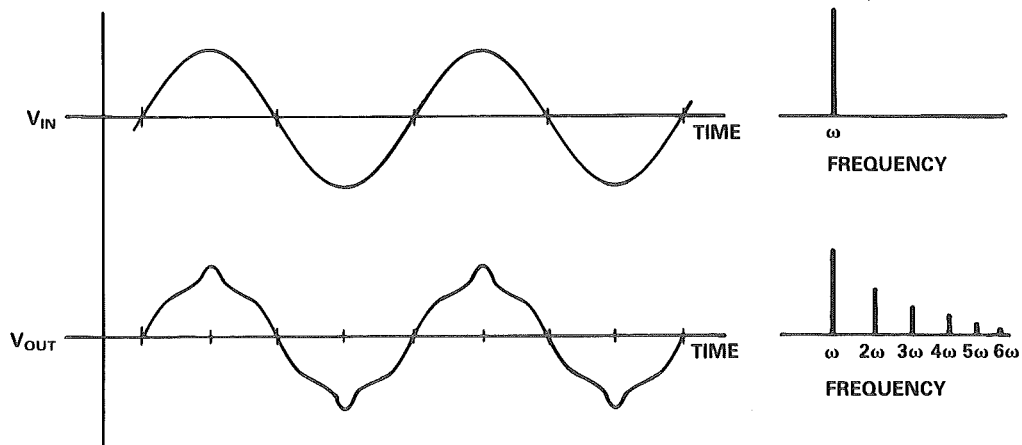


Exponential transfer functions are a bit trickier to analyze, as some are regarded as transcendental. Subsequently the curve fitting equations used are exponential and not quadratic. The result of a logarithmic transfer function is to distort an input signal thus producing both odd and even harmonics in the output signal. This differs fundamentally from the quadratic which produces second harmonics only.

LOGARITHMIC DISTORTION



LOGARITHMIC DISTORTION



SELECTING THE RIGHT AMPLIFIER FOR THE APPLICATION

1. Consider the Signal Source
2. Analyze the Effects of Op Amp Specs on the Application
3. Determine Environmental Variations
4. Don't Spend More (or Less) Than Necessary

The choice of an operational amplifier is generally governed by the application in which it is to be used. The process of selecting the best amplifier for an application involves a rigorous analysis of error sources—both in the amplifier and in the external components. It is important to analyze the characteristics of the input signal, the desired accuracy and the environmental conditions to which the circuit will be subjected. In this section of the seminar we shall consider a series of op amp applications and outline the principles behind the amplifier selection.

"WHAT DO YOU DO WHEN A 741* WON'T DO?"

- A. Impossible — 741s Can Do Anything
- B. Use A 741 Anyway and Hope Nobody Notices
- C. Use A Dual 741 (Two Amps are Better than One)
- D. Give Up and Take Up Computer Programming
- E. Select an Amplifier with Specs Which Match the Application

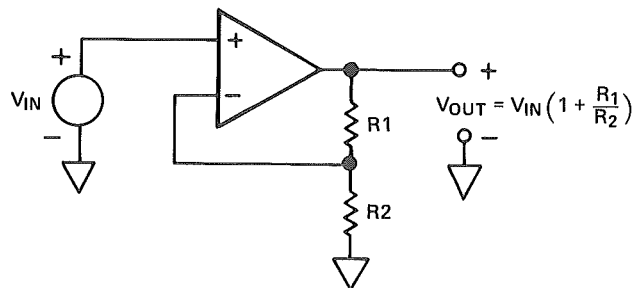
*Or 301A, 308A, 356A, or Whatever General-Purpose Op Amp You Use "Everywhere".

PREAMPLIFIER APPLICATIONS

Operational amplifiers were originally intended for use inside analog computational circuits, where feedback networks allowed them to perform analog operations such as integration and differentiation. They are now widely used as low cost gain stages. The choice of op amps in such applications depends upon the characteristics of the signal source, the gain stage and the desired accuracy.

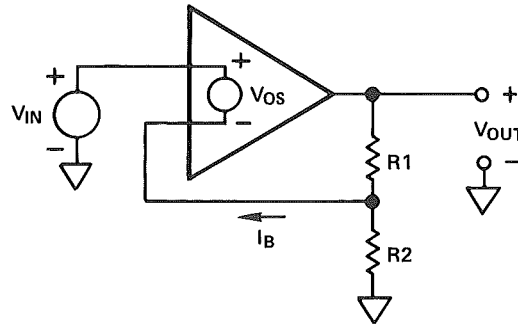
Consider a noninverting preamplifier built with an ideal op amp and ideal resistors, and intended for amplifying a low level signal. The output will be equal to the input times one plus the resistor ratio.

IDEAL NON-INVERTING GAIN STAGE



The input offset voltage of the op amp will appear as a voltage source in series with the signal, and will be amplified along with it. Any bias current in the amplifier input will flow in the source resistance and develop an additional offset voltage which will also be amplified. Interestingly, since the voltage produced by the bias current is equal to $I_B(R1R2/(R1 + R2))$ and the gain applied to this voltage is $(R1 + R2)/R2$ the net effect at the output is a term equal to I_BR1 .

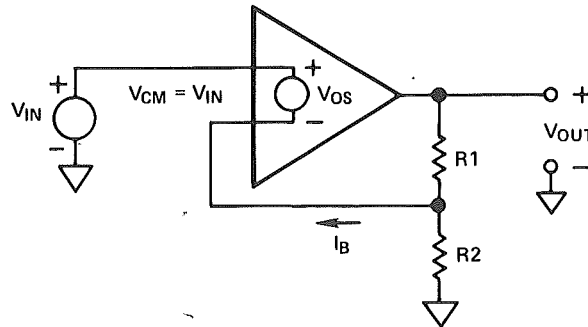
NON-INVERTING AMPLIFIER FIRST-ORDER ERRORS



$$V_{OUT} = \left[V_{IN} \left(1 + \frac{R_1}{R_2} \right) + V_{OS} \left(1 + \frac{R_1}{R_2} \right) + I_B R_1 \right] = \left(1 + \frac{R_1}{R_2} \right) \left[V_{IN} + V_{OS} + I_B \left(\frac{R_1 R_2}{R_1 + R_2} \right) \right]$$

Common mode effects cause an additional error in noninverting amplifiers. Although an ideal op amp is insensitive to common-mode inputs, real amplifiers do respond to them. This error term can be modeled at the input as an additional offset voltage equal to common-mode voltage divided by CMRR. As an offset, it is amplified by the same amount as the signal.

NON-INVERTING AMPLIFIER SECOND-ORDER ERRORS

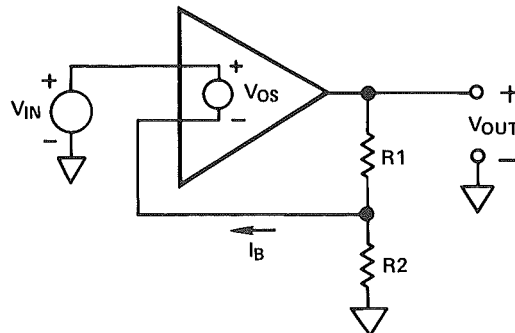


$$V_{OUT} = \left[V_{IN} \left(1 + \frac{R_1}{R_2} \right) + V_{OS} \left(1 + \frac{R_1}{R_2} \right) + I_B R_1 + \left(\frac{V_{IN}}{CMRR} \right) \left(1 + \frac{R_1}{R_2} \right) \right]$$

$$= \left(1 + \frac{R_1}{R_2} \right) \left[V_{IN} + V_{OS} + I_B \left(\frac{R_1 R_2}{R_1 + R_2} \right) + \left(\frac{V_{IN}}{CMRR} \right) \right]$$

Finite open loop gain adds yet another error. It can be shown that the actual gain of the circuit will be equal to the desired gain multiplied by $1 + [1/A_{OL} \cdot \beta]$ where A_{OL} is the amplifier's open loop gain and β is the voltage feedback ratio.

GAIN ERROR INTRODUCED BY FINITE A_{OL}



$$V_{OUT} = \left(\frac{R_1 + R_2}{R_1} \right) \left[V_{IN} + V_{OS} + I_B \left(\frac{R_1 R_2}{R_1 + R_2} \right) + \frac{V_{CM}}{CMRR} \right] \left[\frac{A_{OL} \frac{R_2}{R_1 + R_2}}{1 + A_{OL} \frac{R_2}{R_1 + R_2}} \right]$$

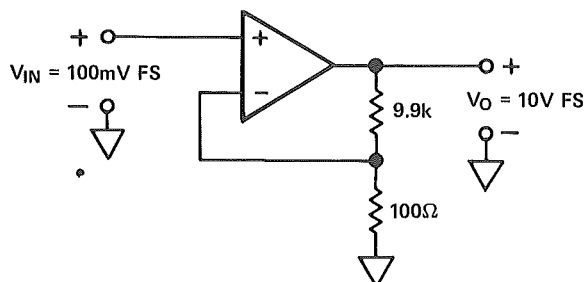
A numerical example may assist in the appreciation of the relative magnitudes of these errors. Consider a noninverting preamplifier with a gain of 100, built with perfect resistor, and intended to amplify 100mV to 10V. We might choose to use an AD741 or an AD OP07.

COMPARISON OF AD741 AND ADOP-07 SPECIFICATIONS

	AD741CH	ADOP-07CH
V_{OS}, Max	6mV	150 μ V
I_B, Max	500nA	7nA
CMR, Min	70dB	100dB
A_{OL}, Min	20,000V/V	1,200,000V/V

The largest error is due to offset voltage. Since this error can be reduced or eliminated by trimming it might be thought that it can be neglected but temperature induced offset drifts are not so easily corrected.

COMPARISON OF 741C AND OP-07C PERFORMANCE



$$V_{OUT}|_{741C} = (100) \left[0.1 - 0.006 - (500\text{nA} \times 100\Omega) - \frac{0.1}{3500} \right] \left[\frac{(20000) \left(\frac{1}{100} \right)}{1 + (20000) \left(\frac{1}{100} \right)} \right]$$

$$= 9.3922 \left[\frac{200}{201} \right] \text{ V (6.08\% ERROR)}$$

$$= 9.3455\text{V (6.54\% ERROR)}$$

$$V_{OUT}|_{OP-07C} = (100) \left[0.1 - 0.00015 - (7\text{nA} \times 100\Omega) - \frac{0.1}{100000} \right] \left[\frac{1.2 \times 10^6 \times \frac{1}{100}}{1 + (1.2 \times 10^6) \frac{1}{100}} \right]$$

$$= 9.98493 \left[\frac{12000}{12001} \right] \text{ V (0.15\% ERROR)}$$

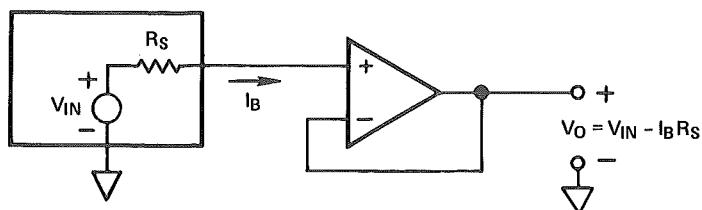
$$= 9.98409\text{V (0.159\% ERROR)}$$

It is also possible to reduce bias current errors by matching the impedance seen at each input of the op amp. The resultant error is now I_{OS} times the source resistance. CMRR and A_{OL} induced errors cannot be compensated because they are not constant. CMRR is often a nonlinear function of common-mode voltage.

Noninverting gain stages are often used to buffer high impedance sources. In these applications bias current may be the primary error source. Consider a pH or other ion selective electrode. The desired signal is a few hundred millivolts but the source impedance can exceed 100 megohms and is neither well defined nor stable.

An FET input op amp is the only practical choice for such applications. Consider a simple voltage follower with a wanted signal of V_{IN} —the actual signal being amplified is $V_{IN} - I_B R_S$.

EFFECT OF BIAS CURRENT IN BUFFER AMPLIFIER



OP AMP
IDEAL
AD741
AD OP-07
AD711
AD549

I_B
0
500nA
7nA
15pA
60fA

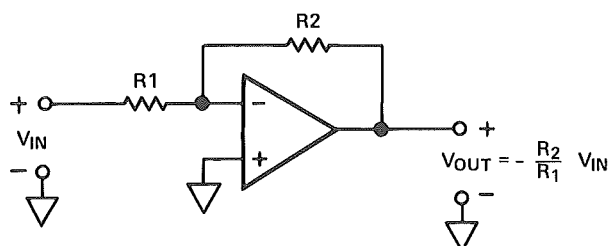
ERROR FOR
 $R_S = 10\text{M}\Omega$
0
5V
70mV
0.15mV
0.6 μ V

With a 10M source resistance the 500nA bias current of an AD741 op amp will produce an error of about 5V whereas the 60fA of an AD549 FET input electrometer amplifier gives an error of 0.6μV. Neither amplifier should be used in the application: the error from the AD741 is intolerable, while the voltage offset of the AD549 is so much greater than the 0.6μA bias related error that use of an AD549 would be overkill. The correct choice is a less expensive FET input amplifier such as the AD711C which has less than 25pA bias current (giving 250μV error) or the AD545 which has 1pA bias current (giving 10μV error at 25°C and [bias doubling every 10°C] under 250μV at 70°C).

THE INVERTING AMPLIFIER

Many circuits require a signal to be inverted as well as amplified. Op amps are easily configured for such negative gain. An inverting amplifier consists of an op amp and two resistors, R1 and R2. The noninverting input of the op amp is grounded, its output is connected to its inverting input by R2 and the input signal is applied to the inverting input through R1. Negative feedback keeps the inverting input at the same potential as the noninverting input. This means that the currents in the two resistors must be equal—so that voltages on them must be in the ratio R1:R2 and the output voltage (V_O) must be $-R_2/R_1$ times the input voltage (V_{IN}). The input impedance is equal to R1 (unlike the noninverting amplifier which has a very high input impedance).

IDEAL INVERTING AMPLIFIER

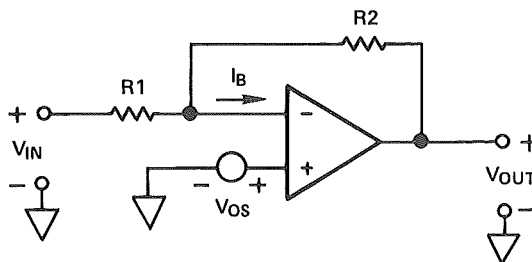


The error analysis of an inverting amplifier differs slightly from that of a noninverting amplifier. Since inverting amplifiers cannot have common-mode voltages CMRR is unimportant but errors due to offset and internal noise are actually amplified more than the signal is. The input offset and noise voltages can be modeled as a voltage source in series with the inputs. The gain seen by this offset/noise voltage is $(1 + R_2/R_1)$ which is the inverting gain plus 1. Where the desired signal gain is low (particularly where the inverting gain is less than unity), this increased offset/noise gain can be troublesome.

NOTE THAT IN AN INVERTING AMPLIFIER:

1. $V_{CM} \approx 0$ so CMRR is Not as Important as it is in a Noninverting Amplifier.
2. "Noise Gain" is Equal to Signal Gain Plus 1. Offset Voltage will be Amplified More than the Signal.

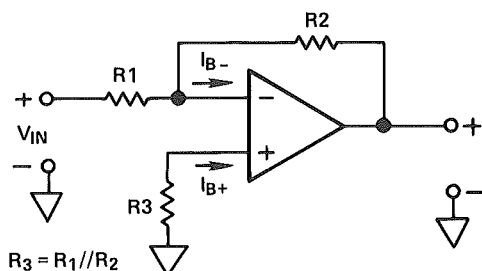
INVERTING AMPLIFIER WITH FIRST-ORDER ERRORS



$$V_{OUT} = (V_{IN}) \left(-\frac{R_2}{R_1} \right) + (V_{OS}) \left(1 + \frac{R_2}{R_1} \right) + (I_B) \left(\frac{R_1 R_2}{R_1 + R_2} \right) \left(\frac{R_1 + R_2}{R_1} \right)$$

Input bias current is another source of error in inverting amplifiers. The bias current flows through a resistance equal to the parallel resistance of R_1 and R_2 creating an offset voltage which is again amplified by $(1 + R_2/R_1)$ which gives us a bias related output error of $I_B R_2$. It is common practice to add a resistance R_3 (equal to the parallel resistance of R_1 and R_2) in series with the noninverting input so that the bias related offset $I_B R_3$ is compensated by an equal and opposite voltage in the noninverting input. As mentioned in an earlier section this resistive bias compensation is only effective in simple bipolar op amps where the bias currents are approximately equal—in bias compensated and FET op amps (whose bias currents may be seriously unequal—i.e., I_{OS} may be as large as or larger than I_B) the use of this type of bias current compensation may actually increase errors. It does also introduce a small common-mode voltage into the system but this is most unlikely to produce significant errors.

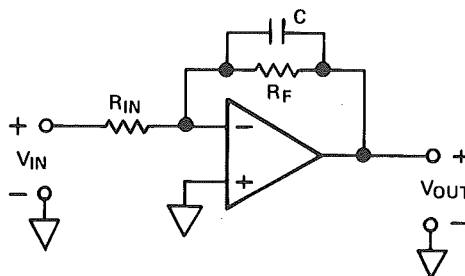
USE OF BIAS CURRENT CANCELLATION RESISTOR



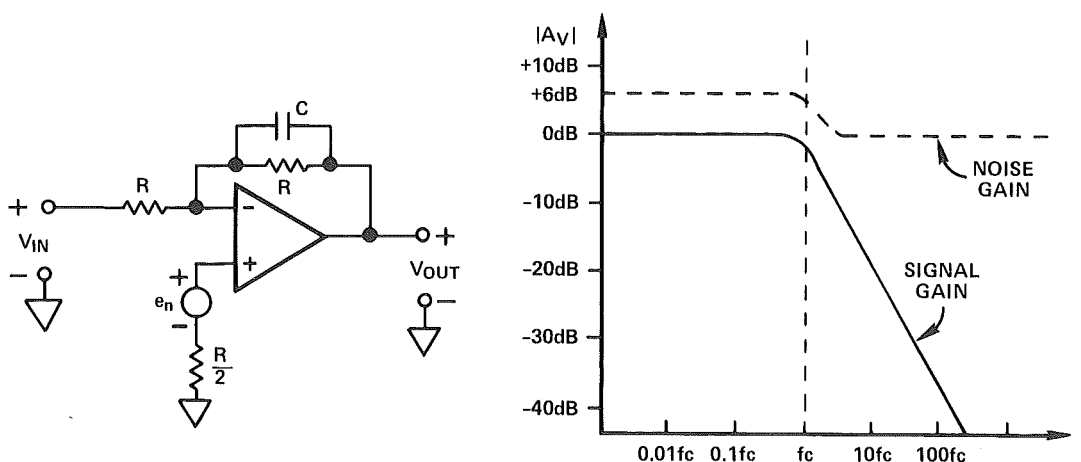
$$\begin{aligned} V_{OUT} &= -V_{IN} \left(\frac{R_2}{R_1} \right) + \left[(I_{B-} R_2) - (I_{B+} R_3) \left(\frac{R_1 + R_2}{R_1} \right) \right] \\ &= -V_{IN} \left(\frac{R_2}{R_1} \right) + \left[(I_{B-} R_2) - \left(I_{B+} \frac{R_1 R_2}{R_1 + R_2} \right) \left(\frac{R_1 + R_2}{R_1} \right) \right] \\ &= -V_{IN} \left(\frac{R_2}{R_1} \right) + [(I_{B-} - I_{B+}) R_2] \\ &= -V_{IN} \left(\frac{R_2}{R_1} \right) + I_{OS} R_2 \end{aligned}$$

The useful dynamic range of any amplifier is limited by the input level at which clipping occurs and by the random noise generated within the circuit. In analyzing the effects of noise it is important to consider the frequency response of the noise gain. The gain of an inverting amplifier with a capacitor in parallel with R_2 will drop at 6dB/octave forever from the frequency at which the reactance of the capacitor equals R_2 —the noise gain of the same circuit drops at the same rate (but the noise gain was higher than the signal gain to begin with) but only drops to unity, after that it remains constant with rising frequency. Thus as the frequency goes up the ratio of signal gain to noise gain degrades at 6dB/octave.

INVERTING AMPLIFIER WITH INTEGRATING CAPACITOR

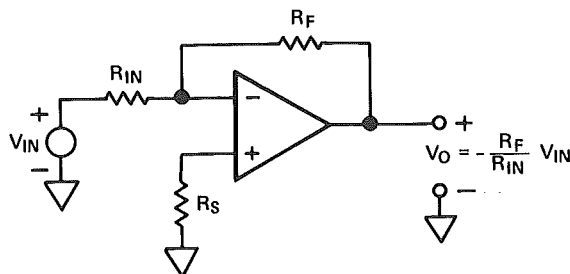


FREQUENCY RESPONSE OF UNITY-GAIN INVERTING AMPLIFIER



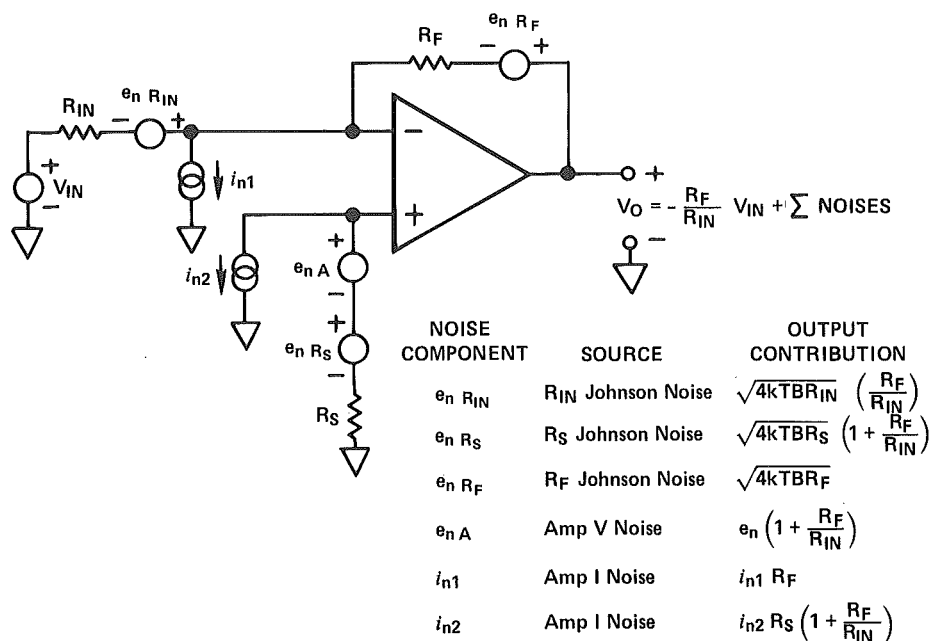
In addition to considerations of noise gain there are other factors which must be taken into account when designing op amp circuits for low noise. If we assume all components to be free of noise an amplifier circuit is easy to model.

NOISELESS INVERTING AMPLIFIER



Such a model is completely unreal since even perfect resistors have Johnson noise at any temperature above absolute zero. The noise model of a real inverting amplifier contains six separate uncorrelated noise sources! These are the three amplifier noise sources (differential voltage noise and the current noises in each input—which develop noise voltages when they flow in the circuit resistors) and the Johnson noise in the input resistor, the bias compensation resistor and the feedback resistor (this last is often overlooked because it is not in the input circuitry but it can be important).

NOISE IN INVERTING AMPLIFIER



As mentioned in an earlier section the Universe is a noisy place and it is not always possible to achieve both the noise performance and the impedance levels that one requires simultaneously. Understanding the theory behind noise can be of great assistance in designing circuits and systems with optimum noise performance.

CHOOSING AN AMPLIFIER FOR LOW NOISE

1. Examine Circuit Impedances.
2. Determine Noise Voltage Gain.
3. Trade Off Between Low Current Noise and Low Voltage Noise.
4. Note that the Desired Noise Performance May Be Physically Impossible if Impedances are too High.

FILTER APPLICATIONS

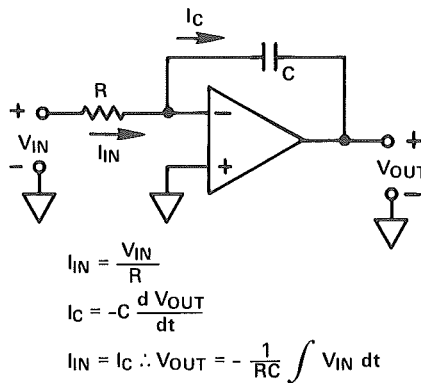
A filter is a circuit which amplifies some of the frequencies applied to its input and attenuates others. There are four common types: high pass, which only amplifies frequencies above a certain value; low pass, which only amplifies frequencies below a certain value; bandpass, which only amplifies frequencies within a certain band; and bandstop, which amplifies all frequencies except those in a certain band.

Filters may be produced using many different methods. These include passive filters which use only passive components such as resistors, capacitors and inductors, filters which use mechanical resonances in quartz or other piezo-electric materials as a means of constructing frequency dependent circuitry, digital filters which use analog-digital converters to convert a signal to digital form and then use high-speed digital computing techniques to filter it, and active filters which use amplifiers in addition to resistors, capacitors and inductors in order to obtain performance impossible with passive filters. Operational amplifiers are frequently used as the gain blocks in active filters. While it is beyond the scope of this text to provide a detailed study of active filter design some important considerations will be discussed.

Integrators (Low Pass Filters)

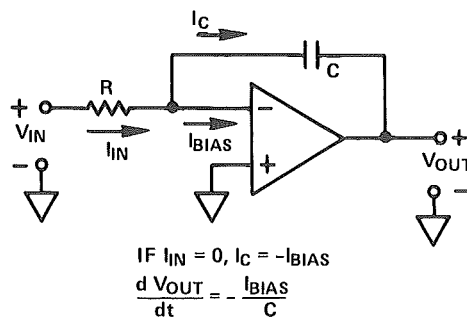
The ideal integrator uses a perfect op amp in the inverting configuration with a capacitor in the feedback path. It provides an output signal which is proportional to the integral of the input signal. The op amp maintains both inputs at zero volts, so the input current, I_{IN} , is equal to V_{IN}/R . Ideally all of this current flows into the capacitor and none into the op amp input.

IDEAL INTEGRATOR/LOW-PASS FILTER



Bias current is always present in a real op amp and it goes to charge the capacitor, causing the output to move in the absence of an input signal. This output voltage will appear as a dc output drift with respect to time, and is equal to $(-I_B/C)$ volts/second. As an example of the relative magnitude of this drift an amplifier with 100nA bias current used in an integrator with 1000pF integrating capacitor has an error due to bias current charging of 100 volts/second. This means that the amplifier will saturate shortly after power is applied and the integrator function will be lost.

INTEGRATOR WITH BIAS CURRENT



$$\frac{dV_{OUT}}{dt} = \frac{I_{BIAS}}{C} \frac{V}{\text{SEC}}$$

$$\text{FOR } C = 1000\text{pF AND } I_{BIAS} = 100\text{nA}$$

$$\frac{dV_{OUT}}{dt} = 100 \frac{V}{\text{SEC}} !$$

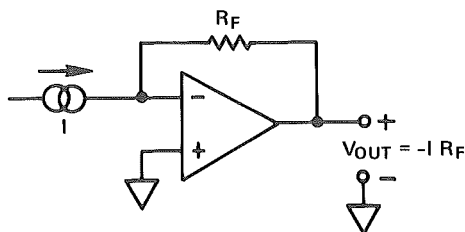
While this is a simple case it does illustrate that dc parameters of op amps can have a significant impact on active filter circuit performance. To select an op amp for a filter application assume that it is ideal and select appropriate pole and zero locations for the desired response characteristics. Once the capacitor and resistor values have been selected, determine the effect of offset and other dc parameters. Keep in mind that even in ac coupled stages large dc offset gains can cause large dc offsets, which cause clipping or saturation.

Current-to-Voltage Converters

Some transducers produce an output voltage, others an output current. Current transducers include photodiodes, some temperature sensors and a variety of biological probes. Often the currents produced are very small—of the order of nanoamps or less. Such currents require amplification before they can be used in a system and the first stage of such amplification is usually a current/voltage converter.

While a resistor is a genuine current-to-voltage converter it is generally impractical to force the current directly through a resistor and measure the voltage produced since most current sources have a limited range of compliance voltage. A better technique for converting transducer currents to voltages uses an op amp and a feedback resistor. The input current source drives a constant terminal voltage (zero volts) and the current flows through the resistor to the op amp output. The output voltage is equal to the input current times the feedback resistor.

CURRENT-TO-VOLTAGE CONVERTER



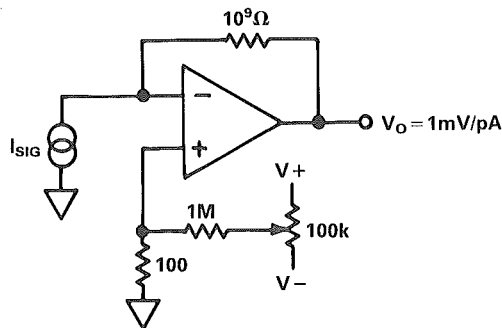
The amplifier input bias current will clearly cause an error so amplifiers chosen for this application have very low bias currents and are generally FET input types. The lowest available bias current is 60fA at room temperature from the monolithic AD549.

AD549 FEATURES

Ultra Low Bias Current	60pA max
Low Offset Voltage	250 μ V
Low Offset Drift	5 μ V/ $^{\circ}$ C
Low Noise	4 μ V p-p (0.1Hz – 10Hz)
Low Power	700 μ A Supply Current

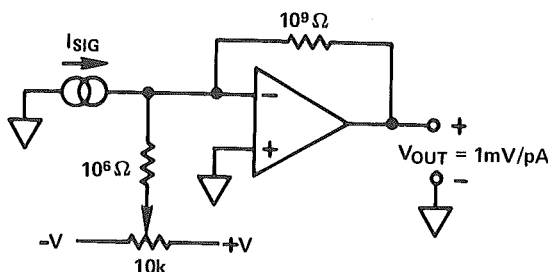
The best way to minimize the offset voltage of current/voltage converters is to use the offset null terminals provided by the manufacturer. Dual op amps, however, tend not to have offset null terminals and other techniques must be used. The correct method is to apply a correction voltage to the noninverting input.

THE CORRECT WAY TO NULL THE OFFSET OF A I/V CONVERTER WITHOUT OFFSET TRIM TERMINALS



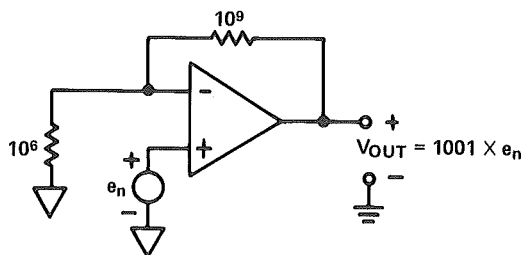
If offset null is applied to the inverting input the noise and drift performance will be drastically degraded. Consider the circuit which might be used—a potentiometer connected to a high value resistor is used to inject a correction current into the summing junction of the amplifier.

NULLING OFFSETS IN CURRENT-TO-VOLTAGE CONVERTER (WRONG WAY)



If we draw the equivalent circuit we see that we have a noise/offset gain of 1001 because of the high ratio of feedback to injection resistors. The circuit is obviously unsatisfactory.

EQUIVALENT CIRCUIT FOR NULLED I/V CONVERTER



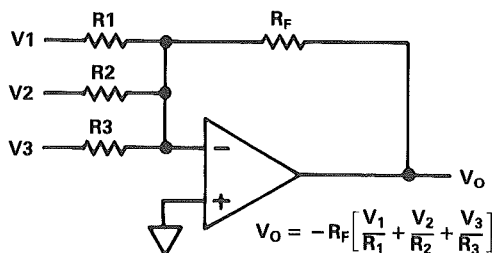
APPLICATIONS

There are innumerable ways in which operational and transimpedance amplifiers may be used in practical applications and whole books have been written on the subject. This section of our seminar is not exhaustive but discusses a number of amplifier applications which will probably be useful in themselves and which will certainly be useful in demonstrating the philosophy behind amplifier applications.

Inverting Amplifier

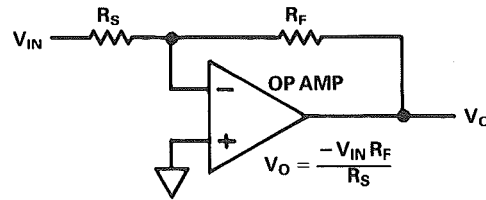
Operational amplifiers were originally intended for use as inverting amplifiers and they are probably more often used in this mode than in any other since it is so flexible. In this mode negative feedback acts to maintain the inverting input at the same potential as the noninverting input. A point which is not ground but is maintained at low impedance ground potential by negative feedback in this way is known as a “virtual ground” or “virtual earth”. The inverting input is therefore a low impedance summing point. If a number of resistors are connected to it and a signal applied to each the signals will be summed without interacting. Furthermore, the scaling factor applied to each signal will depend only upon the resistor through which it is connected to the summing junction.

OPERATIONAL AMPLIFIER USED AS A SUMMING AMPLIFIER



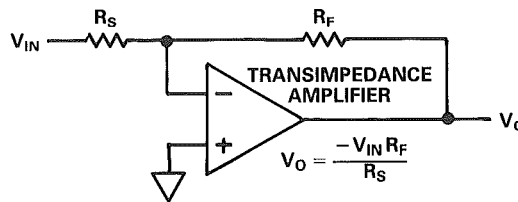
This is a special case of a simple inverting amplifier.

SIMPLE INVERTING AMPLIFIER



The above circuits use an operational amplifier as the active element. Transimpedance amplifiers have several characteristics that make them particularly suitable for use in inverting amplifiers. One is their high slew rate and the other is their feature of maintaining constant bandwidth as their closed loop gain is varied. The constant bandwidth is achieved only if the closed loop gain is adjusted by varying the input resistance, R_S , while holding the feedback resistance, R_F , constant.

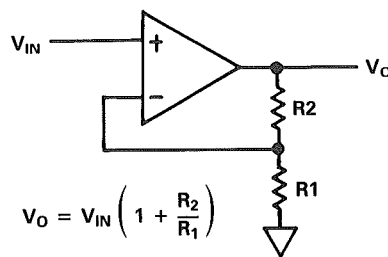
INVERTING AMPLIFIER USING A TRANSIMPEDANCE AMPLIFIER



Noninverting Amplifier

The noninverting amplifier configuration offers several advantages over the inverting amplifier configuration but also certain drawbacks. The major advantage of the noninverting mode is its high input impedance—the common-mode impedance of the op amp being used. Despite the high impedance of such an input the bias current of the amplifier must still be allowed for—it should be modelled as a current source feeding the input terminal.

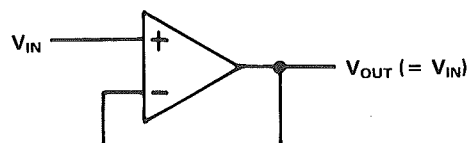
NONINVERTING AMPLIFIER



The high input impedance of a noninverting stage offers minimum circuit loading at a cost of lowered noise immunity. Amplifiers with high input impedance are much more sensitive to E-field noise (electrostatic pickup) and must be guarded against it by attention to circuit layout and shielding.

One of the most common noninverting amplifiers is the unity gain “voltage follower”. It consists of an internally compensated op amp with its output connected directly to its inverting input and uses no other components. It has a very high input impedance (don’t forget the bias current!) and a very low output impedance and is widely used to buffer signals from high source impedances to low load impedances.

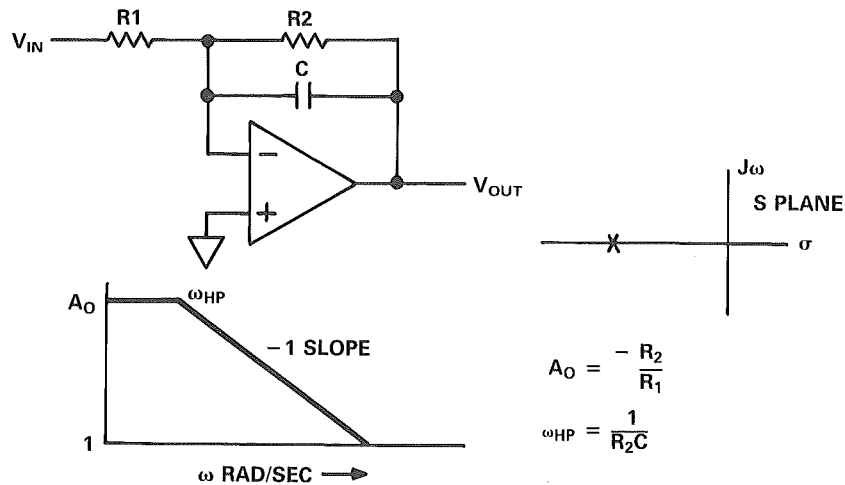
OPERATIONAL AMPLIFIER USED AS A VOLTAGE FOLLOWER



Basic Low Pass Filter

The low pass filter is the most simple one of all. It consists of an inverting amplifier with a capacitor in parallel with its feedback resistor. As the frequency increases the capacitive reactance decreases—hence the gain decreases.

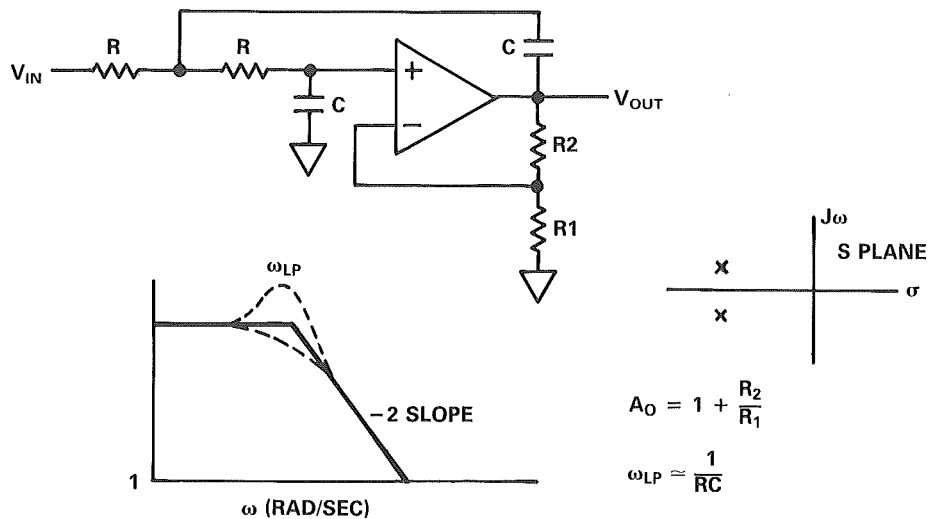
BASIC LOW PASS FILTER



The cutoff frequency, ω_{HP} , is simply $1/(R_2 C)$ and the rolloff is 6dB/octave (20dB/decade). The noise is a factor of $\pi/2$ more than would be encountered if the filter skirt were infinitely steep.

Another low pass filter is the Sallen and Key. This two capacitor filter is easily implemented, as the component selection is not critical and a high performance filter is readily synthesized.

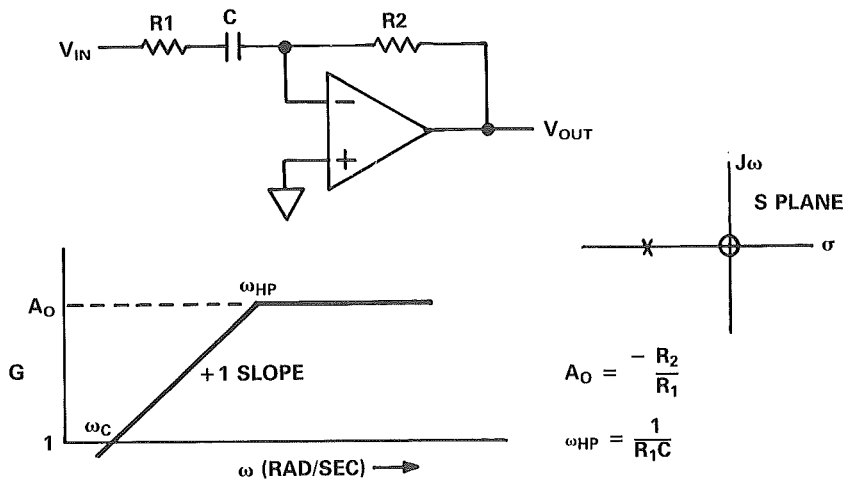
SALLEN AND KEY LOW PASS FILTER (2 POLE)



Basic High Pass Filter

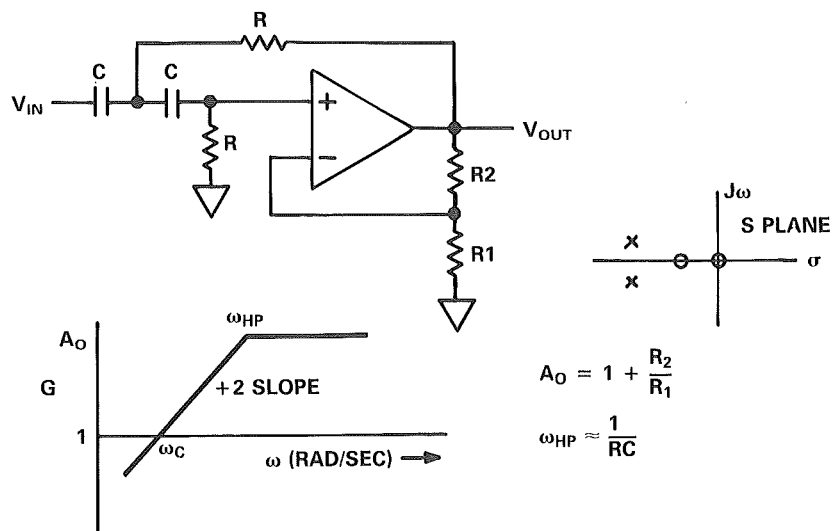
The basic high pass filter uses a capacitor as an input element. The gain increases with frequency until it is limited by the series resistor R_1 . The high frequency gain therefore flattens out at a value of R_2/R_1 . This response shows both a pole and a zero. If the high frequency gain were not limited by R_1 and R_2/R_1 the system would probably oscillate.

BASIC HIGH PASS FILTER



The Sallen and Key structure can also be used to produce a high pass filter. Like the low pass design the component selection is straight forward. Note the zero at the origin, the zero on the real axis and the pair of conjugate poles.

SALLEN AND KEY 2-POLE HIGH PASS FILTER

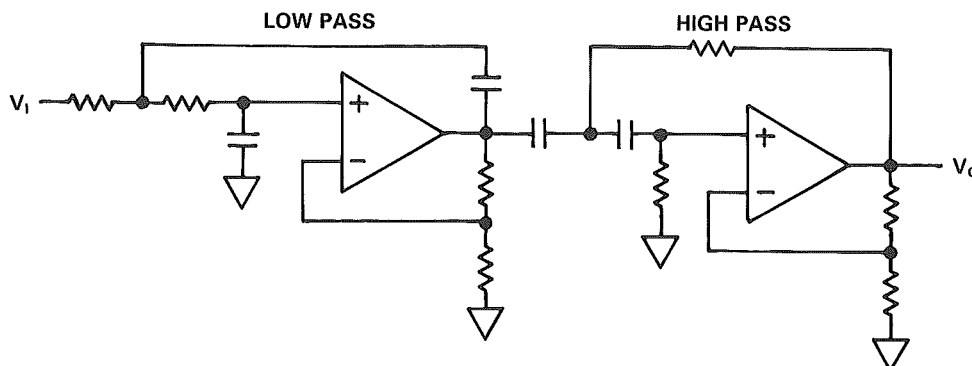


Band Pass Filters

The simplest bandpass filter consists of cascaded high pass and low pass filters. Where quick and predictable results are required there is a great deal to recommend this approach since the upper and lower cutoff characteristics may be manipulated separately.

Integrated bandpass filters where a single amplifier functions in both high pass and low pass modes are considerably more demanding to design and frequently require very inconvenient (and very accurately defined) component values. For details on the design of active filters consult "Electronic Filter Design Handbook".

BAND-PASS FILTER USING LOW PASS AND HIGH PASS SALLÉN AND KEY FILTERS

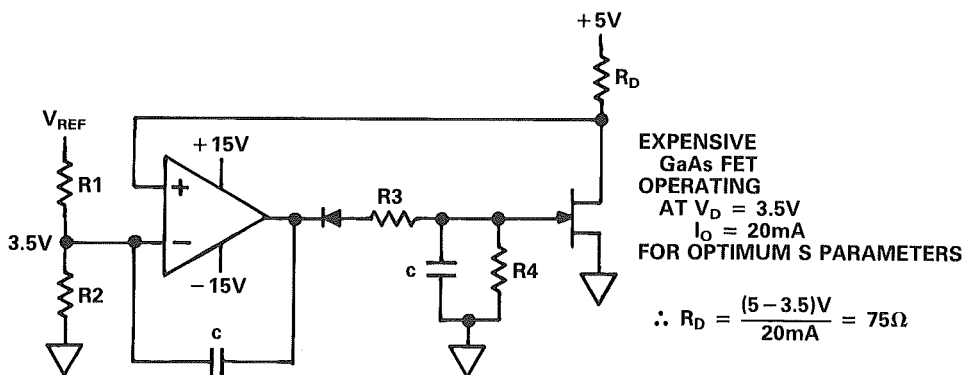


Active Bias Networks

An op amp active bias network is a handy means of reliably establishing precise voltage and current bias conditions for an external component. Consider the following example.

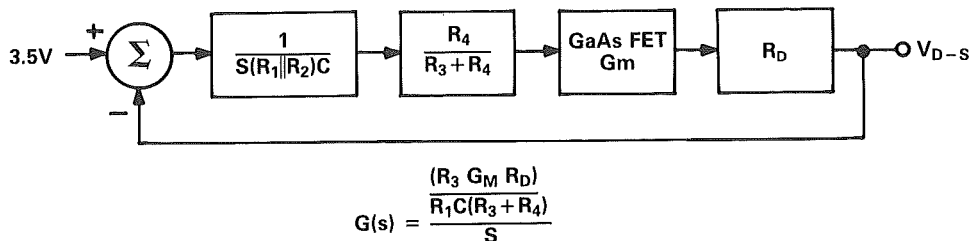
Gallium arsenide field effect transistors (GaAsFETs) are used in amplifiers at microwave frequencies and their high frequency performance is usually specified with S-parameters—which are merely voltage reflection coefficients measured in a 50Ω system. The S-parameters of a GaAsFET are measured at a specific dc bias and if the bias conditions change the amplifier performance will suffer. It is thus necessary to ensure that the bias conditions remain constant over the required environmental temperature range. This is done with an active bias network.

ACTIVE BIAS CIRCUIT FOR UHF AMPLIFIER (ALL RF COMPONENTS OMITTED FOR CLARITY)



The operational amplifier contains an integrator to establish a predictable unity gain crossover frequency. The voltage divider limits the maximum voltage that can be applied to the gate of the GaAsFET and the diode prevents a positive voltage from being applied to the gate (designers should always assume that supply voltages may be applied to a network in random order and should design their circuit so that it will tolerate this).

CONTROL SYSTEM MODEL OF ACTIVE BIAS NETWORK

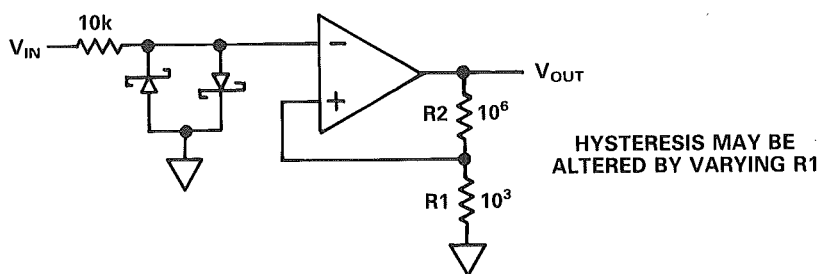


In this application the critical items are the stability of the reference voltage (as this establishes the drain to source voltage) and the +5V supply, and the value of the drain resistor (which sets the drain current). Almost any operational amplifier can be used in this application.

Comparators

Operational amplifiers make perfectly adequate comparators in applications where high speed is not essential. Dedicated comparators frequently have an open collector output for convenient logic interface but in many cases the output stage of an op amp will work as well as necessary—and the op amp will probably be less expensive. If hysteresis is required to minimize noise or oscillation in the transition region it may be provided with two resistors. The circuit shown is an example of an op amp used as a comparator in a zero crossing detector. The design incorporates hysteresis and a diode clamp. This type of circuit is often used to produce a square wave or to clean up a noisy signal.

OPERATIONAL AMPLIFIER USED AS A ZERO CROSSING COMPARATOR

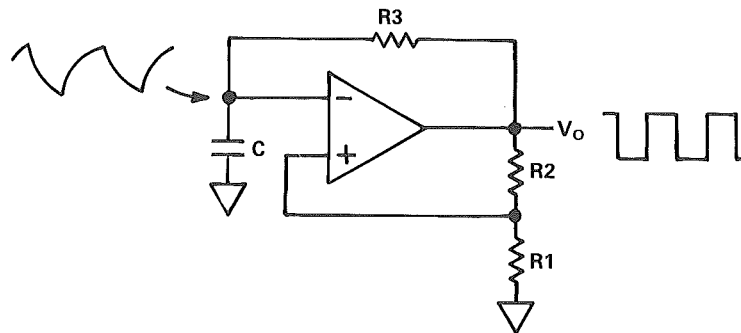


Oscillators

The subject of oscillators is so large that it is impractical to cover it in a seminar of this type. It is important, however, to realize that operational amplifiers are used as gain blocks in oscillators of many different types and that the same care should be devoted to the choice of op amp for oscillator use as for any other application. It is undoubtedly true that with sufficient positive feedback any circuit can be made to oscillate but problems of bias current, noise, phase shift and slew rate will affect the quality of an oscillator's performance just as they would affect any other op amp application. The basic principle of design is to design a system assuming a perfect op amp and then analyze the effects of those parameters likely to affect its operation.

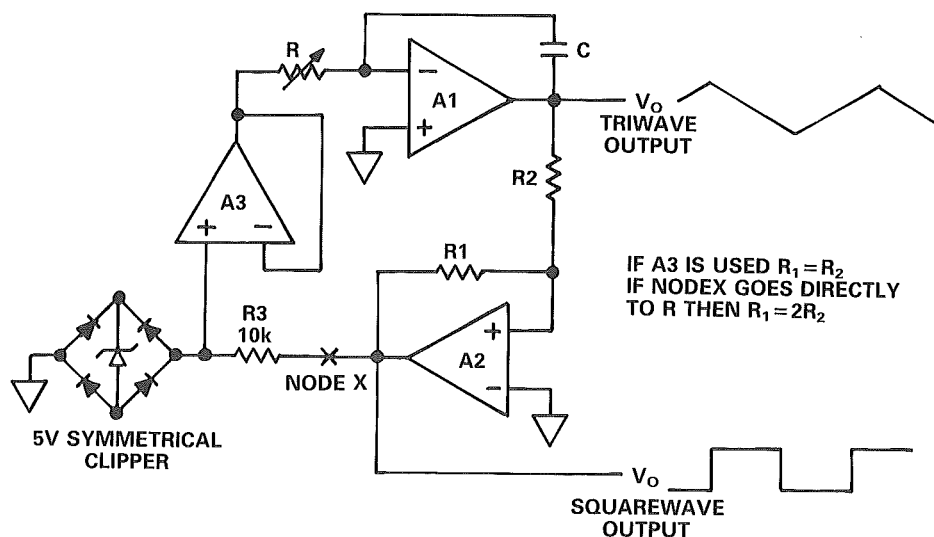
In a relaxation oscillator using an op amp and just three resistors and a capacitor the parameters which affect operation are its output levels (how much "headroom" does the output stage have?) and bias current. The headroom will affect the frequency of oscillation while the bias current will affect the mark:space ratio of the (nominally) square output waveform as well as having a second-order effect on oscillation frequency.

OPERATIONAL AMPLIFIER RELAXATION OSCILLATOR



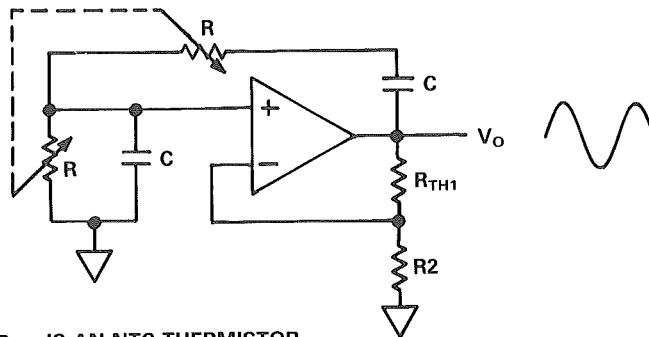
Another oscillator which can be classified as a relaxation oscillator is a triangular wave oscillator using two or three op amps (the third, the unity gain buffer between the clipper and the integrator input, is not essential but improves performance). The frequency is set by the values of R and C and the amplitude of both the square and tri-wave outputs by the bipolar clipping circuit made with a zener diode and a diode bridge. If amplifier A2 limits symmetrically the circuit may be simplified by connecting the node X directly to A2 output and omitting the clipper, A3 and R3—in this case R2 must be smaller than R1 and the amplitude of the triwave will be smaller than that of the square wave.

TRIWAVE OSCILLATOR



No work on op amp oscillators would be complete without a Wien bridge circuit. The Wien bridge contains two equal resistors and two equal capacitors has an attenuation of 3 at zero phase shift. (No bridge that I ever saw in Vienna looks anything like it.) If it is used in a positive feedback loop the circuit will oscillate at this zero phase frequency. The amplifier used in a Wien bridge oscillator must have gain at least 3. The gain is exactly 3 the circuit will take forever to start oscillating, if it is much more than three the amplitude of oscillations will increase until the amplifier limits and the sine wave output becomes distorted. Wien bridge oscillators therefore contain agc circuitry to stabilize their output level at some preset value.

WIEN BRIDGE OSCILLATOR



R_{TH1} IS AN NTC THERMISTOR
 AT AMBIENT TEMPERATURE $R_{TH1} \geq 4R_2$
 WHEN $V(rms)_{TH1} = (.66 \text{ WANTED } V_O)$ THEN R_{TH1} SHOULD BE $2R_2$

