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SECTION XI

OP AMP SUBTLETIES

ADOLFO GARCIA, JAMES WONG, WALT KESTER, CHRIS HYDE, JAMES BRYANT, PAUL BROKAW

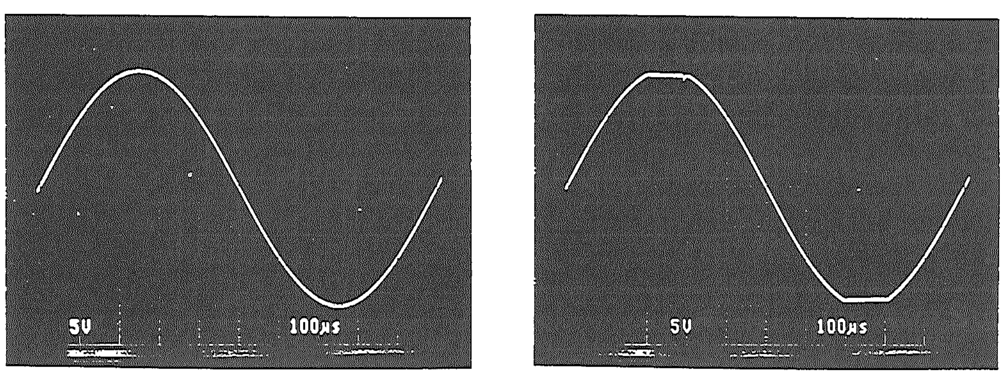
OPERATIONAL AMPLIFIER OUTPUT VOLTAGE PHASE REVERSAL

ADOLFO A. GARCIA

In vibration translation, process control, instrumentation, and servo system applications where sensitive sensors are used for recording and monitoring, oftentimes signals from these sensors can exceed an amplifier's common-mode voltage range during a fault condition or a power-up sequence. In the most extreme cases, fault conditions may generate input signals greater than the amplifier's supply voltages. Although not protected in the latter case, many, if not most all, operational amplifiers have protection built into them to safeguard the device from permanent damage should large differential signals appear at the device's input terminals. In many cases where input signals do exceed an amplifier's common-mode range, the amplifier behaves as a clipping circuit. Shown in Figure 11.1 is an example of the industry-standard OP-07 operational amplifier in a voltage follower configuration where the input voltage has exceeded the amplifier's common-mode range. Note how the amplifier clips the input signal without damage to the amplifier or the load.

BEHAVIOR OF OP-07 WHEN INPUT IS DRIVEN BEYOND LINEAR COMMON MODE RANGE

INPUT OUTPUT



VERTICAL SCALE: 5V / div.
HORIZONTAL SCALE: 100µs / div.

Figure 11.1

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However, not all operational amplifiers behave as clipping circuits. Because of varied designs used in the input and gain stages, some operational amplifiers behave unusually when the input signal voltages exceed the amplifier's input common-mode range. For example, high input impedance amplifiers, such as the OP-42 and the AD711, use ion-implanted, p-channel JFETs in the input stage to reduce the bias currents to the picoamp range and, at the same time, to raise the input impedance levels above $10^9\ \Omega$. As shown in Figure 11.2, the output of these amplifiers exhibit an output voltage phase inversion when the input signal exceeds the amplifier's linear, negative common-mode range.

Note that the output is suddenly driven to the positive output voltage limit and remains there until the input signal level returns within the common-mode range of the amplifier. This occurs most often with BiFET operational amplifiers configured as voltage followers in dual supply applications. Operational amplifiers designed for single supply applications also suffer from this malady and will be discussed elsewhere.

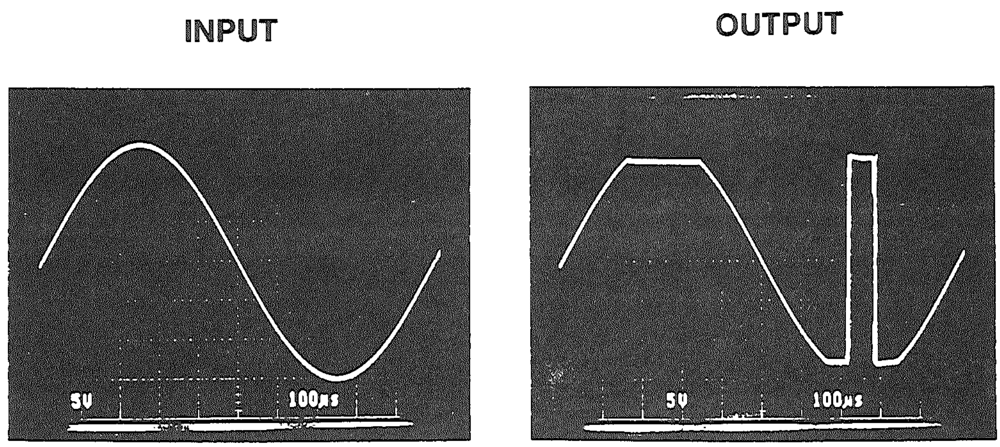
This effect in BiFET operational amplifiers can cause a very dangerous lock-up condition in mechanical system applications where the amplifiers act as elements controlling a system's servo loops. In one particular application, a customer who used BiFET amplifiers in the servo loops of a shaker table controller noted that he could consistently *launch* 100 lb. objects across his lab every time he induced a failure mode in his system. In one way, he was quite pleased that he had stumbled upon a new catapult system and wondered aloud if there would be a military application for it . On the other hand and not surprisingly, he was quite alarmed by the effect. Situations where a system failure could drive the input of the amplifier beyond its common-mode range occurs in roughly 2 to 3 % of all applications that use BiFET amplifiers. It is still important to understand what is happening and prevent the effect from occurring.

Shown in Figure 11.3 is the vertical cross section of an ion-implanted, p-channel JFET which is a good starting point for understanding the cause of amplifier output phase reversal.

In the equivalent circuit shown in Figure 11.3b, there are two parasistic pnp transistors directly under the source and drain diffusions. These parasitic devices typically exhibit very low current gains and contribute very little excess current flow during phase reversal. However, if the input exceeds the negative supply voltage (a condition that must always be avoided), these devices energize and can lead to destruction of the amplifier due to excessive current flow and power dissipation in the die.

At this point, a simplified schematic diagram of a BiFET operational amplifier can be used to examine how the forward-biasing of an input transistor's drain-gate diode leads to the phase reversal effect. Shown in Figure 11.4 is a schematic of one such BiFET amplifier, the OP-42. Note that a highlighted path illustrates the devices involved.

OP-42 OUTPUT PHASE REVERSAL WHEN INPUT DRIVEN WELL BEYOND LINEAR NEGATIVE COMMON MODE RANGE



VERTICAL SCALE: 5V / div.
HORIZONTAL SCALE: 100µs / div.

Figure 11.2

VERTICAL CROSS SECTION AND EQUIVALENT CIRCUIT FOR ION-IMPLANTED, P-CHANNEL JFET

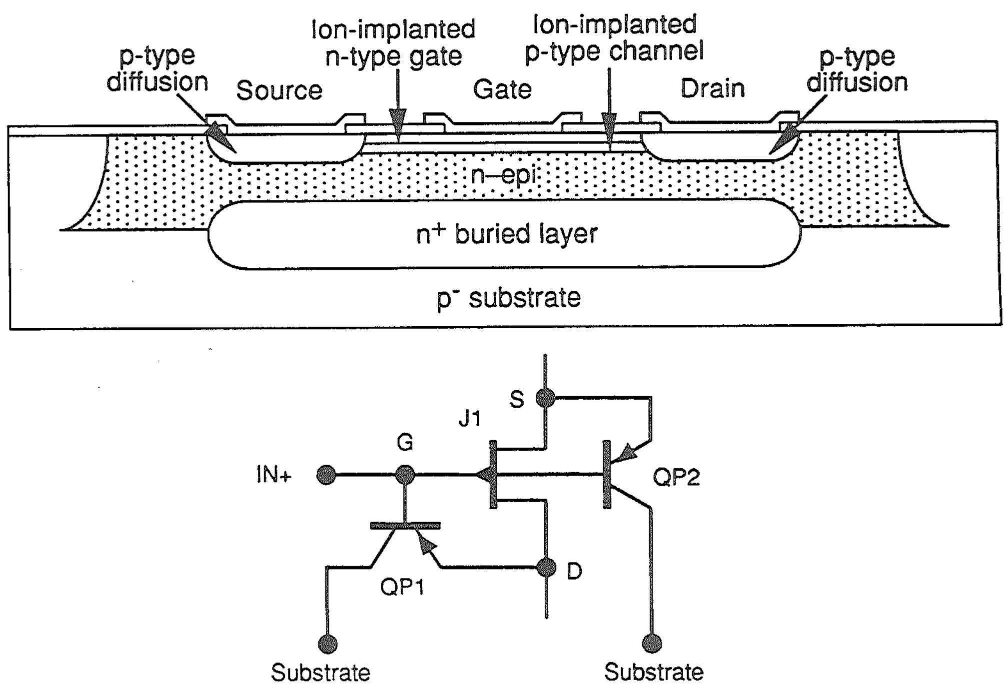


Figure 11.3

OP-42 SIMPLIFIED SCHEMATIC SHOWING DEVICES INVOLVED DURING OUTPUT PHASE REVERSAL

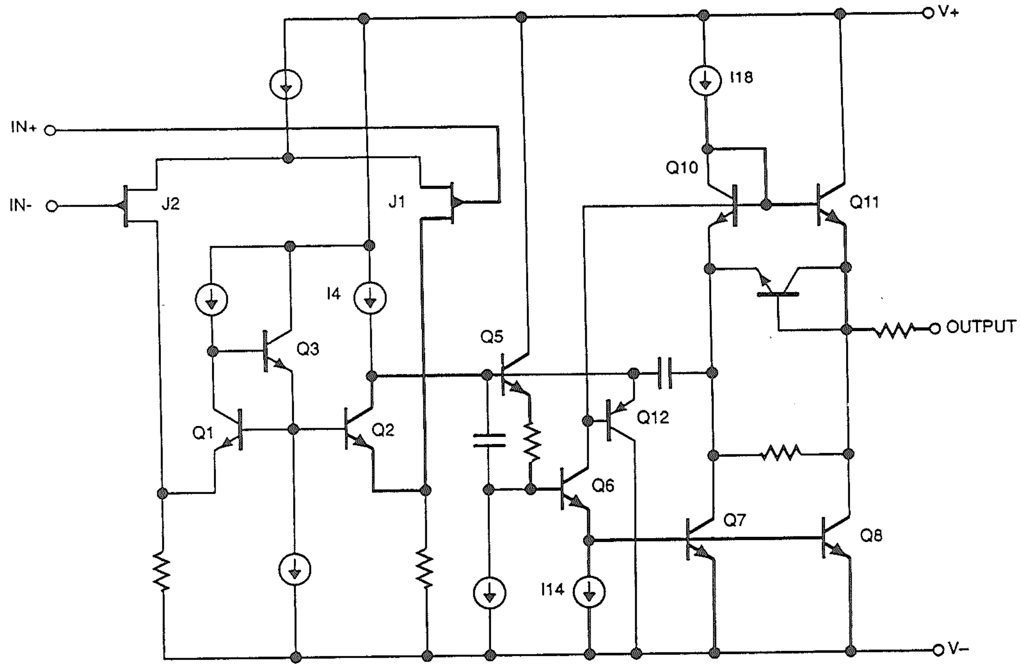


Figure 11.4

Under normal operating conditions in the OP-42, the drain voltages of J1 and J2 are typically 1 V above the negative supply rail. As the input voltage approaches -14 V to -14.5 V, the drain-gate pn junction of J1 forward biases providing a path for current to flow through its gate metal. In fact, when an amplifier exhibits phase reversal, there is a significant increase in the amplifier's input bias current. The phase reversal effect can be localized to the amplifier's second gain stage — a folded cascode circuit comprised of Q1, Q2, Q3, and current sources I3 and I4. The collector potentials of the cascode transistors, Q1 and Q2, are approximately 3.5 V above the negative supply. As the input voltage approaches -11.5 V, a pnp transistor, Q12, energizes and clamps the negative output voltage to approximately -12 V. As the input signal continues towards the negative supply, J1's drain-gate diode begins to forward bias, causing Q2 to supply that current. At this point, Q2's emitter ceases to control J1's drain voltage and begins to track the input voltage. Q2 continues to supply current for J1's forward-biased drain-gate diode until Q2 is forced into saturation because of the limited current available from I4. Once Q2 is saturated, its collector and emitter voltages are nearly equal causing the base potential of Q5 to track the input voltage. As the base of Q5 continues to the negative supply, so does Q6's emitter voltage until the current source, I14, is forced into saturation. When I14 saturates, both Q7 and Q8 are forced into cutoff.

Normally, current source I18 supplies Q10, Q7, and the base current for the output transistor, Q11. When Q7 and Q8 are turned off, all of I18's current is then pumped into Q11's base forcing the output transistor into saturation. The result is (quite dramatically) that the output voltage is pulled towards the positive supply rail; hence, the output voltage reverses phase. The device does not suffer permanent damage as a result of phase reversal provided the input is never allowed to exceed the negative supply rail.

The Fix

Since the effect is caused by the saturation of Q2 due to the forward-biasing of J1's drain-gate pn junction, the solution to the phase reversal effect is preventing Q2 from being driven into saturation. Placing a resistor in series with the gate of J1 (the non-inverting terminal of the amplifier), as shown in Figure 11.5, limits the current flow through Q2. As illustrated in Figure 11.6, a 7.5 kΩ resistor effectively prevents the OP-42's output voltage from phase reversal over the full MIL-temperature range.

For other BiFET amplifiers, Figure 11.7 summarizes the series resistor values required to suppress output voltage phase reversal. Also shown for comparison in the table are the equivalent input voltage noise spectral densities of the amplifiers alone and with the an external series resistor. In using this technique, the increase in overall system noise and the new pole created with the stray input capacitance must be evaluated.

ELIMINATING OUTPUT VOLTAGE PHASE REVERSAL IN THE OP-42

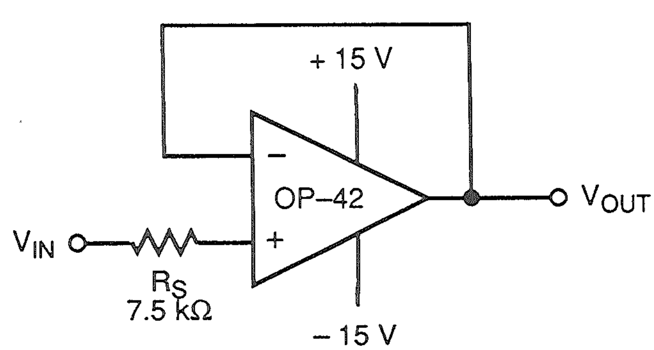


Figure 11.5

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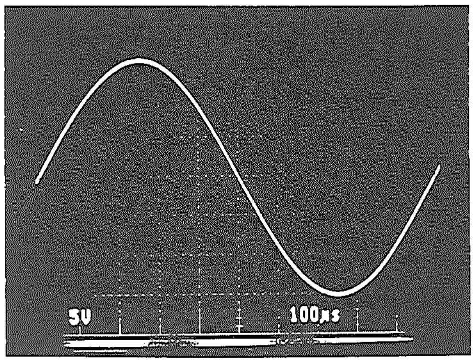
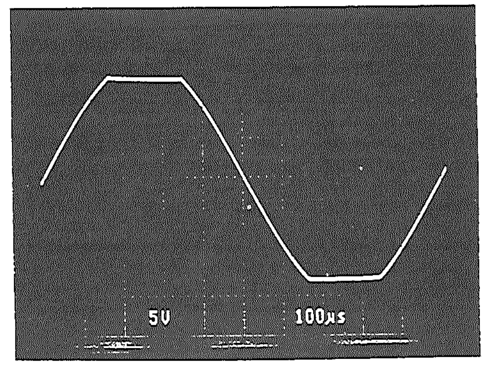
EFFECTIVENESS OF SERIES RESISTOR IN ELIMINATING PHASE REVERSAL IN THE OP-42

INPUT



OUTPUT



VERTICAL SCALE: 5V / div.
HORIZONTAL SCALE: 100µs / div.

Figure 11.6

SERIES RESISTANCE VALUES FOR BIFET AMPLIFIERS TO PREVENT PHASE REVERSAL

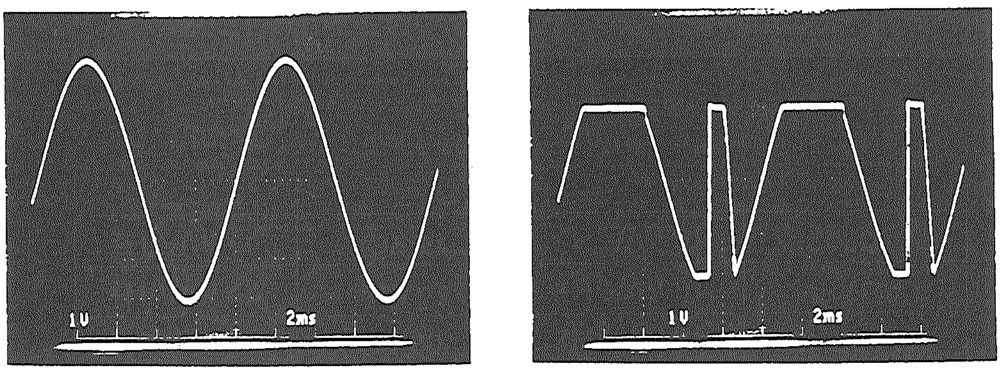
Device	E_n @ 1 kHz	Series Resistor	Total Noise (Amplifier + Resistor)
OP–42	12 nV/$\sqrt{\text{Hz}}$	7.5 kΩ	16 nV/$\sqrt{\text{Hz}}$
OP-249	16 nV/$\sqrt{\text{Hz}}$	36 kΩ	29 nV/$\sqrt{\text{Hz}}$
OP-282/482	36 nV/$\sqrt{\text{Hz}}$	200 kΩ	68 nV/$\sqrt{\text{Hz}}$
AD711	16 nV/$\sqrt{\text{Hz}}$	14 kΩ	22 nV/$\sqrt{\text{Hz}}$
AD712	16 nV/$\sqrt{\text{Hz}}$	14 kΩ	22 nV/$\sqrt{\text{Hz}}$

Figure 11.7

As previously mentioned, BiFET amplifiers are not the only devices susceptible to output phase reversal. The effect also manifests itself in operational amplifiers designed for single-supply operation. In these cases, these amplifiers exhibit a phase reversal when the input signal drops below the negative supply rail. For example, Figure 11.8 illustrates an OP-90 configured as a voltage follower that uses a single + 5V supply. When the input exceeds the negative supply – in this case, ground – by about 0.6V, the OP-90's output voltage goes into phase reversal.

PHASE REVERSAL IN THE OP-90
SINGLE SUPPLY OP AMP

INPUT **OUTPUT**



VERTICAL SCALE: 1V / div.
HORIZONTAL SCALE: 2ms / div.

Figure 11.8

In order for the input common-mode range to include the negative supply in these amplifiers, the input stages use pnp transistors, usually lateral devices. The input stages are intentionally designed to have slightly forward-biased collector-base junctions when the inputs are at zero volts. These junctions could have as much as 80 mV to 120 mV of forward bias across the collector-base junction. There are other active devices to consider in addition to the input transistors under this condition. Consider the vertical cross section of a typical lateral pnp transistors Figure 11.9, there are two parasitic vertical pnp transistors to the substrate: one under the collector diffusion and one under the emitter diffusion.

VERTICAL CROSS SECTION AND EQUIVALENT CIRCUIT FOR LATERAL PNP TRANSISTOR

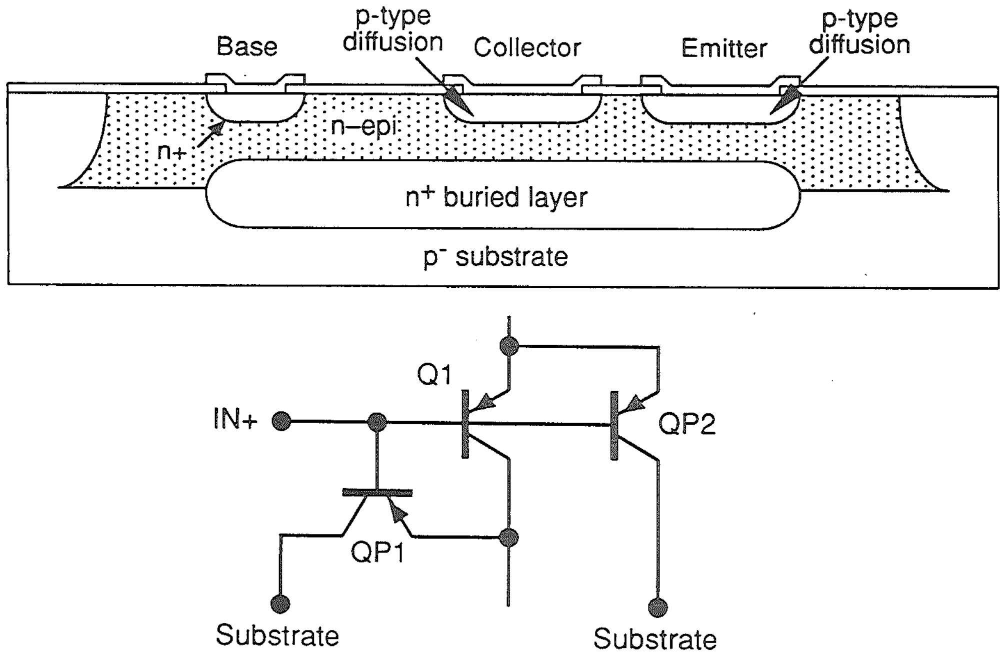


Figure 11.9

OP-90 SIMPLIFIED SCHEMATIC SHOWING DEVICES INVOLVED DURING OUTPUT PHASE REVERSAL

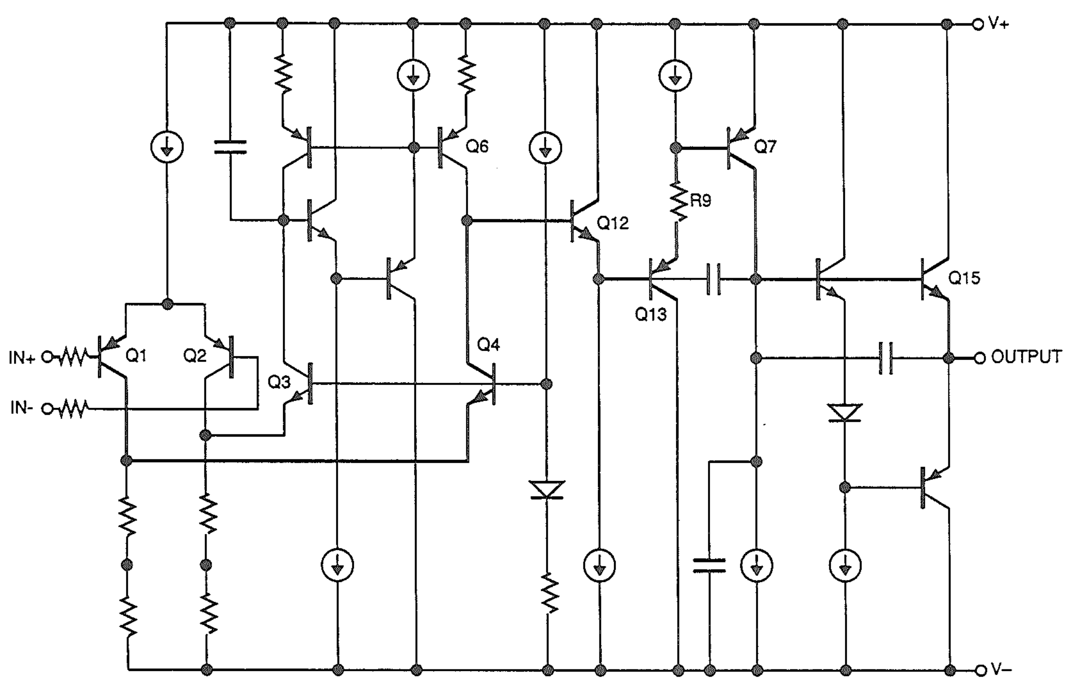


Figure 11.10

Since the collector junction of Q1 is biased slightly above the negative supply, the base-emitter junction of QP1 also begins to conduct current. As the input of the amplifier goes below ground by approximately 0.6 V, both QP1 and QP2 are forced into saturation. Even though three pn junctions are turned on, any increase in input bias current during phase reversal is slight because the highly resistive epitaxial layer limits current flow through the base junctions.

Using the same approach as for the BiFET amplifiers, a simplified schematic of the OP-90 can be used to discover the cause of phase reversal in single-supply amplifiers. Its simplified schematic is shown in Figure 11.10. The devices involved in phase reversal are highlighted.

When Q1's input drops below the negative supply, its collector junction tracks the input causing a transistor, Q4, in the folded cascode gain stage to saturate because of the limited amount of current in the stage provided by Q6. With Q4 in saturation, the base voltage of the complementary darlington pair, Q12 and Q13, also tracks the input voltage. This causes a larger voltage drop across R9 and, in turn, causes larger base currents to flow in Q7. In the design of the OP-90, Q7 is a large lateral pnp transistor whose function is to provide the base drive for the output drive transistor, Q15. As Q7's collector current increases in the absence of a load at the output of the amplifier, Q15 is driven into saturation. The output voltage is then pulled up to the amplifier's positive output voltage limit.

THE FIX FOR THE OP-90

Since there is no marked increase in the input bias current of the OP-90 at the point of phase reversal, the best method for preventing this effect is to clamp the input voltage to ground using a schottky diode, as shown in Figure 11.11. This technique is best used when the amplifier is configured as a voltage follower. Figure 11.12, shows that this technique is quite effective in preventing output phase reversal.

A schottky diode typically exhibits a turn-on voltage of 0.3 V to 0.4 V which is below the voltage required to forward bias Q1's collector-base junction. A 1 kΩ resistor in series with the input of the OP-90 is optional; however, using it is a good idea because it serves to limit the current flow in the schottky diode should the input voltage become significantly larger than the forward voltage drop of the schottky diode. In single-supply + 12 V or + 15 V applications, be mindful of schottky diodes' low breakdown voltages and comparatively high reverse leakage. Large reverse-bias leakage currents may flow through the diode and might appear as input bias current errors in the amplifier circuit.

ELIMINATING OUTPUT PHASE REVERSAL IN THE OP-90

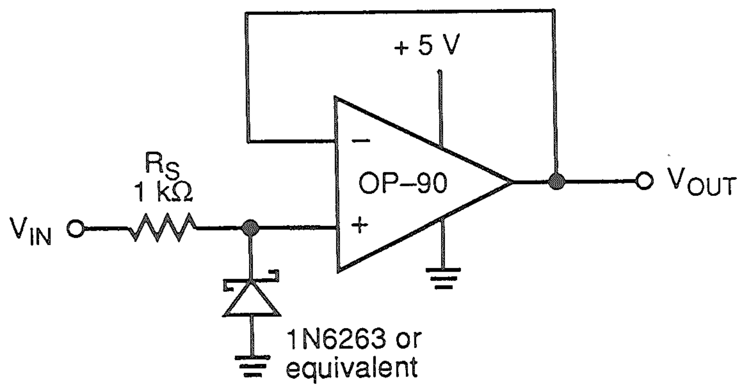
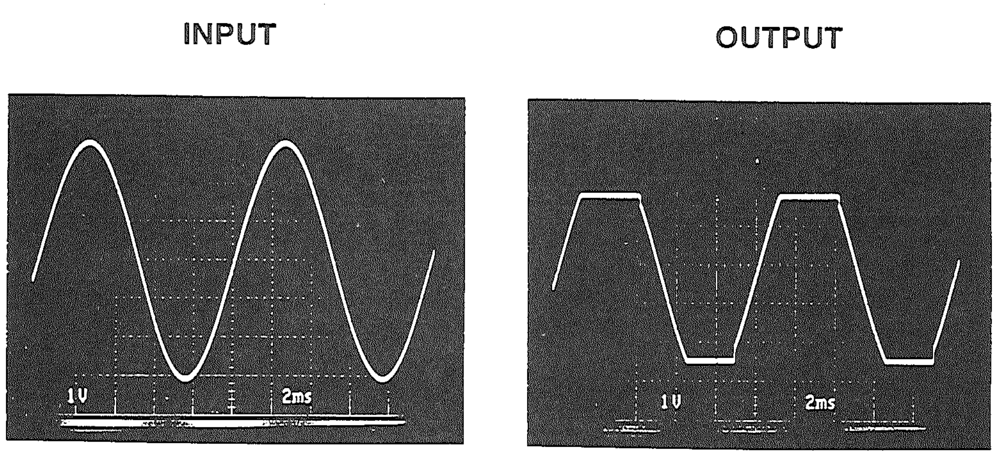


Figure 11.11

EFFECTIVENESS OF SCHOTTKY DIODE IN PREVENTING PHASE REVERSAL IN SINGLE-SUPPLY OP-90



VERTICAL SCALE: 1V / div.
HORIZONTAL SCALE: 2ms / div.

Figure 11.12

Does Op Amp Open-Loop Gain Non-Linearity Affect Accuracy?

James Wong

This question has a two-part answer. In the case of a high gain circuit, the answer is YES, and gain nonlinearity can be a significant source of error. On the other hand, its nonlinear effect is much diminished at lower gains. Unlike input offset voltage and common-mode errors which are generally specified and can be measured at a fixed input voltage or an input voltage range, open-loop gain nonlinearity is a subtle error produced by output swings. Generally this error source is neither specified nor quantified. Yet it may be a real part of an op amp's error source. If present, it degrades the accuracy of a circuit without being noticed.

What Is Open-Loop Gain Nonlinearity?

Nonlinearity exists if an op amp experiences a change in its open-loop gain as the output swings through its linear voltage range. Because gain is the ratio of output to input, any nonlinearity is referred to the input and shows up as a change in input offset. Thus in an amplifier circuit, this translates to a deviation of the output voltage from the expected output because of the offset change. This occurs even though the input common-mode voltage remains relatively constant. Clearly, this can have appreciable effect on the accuracy of an amplifier. The error is especially pronounced if the amplifier is operating at high closed-loop gains.

To be sure, this input offset error change is indistinguishable from any input offset voltage change due to other factors such as self-heating, or bias shifts. Nonetheless, in an actual application, the error is real and directly affects the output accuracy of an amplifier.

Figure 11.13 shows a useful test circuit that can be used to measure the gain linearity behavior of an op amp. The nonlinear behavior can take many different shapes such as bowing, double bowing, or S-curved. Every device type is different.

The linear ramp input causes the output to swing through a +/-10V range. This voltage is measured against the pseudo summing node V_Y which represents the input voltage of the op amp magnified by a gain-factor (that is determined by $R_G/10\ \Omega$). The open-loop gain is the inverse of the slope of the curve, dV_X / dV_Y, multiplied by the gain-factor. If the open-loop gain is linear, V_Y will be linearly proportional to V_X. Thus a straight line will be plotted. Otherwise a nonlinear line will be drawn. An offset adjustment may be necessary to center the curve on the plotter.

Component values shown in the figure are for use with open loop gains from 1 million to 10 million. The value of R_G may be reduced for gains less than 1 million to increase measurement resolution.

Figure 11.14 shows an example of the measured behavior of an industry-standard precision op amp. Of course, every op amp has its own unique characteristic. One should measure this parameter to assure the right op amp is chosen for the application.

OPEN-LOOP GAIN NON-LINEARITY TEST CIRCUIT

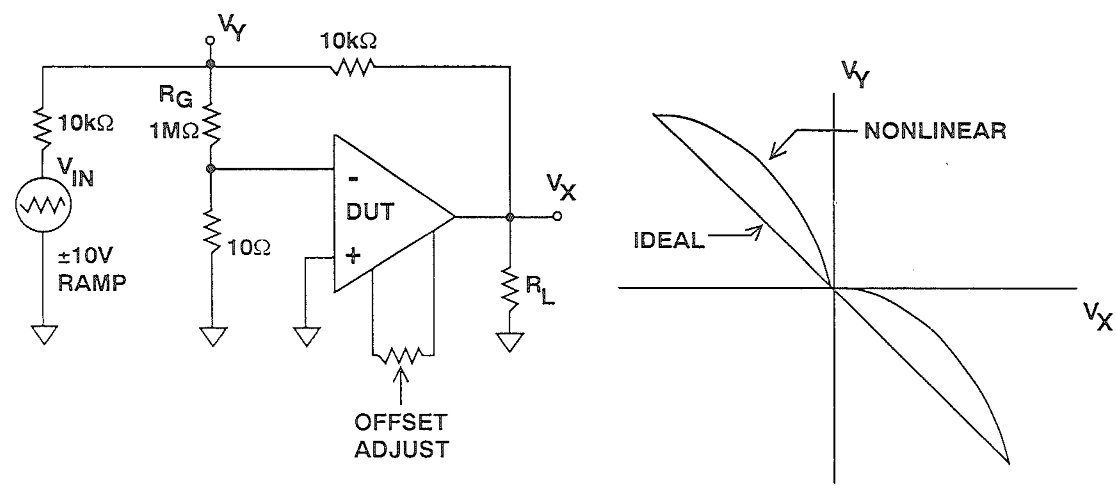


Figure 11.13

MEASURED OPEN-LOOP GAIN BEHAVIOR AS A FUNCTION OF OUTPUT VOLTAGE

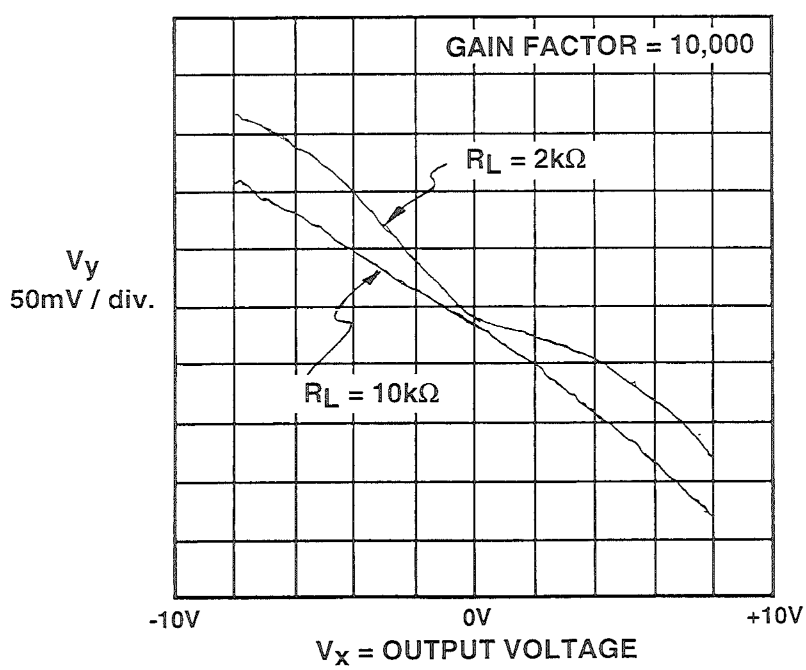


Figure 11.14

What Causes Gain Nonlinearity ?

The severity of the nonlinearity varies widely from device type to device type, and may be dependent on the load it is driving. Although many precision op amps do exhibit some amount of nonlinearity, it is important for the designer to know up-front whether they have high enough open-loop gain to minimize its impact on accuracy. Only a few exhibit no error at all. These are the most desirable and produce the best accuracy.

Gain nonlinearity can come from many sources, depending on the design and layout of the chip. One possible source is thermal feedback. As the output swings, the amount of output current also changes, causing the power dissipation of the output device to change. Should the resulting temperature shift cause the gain stage to drift, a nonlinear gain will result. Similarly, if this temperature shift migrates to the input stage's differential pair in an unbalanced way, an additional offset error will result, which produces an apparent gain shift.

If temperature shift is the sole cause of the nonlinearity error, it can be postulated that minimizing the output loading will help. To verify this, measure the linearity behavior under a light load (>10 kΩ) and compare with the case of a heavy load (<2 kΩ). If nonlinearity is detected with the heavy load and not with the light load, then the solution is to reduce the loading. An example of this is the OP-97 low-power precision op amp, whose open-loop gain curves under different loads are shown in Figure 11.15 below.

GAIN NONLINEARITY INCREASES AS LOAD RESISTANCE DECREASES, SHOWN FOR OP-97 PRECISION OP AMP

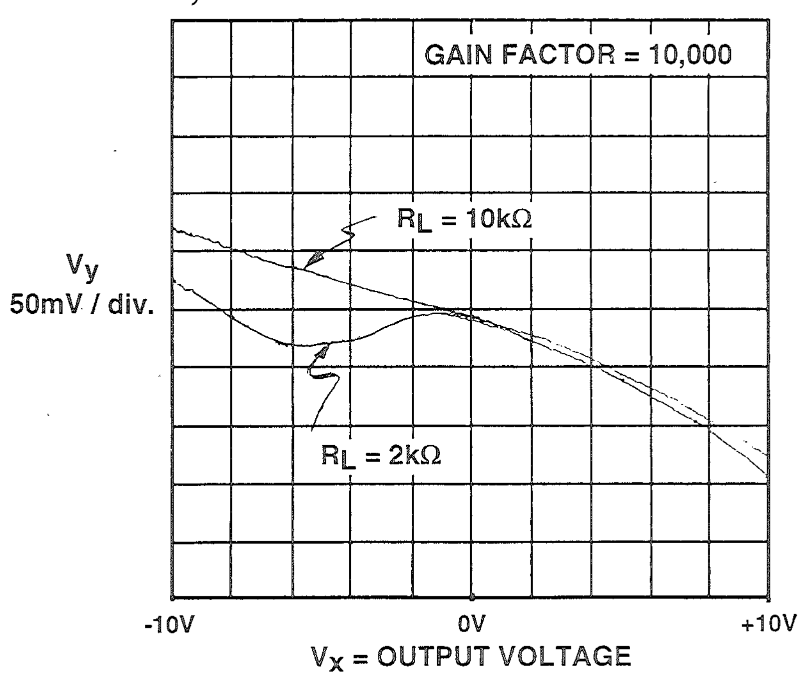


Figure 11.15

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With this knowledge, one should operate the OP-97 with no less than 10 kΩ load, which in most applications is not a problem. However, not all op amps' gain nonlinearity can be cured this way, as some are inherently nonlinear regardless of load. For these devices, there must be other second-order effects inherent of the design that cause the nonlinear behavior. For example, a sample of the LT1077 op amp exhibited rather poor linearity as evidenced in the measured behavior in Figure 11.16. Reducing the load did not help eliminate the problem.

SOME OP AMPS EXHIBIT POOR GAIN LINEARITY REGARDLESS OF LOAD CONDITIONS

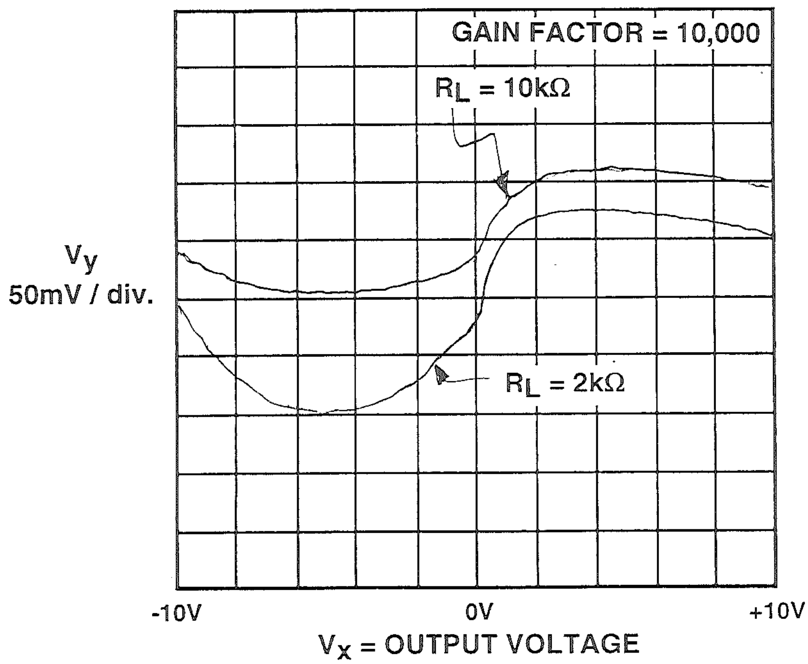


Figure 11.16

HOW DOES GAIN-NONLINEARITY EFFECT ACCURACY ?

From this nonlinearity analysis, we can put this error term into proper perspective. For example, using the op amp in Figure 11.16 which has an average open-loop gain of 2,000,000, assume an application's closed-loop gain is 1,000, then the closed-loop gain error in percent is 0.05% - if the open-loop gain remains constant. However, at the output voltage between 0V and +2V, the gain appears to have fallen off to approximately 300,000. This produces an appreciable amount of inaccuracy which should be included in the worst-case error budget.

GAIN NONLINEARITY EFFECT ON GAIN ERROR

- Assumptions: $A_{VOL} = 2{,}000{,}000$

  $A_{VCL} = 1{,}000$

- Gain Error (closed-loop) $= \dfrac{A_{VCL}}{A_{VOL}} = \dfrac{1000}{2000000} \times 100\%$

  $= 0.05\ \%$

- If A_{VOL} Drops To 300,000 Due To Gain Nonlinearity,

- Gain Error (closed-loop) $= \dfrac{1000}{300000} \times 100\%$

  $= 0.33\%$

Figure 11.17

From this analysis, clearly gain nonlinearity cannot be ignored if absolute precision is to be attained. There are two ways to solve the problem from a system designer's perspective. One is to simply choose an op amp that does not exhibit this nonlinear behavior. Examples of a linear op amp are the OP-177 and the AD707, which are among the industry's highest precision op amps. Their open-loop gain linearities are shown in Figure 11.18 and Figure 11.19, respectively. The OP-177 open loop gain is approximately 8 million, while that of the AD707 is approximately 25 million.

The best way to minimize the nonlinear impact is to make sure the amplifier open loop gain is sufficiently high such that in spite of its presence, there is ample loop-gain to correct for any gain variations.

11

THE OP-177 IS DESIGNED FOR LINEAR GAIN TO INSURE PRECISION PERFORMANCE

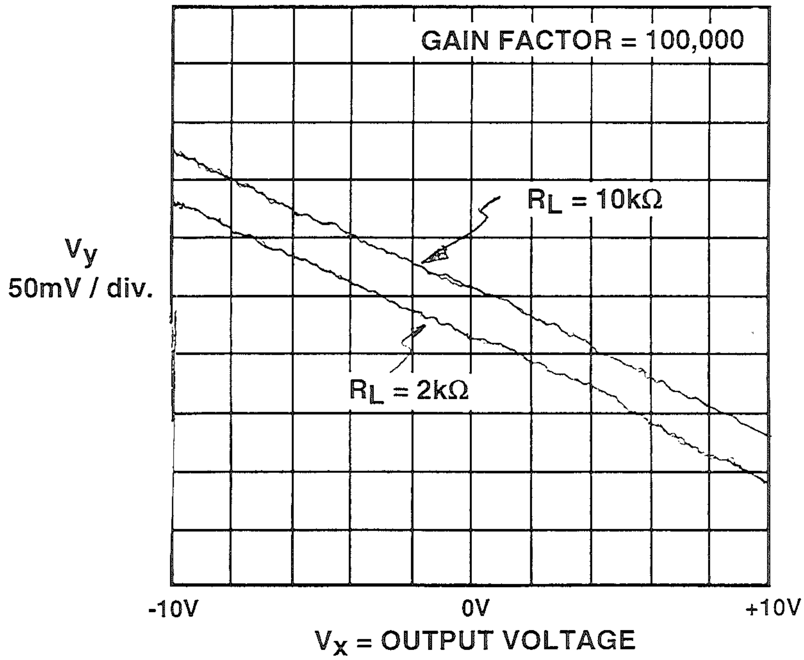


Figure 11.18

THE AD707 EXHIBITS LINEAR GAIN TO INSURE CIRCUIT ACCURACY

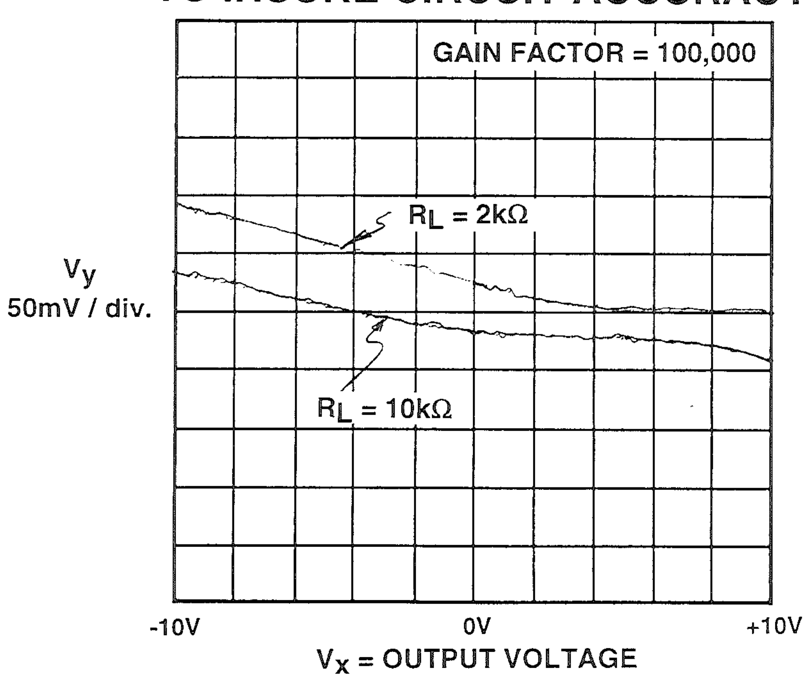


Figure 11.19

HIGH SPEED OP AMP SETTLING TIME MEASUREMENTS

WALT KESTER

Settling time of an op amp is defined as the interval of time from the application of an ideal step function input until the closed-loop amplifier output has entered and remains within a specified error band. This definition encompasses the major components which comprise settling time. They include (1) propagation delay through the amplifier; (2) slewing time to approach the final output value; (3) the time of recovery from the overload associated with slewing and (4) linear settling to within the error band.

The measurement of settling times of high speed amplifiers is obviously a challenge and needs to be done accurately to assure the user that the amplifier is worth consideration for the application.

Measuring settling times of inverting op amps may be accomplished by creating a false summing junction with a resistive divider connected between the amplifier output and the input as shown in Figure 11.20. Care should be taken to insure that the time constant at the false summing junction created by the scope capacitance is less than that of the amplifier being tested. The back-to-back diodes help prevent oscilloscope overdrive.

The circuit shown in Figure 11.21. may be used if extra gain is required between the summing junction and the scope. This circuit is useful for measuring settling times of over 80ns to 12-bit accuracy (0.01%). The diagram shows the test setup for the AD843 high speed FET input op amp which settles to 0.01% in less than 135ns. The settling time characteristics are shown in Figure 11.22.

MEASUREMENT OF INVERTING MODE SETTLING TIME USING A FALSE SUMMING JUNCTION

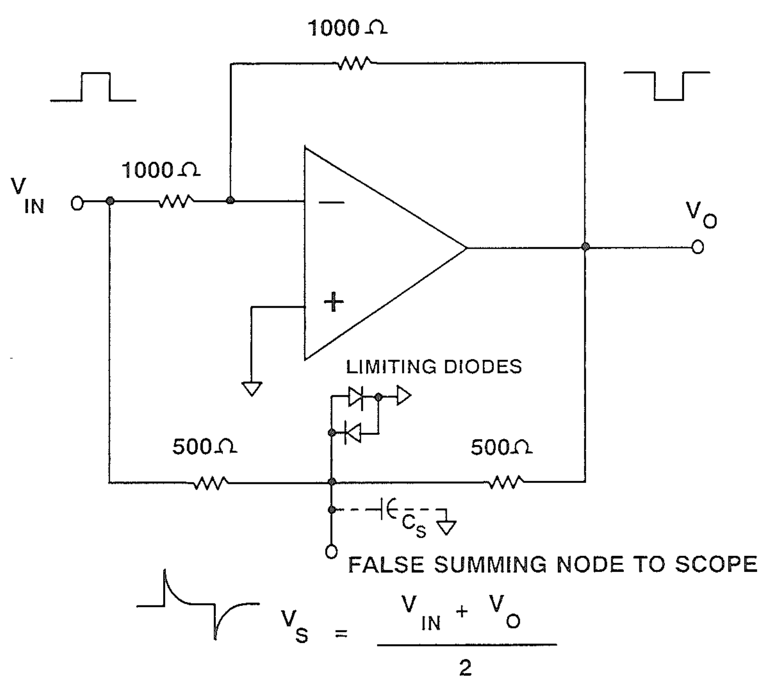


Figure 11.20

GREATER THAN 80ns SETTLING TIME CIRCUIT WITH SUMMING JUNCTION GAIN

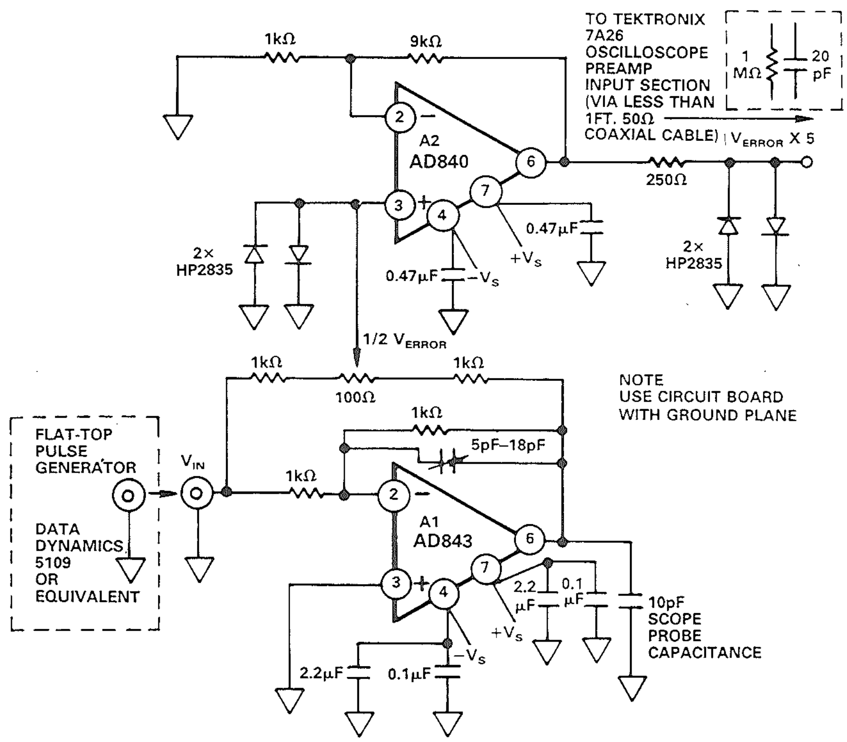
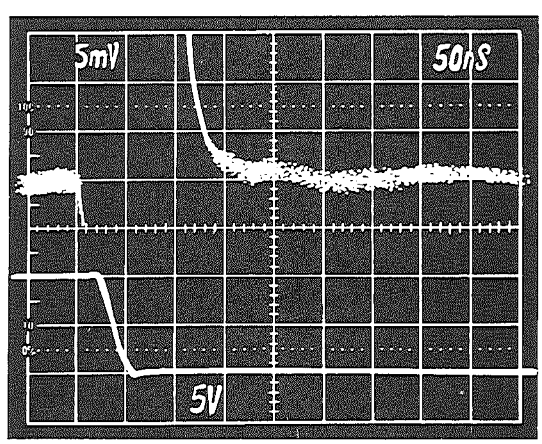


Figure 11.21

AD843 FET INPUT OP AMP SETTLING CHARACTERISTIC FOR +10V TO 0V STEP



UPPER TRACE: AMPLIFIED ERROR VOLTAGE (0.01%/DIV)

LOWER TRACE: OUTPUT OF AD843 (5V/DIV)

Figure 11.22

The false summing junction technique described above will not work with a non-inverting amplifier for obvious reasons. Most oscilloscopes will not allow direct observation of the output waveforms true settling time because they will be severely overdriven when operating at the required resolution. The technique shown in Figure 11.23 uses a diode network on the amplifier output to limit the swing of the 5V output step to 0.75V. This step can be observed directly by the scope with approximately 25mV/division vertical sensitivity without overdriving the scope input. Actual measurements on an amplifier are shown in Figure 11.24. The Schottky diode network on the input of the amplifier is used to generate a flat pulse. This network must be placed less than 0.5 inches from the amplifier input in order to minimize parasitic effects.

When making settling time measurements to 0.02% to less than 15ns, especially in the non-inverting mode, there are problems with all of the above techniques. At these speeds, it is best to use a high-speed sampling plug-in, such as the Data Precision Model 640 (using Data Precision 6100 Waveform Digitizer). This plug-in uses a 16-bit successive approximation ADC with the comparator actually an integral part of the probe tip. Using this architecture, the scope input is not sensitive to overdrive, and the settling time measurement can be made directly as shown in Figure 11.25. for the AD9620 unity-gain buffer. The output pulse is 2V, and the vertical resolution of the digitized output is 0.4mV, or 0.02%/division.

SETTLING TIME TEST CIRCUIT LIMITS OSCILLOSCOPE OVERDRIVE

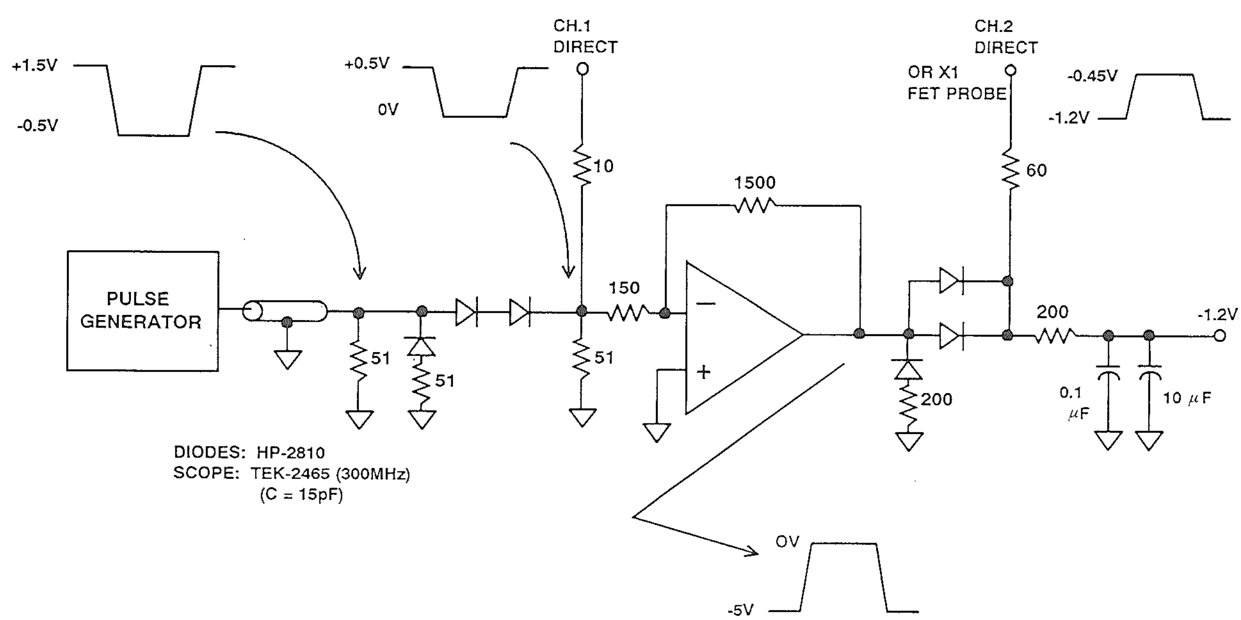
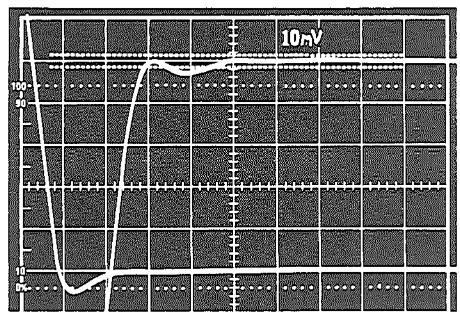


Figure 11.23

SETTLING TIME MEASUREMENT USING OVERDRIVE-LIMITING TEST CIRCUIT



GAIN = −10; 5V OUTPUT; ERROR WINDOW (±5mV) = 0.1%; 5ns/DIV

Figure 11.24

AD9620 BUFFER SETTLING TIME MEASURED DIRECTLY WITH DATA PRECISION 640 DIGITIZING PLUG-IN

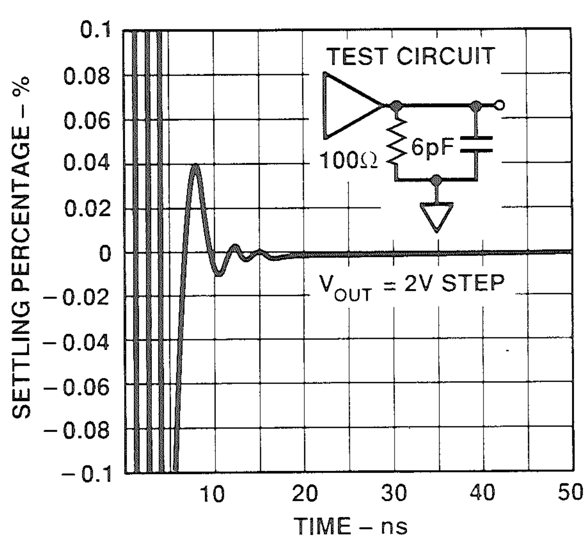


Figure 11.25

In order to duplicate test results such as those described above, excellent high speed layout techniques must be observed! Large-area ground plane, compact layout, appropriate decoupling is mandatory. Molded plastic socket assemblies are absolutely forbidden! If sockets must be used, individual pin sockets on each IC pin may be acceptable, provided the IC is mounted flush on the PCB. There must be no excess lead length between the tops of the pin sockets and the bottom of the IC.

OP AMP NOISE AND HOW TO AVOID IT

JAMES M. BRYANT

INTRODUCTION

This section of the seminar considers the two categories of noise found in op-amp circuitry and discusses how they can be minimized in practical applications.

These two categories of noise are the internal noise of the circuits, that is the op-amp itself and the components of its immediate circuitry, and external noise from elsewhere in the system. We shall consider DC error sources in this section as well, since it is quite reasonable to treat DC errors as noise.

TYPES OF NOISE

- NOISE GENERATED WITHIN A CIRCUIT
- EXTERNAL NOISE (Which is a signal which is wanted, in a place where it isn't.)

Figure 11.26

UNCORRELATED NOISE VOLTAGES ADD AS THE ROOT OF THE SUM OF THE SQUARES

- For N uncorrelated noise voltages V_1, V_2,....V_N the total noise voltage Vn is given by

$$V_n = \sqrt{V_1^2 + V_2^2 + \ldots V_N^2}$$

- If any noise voltage is more than 3-5 times another we may generally ignore the smaller.

Figure 11. 27

The general principle of noise calculation is that uncorrelated noise sources add in a root sum of squares manner, which means that if a noise source has a contribution to the output noise of a system which is less than 20% of the amplitude of the noise from some other noise source in the system then its contribution to the total system noise will be less than 2% of the total and that noise source can almost invariably be ignored - in many cases noise sources smaller than 33% of the largest can be ignored. Remember, though, that design changes can change the relative amplitudes of noise sources so that a previously unimportant noise source becomes dominant. This means that all noise sources must be re-evaluated when a design is changed.

INTERNAL CIRCUIT NOISE

The three noise sources in an op-amp circuit are the voltage noise of the op-amp, the current noise of the op-amp (there are two uncorrelated sources, one in each input), and the Johnson noise of the resistances in the circuit.

THERE ARE THREE SOURCES OF NOISE WITHIN A CIRCUIT

- Voltage Noise of the Op-Amp
- Current Noise of the Op-Amp
- Johnson Noise of the Resistors in the Circuit

Figure 11.28

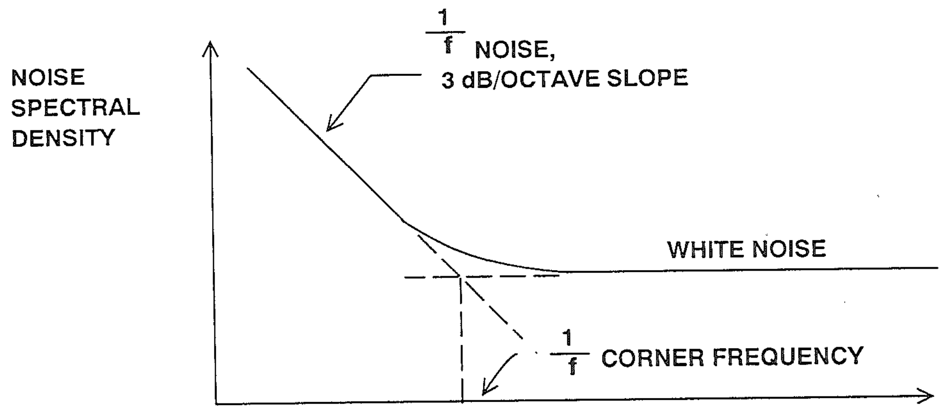
Operational amplifier noise has two components - white noise at medium frequencies and low frequency noise whose spectral density is inversely proportional to the square root of the frequency. It should be noted that, though both the voltage and the current noise may have the same characteristic behavior, in a particular amplifier the 1/f corner frequency is not necessarily the same for voltage and current noise (it is usually specified for voltage noise).

The low-frequency noise is generally known as 1/f noise (the noise *power* obeys a 1/f law - the noise voltage or noise current is proportional to $1/\sqrt{f}$). The frequency at which the 1/f noise spectral density equals the white noise is known as the "1/f Corner Frequency" and is a figure of merit for the op-amp, with low values indicating better performance. Values of 1/f corner frequency vary from a few Hz for the most modern low noise low frequency amplifiers to several hundreds, or even thousands, of Hz for some high speed op-amps where process compromises favor high speed rather than low frequency noise.

Voltage noise, which appears in series with the two input terminals of the op-amp, can have spectral densities of under $1nV/\sqrt{Hz}$ in the best low noise op amps.

11

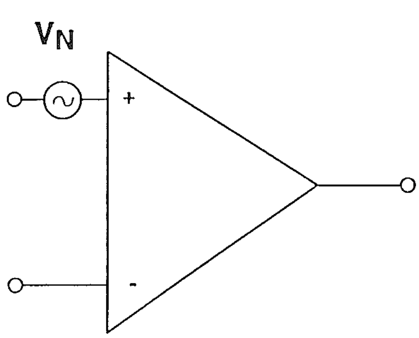
OP AMP NOISE CHARACTERISTICS



■ Voltage & Current Noise have the same characteristics but the 1/f Corner Frequencies for Voltage and Current Noise in the same Amplifier are not necessarily the same.

Figure 11.29

VOLTAGE NOISE IN OP AMPS



- The best low-noise Op-Amps have voltage noise

  $<1 nV/\sqrt{Hz}$

- 1/f Corner Frequencies can be as low as 2.7 Hz (OP-27) or as high as 1-2 KHz in high speed op-amps where low 1/f noise is sacrificed for better HF performance.

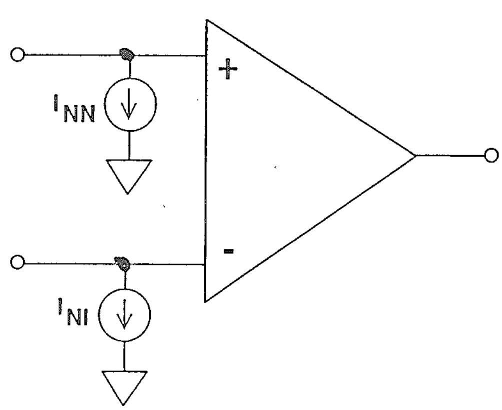
Figure 11.30

Current noise appears at each of the input terminals of an op-amp. In conventional (voltage feedback) op-amps the noise currents in each input are approximately equal in amplitude, though not strongly correlated (but they may be weakly correlated to the voltage noise). The noise currents in the inverting and non-inverting inputs of current feedback (transimpedance) op-amps are not matched in any way - they are neither equal nor correlated. Current noise varies very widely from type to type of amplifier, by at least 10^4:1. from under 0.1 fA/√Hz in the lowest bias current BIFET op-amps to several pA/√Hz in bias-compensated bipolar op-amps and in high speed op-amps.

Current noise is often inadequately specified. In op-amps where the input stage is a simple long-tailed pair of bipolar or field-effect transistors the whole of the bias current flows in the input junction and the current noise floor is equal to the Schottky noise (or "shot noise") of the bias current which can be calculated from the bias current. But in bias compensated amplifiers and current feedback amplifiers there are several currents flowing internally through each input node and it is not possible to predict the noise current from the external bias current and if the noise current is not specified it cannot be calculated. It is therefore important to be aware of the structure of an op-amp if its noise current is likely to be of importance.

The noise current of an op-amp is important if it flows in an impedance to produce a voltage, which the op-amp then amplifies. Consequently the impedance (Z) seen by each input of an op-amp produces a noise voltage of $I_n Z$, where I_n is the noise current at that input.

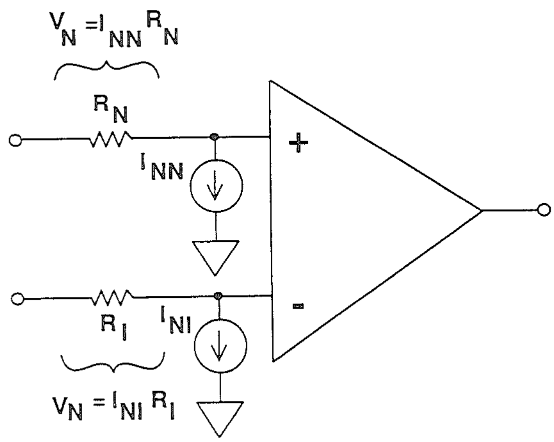
CURRENT NOISE IN OP AMPS



- In conventional (voltage feedback) op-amps the current noise in the inverting and non-inverting inputs are equal, but are not correlated.
- In current feedback (transimpedance) op-amps the current noise in the inverting and non-inverting inputs may be totally different (and is certainly uncorrelated).

Figure 11.31

INPUT NOISE CURRENT



- Conventional (Voltage Feedback) Op-Amps without bias-compensated input structures have white noise approximately equal to the Schottky Noise (or Shot Noise) of their Bias Current.

$$I_n = \sqrt{2IqB}$$

(I is the bias current, q is the electron charge and B is bandwidth)

- Current Feedback (Transimpedance) and Bias-Compensated Op-Amps do not.
- Current Noise flows in Op-Amp Input Source Impedances to produce noise voltages.

Figure 11.32

JOHNSON NOISE

If the impedance is resistive there is a second noise voltage to be reckoned with. All resistors have a noise voltage, known as thermal, or Johnson, noise, of $\sqrt{4kTBR}$, where k is Boltzmann's constant (1.38×10^{-23} J/K), T is the absolute temperature, B is the bandwidth, and R is the resistance.[1] Every resistance in a circuit contributes noise to it. Although the noise contributions from most resistors are insignificant the noise contribution of each resistor in a noise-critical circuit should be checked at some point during its design.

JOHNSON NOISE

- ALL RESISTANCES (BUT *NOT* REACTIVE IMPEDANCES) HAVE THERMAL NOISE OF

$$\sqrt{4kTBR}$$

- WHERE k IS BOLTZMANN'S CONSTANT (1.38×10^{-23} J/K)
- T IS THE ABSOLUTE TEMPERATURE
- B IS THE BANDWIDTH & R IS THE RESISTANCE
- EVERY RESISTANCE IN A CIRCUIT WILL CONTRIBUTE TO ITS NOISE
- A 1000 Ω RESISTOR GENERATES $4nV/\sqrt{Hz}$ @ 25°C

Figure 11.33

DOMINANT NOISE SOURCES

We have already pointed out that any noise source which produces less than one fifth of the noise of some other source can usually be ignored. (Both noise voltages must be referred to the same point in the circuit.) To analyze the noise performance of an op-amp circuit we must assess the noise contributions of each part of the circuit and determine which are significant. To simplify the following calculations we shall work with noise spectral densities rather than actual voltages to leave bandwidth out of the expressions (the noise spectral density, which is generally expressed in $nV/\sqrt{Hz}$, is equivalent to the noise in a 1 Hz bandwidth).

If we consider the circuit in Figure 11.34, which is an amplifier consisting of an op-amp and three resistors (R_1 represents the source resistance at node A), we can find six separate noise sources: the Johnson noise of the three resistors, the op-amp voltage noise, and the current noise in each input of the op-amp. Each has its own contribution to the noise at the amplifier output. (Noise is generally specified r.t.i., or referred to the input, but it is often simpler to calculate the noise at the output and then divide it by the *signal* gain (not the *noise* gain) of the amplifier to obtain the r.t.i. noise).

A signal applied to input A (B being grounded) sees a gain of

$$\frac{R_1 + R_2}{R_1} \quad \text{or} \quad 1 + \frac{R_2}{R_1} \qquad [1]$$

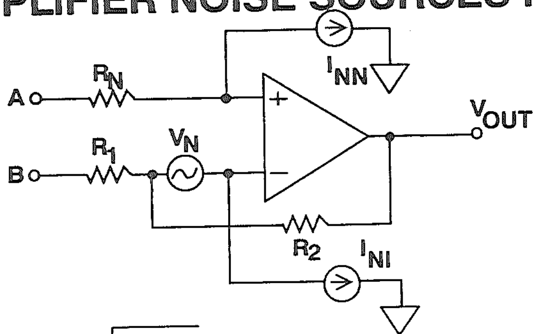
Whereas a signal applied to B (A being grounded) sees a gain of

$$\frac{-R_2}{R_1} \qquad [2]$$

These are the non-inverting and inverting gains, respectively, of the amplifier.

11

AMPLIFIER NOISE SOURCES REFERRED TO OUTPUT



GAIN TO OUTPUT

A: $\frac{R_1+R_2}{R_1} = 1 + \frac{R_2}{R_1}$ [1]

B: $\frac{-R_2}{R_1}$ [2]

$\frac{(R_1 + R_2)\sqrt{4kTR_N}}{R_1}$	[A] JOHNSON NOISE IN R_N (AMPLIFIED BY [1])
$\frac{I_{NN}R_N(R_1 + R_2)}{R_1}$	[B] CURRENT NOISE (NON-INVERTING INPUT, FLOWING IN R_N, AMPLIFIED BY [1])
$\frac{V_N(R_1 + R_2)}{R_1}$	[C] AMPLIFIER VOLTAGE NOISE (AMPLIFIED BY [1])
$\frac{R_2\sqrt{4kTR_1}}{R_1}$	[D] JOHNSON NOISE R_1 (AMPLIFIED BY [2])
$\sqrt{4kTR_2}$	[E] JOHNSON NOISE R2
$I_{NI}R_2$	[F] CURRENT NOISE (INVERTING INPUT, FLOWING IN R_2)

Figure 11. 34

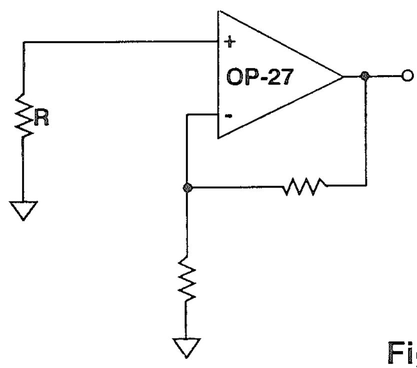
The current noise of the non-inverting input, I_{NN}, flows in R_N and gives rise to a noise voltage of $I_{NN}R_N$, which is amplified by [1], as are the op-amp noise voltage, V_N, and the Johnson noise of R_N, which is $\sqrt{4kTR_N}$. The Johnson noise of R_1 is amplified by [2], and the Johnson noise of R_2 is not amplified at all but is buffered directly to the output. The current noise of the non-inverting input, I_{NI}, does not flow in R_1, as might be expected - negative feedback around the amplifier works to keep the potential at the inverting input unchanged, so that a current flowing from that pin is forced, by negative feedback, to flow in R_2 only, resulting in a voltage at the amplifier output of $I_{NI}R_2$ (we could equally well consider the voltage caused by I_{NI} flowing in the parallel combination of R_1 & R_2 and then amplified by the noise gain of the amplifier (see below) but the results are identical - only the calculations are more involved).

If we consider these six noise contributions we see that if R_N and R_2 are low then the effect of current noise and Johnson noise will be minimized and the dominant noise will be the op-amp's voltage noise. As we increase resistance both Johnson noise and the voltage noise produced by noise currents will rise. If noise currents are low then Johnson noise will take over from voltage noise as the dominant contributor. Johnson noise, however, rises with the square root of the resistance, while the current noise voltage rises linearly with resistance, so ultimately, as the resistance continues to rise, the voltage due to noise currents will become dominant. An example is shown in Figure 11.35.

These noise contributions we have analyzed are not affected by whether the input is connected to node A or node B (the other being grounded or connected to some other low-impedance voltage source), which is why the non-inverting gain $(1 + R_2/R_1)$, which is seen by the voltage noise of the op-amp, V_N, is some-

EFFECT OF SOURCE RESISTANCE ON NOISE

Example: OP-27
Voltage Noise = 3nV/$\sqrt{\text{Hz}}$
Current Noise = 1pA/$\sqrt{\text{Hz}}$
(T = 25°C)



CONTRIBUTION FROM	VALUES OF R		
	0	3kΩ	300kΩ
AMPLIFIER VOLTAGE NOISE	$\boxed{3}$	3	3
AMPLIFIER CURRENT NOISE FLOWING IN R	0	3	$\boxed{300}$
JOHNSON NOISE OF R	0	$\boxed{7}$	70

RTI NOISE (nV/$\sqrt{\text{Hz}}$)
Dominant Noise is Highlighted

Figure 11.35

times known as the "noise gain" of the amplifier. The difference between the noise gain and the inverting gain to signals applied to node B can be important at low gains - if we program a gain of -1 by making $R_1 = R_2$ then the amplifier noise will see a gain of 2 and the signal/noise ratio will be worsened by a factor of 2, if the gain is smaller than this the degradation will be even worse. This can be important in the design of low-pass filters, and also in determining the stability of op-amps without full frequency compensation (see below).

We thus see that choice of an op-amp for low noise depends greatly on the impedances involved in the circuit. For low impedance circuitry amplifiers with low voltage noise, such as the OP-27, will be the obvious choice, since they are inexpensive and their comparatively large current noise will not affect the application. At medium resistances, unless we use an absurdly noisy amplifier, we will find that the Johnson noise of our resistors is dominant, while at very high resistances we must choose an op-amp with the smallest possible current noise, such as the AD549 or AD645.

Until recently BIFET amplifiers tended to have comparatively high voltage noise, though very low current noise, and were thus more suitable for low noise applications in high impedance circuitry than at low impedance. Recently the AD645 and AD743/AD745 have become available, which have very low values of both voltage and current noise. The AD645 specifications at 10kHz are 10nV/$\sqrt{\text{Hz}}$ and 0.6fA/$\sqrt{\text{Hz}}$, and the AD743/AD745 specifications at 10 kHz are 2.9nV/$\sqrt{\text{Hz}}$ and 6.9fA/$\sqrt{\text{Hz}}$. These make possible the design of low-noise amplifiers which have low noise over a wide range of source impedances.

DIFFERENT AMPLIFIERS ARE BEST AT DIFFERENT IMPEDANCE LEVELS

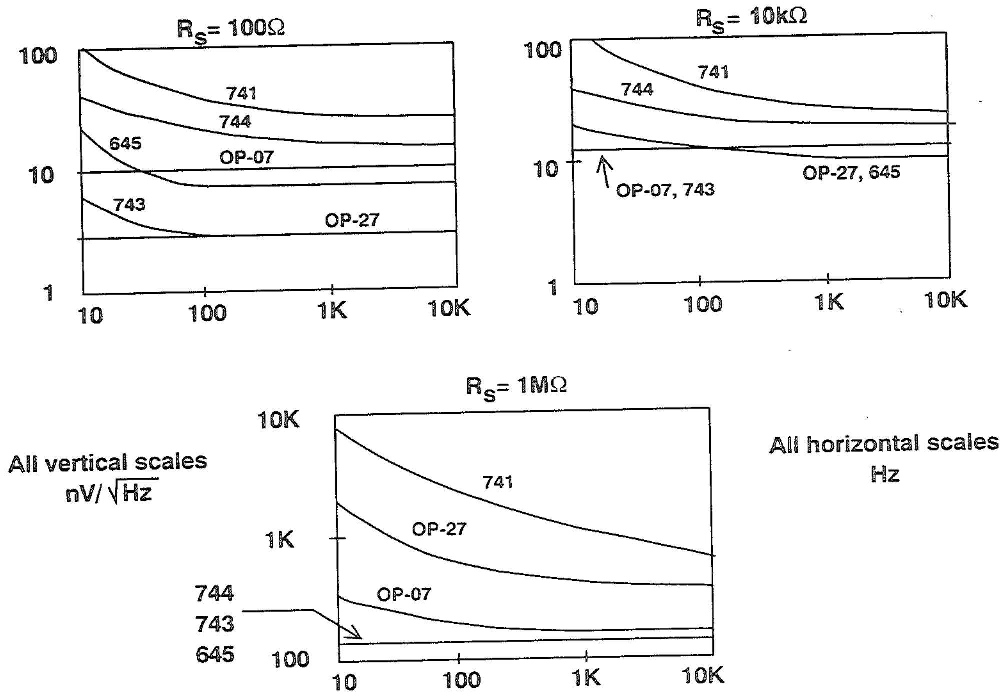


Figure 11.36

NOISE FIGURE

The "noise figure" of an amplifier is defined at a particular source resistance. It is the ratio, expressed in dB, of the noise of the amplifier when driven by a perfect resistance of the particular value to the noise of a perfect (noise-free) amplifier operated under the same conditions. Noise figure is a useful parameter for RF amplifiers, which are generally operated with resistive termination of 50 Ω or 75 Ω, but since operational amplifiers are operated over a very wide range of conditions and termination impedances noise figure is not a useful op-amp parameter and it is never specified. (Whatever resistance is chosen would be wrong, and the lack of correlation of voltage and current noise in an op-amp make it impossible to calculate the noise figure at one resistance from its value at another - if it is needed it may be calculated from the voltage and current noise figures.)

INSTABILITY & ITS CURES

A common cause of unexpectedly large noise in op-amp circuitry is high frequency oscillation. Integrated circuit chips contain transistors with f_t of hundreds or thousands of MHz and if they are inadequately decoupled they may oscillate at VHF or UHF. Such oscillation may not prevent low frequency operation of the circuit, but it will certainly prevent the achievement of low noise and high DC accuracy. It can also be hard to detect, since few oscilloscopes have sufficient bandwidth to detect a signal of a few hundred MHz - and, indeed, the capacity of the oscilloscope probe may be sufficient to stop the oscillation, so the circuit works properly when an oscilloscope is attached to it!

The best way to detect such VHF oscillation is to use a broadband spectrum analyzer (to at least 1 GHz) since the probe need not actually contact the circuit to detect an oscillation. It is best to search

NOISE FIGURE

- IS *ALWAYS* SPECIFIED AT A PARTICULAR SOURCE RESISTANCE
- IT IS THE RATIO OF THE NOISE OF AN AMPLIFIER OPERATED AT THAT SOURCE RESISTANCE TOTHE NOISE OF A PERFECT AMPLIFIER OPERATED AT THE SAME SOURCE RESISTANCE
- IT IS GENERALLY GIVEN IN dB
- IT IS NOT A PRACTICAL SPECIFICATION FOR OP-AMPS AND IS RARELY SPECIFIED FOR THEM

Figure 11.37

PARASITIC OSCILLATION

- I.C. OP-AMPS CONTAIN TRANSISTORS WITH f_t OF HUNDREDS OF MHz
- THEY CAN OSCILLATE AT HUNDREDS OF MHz

Figure 11.38

DETECTING PARASITIC OSCILLATION

- THE CAPACITANCE OF AN OSCILLOSCOPE PROBE MAY BE SUFFICIENT TO STOP PARASITIC OSCILLATION
- DETECT IT WITH A BROADBAND SPECTRUM ANALYZER WITH ITS PROBE CLOSE TO, BUT NOT TOUCHING,THE NODE BEING ANALYZED
- DRIVE THE INPUT WITH A FULL-SCALE L.F SIGNAL DURING THIS TEST BECAUSE SOME PARASITICS OCCUR ONLY AT CERTAIN INPUT LEVELS

Figure 11.39

for parasitic oscillations with a full-scale input signal to the circuit being tested since sometimes parasitic oscillation only occurs at a particular input level and not over the whole range.

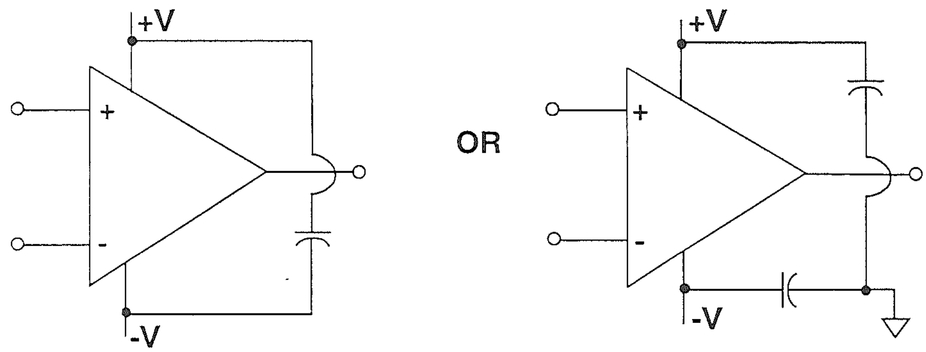
The commonest cause of HF instability is inadequate HF supply decoupling. Internal HF currents within the supplies of an op-amp should be short-circuited by decoupling the supplies with a capacitor having very low HF impedance - this means that the capacitor itself must have low inductance, and that its leads and the PC tracks to them must also have low inductance. (A common layout mistake is to decouple the supply of an IC in one row to the ground bus of the ICs in an adjacent row - the inductance from the supply to the grounding point on the bus may well be low - but the inductance of the length of ground bus from the capacitor grounding point to the ground pin of the IC being grounded may be intolerable.)

Paper, foil, electrolytic and plastic film capacitors all tend to be inductive. The best HF decoupling capacitors are low ESR and ESL monolithic ceramic types, preferably surface mounted, but otherwise with leads of less than 1.5 mm.

Not only is it important to ensure that the internal device currents are properly decoupled at HF, it is also important to ensure that HF load currents have the lowest possible impedance of return path.

Many IC processes make NPN transistors which are much faster than their PNP transistors (or vice versa), although this is not the case with Analog Devices' "CB" process, which makes complimentary bipolar transistors of roughly equal speed.

AMPLIFIER DECOUPLING



- TO PREVENT PARASITIC OSCILLATION DECOUPLING CAPACITORS MUST HAVE LOW INDUCTANCE TO PROVIDE A MINIMUM IMPEDANCE PATH TO H.F. CURRENTS IN THE I.C.
- This means low inductance capacitors (monolithic ceramic and NOT foil or electrolytic types).
- It also means low inductance leads and PC tracks (surface mount or lead lengths <1.5 mm, PC tracks as wide as possible and no longer than a few mm)

Figure 11.40

SOME OP AMPS HAVE ONE SUPPLY WHOSE DECOUPLING IS MORE CRITICAL THAN THE OTHER. EXTRA CARE SHOULD BE TAKEN TO DECOUPLE THE "ACTIVE SUPPLY"

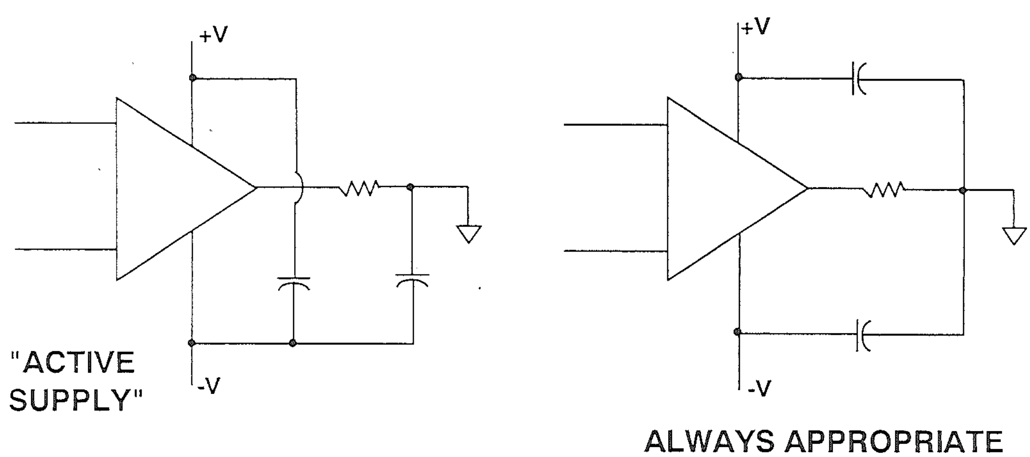


Figure 11.41

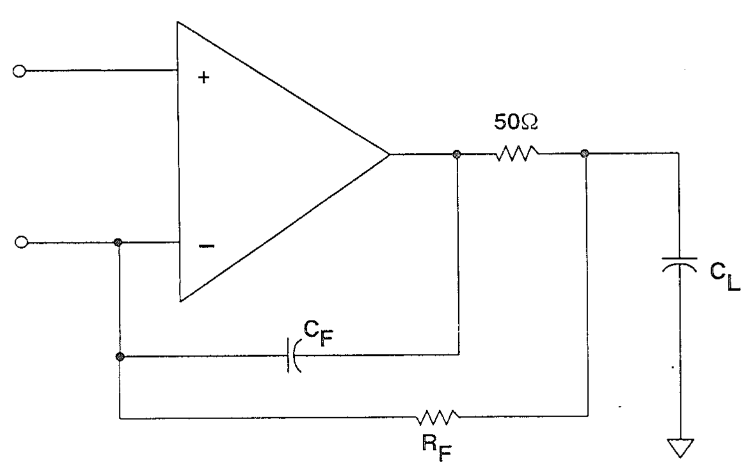
Where transistor speeds are unequal it is quite common for the majority of the HF output current to flow in only one of the supplies, the other supply current varying only slowly. In such cases the supply which is active at HF should be decoupled to the ground of the load by a capacitor with minimum possible inductance.

It is also important that the supply to which any integrator or feed-forward capacitor in the op amp is referred should be decoupled to the load ground in the same way. The critical supply of many op amps is listed in the application note at the end of this section.[2]

Two other sources of instability are reactive loads and insufficient frequency compensation. Many operational amplifiers oscillate if required to drive capacitances of more than a few picofarads, although some (such as the AD847) are specially designed to drive any value of capacity without instability (the bandwidth is, of course, reduced as the capacity increases). In addition to the obvious measure of reducing stray capacity the problem can often be eased by placing a small resistance in series with the output of the op-amp.

Most operational amplifiers have a single dominant pole in their frequency response. It is set by a capacitor which may be integrated in the amplifier or connected externally. If the 6 dB/octave low-pass frequency response caused by this pole continues down to gains of 1 or less the amplifier is said to be fully compensated and may be used in closed loop applications at any gain. If the frequency of the pole is set higher one of the other poles in the amplifier response will take effect at a gain of G, before the gain has dropped to unity, and the gain then rolls

CAPACITIVE LOADS CAUSE INSTABILITY: ISOLATE THEM WITH A SMALL SERIES RESISTOR



Some amplifiers, such as the AD847, are designed to drive capacitive loads without this problem

Figure 11.42

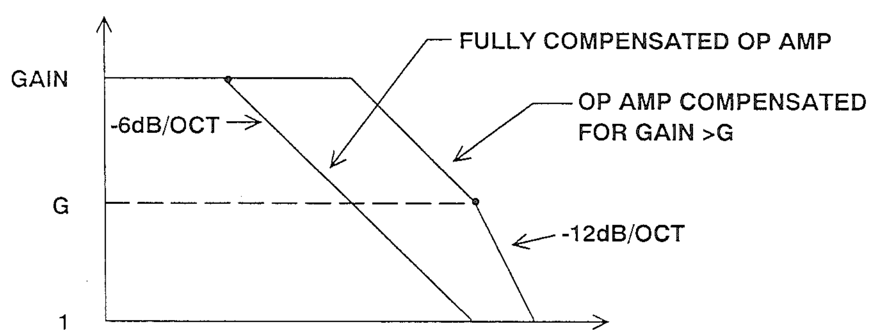
off at 12 dB/octave. In this case the amplifier has a higher gain/bandwidth product but is not stable at closed-loop gains of much less than G because the additional phase shift in the negative feedback makes it positive feedback.

If this incomplete compensation is internal the op-amp data sheet will specify "Stable at gains ≥ X" with some value of X (usually between 2 and 25). It is quite common to find families of amplifiers which are identical except for the value of the internal compensation capacitor (an example is the family of the AD847, AD848 and AD849). If an amplifier which has two pole roll-off is used at too low a closed-loop gain it will oscillate. This effect is easily overlooked during design and the resulting problems are quite commonly referred to Analog Devices' Applications Department for solution.

The simplest solution is to use an op-amp which is stable at the gain required and has sufficient bandwidth. However, if, for whatever reason, a particular op-amp must be used at a lower gain than is stable, there is a solution which is practical (but has the disadvantage of worsened noise performance).

Consider Figure 11.44. If R_1 is omitted the amplifier has a gain of 1 to a signal and a noise gain of 2 (see the formulas [1] and [2] earlier in this section) and if it uses an op-amp which is stable only at gains ≥5 it will probably oscillate. If R_1 is present, however, the signal gain is unaffected but the noise gain is increased to 12 and the circuit becomes stable. *But* the signal to noise ratio becomes six times worse (if the dominant noise is the op-amp voltage noise) since the noise gain has increased by a factor of 6. In many applications this is unimportant, so the technique is a useful one.

FULLY AND PARTIALLY COMPENSATED OP AMPS



- A fully compensated op amp has a roll-off of 6 dB/octave to unity gain
- A partially compensated amplifier has a larger gain/bandwidth product (for the same basic amplifier) but the roll-off increases to 12 dB/octave above a certain frequency (when the gain is less than G). As a consequence, such amplifiers are not stable in closed-loop applications with gains < G.

Figure 11.43

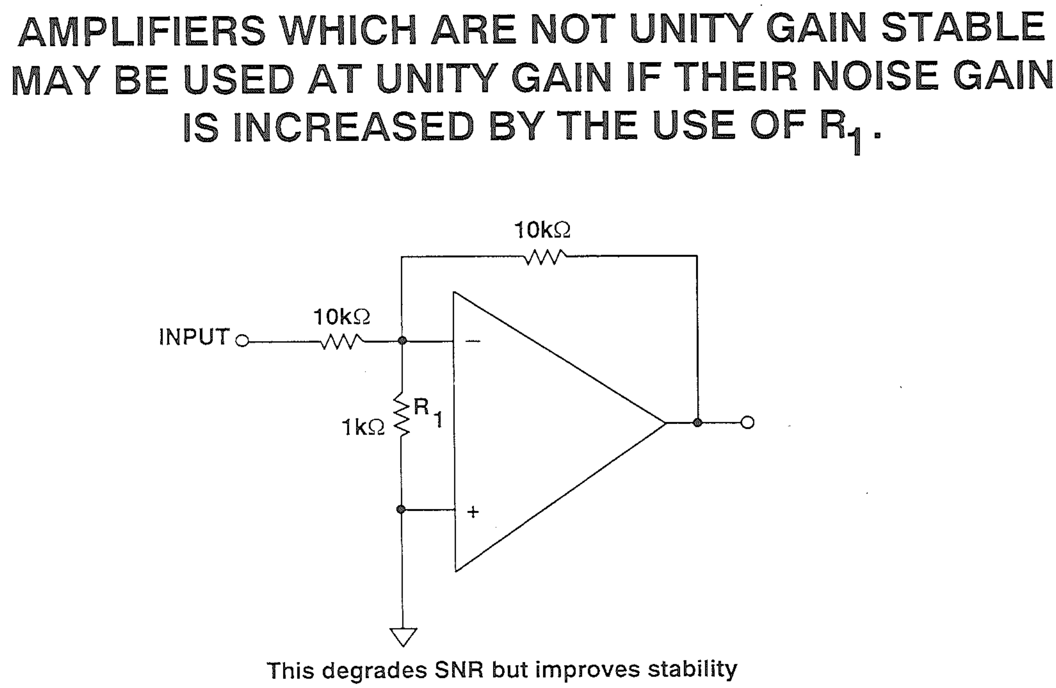


Figure 11.44

CIRCUIT NOISE FROM COMPONENTS - CAPACITORS

Noise and error can be introduced into op-amp circuits by imperfections in the components used. We have already pointed out the need for non-inductive decoupling capacitors. The other capacitor problems in analog circuits are leakage, thermal stability, dielectric absorption, and microphonics.

None of them is very serious, since capacitors are available which are substantially free from these problems - the important consideration for the circuit designer is to ensure that he considers the effects of the errors on his circuit and chooses a component whose performance is adequate to its task.

The only modern capacitors with appreciable leakage are electrolytic ones (aluminium and tantalum), and the only common application where leakage presents a severe performance problem is long delay timers. Since CMOS dividers are now so cheap it is probably better to use a higher frequency oscillator and a divider than a long time constant RC circuit, but where long time constants are produced with RC circuitry using electrolytic capacitors it is important to calculate the effect of the specified leakage (at the extremes of operating temperature) on the performance of the circuit.

Dielectric absorption is the phenomenon which causes a discharged capacitor to recover a small amount of charge if it open-circuited - the amount depends on the previous state of charge. The normal application where dielectric absorption must be avoided is in the design of sample and hold circuitry, since dielectric absorption causes the hold state of a SHA to depend not only on the voltage present at the time of the sample/hold transition but also on the previous charge history of the

THE PRINCIPLE PROBLEMS WITH CAPACITORS

- INDUCTANCE
- LEAKAGE
- TEMPERATURE COEFFICIENT
- DIELECTRIC ABSORPTION
- With modern capacitors none of them is very serious but they must be considered when designing circuits.

Figure 11.45

capacitor. Today most SHAs contain integral capacitors and the designer need not specify a hold capacitor. However, on the occasion when a SHA is built with a discrete hold capacitor it is critically important that a capacitor is chosen with *guaranteed* dielectric absorption. This ensures that the manufacturer has tested this particular parameter. The problem is that dielectric absorption can vary quite widely from batch to batch, even with types of capacitor which are normally good. If the dielectric absorption is not tested parts from a batch with bad dielectric absorption will be found from time to time among parts of a type which is normally quite good. A guaranteed and tested part eliminates this risk.

Certain analog designs are affected by small variations of capacitance with temperature. In particular, high-K ceramic and electrolytic capacitors should not be used in accurate filter circuits since their TCs can be as high as 1000 ppm/°C. For the timing capacitors in VFC and other precision oscillators mica, ceramic and polyester parts with guaranteed TCs of a few ppm/°C are available - but use *guaranteed* parts, not generic ones or, still worse, anonymous objects from that old parts box on the workbench.

FEATURES OF COMMON CAPACITORS

TYPE	TYPICAL DIELECTRIC ABSORPTION	ADVANTAGES	DISADVANTAGES
NPO Ceramic	0.1%	Small Case Size Inexpensive Good Stability Wide Range of Values Many Vendors Low Inductance	DA too High for More than 8-Bit Applications
Polystyrene	0.001% to 0.02%	Inexpensive Low DA Available Wide Range of Values Good Stability	Destroyed by Temperature > +85°C Large Case Size High Inductance
Polypropylene	0.001% to 0.02%	Inexpensive Low DA Available Wide Range of Values	Destroyed by Temperature > +105°C Large Case Size High Inductance
Teflon	0.003% to 0.02%	Low DA Available Good Stability Operational Above +125°C Wide Range of Values	Relatively Expensive Large High Inductance
MOS	0.01%	Good DA Small Operational Above +125°C Low Inductance	Limited Availability Available only in Small Capacitance Values
Polycarbonate	0.1%	Good Stability Low Cost Wide Temperature Range	Large DA Limits to 8-Bit Applications High Inductance
Polysulfone	0.1%	Good Stability Low Cost Wide Temperature Range	Large DA Limits to 8-Bit Applications High Inductance
Monolithic Ceramic	>0.2%	Low Inductance Wide Range of Values	Poor Stability Poor DA
Mica	>0.003%	Low Loss at HF Low Inductance Very Stable Available in 1% Values or Better	Quite Large Low Values (<10nF) Expensive
Aluminium Electrolytic	High	Large Values High Currents High Voltages Small Size	High Leakage Usually Polarized Poor Stability Poor Accuracy Inductive
Tantalum Electrolytic	High	Small Size Large Values Medium Inductance Reliable	Quite High Leakage Usually Polarized Expensive Poor Stability Poor Accuracy

Figure 11.46

Circuit Noise from Components - Resistors

In the past many resistors were noisy. Today better resistor materials and structures are available and most resistors have noise which is quite close to the Johnson noise of an ideal resistor of the same value. (This excess noise is related to the current flowing in the resistor.) Solid carbon composition resistors and ultra-high value resistors are exceptions. In general it is best to avoid cheap carbon composition resistors (but *not* carbon film resistors, which use a completely different technology).

Unfortunately it is sometimes impracticable to avoid the use of very high value resistors, although it is good practice to avoid resistors above 10 MΩ where possible. Where their use is inevitable study the data on available types and ensure that the performance of the type you choose does not compromise the performance of your system. You should also note that many types of very high value resistor have rather poor linearity (i.e. they do not (quite) obey Ohm's Law). This, again, should be checked before choosing a component.

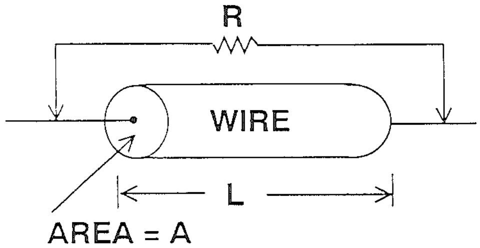
In addition to noise there are other resistor characteristics which affect the accuracy which may be achieved by amplifier circuits. We shall consider them in turn, with the lowest level effects first.

RESISTOR NOISE

- IN GENERAL MODERN RESISTORS HAVE NOISE WHICH IS NOT MUCH GREATER THAN THEIR JOHNSON NOISE
- Carbon composition resistors are an exception and should be avoided where noise matters.
- Ultra-high resistors (>10 MΩ) can be noisy, too. They may also have other nasty habits, like non-linearity. Avoid them if you can and if you can't, read the data sheet VERY CAREFULLY.

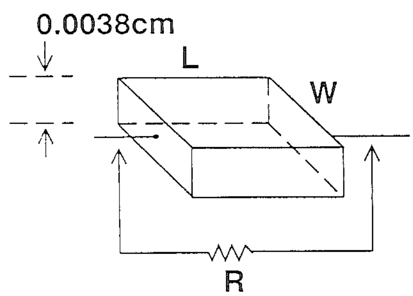
Figure 11.47

RESISTANCE OF CONDUCTORS



$R = \frac{\rho L}{A}$, WHERE

ρ = BULK RESISTIVITY
= $1.724 \times 10^{-6}\ \Omega$cm FOR COPPER
A = AREA (cm2)
L = LENGTH



THICKNESS OF STANDARD PC FOIL - 0.0038cm
FOR A SQUARE OF FOIL, L = W

$\therefore R = \frac{\rho}{0.0038} \approx 0.45\text{m}\Omega$

$\therefore$ SHEET RESISTIVITY OF PC FOIL $\approx \frac{0.45\text{m}\Omega}{\text{SQUARE}}$

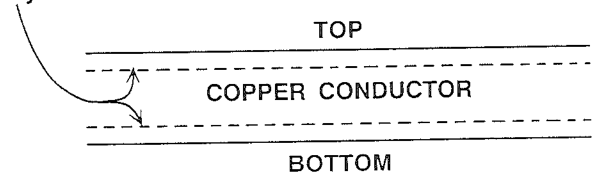
Figure 11.48

Every conductor in a circuit, including wires and PC tracks, has some resistance. Voltage drops in this resistance, though generally small, can be large enough to affect system accuracy and if there is the slightest reason to suspect that this effect will be significant it should certainly be calculated. At high frequencies the resistance of conductors rises due to "skin effect" - and this, too, should be considered.

Wherever there is a junction between two different conductors we have a voltage. If two junctions are present in a circuit we have a thermocouple, and if those two junctions are at different temperatures there will be a net voltage in the circuit. This effect is used to measure temperature, but is a potential source of inaccuracy in low level circuits, since wherever two different conductors meet we have a thermocouple (whether we think "This is a thermocouple." or not). This will cause errors if the various junctions are at different temperatures. The effect is hard to avoid since, even if we use copper conductors everywhere, every resistor contains two thermocouples. Therefore, if there is a temperature difference between the ends of a resistor (or between any other two junctions in a circuit) there will be a potential difference. Values for typical wirewound resistors are of the order of 40 μV/°C.

SKIN EFFECT

- HF Current flows only in thin surface layers



- Skin Depth: $6.61/\sqrt{f}$ cm, f in Hz
- Skin Resistance: $2.6 \times 10^{-7}\sqrt{f}$ ohms per square, f in Hz
- Since skin currents flow in both sides of a PC track, the value of skin resistance in PCBs must take account of this

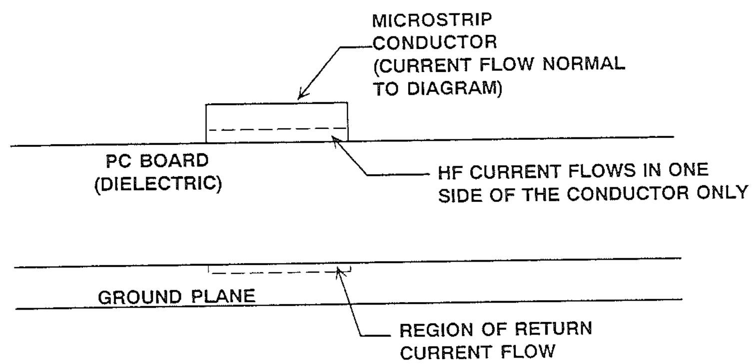
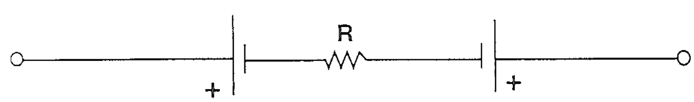


Figure 11.49

RESISTORS CONTAIN THERMOCOUPLES



- If their ends are at different temperatures they may introduce significant errors into low-level DC circuits. (A typical value for wirewound resistors is 40μV/°C temperature difference.)
- The same effect occurs wherever dissimilar conductors meet in a circuit.
- Minimize the problem by planning to minimize temperature differences at critical junctions by careful layout and avoiding heat sources.

Figure 11.50

The effect is minimized by careful design. The problem will generally be kept under control if all critical resistors have leads of equal length, and they, and any other potential thermocouples in the circuit, are positioned and orientated with respect to any heat sources so as to have minimum temperature differential from end to end. It is possible to buy resistors which are made of materials with very low thermoelectric e.m.f. These ease the problem still further but are somewhat expensive.

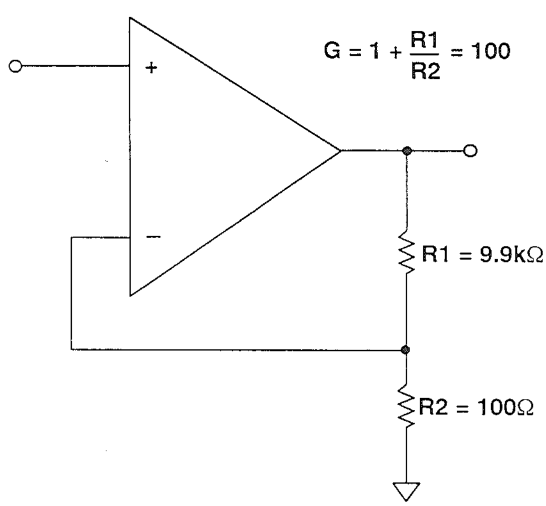
The temperature coefficient of resistors can also be a source of errors. Consider a non-inverting amplifier with a gain of exactly 100. It uses an op-amp and two resistors, R_1 is 9.9 kΩ and R_2 is 100Ω, and they are perfectly matched at 25°C.

If the temperature coefficients of R_1 and R_2 differ by only 25 ppm, the gain of the amplifier will change by 250 ppm for a 10°C temperature change. This is about 1 lsb in a 12-bit system, and a major disaster in a 16-bit system.

Even if the temperature coefficients are identical there are problems. Suppose R_1 and R_2 have identical temperature coefficients of 25 ppm and are both ¼ W resistors. If the signal input in Figure 11.52 is zero the resistors will dissipate no heat, but if it is 100 mV there will be 9.9 V across R1, it will dissipate 9.9 mW and experience a temperature rise of 1.24°C, which will change its resistance and the gain by 31 ppm. R2, with only 100 mV across it, is only heated 0.0125°C, which is negligible. This change in resistor matching represents a full-scale error of ¾ lsb at 16-bits.[3]

These, and similar, problems are avoided by having critical resistors with accurate matching, accurately matched temperature coefficients, and tight thermal coupling between resistors whose matching is important. This is best achieved by using a resistor network on a single substrate - such a network may be within an IC or may be a separately packaged thin-film resistor network.

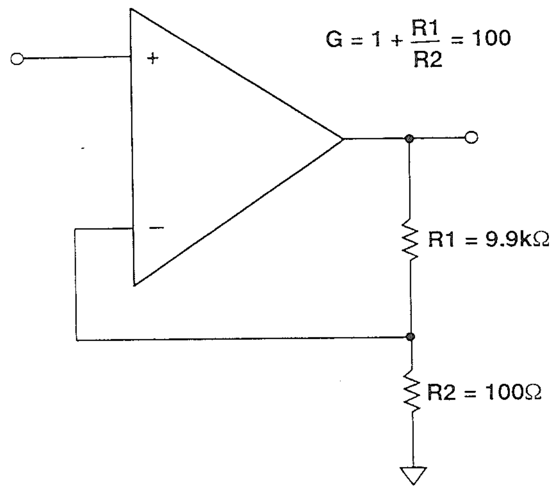
RESISTOR TEMPERATURE COEFFICIENT MISMATCHES CAUSE GAIN VARIATION WITH TEMPERATURE



If the TCs of R_1 & R_2 differ by only 25 ppm/°C a temperature change of 10°C will cause of gain change of 250 ppm - this is 1 lsb in a 12-bit system and a major disaster in a 16-bit system.

Figure 11.51

RESISTOR SELF-HEATING, EVEN IN RESISTORS WHOSE TCs ARE PERFECTLY MATCHED, CAN CAUSE GAIN VARIATION WITH INPUT LEVEL.



If R1 & R2 are ¼ W resistors with a resistance TC of 25 ppm/°C, an input of 100 mV will heat R1 by 1.24°C and R2 by a negligible amount. This will alter the gain by 31 ppm, or ¾ lsb at 16-bits.

Figure 11.52

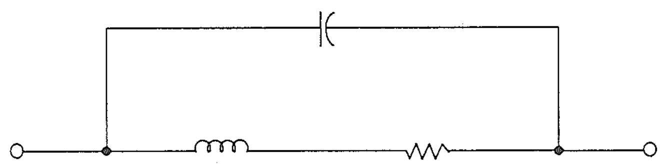
RESISTOR MATCHING

- MOST RESISTOR MATCHING PROBLEMS CAN BE ELIMINATED BY MAKING ALL RESISTORS WHOSE MATCHING IS CRITICAL ON A SINGLE THIN-FILM SUBSTRATE.
- THE NETWORK HAS TIGHT THERMAL COUPLING AND EXCELLENT MATCHING OF CHARACTERISTICS.
- This substrate may be an IC chip or a glass substrate depending on circumstances.
- This does not remove the necessity of care with layout

Figure 11.53

11

RESISTOR MODEL



ALL RESISTORS ARE INDUCTIVE
AND HAVE STRAY CAPACITANCE
Wirewound ones (even "non-inductive" wirewound ones) are bad, but even cermet types have L of a few nH and C of a few tenths of a pF. Verify that these parasitic components will not affect your design.
They probably won't - *but make sure!*

Figure 11.54

A final consideration with resistors is their stray reactances. Wirewound resistors, even the so-called "non-inductive" types which have half their turns wound clockwise and half anticlockwise to minimize their inductance, have quite considerable inductance and capacitance (several μH and pF or even more) and must be treated with respect even at medium frequencies.

But even cermet resistors have inductance of a few nanohenries and stray capacitance of a few tenths of a pF. These stray reactances may need to be considered in higher frequency designs, as should the stray reactances of the tracks of the PCB.[4] While such issues are rarely important in high accuracy LF designs it is important that engineers are aware that resistors have such characteristics and that they may affect the performance of their circuits, since, as we have mentioned before, even low frequency ICs may contain transistors with very high frequency performance.

RESISTOR COMPARISON CHART

	TYPE	ADVANTAGES	DISADVANTAGES
DISCRETE	Carbon Composition	Lowest Cost High Power/Small Case Size	Poor Tolerance (5%) Poor Temperature Coefficient (1500ppm/°C)
	Wire-Wound	Excellent Tolerance (0.01%) Excellent TC (1ppm/°C) High Power	Reactance May be a Problem Large Case Size Most Expensive
	Metal Film	Good Tolerance (0.1%) Good TC (<1 to 100ppm/°C) Moderate Cost	Must be Stabilized with Burn-In Low Power
	Bulk Metal or Metal Foil	Excellent Tolerance (to 0.005%) Excellent TC (to <1ppm/°C) Low Reactance	Low Power Very Expensive
	High Megohm	Very High Values (10^8 - $10^{14}\,\Omega$) Only Choice for Some Circuits	High Voltage Coefficient (200ppm/V) Fragile Glass Case Expensive
NETWORKS	Thick Film	Low Cost High Power Laser-Trimmable Readily Available	Fair Matching (0.1%) Poor TC (>100ppm/°C) Poor Tracking TC (10ppm/°C)
	Thin Film on Glass	Good Matching (<0.01%) Good TC (<100ppm/°C) Good Tracking TC (2ppm/°C) Moderate Cost Laser-Trimmable Low Capacitance	Delicate Often Large Geometry Low Power
	Thin Film on Ceramic	Good Matching (<0.01%) Good TC (<100ppm/°C) Good Tracking TC (2ppm/°C) Moderate Cost Laser-Trimmable Low Capacitance Suitable for Hybrid IC Substrate	Often Large Geometry
NETWORKS	Thin Film on Silicon	Good Matching (<0.01%) Good TC (<100ppm/°C) Good Tracking TC (2ppm/°C) Moderate Cost Laser-Trimmable Suitable for Monolithic IC Construction	Some Capacitance to Substrate Low Power
	Thin Film on Sapphire	Good Matching (<0.01%) Good TC (<100ppm/°C) Good Tracking TC (2ppm/°C) Laser-Trimmable Low Capacitance	Higher Cost Low Power

Figure 11.55

11

EXTERNAL NOISE

A good definition of this type of noise is a wanted signal in a place where it is not wanted. This section considers ways in which unwanted signals can invade small-signal circuitry and damage its performance. There are four basic mechanisms: conduction, electrostatic coupling, magnetic coupling and electromagnetic coupling. If we understand the coupling mechanism responsible for a particular problem we are well on our way to solving it.

EXTERNAL NOISE

IS A WANTED SIGNAL IN A PLACE WHERE IT'S NOT WANTED

COUPLING METHODS ARE

CONDUCTION	**INDUCTION**
ELECTROSTATIC	**ELECTROMAGNETIC**

Figure 11.56

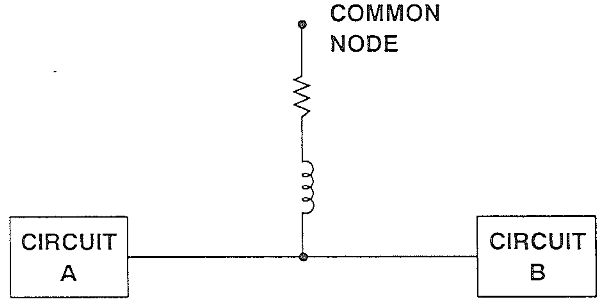
NOISE COUPLING - CONDUCTION

If two circuits share a common impedance, voltages caused by one will affect the other. This is a common mechanism for noise coupling.

If the common impedance is a power or ground conductor the problem may be cured by using separate conductors for the two circuits, or by using a shorter, fatter conductor with lower resistance and inductance. Properly decoupling of power supplies at both HF and LF is, of course, essential.

On a PCB power and ground tracks should never be minimized, but should be as wide as practicable. The ideal is to use ground planes and power planes wherever possible. These are whole layers of a PCB which are devoted to virtually continuous conductor, which is used for ground or power respectively. Ground planes offer the lowest possible ground impedance on a PCB.[5]

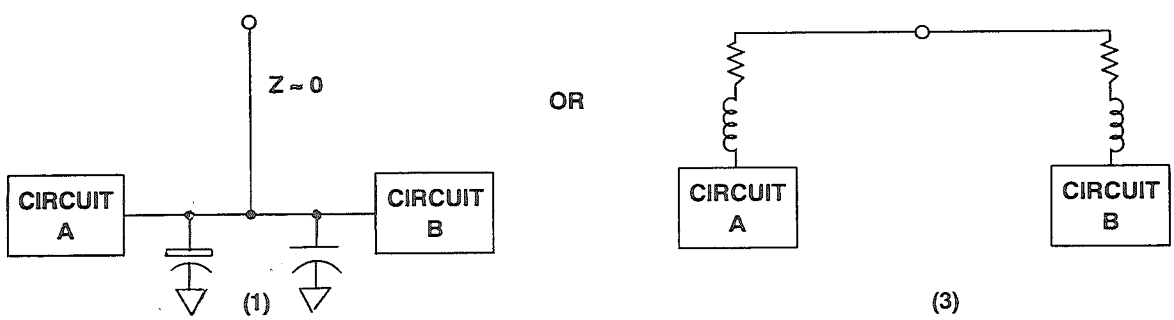
IF TWO CIRCUITS SHARE A COMMON IMPEDANCE, CURRENT FROM ONE WILL CAUSE A VOLTAGE DROP WHICH MAY AFFECT THE OTHER



Common impedances are most often ground or power lines.
Cures include removing common impedances or reducing them.

Figure 11.57

SOLUTIONS INCLUDE:



DECOUPLING POWER SUPPLIES AT HF & LF

(2)

REDUCING THE COMMON IMPEDANCE

ELIMINATING THE SHARED PATH

- Sometimes these are not possible or fully effective and more complex solutions are required.

Figure 11.58

GROUND PLANES AND POWER PLANES

- ARE CONTINUOUS LAYERS OF CONDUCTOR IN A PCB
- They offer the lowest practicable resistance and inductance
- THEY ARE NOT SUPERCONDUCTORS!
- Room temperature superconductors have not yet been invented

Figure 11.59

Room temperature superconductors have not yet (1-1-92) been invented, so even unbroken planes cannot offer zero resistance, let alone zero inductance. There are many applications where common impedances cannot be avoided - but it is often possible to design around them. Where a noisy common impedance causes signal transmission problems these may often be solved by sending the low-level signal differentially (so that common-mode noise is ignored) or reconfiguring the system so that the signal is sent as a current, at a high level, or in digital form - all these modes are, relatively, noise immune.

WHERE COMMON GROUND IMPEDANCES OR GROUND NOISE CANNOT BE REDUCED SEND SIGNALS AS CURRENTS, WORK AT HIGHER VOLTAGE LEVELS, CONVERT SIGNALS TO DIGITAL FORM BEFORE TRANSMISSION, OR SEND LOW-LEVEL SIGNALS DIFFERENTIALLY

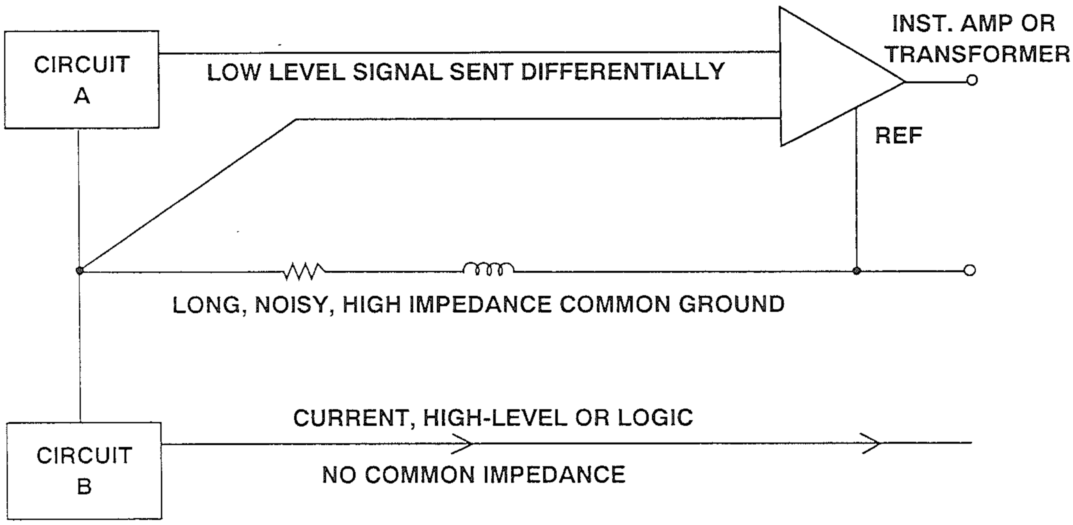


Figure 11.60

NOISE COUPLING - CAPACITIVE

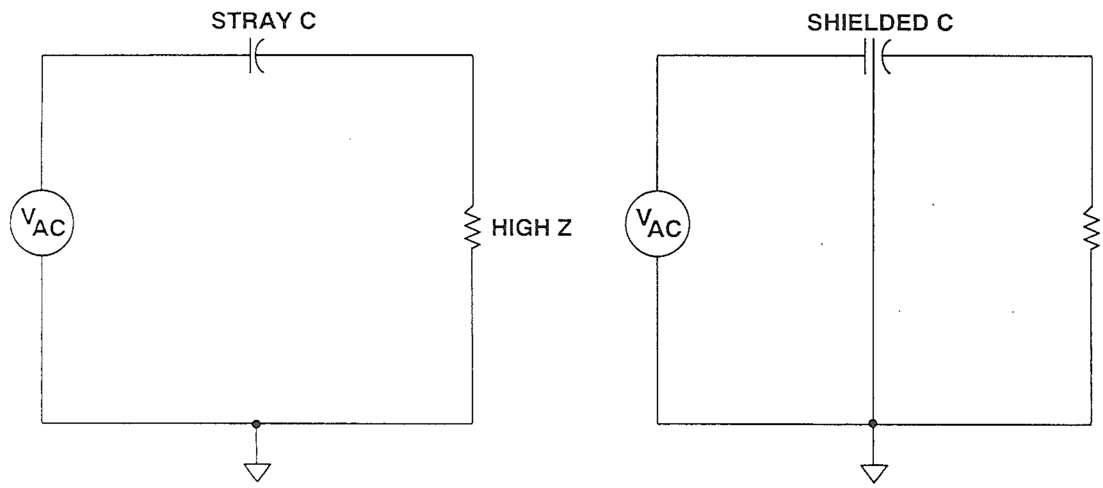
In addition to the capacitors which are designed into the system, practical circuits are full of small stray capacitances. If a high impedance low-level point in a circuit is separated from a high voltage AC signal by a small distance it is probably that the stray capacitance will couple a noise signal into the low-level circuit.

Fortunately this particular source of noise is very easily cured. If a conductive shield is inserted between the plates of the stray capacitor, *and grounded*, the displacement current will flow to ground through the shield and not through the signal circuits. This conductive shield is known as a "Faraday Shield" and is easily constructed and used. It is generally very efficient, but a common fault is to fail to ground it - in which case the effective stray capacity is generally increased and the original problem is exacerbated.

Some integrated circuit packages have unconnected metal lids, which can act as ungrounded Faraday shields and thus pick up noise signals. These lids are generally made of gold-plated kovar and can be soldered carefully without risk to the chip in the package - they should generally be grounded, but it is sometimes better to connect them to other points in the circuit. They should never be left unconnected! [6]

11

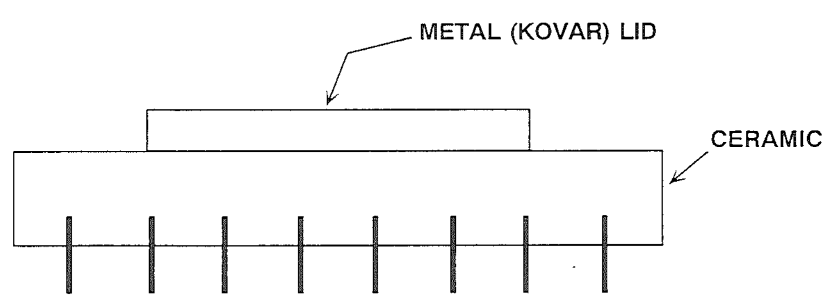
CAPACITIVE NOISE COUPLING IS EASILY MODELLED AND EASILY CURED



The commonest mistake is to omit to ground the Faraday Shield

Figure 11.61

SOME INTEGRATED CIRCUITS HAVE UNCONNECTED METAL LIDS WHICH BEHAVE AS UNGROUNDED FARADAY SHIELDS AND PICK UP NOISE



They should be grounded - they can be soldered without risk.

DO NOT GROUND LIDS WHICH ARE CONNECTED TO POWER!

Figure 11.62

Noise Coupling - Mutual Inductance

If a current flows in a conductor it produces a magnetic field. If the current changes the magnetic field strength changes. If the magnetic field around a conductor changes it produces an e.m.f. in that conductor. It therefore follows that if the current in a conductor changes voltages will be induced in nearby conductors. The phenomenon is known as mutual inductance. (A voltage is also induced in the original conductor by the related phenomenon of *self inductance* but we shall not consider that here.)

Mutual inductance is a powerful cause of noise. Consider a PCB where an AC current is flowing from a source to a load. If the outward and return signal paths are separated we can consider the total current path as a single-turn coil - the larger the loop area, the greater the inductance. If there is another such loop nearby we find that current changes in one loop produce voltages in the other - the two are coupled by mutual inductance.

Once the problem is stated in this way it is obvious that it may be attacked by reducing the areas of both loops and increasing their separation.[7]

Magnetic shielding is also possible. At high frequency a conductor will act as a magnetic shield provided that the conductor thickness is greater than the skin depth at that frequency - this tells us that standard PC foil acts as a magnetic screen above 12 MHz or so and that a room-temperature superconductor would be very useful! In the absence of room-temperature superconductors a high permeability material such as mu-metal will provide magnetic shielding at DC and LF but it is best avoided, where possible, because of its cost, weight, and fragile magnetic characteristics (a single blow will affect its grain-orientation and reduce its permeability).

MUTUAL INDUCTANCE: THE PROBLEM

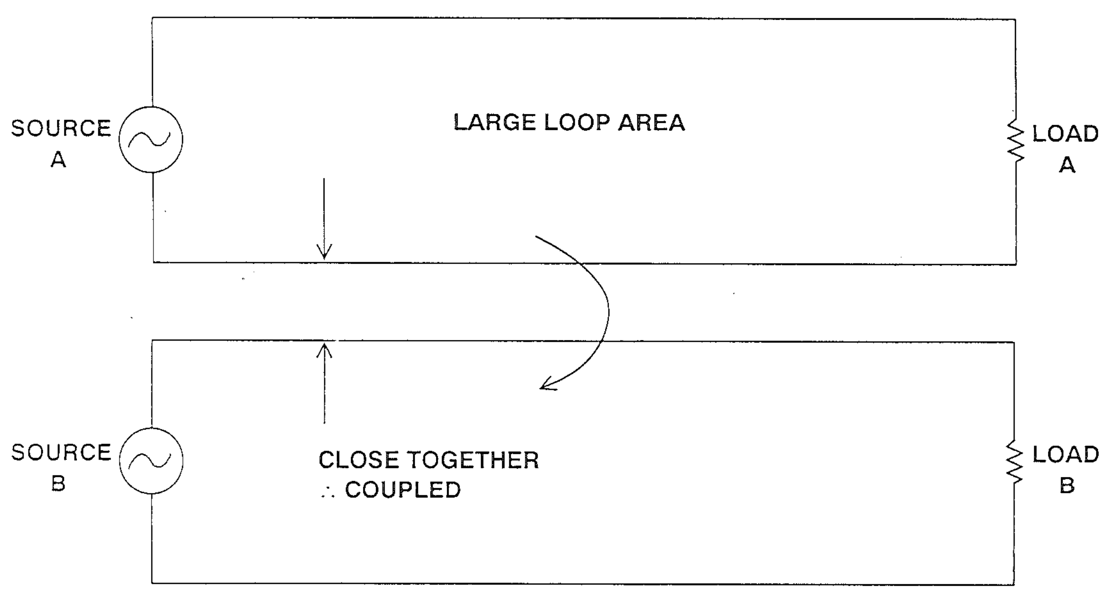


Figure 11.63

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MUTUAL INDUCTANCE: THE CURE

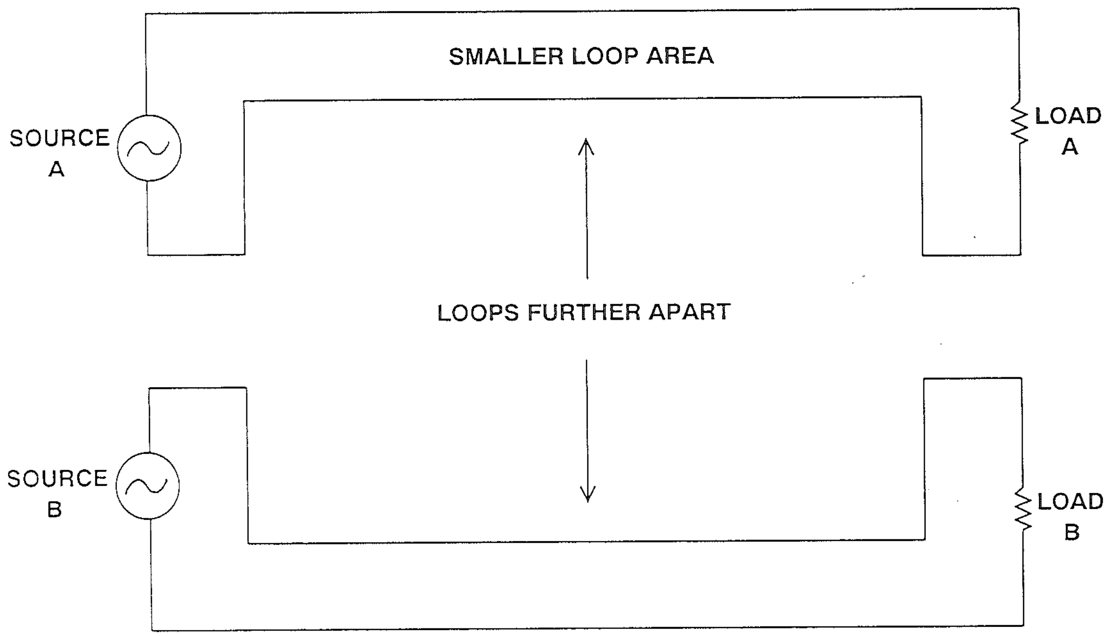


Figure 11.64

MAGNETIC SHIELDING

- THE EFFECTS OF MUTUAL INDUCTANCE CAN BE REDUCED BY SHIELDING
- BUT IT IS RARELY WORTHWHILE EXCEPT AT HIGH FREQUENCY
- At low frequency the high μ material required is expensive, heavy and easily damaged.

Figure 11.65

NOISE COUPLING - ELECTROMAGNETIC INTERFERENCE (EMI)

The World is full of radio transmitters. In addition to radio hams, CBers, policemen, security men, pizza delivery trucks, yuppies with portable telephones, radio and TV transmitters, garage door openers, microwave proximity detectors on supermarket doors and traffic lights, radars in ships, aeroplanes and speed traps, and the odd international terrorist - all of which deliberately produce and radiate electromagnetic radiation in the RF spectrum - there are also innumerable natural and non-deliberate sources of such radiation: lightning, the Sun, microwave ovens, medical diathermy, industrial heating and many more. It is folly to assume that your electronic equipment will never be required to perform in an environment where there is a field strength of a few volts per meter.

It is not difficult to shield a circuit against electromagnetic fields of this magnitude. If every conductor entering the circuit is decoupled with a low inductance capacitor that may well be sufficient - if a ferrite bead is used as well to produce a low-pass L filter there will be even more protection.

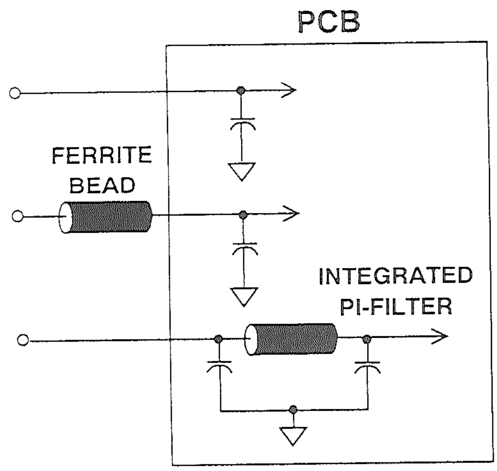
ELECTROMAGNETIC NOISE INTERFERENCE

- The world is full of radio transmitters.
- Police, taxis, broadcast, amateur, CB, cellular and cordless telephones, telemetry and garage door openers.
- Do not imagine that your circuit will never encounter one.

Figure 11.66

11

EMI PREVENTION



- SINGLE CERAMIC CAPACITOR
- FERRITE BEAD AND CAPACITOR
- INTEGRATED PI-FILTER

WHERE HIGH E-M FIELDS ARE ENCOUNTERED, CIRCUITRY SHOULD BE SCREENED BY A GROUNDED CONDUCTING ENCLOSURE

Figure 11.67

If any port turns out to be more vulnerable it is possible to buy, quite cheaply, an integrated π-filter, consisting of two ceramic capacitors and a ferrite bead in a single small "epoxy-blob" package.

If HF signals must enter or leave the board they should do so on a co-axial cable, which provides shielding. It may also be necessary to provide a bandpass filter to keep out of band signals from entering the board by this path.

If a circuit is particularly sensitive to EMI, or is to be used in an environment where high RF fields are likely to be encountered, it is good practice to enclose the entire circuit, or the sensitive parts of it, in a grounded conducting shield or can, which may be made of tin-plated steel or of copper foil.

PHOTOELECTRIC EFFECTS

Light is also an electromagnetic radiation, and it affects silicon junctions - all silicon diodes are photodiodes. Few amplifiers are not encapsulated in opaque packages[8] so it rarely necessary to worry about the effect of light on an IC, but diodes used in protective circuitry are often glass-encapsulated and may cause problems.

Figure 11.68 shows an example. A high voltage sensor with a low current output (a photocell or an ion detector or some similar transducer) is connected to a low-bias current op-amp configured as a current-voltage converter. To protect the op-amp against possible breakdown of the transducer there is a current-limiting resistor and a diode, which clamps the op-

PHOTOCURRENT IN GLASS DIODES CAN CAUSE HUM IN AMPLIFIERS

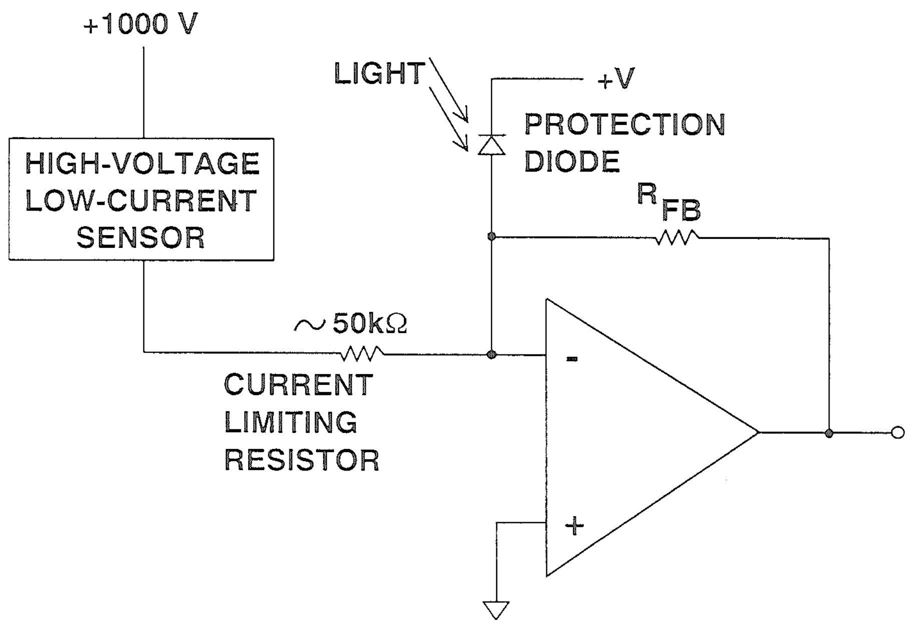


Figure 11.68

amp input to its positive supply. If the diode is a glass-encapsulated part it will be photosensitive. In most such diodes the sensitivity will be very low and may be disregarded, but occasionally a diode will be found which has higher photosensitivity. The DC photocurrent may be a problem, but much more commonly the 100 Hz (120 Hz in the USA) modulated light from fluorescent lamps modulates the photocurrent and causes hum in the amplifier. The solution is either to paint the diode black, or to use a plastic encapsulated diode.

NOISE FROM SWITCHING POWER SUPPLIES

The switching power supply offers low cost, small size, high efficiency, high reliability and the possibility of operating from a wide range of input voltages without adjustment. Unfortunately, it is very dangerous because it generates every form of noise known to mankind and several more which are not.

Switching power supplies work by chopping their input with varying mark/space ratios at frequencies which generally lie between 3 and 200 kHz and then using quite small inductors or transformers to transform to the final output voltage. The faster the switching can be accomplished, the higher the efficiency - but the more harmonics will be produced. Switching supplies produce noise over a broad band of frequencies - and it occurs as conducted noise, radiated noise, and electric and magnetic fields.

SWITCHING MODE POWER SUPPLIES

- Generate every imaginable type of noise and some inconceivable ones as well
- DO NOT USE THEM WHERE NOISE IS IMPORTANT
- If their use is unavoidable do not relax and enjoy it, but take extreme precautions against all forms of noise
- Remember that a manufacturer's design change in a bought-in switching mode power supply may alter its effects on your system noise without altering its published specification.
- When developing a system using a switching mode supply it is instructive and often frightening to temporarily replace the switching supply with a battery or a linear supply and to remeasure the system noise!

Figure 11.69

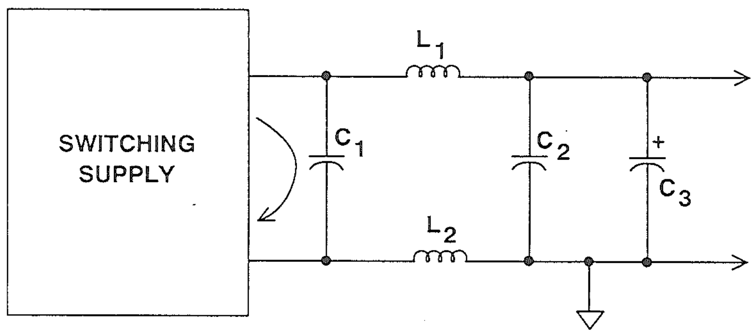
More practical, though, is to take extreme precautions when using a switching supply, and to recognize that in a very low noise system it may be better to use a less efficient linear supply. It is always wise, when designing a system using a switching supply, to measure the system noise, and then to remove the switching supply and, temporarily, replace it with a battery and measure the noise again. The results of such an experiment are always instructive and may be startling!

The conducted noise on a switching supply is dealt with by filtering. Remember that the capacitive reactance formula is only approximately true for an electrolytic capacitor at high frequencies because of the inductance of its structure, and that both electrolytic and ceramic capacitors and, probably, a choke (inductor) or two will be necessary to give an adequately low level of noise on the output of a switching supply.

The manufacturer will often include a filter which is adequate for supplies for digital systems, but for precision analog and, especially, for mixed signal sampled data systems further filtering should be taken for granted.

Switching supplies generate HF electric and magnetic fields (they generate electromagnetic fields as well, but these can generally be ignored). The electric fields are easily suppressed by a *grounded* Faraday shield (screen) , the HF magnetic fields, but not LF ones, will also be attenuated by such a shield (see comments earlier in this section on magnetic shielding), and both will cause fewer problems if the supply is physically located as far as possible from sensitive circuitry.

DECOUPLING A SWITCHING SUPPLY



- C_1 MUST HAVE LOW INDUCTANCE AND BE CLOSE TO THE SUPPLY TO MINIMIZE HF CURRENT LOOPS AND RESULTANT HF MAGNETIC FIELDS
- C_2 IS ALSO LOW INDUCTANCE, C_3 IS ELECTROLYTIC
- IF THE SWITCHING SUPPLY IS INTERNALLY GROUNDED, L_2 SHOULD BE OMITTED

Figure 11.70

MINIMIZING THE EFFECTS OF SWITCHING POWER SUPPLY ELECTRIC AND MAGNETIC FIELDS

- SHIELD THEM AND LOCATE THEM WELL AWAY FROM SENSITIVE CIRCUITRY
- THIS IS A GENERAL PRINCIPLE
- Electromagnetic radiation is not usually too bad

Figure 11.71

11

PHYSICAL SEPARATION

MINIMIZES NOISE

Figure 11.72

This is really the global solution. It is bad practice to place switching supplies too close to sensitive circuitry - the design of the supply may change, without changing its data sheet specifications, and the external electric and magnetic fields may be completely altered. If sensitive circuitry is located close to a switching supply its performance may be devastated - if it is further away problems will be minimized.

NOISE REDUCTION BY SEPARATION

The idea of minimizing external noise by physical separation does not apply only to switching supplies. Wherever noise from one circuit corrupts another, physical separation will reduce the effect. This separation can be by means of a shield, as discussed earlier, or it can be by sheer physical distance.

Modern electronic layout is costed to encourage minimal PCB area and components are placed as close together as possible. This may save money, but if it degrades the performance of circuits it is poor economy. It is often worthwhile to use separations of 4-10 mm of unpopulated board between potentially interfering parts of a system. This will allow a "moat" or "guard ring" of grounded copper to act as a shield between the parts. Production managers will howl about "wasted space" but the improvement in performance often justifies the cost.

MEASURING EXTERNAL NOISE

When a system is troubled with external noise it is often helpful to be able to measure the noise at various points in ground. It is rarely possible to do this with a simple oscilloscope probe but an instrumentation amplifier simplifies the job enormously.

If an instrumentation amplifier is set to a gain of 1000 and drives an oscilloscope with a sensitivity of 5 mV/cm then the sensitivity from its input to the oscilloscope will be 5 μV/cm. It will also have a common-mode rejection of over 100 dB and is thus ideal for measuring ground noise in PCBs and elsewhere.

There are two caveats: the external ground pin of the circuit board must be connected to the reference pin of the in-amp, the ground of the in-amp supply, and the ground of the oscilloscope in order to allow bias current to flow in the in-amp inputs, and, since if changing magnetic fields are present they can induce e.m.f.s in the loop formed by the input leads, giving spurious readings, their presence should be checked by shorting the probes to the board to provide bias current and moving their leads about and seeing if an AC signal, due to magnetic coupling, is detected.

This simple test system is extremely useful. Analog Devices once supplied their sales force with test boxes containing an instrumentation amplifier and batteries for performing this test. It was intended that they should lend them to customers who had ground noise problems. Within three months some 500 of these test boxes had been appropriated by customers who found them too useful to return![9]

GROUND NOISE DOWN TO 5μV
FROM DC TO 50 kHz
MAY BE MEASURED WITH AN OSCILLOSCOPE
AND AN INSTRUMENTATION AMPLIFIER

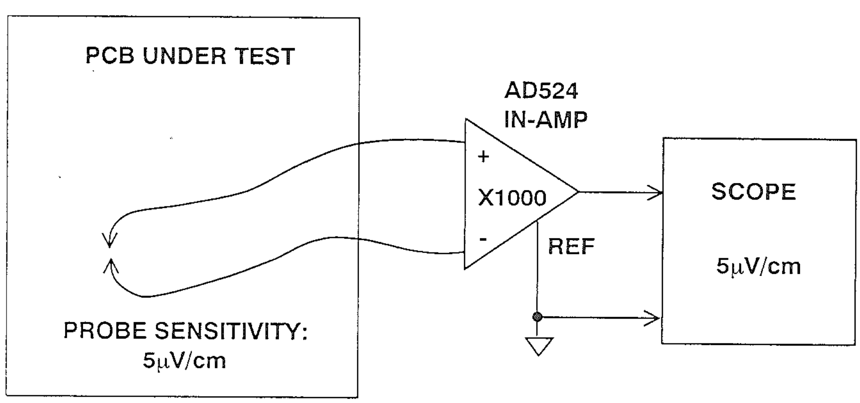


Figure 11.73

GROUND NOISE DOWN TO 0.5μV FROM 0.5 TO 500 MHz (OR MORE) MAY BE MEASURED WITH A BROADBAND TRANSFORMER AND A SPECTRUM ANALYZER

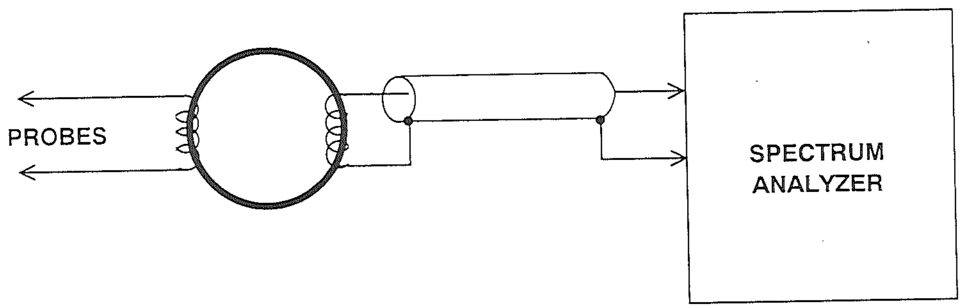


Figure 11.74

Many in-amps have a bandwidth of DC to 50 KHz. At higher frequencies there is another technique for ground noise measurement - a broadband transformer. Two lengths of enamelled copper wire are twisted together and the resulting twisted pair wound on a ferrite toroidal core, and one wire used as the primary and the other as the secondary of a transformer. Such a "transmission line transformer" or "Ruthroff transformer" is efficient over a very wide bandwidth.[10] If the leads to the primary are used as the test probes and the secondary is connected (by a co-axial cable) to the input of a broadband spectrum analyzer the system will measure ground noise down to a few microvolts from a few hundred KHz to over 1 GHz. The display, of course, will be in the frequency domain, rather than the time domain display of an oscilloscope, but this is quite easily interpreted and the additional 60-70 dB of dynamic range is worthwhile (a broadband oscilloscope is unlikely to have a sensitivity much better than 5 mV/div while a spectrum analyzer, with its logarithmic display, will certainly go down to a μV or less).

Conclusion

Noise has a reputation as a subject that is more black art than science. In fact the subject is quite simple - but extensive. Noise can come at you from all directions and it is important to consider all the possibilities, since Murphy's Law says that the one you don't consider is the cause of the problem.

This section of our seminar has attempted to lay the groundwork of noise in amplifier circuits, both the internal noise of the circuits and components and the external noise or interference. If all the sources are considered and suitable tests are performed it is usually comparatively easy to determine the source or sources of the noise and error in a system. If the source is understood the cure is much more easily accomplished.

If we must summarize the message of this section in a sentence, it is

THOUGHT BEFORE ACTION[11]

THOUGHT BEFORE ACTION

Figure 11.75

NOTES

1 All resistors have this noise, which is fundamental. It is possible to reduce it by reducing the temperature (but, being a square root law, quite low temperatures produce comparatively little improvement), the bandwidth or the resistance. It is not possible to reduce Boltzmann's Constant as he is dead, having committed suicide in Vienna in 1906.

2 For much more detail on this topic see "The Integrated Circuit Amplifier User's Guide to Grounding, Decoupling, and Making Things go Right for a Change" by A. Paul Brokaw (Analog Devices Application Note).

3For more detailed analysis of these two examples consult "Linear Design Seminar" Notes, Chapter 1. Available from Analog Devices.

4 For more detailed discussion of PCB layout issues see "Mixed Signal Seminar" Notes, Section XI. Available from Analog Devices.

5 But see the "Mixed Signal Seminar" Notes, pp XI-39 to XI-42 for more detailed discussion of problems with ground planes.

6 These lids are not ungrounded through the whim or malice of the analog IC manufacturer, but because the *package* manufacturer will only supply such packages with the lid connected to a corner pin, and analog ICs may not be able to use corner pins as ground.

7 See "Linear Design Seminar" Notes, Chapter 10, and "Mixed Signal Seminar" Notes, Section XI. Both available from Analog Devices.

8 Several manufacturers of optical remote control ICs manufacture transparent packages (sometimes packages transparent to infra-red only) containing a photocell and its associated amplifier on a single silicon chip. Such circuits are a special case.

9Although the boxes are no longer available (they've all been stolen) the data sheet on the box, which describes how it is made and used, is still available from Analog Devices Applications Department.

10 C. L. Ruthroff, "Some Broadband Transformers," **Proc. I.R.E.**, Vol 47, pp.1337-1342, August, 1959.

11 The full quotation comes from "The Edge" by Dick Francis and is "Thought before action - if there's time." In the analysis of noise in electronic circuits there is usually time.

ANALOG DEVICES

APPLICATION NOTE

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An I.C. Amplifier Users' Guide To Decoupling, Grounding, And Making Things Go Right For A Change

by Paul Brokaw

"There once was a breathy baboon
Who always breathed down a bassoon,
For he said "It appears
that in billions of years
I shall certainly hit on a tune"
(Sir Arthur Eddington)

This quotation seemed a proper note with which to begin on a subject which has made monkeys of most of us at one time or another. The struggle to find a suitable configuration for system power, ground, and signal returns too frequently degenerates into a frustrating glitch hunt. While a strictly experimental approach can be used to solve simple problems, a little forethought can often prevent serious problems and provide a plan of attack if some judicious tinkering is later required.

The subject is so fragmented that a completely general treatment is too difficult for me to tackle. Therefore, I'd like to state one general principle and then look a bit more narrowly at the subject of decoupling and grounding as it relates to integrated circuit amplifiers.

. . . Principle: Think—where the currents will flow.

I suppose this seems pretty obvious, but all of us tend to think of the currents we're interested in as flowing "out" of some place and "through" some other place but often neglect to worry how the current will find its way back to its source. One tends to act as if all "ground" or "supply voltage" points are equivalent and neglect (for as long as possible) the fact that they are parts of a network of conductors through which currents flow and develop finite voltages.

In order to do some advance planning it's important to consider where the currents originate and to where they will return and to determine the effects of the resulting voltage drops. This in turn requires some minimum amount of understanding of what goes on inside the circuits being decoupled and grounded. This information may be lacking or difficult to interpret when integrated circuits are part of the design.

Operational amplifiers are one of the most widely used linear I.C.'s, and fortunately most of them fall into a few classes, so far as the problems of power and grounding are concerned. Although the configuration of a system may pose formidable problems of decoupling and signal returns, some basic methods to handle many of these problems can be developed from a look at op-amps.

OP AMPS HAVE FOUR TERMINALS:

A casual look through almost any operational amplifier text might leave the reader with the impression that an ideal op-amp has three terminals: a pair of differential inputs and an output as shown in Figure 1. A quick review of fundamentals, however, shows that this can't be the case. If the amplifier has an output voltage it must be measured with respect to some point . . . a point to which the amplifier has a reference. Since the ideal op-amp has infinite common mode rejection, the inputs are ruled out as that reference so that there must be a fourth amplifier terminal. Another way of looking at it is that if the amplifier is to supply output current to a load, that current must get into the amplifier somewhere. Ideally, no input current flows, so again the conclusion is that a fourth terminal is required.

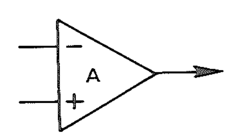

Figure 1. Conventional "Three Terminal" Op Amp

A common practice is to say, or indicate in a diagram, that this fourth terminal is "ground." Well, without getting into a discussion of what "ground" may be we can observe that most integrated circuit op-amps (and a lot of the modular ones as well) don't have a "ground" terminal. With these circuits the fourth terminal is one or both of the power supply terminals. There's a temptation here to lump together both supply voltages with the ubiquitous ground. And, to the extent that the supply lines really do present a low impedance at all frequencies within the amplifier bandwidth, this is probably reasonable. When the impedance requirement isn't satisfied, however, the door is left open to a variety of problems including noise, poor transient response, and oscillation.

11

DIFFERENTIAL TO SINGLE-ENDED CONVERSION:

One fundamental requirement of a simple op-amp is that an applied signal which is fully differential at the input must be converted to a single-ended output. Single ended, that is, with respect to the often neglected fourth terminal. To see how this can lead to difficulties, take a look at Figure 2.

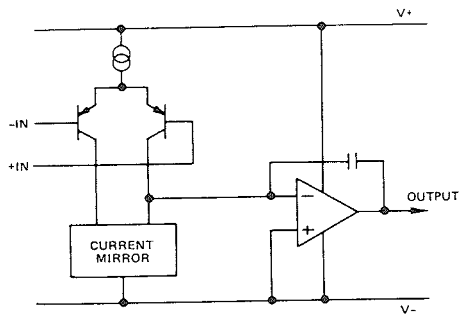

Figure 2. Simplifited "Real" Op Amp

The signal flow illustrated by Figure 2 is used in several popular integrated circuit families. Details vary, but, the basic signal path is the same as the 101, 741, TLO72, LF411,712, 515, and other integrated circuit amplifiers. The circuit first transforms a differential input voltage into a differential current. This input stage function is represented by PNP transistors in Figure 2. The current is then converted from differential to single-ended form by a current mirror which is connected to the negative supply rail. The output from the current mirror drives a voltage amplifier and power output stage which is connected as an integrator. The integrator controls the open-loop frequency response, and its capacitor may be added externally, as in the 101, or may be self-contained, as in the 741. Most descriptions of this simplified model don't emphasize that the integrator has, of course, a differential input. It's biased positive by a couple of base emitter voltages, but, the non-inverting integrator input is referred to the negative supply.

It should be apparent that most of the voltage difference between the amplifier output and the negative supply appears across the compensation capacitor. If the negative supply voltage is changed abruptly the integrator amplifier will *force* the output to follow the change. When the entire amplifier is in a closed loop configuration the resulting error signal at its input will tend to restore the output, but, the recovery will be limited by the slew rate of the amplifier. As a result, an amplifier of this type may have outstanding low frequency power supply rejection, but, the negative supply rejection is fundamentally limited at high frequencies. Since it is the feedback signal to the input that causes the output to be restored, the negative supply rejection will approach zero for signals at frequencies above the *closed loop* bandwidth. This means that high-speed, high-level circuits can "talk to" low-level circuits through the common impedance of the negative supply line.

Note that the problem with these amplifiers is associated with the negative supply terminal. Positive supply rejection may also deteriorate with increasing frequency, but, the effect is less severe. Typically, small transients on the positive supply have only a minor effect on the signal output. The difference between these sensitivities can result in an apparent asymmetry in the amplifier transient response. If the amplifier is driven to produce a positive voltage swing across its rated load it will draw a current pulse from the positive supply. The pulse may result in a supply voltage transient, but, the positive supply rejection will minimize the effect on the amplifier output signal. In the opposite case, a negative output signal will extract a current from the negative supply. If this pulse results in a "glitch" on the bus, the poor negative supply rejection will result in a similar "glitch" at the amplifier output. While a positive pulse test may give the amplifier transient response, a negative pulse test may actually give you a pretty good look at your negative supply line transient response, instead of the amplifier response!

Remember that the impulse response of the power supply itself is not what is likely to appear at the amplifier. Thirty or forty centimeters of wire can act like a high Q inductor to add a high-frequency component to the normally overdamped supply response. A decoupling capacitor near the amplifier won't always cure the problem either, since the supply must be decoupled *to* somewhere. If the decoupled current flows through a long path, it can still produce an undesirable glitch.

Figure 3 illustrates three possible configurations for negative supply decoupling. In 3a the dotted line shows the negative signal current path through the decoupling and along the ground line. If the load "ground" and decoupled "ground" actually join at the power supply the "glitch" on the ground lines is similar to the "glitch" on the negative supply bus. Depending upon how the feedback and signal sources are "grounded" the effective disturbance *caused* by the decoupling capacitor may be larger than the disturbance which it was intended to prevent. Figure 3b shows how the decoupling capacitor can be used to minimize disturbance of V– and ground busses. The high-frequency component of the load current is confined to a loop which doesn't include any part of the ground path. If the capacitor is of sufficient size and quality, it will minimize the glitch on the negative supply without disturbing input or output signal paths. When the load situation is more complex, as in 3c, a little more thought is required. If the amplifier is driving a load that goes to a virtual ground, the actual load current does not return to ground. Rather, it must be supplied by the amplifier creating the virtual ground as shown in the figure. In this case, decoupling the negative supply of the first amplifier to the positive supply of the second amplifier closes the fast signal current loop without disturbing ground or signal paths. Of course, it's still important to provide a low impedance path from "ground" to V– for the second amplifier to avoid disturbing the input reference.

The key to understanding decoupling circuits is to note where the actual load and signal currents will flow. The key to optimizing the circuit is to bypass these currents around ground and other signal paths. Note, that as in figure 3a, "single point grounding" may be an oversimplified solution to a complex problem.

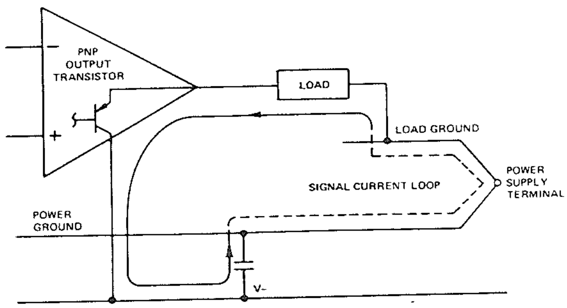

Figure 3a. Decoupling for Negative Supply Ineffective

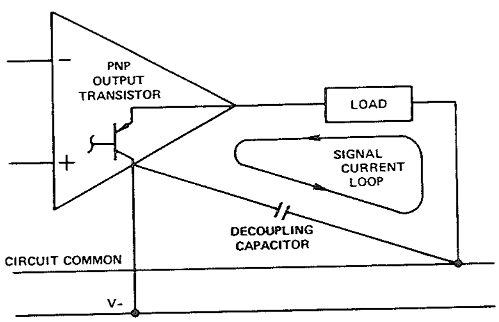

Figure 3b. Decoupling Negative Supply Optimized for "Grounded" Load

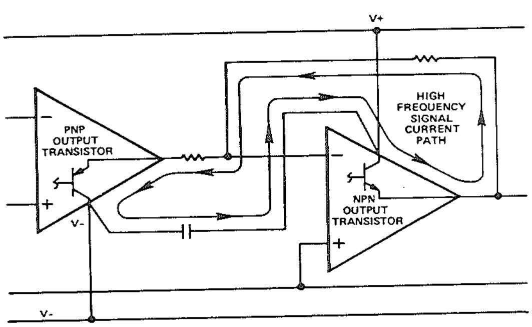

Figure 3c. Decoupling Negative Supply Optimized for "Virtual Ground" Load

Figure 3b and 3c have been simplified for illustrative purposes. When an entire circuit is considered conflicts frequently arise. For example, several amplifiers may be powered from the same supply, and an individual decoupling capacitor is required for each. In a gross sense the decoupling capacitors are all paralleled. In fact, however, the inductance of the interconnecting power and ground lines convert this harmless-looking arrangement into a complex L-C network that often rings like the "Avon Lady". In circuits handling fast signal wavefronts, decoupling networks paralleled by more than a few centimeters of wire generally mean trouble. Figure 4 shows how small resistors can be added to lower the Q of the undesired resonant circuits. The resistors can generally be tolerated since they convert a bad high-frequency jingle to a small damped signal at the op amp supply terminal. The residual has larger *low frequency* components, but, these can be handled by the op-amp supply rejection.

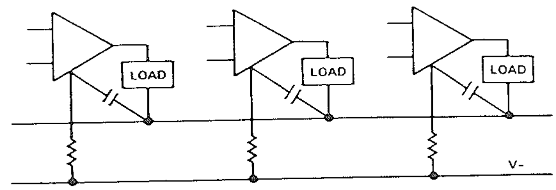

Figure 4. Damping Parallel Decoupling Resonances

FREQUENCY STABILITY:

There's a temptation to forget about decoupling the negative supply when the system is intended to handle only low-frequency signals. Granted that decoupling may not be required to handle low-frequency signals, but it may still be required for frequency stability of the op-amps.

Figure 5 is a more-detailed version of Figure 2 showing the output stage of the I.C. separated from the integrator (since this is the usual arrangement) and showing the negative power supply and wiring impedance lumped together as a single constant. The amplifier is connected as a unity gain follower. This makes a closed-loop path from the amplifier output through the differential input to the integrator input. There is a second feedback path from the collector of the output PNP transistor back to the other integrator input. The net input to the integrator is the difference of the signals through these two paths. At low frequencies this is a net, negative feedback. The high-frequency feedback depends upon both the load reactance and the reactance of the V– supply.

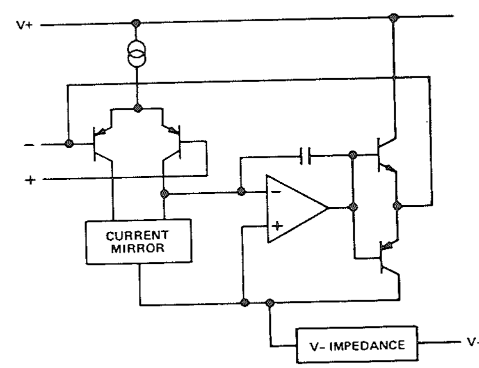

Figure 5. Instability Can Result from Neglecting Decoupling

When the supply lead reactance is inductive, it tends to destabilize the integrator. This situation is aggravated by a capacitive load on the amplifier. Although it's difficult to predict under exactly what circumstances the circuit will become unstable, it's generally wise to decouple the negative supply if there is any substantial lead inductance in the V– lead *or* in the common return to the load and amplifier input signal source. If the decoupling is to be effective, of course, it must be with respect to the *actual* signal returns, rather than to some vague "ground" connection.

POSITIVE SUPPLY DECOUPLING:

Up to this point we haven't considered decoupling the positive supply line, and with amplifiers typified by Figures 2 and 5 there may be no need to. On the other hand, there are a number of integrated circuit amplifiers which refer the compensating integrator to the positive supply. Among these are the 108, and similar families. When these circuits are used, it's the positive supply which requires most attention. The considerations and techniques described for the class of circuits shown in Figure 2 apply equally to this second class, but, should be applied to the positive supply rather than the negative.

FEED-FORWARD:
A technique which is most frequently used to improve bandwidth is called feed-forward. Generally, feed-forward is used to bypass an amplifier or level translator stage which has poor high frequency response. Figure 6 illustrates how this may be done. Each of the amplifiers shown is really a subcircuit, usually a single stage, in the overall amplifier. In the illustration, the input stage converts the differential input to a single-ended signal. The signal drives an intermediate stage (which in practice often includes level translator circuitry) which has low-frequency gain, but, limited bandwidth. The output of this stage drives an integrator-amplifier and output stage. The overall compensation capacitor feeds back to the input of the second stage and includes it in the integrator loop. The compromises necessary to obtain gain and level translation in the intermediate stage often limit its bandwidth and slow down the available integrator response. A feed-forward capacitor permits high-frequency signals to bypass this stage. As a result, the overall amplifier combines the low-frequency gain available from 3 stages with the improved frequency response available from a 2-stage amplifier. The feed-forward capacitor also feeds back to the non-inverting input of the intermediate stage. Note that the second stage is not an integrator, as it may appear at first glance, but actually has a positive feedback connection. Fed-forward amplifiers must be carefully designed to avoid internal oscillations resulting from this connection. Improper decoupling can upset this plan and permit this loop to oscillate.

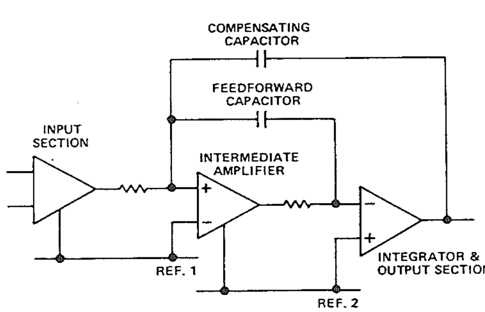

Figure 6. Fast Fed-Forward Amplifier

Note that the internal input stages are shown as being referred to separated reference points, Ideally, these will be the same reference so far as signals are concerned, although they may differ in bias level. In practice this may not be the case. Examples of fed-forward amplifiers are the AD518 and the AD707. In these amplifiers, signal Reference 1 is the positive supply, while signal Reference 2 is the negative supply. Signals appearing between the positive and negative supply terminals are effectively inserted inside the integrator loop!

Obviously, while feed-forward is a valuable tool for the high-speed amplifier designer, it poses special problems in application. A thoughtful approach to decoupling is required to maximize bandwidth and minimize noise, error, and the likelihood of oscillation.

Some fed-forward amplifiers have other arrangements, which include the "ground" terminal in inverting only amplifiers. Almost without exception, however, signals between some combination of the supply terminals get "inside" the amplifier. It is vital to proper operation that the involved supply terminals present a common low impedance at high frequencies. Many high-speed modular amplifiers include appropriate capacitive decoupling within the amplifier, but, with I.C. op amps this is impossible. The user must take care to provide a cleanly decoupled supply for fed-forward amplifiers. Figure 7 shows a decoupling method which may be applied to the AD518 as well as to other fast fed-forward amplifiers such as the 118. One capacitor is used to provide a low-impedance path between the supply terminals at high frequencies. The resistor in the V+ lead insures that noise on the supply lines will be rejected *and* prevents the establishment of resonances with other decoupling circuits. The second capacitor decouples the low side of the integrator to the load.

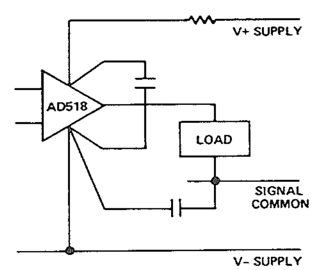

Figure 7. Decoupling for a Fed-Forward Amplifier

Alternatives include a resistor in both supply leads and/or decoupling from V+ to the load. In principle, the positive and negative supply should be tied in a "tight knot" with the signal return. To the extent that this cannot be done, there is a slight advantage to favoring the negative supply due to the high frequency limitations of PNP transistors used in junction-isolated I.C.'s.

OTHER COMPENSATION:
While most integrated circuit amplifiers use one of the three compensation schemes already described, a significant fraction use some other plan. The 725 type amplifiers combine a V– referred integrator with a network which the manufacturers recommend to be connected from signal ground to the integrator input. This makes the circuit extremely liable to pick up noise between V– and ground. In many circumstances it may be wiser to connect the external compensation to the negative supply, rather than to signal ground.

One more class of amplifiers is typified by the Analog Devices AD829 and AD847 . In these circuits, a single capacitor may be used to induce a dominant pole of response without resorting to an integrator connection. The high-frequency response of the amplifier will appear with respect to the "ground" end of the compensation capacitor. In these amplifiers a small internal capacitance is connected between V+ and the compensation point. Unity gain compensation can be added in parallel and the pin-out is arranged to make this simple. The free end of the compensation capacitor can also be connected either to V− or signal common. It is extremely important that the signal common and the compensation connect directly or through a low-impedance decoupling.

Although the main signal path of these amplifiers can be compensated in a variety of ways, some care is required to insure the stability of internal structures. It's always wise to use extra care in decoupling wideband amplifiers to avoid problems with the output stage and other subcircuits which are similar to the main integrator problem illustrated by Figure 5. An effective compensation and decoupling circuit for the AD509 is shown in Figure 8. This arrangement is similar to Figure 7, and one of these two circuits is likely to be suitable for many types of wideband amplifier. Depending upon the power distribution, a small (10Ω to 50Ω) resistor may be appropriate in both of the supply leads to reduce power lead resonance and interference both to and from circuits sharing the power supply.

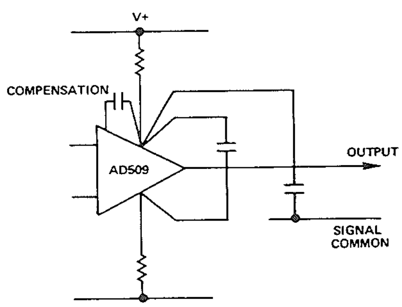

Figure 8. - Decoupling a Wideband Amplifier

GROUNDING ERRORS:

Ground in most electronic equipment is not an actual connection to earth ground, but a common connection to which signals and power are referred. It is frequently immaterial to the function of the equipment whether or not the point actually connects to earth ground. I myself prefer some distinguishing name or names for these common points to emphasize that they must be *made* common. The term "ground" too often seems to be associated with a sort of cure-all concept, like snake oil, money or motherhood. If you're one of those who regards ground with the same sort of irrational reverence that you hold for your mother, remember that while you can always trust your mother, you should *never* trust your "ground." Examine and think about it.

It's important to have a look at the currents which flow in the ground circuit. Allowing these currents to share a path with a low-level signal may result in trouble. Figure 9 illustrates how careless grounding can degrade the performance of a simple amplifier. The amplifier drives a load which is represented by the load resistor. The load current comes from the power supply and is controlled by the amplifier as it amplifies the input signal. This current must return to the supply by some path; suppose that points A and B are alternative power supply "ground" connections. Assuming that the figure represents the proper topology or ordering of connections along the "ground" bus, connecting the supply at A will cause the load current to share a segment of wire with the input signal connection. Fifteen centimeters of number 22 wire in this path will present about 8 milliohms of resistance to the load current. With a 2k load, a 10-volt output signal will result in about 40 microvolts between the points marked "ΔV." This signal acts in series with the non-inverting input and can result in significant errors. For example, the typical gain of an AD707 amplifier is 8 million so that only 1¼μV of input signal is required to produce a 10 volt output. The 40μV ground error signal will result in a 32 times increase in the circuit gain error! This degradation could easily be the most serious error in a high-gain precision application. Moreover, the error represents positive feedback so that the circuit will latch up or oscillate for large closed-loop gains with R_f/R_i greater than about 250k.

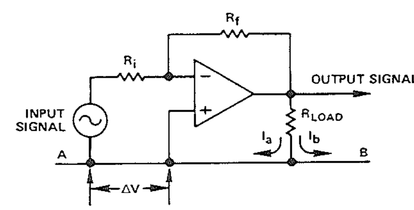

Figure 9. Proper Choice of Power Connections Minimizes Problems

Reconnecting the power supply to point B will correct the problem by eliminating the common impedance feedback connection. In a real system, the problem may be more complex. The input signal source, which is represented as floating in Figure 9, may also produce a current which must return to the power supply. With the supply at point B, any current which flows in additional loads (other than R_i) may interfere with the operation of the amplifier shown. Figure 10 illustrates how amplifiers can be cascaded and still drive auxiliary loads without common impedance coupling. The

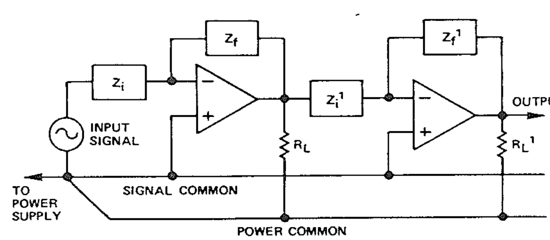

Figure 10. Minimizing Common Impedance Coupling

output currents flow through the auxiliary loads and back to the power supply through power common. The currents in the input and feedback resistors are supplied from

11

the power supply by way of the amplifiers as previously illustrated in Figure 3c. The only current flowing in signal common is the amplifier's input current, and its effect is generally negligibly small.

Having given an example of a simple "grounding error" and its solution, I will now get back on my soap box and say that grounding errors result from neglect based on the assumption that a ground, is a ground, is a ground. *Some* impedance will be present in any interconnection path, and its effect should be considered in the overall design of a system. Quantitative approaches are quite useful in specialized applications. In fast TTL and ECL logic circuitry the characteristic impedance of interconnections is controlled so that proper terminations can reduce problems. In RF circuitry the unavoidable impedances are taken into account and incorporated into the design of the circuit. With op-amp circuitry, however, impedance levels do not lend themselves to transmission line theory, and the power and ground impedances are difficult to control or analyze. The most expedient procedure, short of difficult and restrictive quantitative analysis, seems to be to arrange the unavoidable impedances so as to minimize their effects and arrange the circuitry to overcome the effects. Figures 9 and 10 illustrate the sort of simple considerations which can substantially reduce practical ground problems. Figure 11 illustrates how circuitry can be used to reduce the effect of ground problems which can't be corrected by topological tricks.

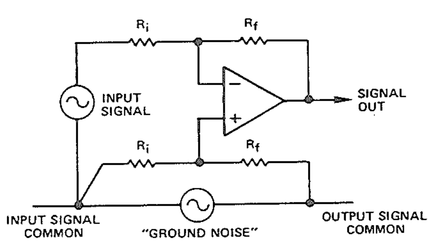

Figure 11. Subtractor Amplifier Rejects Common Mode Noise

GETTING AROUND THE PROBLEM:

In Figure 11 a subtractor circuit is used to amplify a normal mode input signal and reject a ground noise signal which is common to both sides of the input signal. This scheme uses the common-mode rejection of the amplifier to reduce the noise component while amplifying the desired signal. An important aspect of this arrangement, which is often overlooked, is that the amplifier should be powered with respect to the *output* signal common. If its power pins are exposed to the high-frequency noise of the input common, the compensation capacitor will direct the noise right to the output and defeat the purpose of the subtractor. It's just this kind of effect which makes it important to use care in grounding and decoupling. A subtractor or dynamic bridge, like Figure 11, will be ineffective in correcting a grounding problem if the amplifier itself is carelessly decoupled. In general, an op-amp should be decoupled to the point which is the reference for measuring or using its output signal. In "single-ended" systems it should also be decoupled to the input signal return as well. When it is impossible to satisfy both these requirements at once, there's a high probability of either a noise or oscillation problem or both. Frequently the difficulty can be resolved with a subtractor, like Figure 11, where a network like the single-ended feedback network (which needn't be all resistive) joins the input and output signal reference points and provides a "clean" reference point for the non-inverting input of the amplifier.

A problem with the subtractor is that it uses a balanced bridge to reject the common mode signal between the input and output reference points. The arms of the network must be carefully balanced, since to the extent they don't match, the unwanted signal will be amplified. Although even a poorly matched network will probably eliminate oscillation problems, noise rejection will suffer in direct proportion to any mismatches. An easier way to reject large "ground noise" signals is to use a true instrumentation amplifier.

INSTRUMENTATION AMPLIFIERS:

A true instrumentation amplifier has a very visible "fourth terminal." The output signal is developed with respect to a well defined reference point which is usually a "free" terminal that may be tied to the output signal common. The instrumentation amplifier also differs from an op amp in that the gain is fixed and well defined, but there is no feedback network coupling input and output circuits. Figure 12 shows how an instrumentation amplifier can be used to translate a signal from one "ground reference" to another. The normal mode input signal is developed with respect to one reference point which may be common to its generating circuits. The signal is to be used by a system which has an interfering signal between its own common and the signal source. The instrumentation amplifier has a high impedance differential input to which the desired signal is applied. Its high common mode rejection eliminates the unwanted signal and translates the desired signal to the output reference point. Unlike the dynamic bridge circuit, the gain and common mode rejection don't depend on a network connecting the input and output circuits. The gain is set, in Figure 12, by the ratio of a pair of resistors which are connected inside the amplifier. The amplifier has a very high input impedance, so that gain and common mode rejection are not greatly affected by variations or unbalance in source impedance.

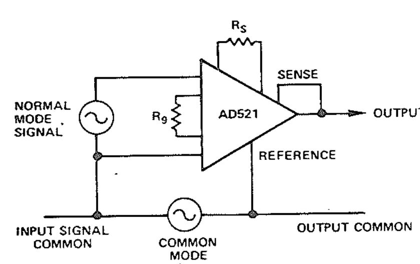

Figure 12. Applying an In-Amp

Since instrumentation amplifiers have a reference or "ground" terminal, they have the potential to be free of the power supply sensitivities of op amps. In practice, however, most instrumentation amplifiers have internal frequency compensation which is referred to the power supply. In the case of the AD521, the compensation integrator is referred to the negative supply terminal. The decoupling of this terminal is particularly important, and it should be decoupled with respect to the output reference terminal, or actually to the point to which this terminal refers.

THE "OTHER" INPUT:

Most I.C. op-amps and in-amps include offset voltage nulling terminals. These terminals generally have a small voltage on them and by loading the terminals with a potentiometer the amplifier offset voltage can be adjusted. While their impedance level is much lower than the normal input, the null terminals can act as another differential input to the amplifier. Although the null terminals aren't generally looked at as inputs, most amplifiers are quite sensitive to signals applied here. For example, in 741 family amplifiers the output voltage gain from the null terminals is greater than the gain from the normal input!

An illustration of the type of problems that can arise with the "other" input is shown in Figure 13. The figure is an op-amp circuit with some of the offset null detail shown.

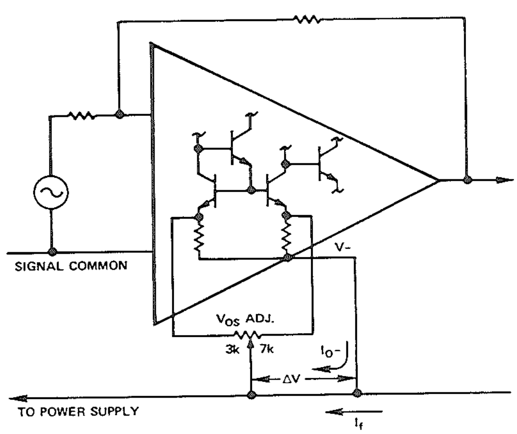

Figure 13. Details of V_{OS} Nulling — the "Other" Input

As it's drawn, the V_{OS} null pot wiper connects to a point along a V— "clothesline" which carries both the return current from the amplifier and currents from other circuits back to the power supply. These currents will develop a small voltage, ΔV, along the conductor between the amplifier V— terminal and the null pot wiper. If the null pot is set on center, the equal halves will form a balanced bridge with the resistors inside the amplifier. The effect of the voltage generated along the wire is balanced at the V_{OS} terminals and will have little effect on the amplifier output. On the other hand, if the null pot is unbalanced, to correct an amplifier offset, the bridge will no longer balance. In this case voltages developed along the "clothesline" will result in a difference voltage at the V_{OS} terminals. For instance, suppose that a 10k null pot balances out the op amp offset when it is set with 3k and 7k branches as shown in the figure. In a 741 the internal resistors are about 1k so that the difference signal at the V_{OS} terminals will be about 1/8 ΔV. The gain from these terminals is about twice the gain from the normal input, so that the disturbance acts as if it were an input signal of about 1/4 ΔV. Using the same assumptions as in the discussion of Figure 9. the current I_O— will result in a 10 microvolt input error signal. In this case, however, the error will appear *only* when the amplifier load current comes from the negative supply. When the load is driven positive the error will disappear. As a result, the V_{OS} input signal will result in distortion rather than a simple gain error!

An additional problem is created by I_f, a current returning to the power supply from other circuits. The current from other circuits is not generally related to the op amp signal, and the voltage developed by it will manifest itself as noise. This signal at the null terminals can easily be the dominant noise in the system. A few milliamps of V— current through a few centimeters of wire can result in interference which is orders of magnitude larger than the inherent input noise of the amplifier. The remedy is to make the connection from the null pot wiper direct to the V— pin of the amplifier, as shown in Figure 14. Some amplifiers such as the AD707 and AD840 refer to the null offset terminals to V+. Obviously, the pot wiper should go to the V+ terminal of this type of amplifier. It's important to connect the line directly to the op amp terminal so as to minimize the common impedance shared by the op amp current and the null pot connection.

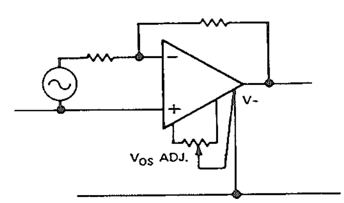

Figure 14. Connecting the Null Pot for Trouble Free Operation

The considerations for op-amp null pots also apply to the similar trimmers on almost all types of integrated circuits. For example, the AD521 In-Amp null terminals exhibit a gain of about 30 to the output. Although this is much less than in the case of most op-amps, it still warrants care in controlling the null pot wiper return. Table I lists the integrated circuits manufactured by Analog Devices, including some popular second-source families, and indicates how internal conversions from differential to single ended are referred. That is, the signals are made to appear with respect to the terminal(s) listed.

	Internal Integrator Referred to:	Comment
AD OP 07/ 27/37	V+, V−	Internal Feedforward Cap V+ to V− and Integrator V− to Output
AD380	V+	
AD390	V−	Output and Reference Amplifier
AD394/AD395	V−	Output Amplifiers
AD396	V−	Output Amplifiers
AD507	−	External Cap to Signal Common or V+
AD508	−	External Cap to Signal Common or V+
AD510	V+	
AD517	V+	
AD518	V+, V−	Internal Feedforward Cap V+ to V− and Integrator V− to Output
AD521	V−	Output Amplifier Integrator
AD524	V−	Output Amplifier Integrator
AD526	V−	Output Amplifier Integrator
AD532/AD533	V+	Multiplier Output Amplifier Integrator
AD534/AD535	V−	Output Amplifier
AD536A	V−, V+ Common	External Integrator to V+, Internal Feedforward V− to Common
AD538	V−	Internal Amplifiers
AD542/AD642	V−	
AD544/AD644	V−	
AD545A	V−	
AD546	V−	
AD547/AD647	V−	
AD548/AD648	V−	
AD549	V−	
AD557/AD558	Common	Output Amplifier and DAC Control Loop Integrator Referred to Common
AD561	V−, Common	DAC Control Loop Integrator and Ref. Amp Referred to Common and Ref. Bias Amplifier Referred to V−
AD565A/ AD566A	V−	DAC Control Loop Integrator Referred to V−. Reference Input Common to Control Loop Isolated from DAC Output Common
AD568	V+	Reference Amplifier
AD580	V−	Output Amplifier
AD581	V−	Output Amplifier
AD582	V−	Output Amplifier
AD584	V−	Output Amplifier
AD586/AD587	V−	Output Amplifier
AD588	V−	Output Amplifier
AD624/AD625	V−	Output Amplifier Integrator
AD636	V−, V+, Common	External Integrator to V+, Internal Feedforward V− to Common
AD637	V−, Common	Internal Feedforward V− to Common
AD645	V−	
AD650/AD652	V+	Internal Amplifier
AD662	Common	DAC Control Loop Integrator and Reference Amplifier Referred to Common
AD664	V−	Output Amplifiers
AD667	V−, Common	Output Amplifier Referred to V− and Reference Amplifier Referred to Common
AD668	V+	Reference Amplifier
AD688	V−	Output Amplifier
AD689	V−	Output Amplifier
AD704/AD705/ AD706	V+	
AD707/AD708	V+, V−	Internal Feedforward Cap V+ to V− and Integrator V− to Output
AD711/AD712/ AD713	V−	
AD736/ AD737	V−, Common	External Integrator to V− Internal Feedforward V− to Common
AD741	V−	
AD744/AD746	V−	
AD766	V−	Output and Reference Amplifier
AD767	V−, Common	Output Amplifier Referred to V− and Reference Amp Referred to Common
AD840/AD841/ AD842	V+, V−	
AD843	V+, V−	
AD844/AD846	V+, V−	
AD845	V+	
AD847/AD848/ AD849	V+, V−	
AD1856/AD1860	V−	Output and Reference Amplifier
AD1864	V−	Output and Reference Amplifier
AD2700/AD2710	Common	Output Amplifier
AD2701	V−	Output Amplifier
AD2702/ AD2712	V−, Common	Output Amplifiers
AD7224/AD7225	V−	Output Amplifiers
AD7226/AD7228	V−	Output Amplifiers
AD7237/ AD7247	V+, Common	Reference Amplifier to Common Output Amplifier to Both V+ and Common
AD7245/ AD7248	V+, Common	Reference Amplifier to V+ Output Amplifier to Both V+ and Common
AD7569/AD7669	V−	All Amplifiers
AD7769	Common	All Amplifiers
AD7770	Common	All Amplifiers
AD7837/AD7847	V+	All Amplifiers
AD7840	V+, Common	Output Amplifiers to V+ Reference Amplifier to Common
AD7845	V+	All Amplifiers
AD7846	V+	All Amplifiers
AD7848	V+, Common	Output Amplifier to V+ Reference Amplifier to Common

Table I.

This collection of examples won't solve all your potential grounding problems. I hope that it will give you some good ideas how to prevent some of them, and it should also give you some of the "inside story" on I.C.'s which you can put to work in very practical ways. There is no general grounding method which will prevent all possible problems. The only generally applicable rule is attention to detail, and remember that you can always trust your mother, but . . . .